



The 3rd International Conference on Advances in Electrical Engineering (ICAEE)

17-19 December, 2015

Venue: Independent University, Bangladesh

Conference Proceedings



Organized by

Independent University, Bangladesh



Technical Co-sponsor

The Institution of Electrical & Electronics Engineers

Proceedings of the 3rd International Conference on Advances in Electrical Engineering (ICAEE), 2015

17-19 December, 2015

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Mr. M. A. Matin Chowdhury

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Independent University, Bangladesh (IUB)

Message

Independent University, Bangladesh (IUB) is organizing the 3rd International Conference on Advances in Electrical Engineering (ICAEE), 2015 during 17-19 December, 2015 at the IUB campus in Dhaka, Bangladesh. I am particularly happy that IUB will host this prestigious international event biennially, every alternate year.

I hope that with the huge success of the 1st and 2nd events held in December 2011 and 2013, a large number of participants from home and abroad will attend the ICAEE 2015.

I also believe that the participants from both academia and industry will find this conference a solid platform for professional development and collaboration among their institutions.

I wish every success of ICAEE 2015.

M. A. Matin Chowdhury

Chairman

Education, Science, Technology and Cultural Development Trust (ESTCDT)

and

Member, Board of Trustees

Independent University, Bangladesh (IUB)



Mr. Rashed Chowdhury

Chairman, Board of Trustees

Independent University, Bangladesh (IUB)



Message

I am very pleased to know that the 3rd International Conference on Advances in Electrical Engineering (ICAEE), 2015 is going to be organized by the Independent University, Bangladesh (IUB) during 17-19 December, 2015 at IUB campus in Bashundhara, Dhaka, Bangladesh. I am also very happy to know that IUB organizes this international event every two years.

Starting as one of the earliest private universities in the country, IUB has emerged as a leading institution for research and scholarly growth in business, engineering, telecommunication and ICT. It is this spirit of intellectual growth that has traditionally been fostered by IUB.

I wish IUB a successful ICAEE 2015.

Rashed Chowdhury

Chairman, Board of Trustees

Independent University, Bangladesh (IUB)



Prof. M. Omar Rahman
Vice Chancellor
Independent University, Bangladesh



Message

It is an honor for Independent University, Bangladesh (IUB) to host the 3rd International Conference on Advances in Electrical Engineering (ICAEE), to be held during 17-19 December, 2015. It is indeed a great pleasure that IUB hosts this conference every other year at its own permanent campus in Bashundhara, Dhaka.

I, on behalf of IUB, welcome all the participants from far and wide to this common forum for sharing ideas and promoting intellectual growth.

We are all aware that electrical engineering is the driving force behind the development of a nation. With its large population, rapidly growing economy, and geographic and demographic characteristics, the people of the country look to the participants of ICAEE, to come forward to face the challenges and promote national development. I believe that the conference will reflect the recent growth and advancements in electrical & electronic engineering and information & communication technology, and will permeate almost every aspect of our daily lives.

I wish ICAEE 2015 a great success.

Prof. M. Omar Rahman
Chair, Advisory Committee
ICAEE 2015



Prof. Shahriar Khan

Interim Dean

School of Engineering & Computer Science
Independent University, Bangladesh (IUB)

Message

Welcome to the Third ICAEE, on behalf of the School of Engineering and Computer Science (SECS). It is indeed an honor and privilege for SECS to be the host of this prestigious conference; one of the few conferences in Bangladesh, that are co-sponsored by the IEEE. Both SECS and ICAEE stand for research, innovation and development. We take pride in hosting the ICAEE in our quest to become a flagship research-oriented school in the country.

Only a small percentage of research may find immediate application and may generate immediate revenue. But research has led to the progress of civilization and the advent of the modern age. The research of today will become the knowledge of tomorrow. The world keeps moving with inventions and discoveries and the conveniences of modern life. Innovation prevents the stagnation of scientific minds, stimulating them with new ideas, both successful and unsuccessful. Research in the US, Europe, and Japan, China, etc. is a major factor in their economical success. It is important to promote and support research and innovation through conferences like ICAEE. The ICAEE rapidly brings scientific findings to the scientific community, with each participant having their own contribution towards the creation of knowledge.

At ICAEE, interactions between academicians, specialists, practitioners and students promote collaboration and the exchange of ideas. But ICAEE is more than an opportunity to encourage and stimulate research, and to congratulate good work. It is an opportunity to critically look at the findings and to ask the hard questions, which will help us to develop the research even further.

We look to ICAEE as a time to make new friends and to renew old friendships. Let us enjoy the camaraderie of those who share our common goals of research and studies. Let us leave with fond memories to be cherished in the years to come.

Dr. Shahriar Khan

Professor

Dept. of Electrical and Electronic Engineering
Independent University, Bangladesh



Prof. Saifur Rahman

Joseph Loring Professor at Virginia Tech

Dept. of Electrical & Computer Engineering

Director

Virginia Tech Advanced Research Institute

Message

On behalf of the *2015 International Conference on Advances in Electrical Engineering* (ICAEE 2015) program committee, I would like to welcome and thank you all for participating in this conference.

Over the past few years tremendous advances have been made in the general area of electrical engineering. The main purpose of ICAEE 2015 is to bring together researchers and practitioners from academia, industry and government research labs to share their expertise on the latest trends and advancements in various areas of electrical engineering, such as signal processing, energy, circuits, devices and systems.

At ICAEE, many renowned speakers from industry and academia locally and overseas will explore in detail the various issues of electrical & related engineering. We also look forward to hearing the ideas and views of the participants.

With the huge success of the 1st and 2nd events of ICAEE held on December 2011 & 2013, this time we are expecting roughly 200 participants from academia, research organizations and industries from 15-20 countries around the globe including Bangladesh to exchange their ideas and innovations. On behalf of the international program committee, I welcome the keynote speakers, authors, session chairs, and participants from home and abroad to the conference. I extend my thanks to the supporting universities and the sponsors for their financial support and other helps in organizing the event. Finally my hearty thanks go to the chair and members of the Technical Committee, Organizing Committee and honorable reviewers.

I hope that you will take advantage of the various papers presented at this conference and explore collaborative research opportunities with fellow colleagues. I also hope that you will enjoy the conference and your participation contributes to your professional development and relationships.

Finally, warmest welcome to Dhaka, the city of Muslin and Mosques, and enjoy its cultural heritage.

A handwritten signature in black ink, appearing to read "Saifur Rahman".

Prof. Saifur Rahman, President-elect, PES-IEEE (2016-2017) and Fellow-IEEE
Chair, International Program Committee, ICAEE 2015



Dr. Shaikh Anowarul Fattah

Professor

Dept. of Electrical & Electronic Engineering
Bangladesh University of Engineering & Technology

Message

It is my pleasure to inform you that IEEE Bangladesh Section (IEEE-BDS) provided Technical Co-sponsorship for the **3rd International Conference on Advances in Electrical Engineering (ICAEE) to be organized by Independent University, Bangladesh (IUB) during 17–19 December 2015**. As a Technical Co-sponsor, IEEE-BDS provided essential guidelines to enhance the quality of the conference.

IEEE is world's largest professional association and dedicated to advancing technological innovation and excellence for the benefits of humanity. From its inception in 1993, IEEE-BDS is working to promote engineering education and advanced research in Bangladesh. Currently it has around 1500 members, chapters of four technical societies (communication society (COMSOC), power and energy society (PES), electron device/solid state circuit society (EDS/SSCS)), Engineering in Medicine and Biology Society (EMBS), two affinity groups (women in engineering (WIE) and young professional (YP)), two student chapters (Industrial Application Society (IAS) and Microwave Theory and Techniques (MTT) society), two SIGHT groups (FLASH and CARG) and 21 student branches from 21 engineering universities. Petition for opening of IEEE Computer Society is at the final stage. IEEE-BDS and its societies, affinity groups, chapters, and student branches regularly organize technical seminars, workshops, congress, contests, Olympiads, short courses, industrial tours, project competitions, outreach programs, humanitarian activities, robotic challenges, award programs, industrial, educational and professional activities. Because of outstanding performance, our WIE Chair has own prestigious awards from region 10 and international WIE. This year some of our dedicated volunteers have been selected in different R10 committees, which is a great honor for us. IEEE BDS has won project funds from R10 in six different categories through very tough competitions. This year, volunteers from IEEE BDS have won prizes in IEEE R10 SYW Congress held in Sri Lanka. My best wishes to all award winners and project grant winners. Last year total number of members was about 1000 and in this year, it is now more than 1500. This year IEEE-BDS is awarded **"2015 Outstanding Section Membership Recruitment Performance"**. I hope participants in this conference will join in IEEE if not yet joined and actively participate in our upcoming events.

This year apart from ICAEE 2015, several international conferences, namely ICCIT, ICEEICT, ICCIE, EICT, NSYSS, and ICEEE are technically co-sponsored by IEEE-BDS and IEEE ICTP is sponsored by COMSOC BD Chapter. For the first time in history, IEEE-BDS is going to organize WIE technical conference **IEEE WIECON-ECE 2015** in collaboration with WIE Affinity Group of IEEE-BDS. For the first time, IEEE-BDS has launched **IEEE ProTalks 2015**, a workshop focusing technical leadership, management and entrepreneurship. This year IEEE-BDS has successfully organized the **IEEE Region 10**

Meet 2015 where 110 foreign delegates from 17 countries including USA, Australia, Japan, India, and China participated. Since its inception, this is the first time IEEE BDS has got this opportunity.

The conference organizers were requested to follow the guidelines provided by IEEE and IEEE-BDS. In this regard, I appreciate their willingness and effort. I believe, because of hard work of all the members of the tracks and concerned reviewers, including checking of plagiarism and conflict of interest, the quality of the review process is enhanced. It is expected from the organizer that, being an international conference, a significant number of authors from abroad will attend the conference for presenting their papers. I express my sincere gratitude to all the authors, speakers, reviewers, advisers and other members whose sincere efforts are the key factors for the success of this conference.

I appreciate feedback and suggestions from all the participants regarding the technical issues of this conference, which will help us to achieve more success in enhancing the quality of the upcoming conferences. I wish ICAEE 2015 all the success



Prof. Shaikh Anowarul Fattah
Chair, IEEE Bangladesh Section (2015-2016)



Dr. M. Abdur Razzak

Associate Professor

Dept. of Electrical & Electronic Engineering
Independent University, Bangladesh (IUB)

Message

On behalf of the organizing committee, it is my pleasure to welcome you all to the 3rd International Conference on Advances in Electrical Engineering (ICAEE), 2015, organized by Independent University, Bangladesh (IUB). Electrical Engineering is the driving force behind the recent technological developments. ICAEE aims to bring together academicians, scientists, engineers, researchers and students to share their experiences on the latest research, current practices and future trends in the fields of electrical and related engineering, and discuss the practical challenges encountered and the solutions adopted.

The 90 papers that were finally accepted for this conference are well balanced and report on the many ongoing and recent research results in various areas of electrical engineering and closely related fields. All papers submitted to ICAEE 2015 were peer reviewed by up to 4 reviewers with experienced professionals from 9 countries around the globe before final acceptance decisions were made. The acceptance rate is nearly 33% which reflects the quality maintained in this conference. These papers summarize the research of 234 researchers from 11 countries thus providing true international flavor to the conference.

I would like to thank the authors for the contribution of their papers, the reviewers for their dedication and timely review, members of the Advisory Committee and International Program Committee for their advice & guidance, members of the technical committee for their commitment, members of the local organizing committee for making local arrangements, IUB administration for providing the facilities, and the ICAEE 2015 staff to ensure timely production of the ICAEE 2015 conference proceedings. I must profusely thank Dr. Mustafa Habib Chowdhury, Conference Secretary, ICAEE 2015 for spending countless hours to connect with the authors and answer author queries. Thanks are due to the Chief Guest, the Guest of Honor, Special Guests, Keynote Speakers, and Invited Speakers for attending and enlightening ICAEE 2015. Special thanks are also due to IEEE Bangladesh Section for serving as technical co-sponsor of ICAEE 2015. Finally, I am grateful to Prof. Muhammad Harunur Rashid, a Life Fellow of IEEE, for sponsoring the "Dr. Fatema Rashid Best Paper" awards.

I sincerely hope that you will take advantage of the various papers presented at this conference and explore collaborative research opportunities with fellow colleagues who are attending ICAEE 2015 from all over the globe.

Dr. M. Abdur Razzak

Chair, Organizing Committee, ICAEE 2015

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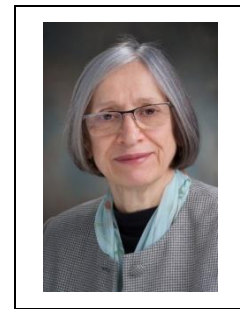
**Abstract
of
Invited and Keynote Papers**

Keynote Paper

Communication Networks: Traffic Data, Network Topologies, and Routing Anomalies

Dr. Ljiljana Trajkovic, Fellow IEEE

Professor
School of Engineering Science
Simon Fraser University, Canada



Abstract

Understanding modern data communication networks such as the Internet involves collection and analysis of data collected from deployed networks. It also calls for development of various tools for analysis of such datasets. Collected traffic data are used for characterization and modeling of network traffic, analysis of Internet topologies, and prediction of network anomalies. In this talk, I will describe collection and analysis of real-time traffic data using special purpose hardware and software tools. Analysis of such collected datasets indicates a complex underlying network infrastructure that carries traffic generated by a variety of the Internet applications. Data collected from the Internet routing tables are used to analyze Internet topologies and to illustrate the existence of historical trends in the development of the Internet. The Internet traffic data are also used to classify and detect network anomalies such as Internet worms, which affect performance of routing protocols and may greatly degrade network performance. Various statistical and machine learning techniques are used to classify test datasets, identify the correct traffic anomaly types, and design anomaly detection mechanisms.

Short Biography

Ljiljana Trajkovic received the Dipl. Ing. degree from University of Pristina, Yugoslavia, in 1974, the M.Sc. degrees in electrical engineering and computer engineering from Syracuse University, Syracuse, NY, in 1979 and 1981, respectively, and the Ph.D. degree in electrical engineering from University of California at Los Angeles, in 1986. She is currently a Professor in the School of Engineering Science at Simon Fraser University, Burnaby, British Columbia, Canada. From 1995 to 1997, she was a National Science Foundation (NSF) Visiting Professor in the Electrical Engineering and Computer Sciences Department, University of California, Berkeley. She was a Research Scientist at Bell Communications Research, Morristown, NJ, from 1990 to 1997, and a Member of the Technical Staff at AT&T Bell Laboratories, Murray Hill, NJ, from 1988 to 1990. Her research interests include high-performance communication networks, control of communication systems, computer-aided circuit analysis and design, and theory of nonlinear circuits and dynamical systems. Dr. Trajkovic serves as President (2014 - 2015) of the IEEE Systems, Man, and Cybernetics Society and served as President-Elect (2013), Vice President Publications (2012 - 2013 and 2010 - 2011), Vice President Long-Range Planning and Finance (2008 - 2009), and a Member at Large of its Board of Governors (2004 - 2006). She served as 2007 President of the IEEE Circuits and Systems Society. She was a member of the Board of Governors of the IEEE Circuits and Systems Society (2001 - 2003 and 2004 - 2005). She is Chair of the IEEE Circuits and Systems Society joint Chapter of the Vancouver/Victoria Sections. She was Chair of the IEEE Technical Committee on Nonlinear Circuits and Systems (1998). She is General Co-Chair of SMC 2016 and served as General Co-Chair of HPSR 2014, Technical Program Co-Chair of ISCAS 2005, and Technical Program Chair and Vice General Co-Chair of ISCAS 2004. She served as an Associate Editor of the IEEE Transactions on Circuits and Systems (Part I) (2004 - 2005 and 1993 - 1995), the IEEE Transactions on Circuits and Systems (Part II) (1999 - 2001 and 2002 - 2003), and the IEEE Circuits and Systems Magazine (2001 - 2003). She was a Distinguished Lecturer of the IEEE Circuits and Systems Society (2010 - 2011 and 2002 - 2003). She is a member of IEEE-HKN and a Fellow of the IEEE.

Keynote Paper

Roadmap towards 5G Mobile Communication System

Tharek A. Rahman

Professor
Faculty of Electrical Engineering
Universiti Teknologi Malaysia



Abstract

Abstract: Over the past few decades, there has been significant progress made to mobile communication networks. The promise of LTE Advanced (4G) as next generation of mobile broadband offers huge potential to service providers, application developers and wireless consumers. Beyond LTE Advanced which also known as 5G will enable the long term Networked Society and realized the vision of unlimited to information and shared data available anywhere and anytime to anyone and anything. The presentation will cover the evolution of mobile communication towards 5G that will support massive traffic volume and massive connected devices which enable to realize the Networked Society.

Short Biography

Prof. Dr. Tharek A. Rahman currently is a professor in wireless communication at faculty of electrical engineering, Universiti Teknologi Malaysia. He obtained his BSc (Hons) in Electrical Engineering from University of Strathclyde, UK, MSc in Communication Engineering from UMIST, Manchester, UK and PhD in Mobile Communication from University of Bristol, UK. He is the Director of Wireless Communication Centre (WCC), Faculty of Electrical Engineering, Universiti Teknologi Malaysia and currently conducting research related to mobile communications, antenna and propagation. He has also conducted various short courses related to mobile and satellite communication to the telecommunication industry and government agencies since 1988. Prof. Tharek has published more than 300 scientific papers in journals and conferences and obtained many national and international awards. He is also a consultant for many communication companies and an active member in several research academic entities.

Keynote Paper

The Changing Scenario in Electrical Power Generation and Grid Modernization for Smarter and Sustainable Solutions

Syed Mofizul Islam

Director, Centre for Smart Grid and Sustainable Power Systems
Professor, Curtin University, Australia



Abstract

The Electrical power and energy sector is currently going through rapid transformational changes. Variable power generation is now considered a viable option due to significant technological advancements. Open access, dynamic line ratings, transmission and distribution infrastructures embedded with ICT infrastructure, automated metering infrastructure enabled demand response techniques, development Home Area Network (HAN), Local Area Network (LAN), Wide Area Network (WAN) is changing the way electricity is generated, transmitted and distributed. Challenges in protection, grid support, and storage are giving rise to solutions such as self- healing techniques, micro-grid opportunities, and development of power electronic based enabling techniques in meeting grid standards. For remote communities, technologies are now in place for low carbon more affordable energy solutions. Electric Vehicles (EVs) in G2V and V2G, roof top PV power, community based clustered micro-grids are threatening the bulk power grid's economic existence. This talk is going to present some experiences from around the World on the new paradigm shift in power generation, transmission and distribution for grid connected and off-grid systems.

Short Biography

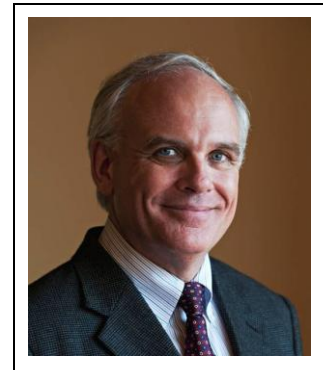
Syed Mofizul Islam received the B.Sc. degree in Electrical Engineering from Bangladesh University of Engineering and Technology, Bangladesh in 1979, the M.Sc. and PhD degree in electrical power engineering from the King Fahd University of Petroleum and Minerals, Dhahran, Saudi Arabia, in 1983, and 1988 respectively. He is currently the John Curtin Distinguished Professor in Electrical Power Engineering and the Director of Centre for Smart Grid and Sustainable Power Systems at Curtin University, Perth, Australia. He received the Dean's medallion for research at Curtin University in 1999. He received the IEEE T Burke Haye's Faculty Recognition award in 2000. He received the Curtin University inaugural award for Research Development in 2012. He received the Sir John Madsen medal in 2011 and 2014 for best electrical engineering paper in Australia. He has published over 270 technical papers in his area of expertise. His research interests are in Condition Monitoring of Transformers, Wind Energy Conversion, and Smart Power Systems. He has been a keynote speaker and invited speaker at many international workshops and conferences. He has been a Visiting Professor at Shanghai University of Electrical Power, China. Prof Islam is also the Dean International for the Faculty of Science and Engineering at Curtin University. He is a member of the steering committee of the Australian Power Institute and a member of the WA EESA board. He is a Fellow of the Engineers Australia, a Senior Member of the IEEE IAS, PES and DEIS, a Fellow of the IET and a chartered engineer in the United Kingdom. He is an Editor of the IEEE Transaction on Sustainable Energy and an Associate Editor of the IET Renewable Power Generation. He was the Guest Editor in Chief for the IEEE Transaction on Sustainable Energy special issue on Variable Power Generation Integration into Grid.

Invited Paper

Today's Industry Applications of Power Electronics and Update on IEEE's Industry Applications Society

David B. Durocher

Senior Member IEEE
President, IEEE Industry Applications Society
Eaton Corporation, USA



Abstract

This presentation will focus on industry applications for a) A medium voltage adjustable frequency drive using high-efficiency IGBTs installed as an upgrade for a steel mill runout table transfer system resulting in monthly energy savings of US \$45,000, b) Application of large scale three-phase uninterruptible power supply (UPS) systems applied in a Greenfield Data Center, delivering high-efficiency power conversion in one of the world's highest energy density systems and c) The role that power electronics will play in tomorrow's Microgrid applications for industry and how users can capitalize on the future of open energy markets. This will be followed by a review of the IAS's last 50 years of accomplishments and a video outlining the Society's path forward into the future.

Short Biography

David B. Durocher (SM '77 M '97-SM '99) received a B.S.E.E. degree from Oregon State University, Corvallis, in 1978. He is currently Global Mining, Metals & Minerals Industry Manager, with Eaton Corporation in Wilsonville. He has over 35 years of experience with Westinghouse and Eaton serving in a variety of product engineering, sales and global marketing roles, authoring numerous technical papers that have been presented at conferences around the world and published in IEEE Industry Applications, Plant Engineering and EC&M Magazine. Dave is a Senior Member IEEE and presently serves as a member of the IEEE IAS Mining Industry Committee, Cement Industry Committee, Pulp & Paper Industry Committee and as President of the IEEE Industry Applications Society.

Invited Paper

IAS Chapters and Membership – Overview – Products and Services

Dr.-Ing. Peter Magyar

Fellow IEEE
Chair, Chapter and Membership Development,
IEEE Industrial Application Society



Abstract

In this talk, about IEEE Industrial Application Society chapters and membership - overview, product and services will be discussed.

Short Biography

Dr. Peter Magyar (IEEE M'91, SM'04, F'07) received the Dipl. eng. and the Dr. tech. degrees from the Budapest University of Technology, Hungary, in 1967 and 1975, respectively. From 1967 to 1991, he was with the Department of Automation of the same university as an assistant professor and a research associate. From 1982, he led the Laboratory of Microprocessor Controlled Drives of the same department. The profile of the laboratory was education and industrial research and development in the field of digitally controlled power electronics systems. Dr. Magyar has been author and co-author technical papers and books and filed more than 30 patents. His research interest focused on control and mathematical modelling of electrical drives and power converters. In 1981, he was recipient of the Alexander von Humboldt fellowship, Bonn, Germany, and spent a research year at the Braunschweig University of Technology, Germany. From 1991 to 2003, he was head of development power electronics and drives with Hanning-Works and with D-Tech Ltd., both Germany. Currently, he is retired from Hella KGaA Hueck&Co., Germany, where he was with the Advanced Development Department from 2004 to 2009. His working field has been electrical drives, car steering control and automatic car parking manoeuvre systems. He is working as consultant in the field of electric cars and electrical drives. He has been expert evaluator of the European Commission (2003, 2010-). Dr. Magyar has been member of the Hungarian Association of Electrical Engineering (1964-) and of the IEEE Societies CS, IA, IE, PEL (1992-) and PE (2008-) as well as the IAS Technical Committees Industrial Drives (2003-) and Industrial Power Converter (2005-). He is recipient of the IAS Distinguished Service Award 2006. Dr. Magyar has been elected to the grade of IEEE Fellow for contributions to digital control of electrical drive systems (2007). Dr. Magyar has been IEEE volunteer in various positions. In the *Germany Section*, he was secretary (1998-2000), vice chair (2001-2002) and chair (2003-2004) of the IAS/PELS/IES German Chapter, chapter coordinator (2005-2010) and nominations and appointments officer (2007-2012) of the Section's Executive Committee. In the *R8 Committee*, he was Division II representative (2007-2008) and chair (2009-2010) of the R8 Chapter Coordination Subcommittee. At *IEEE TAB*, he was member (2010) of the IEEE TAB Nominations & Appointments Committee. In the *IA-Society*, he was IAS chapters area chair R8 East & South (2005-2008), member of the IAS Executive Board as member-at-large (2005) and chair of the IAS Inter Society Cooperation Committee (2006-2008). Currently, he has been chair (2009 -) of the IAS Chapters and Membership Development (CMD) Department (<http://ias.ieee.org/chapters-a-membership.html>) and IAS liaison (2011-) of the R8 Chapter Coordination Subcommittee. His IEEE volunteer working field covers the chapter and member activity of IAS in general, especially leading 8 CMD committees related to conference support, chapter development and promotion, membership development and member promotion for ~12,500 members, 142 section chapters and 123 student branch chapters.

Abstract of Contributed Papers

DESIGN & SIMULATION OF PHOTOVOLTAIC, WIND, BATTERY, GRID CONNECTED HYBRID POWER SYSTEM FOR PATENGA, CHITTAGONG, BANGLADESH

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Abstract— Bangladesh is a developing country. Electricity shortage is the major obstacles for a developing nation. Currently, about 68% of the total population has access to electricity including public & private sectors. Electricity demand is increasing day by day. To overcome this situation, government has taken necessary steps. For instance, establishing various power plants which are non-renewable; such as oil, gas, coal, and many more. In addition, government has taken steps to establish a nuclear power plant in Ruppur and thermal coal based plant in Rumpal & Moheshkhali. On the other hand, due to geographical position, nuclear power plant is not suitable for Bangladesh. Because of geographical position, nuclear power needs a huge technological safety. Bangladesh is far behind this safety. Therefore, renewable energy is very efficient for overcoming this power crisis.

We proposed grid connected hybrid system by utilizing two different energy sources to mitigate the rising demand of electricity, especially in rural or remote areas of Bangladesh which can be used for domestic as well as commercial applications. In Bangladesh, about 72% of the populations live in the rural area and renewable energy is the right choice for providing clean energy.

Keywords— Hybrid system; Photovoltaic; Wind energy; Battery; Grid Connection; HOMER

I. INTRODUCTION

We didn't create energy. It can only be transformed from one form to another form. The major power resource in Bangladesh are natural gas it's about 66% and the other are oil, hydropower and coal. The gas reserve has fallen after 10 year. So, it is very necessary for us to create different scope for producing electricity. Bangladesh has a lot of resource for renewable energy and the natural availability of alternative energy creates opportunities of improve in power crisis. In Bangladesh, only 68% of the population has access to electricity with a per capita availability of 136kWh per annum[1]. In 2015 our total daily generation are 8500MW. In summer load shedding is about 500MW[2]. After 15 year in 2030 our demand will be 30000MW[3]. If we won't take proper step, we can't solve the problem. Most of the area's in Bangladesh are rural area. There are no grid connection in this area. we choose two renewable

energy as a primary source because the resource are environment friendly and won't finished. We proposed a renewable plant & grid connected system to overcome our power demand. There are a lot of private organization in Bangladesh like as GS(Grameen Shakti), IDCOL (Infrastructure Development Company Limited), Rahim Afroz, CDR (Center for community Development & Research) foundations, BRAC (Bangladesh Rural Advancement Committee), are working to develop our power sector[4].

II. RESOURCE DATA ANALYSIS IN PATENGA

Patenga is located at 22 degrees 14.73 minutes North latitude and 91 degrees 47.24 minutes East longitude. We collect one year 2014 climate data and wind data from Bangladesh Meteorological Department, Patenga airport branch [5].also collect one year solar radiation data from NASA

Table I. GHI value for Patenga, Chittagong

Month	Clearness Index	NASA Daily Radiation (KWh/m ² /day)	Weather Office Daily Radiation (KWh/m ² /day)
January	0.593	4.597	4.240
February	0.775	5.126	6.350
March	0.824	5.634	7.800
April	0.774	5.760	8.110
May	0.706	5.786	7.750
June	0.454	4.335	5.040
July	0.508	4.048	5.590
August	0.407	4.224	4.320
September	0.478	4.083	4.670
October	0.489	4.725	4.180
November	0.938	4.409	6.910
December	0.772	4.391	5.250
Annual Average	0.643	4.76	5.85

Table II. Wind speed data for Patenga, Chittagong

Month	Weather Office Average wind speed (m/s) Height(30 feet)
January	2.220
February	2.220
March	2.500
April	2.500
May	3.050
June	2.780
July	3.050
August	2.780
September	2.500
October	1.390
November	1.940
December	1.670
Annual Average (m/s)	2.38

III. PROPOSED RENEWABLE HYBRID SYSTEM

We propose a hybrid system consists of two renewable sources (solar, wind), load , battery , converter & grid connected system.

Fig. 1 shows the complete system model.

Our target is producing 50KW electrical power in an hour. In addition, the power will be supplied to load and extra power will be supplied to national grid to contribute national electricity demand.

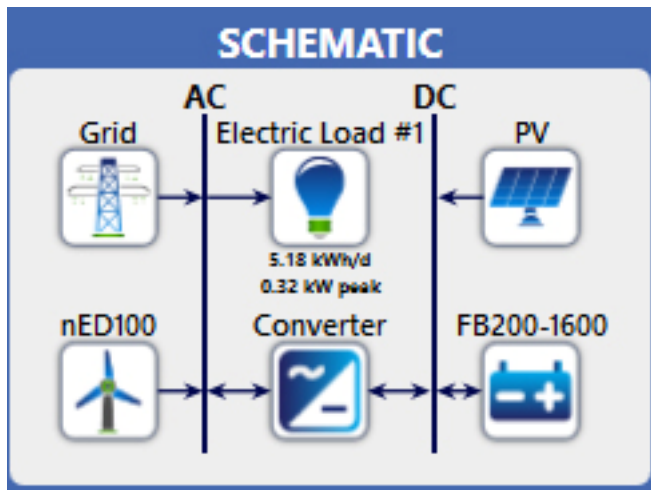


Fig.1 Hybrid Renewable System

IV. System Component Specification For Simulation

A. Solar Energy

We collect January 2014 to December 2014 monthly global radiation data from Bangladesh Meteorological Department Patenga airport branch. Fig.2 shows monthly average radiation.



Fig.2 solar radiation 2014

B. Wind Energy

A wind turbine is a machine that converts the kinetic energy in wind into mechanical energy [6]. we setup a wind turbine which is usually 20m high. Wind comes from atmospheric changes; changes in temperature and pressure makes the air move around the surface of the earth; all of which is triggered by the sun [7]. If wind speed increases, power output increases dramatically. We cannot convert all the wind energy into electricity: we can convert only 59%, according to Betz limit [8].

The output equation for a wind generator is given by:

$$P = (1/2) \times \rho \times A \times v^3 \text{ (in Watts)} \dots \dots \dots (1)$$

Where,

A=area perpendicular to the direction of flow (in m²), v=wind velocity (m/s), ρ =density of air (in Kg/m³) and P=power generation. Fig.3 shows monthly average wind speed data.

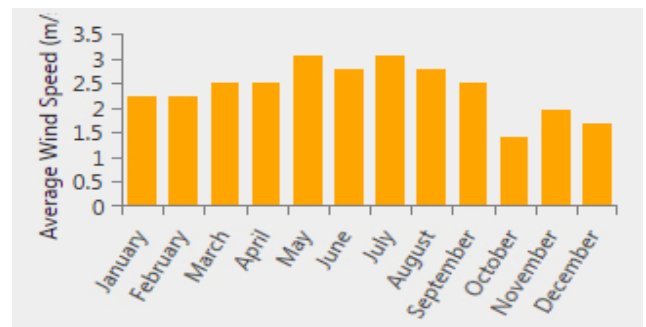


Fig.3 Average monthly wind speed data

C. Battery [9]

Nominal Voltage (V): 700
Nominal Capacity (Ah): 2285.71
Nominal Capacity (kWh): 1,600.00
Round Trip Efficiency (%): 65
Float Life (years): 20
Maximum Capacity (Ah): 3,297.384
Capacity Ratio, c: 0.505
Rate Constant, k: 0.801
Suggested Life Throughput (kWh): 17520000
Max. Charge Rate (A/Ah): 1
Max. Charge Current (A): 230.35
Max. Discharge Current (A): 354.385

D. Electric load

Our target is to produce electricity 50KW per hours, but we used only a household load demand, which collected from monthly energy bill.

Metric	Baseline	Scaled
Average (kWh/d)	5.18	5.18
Average (kW)	.22	.22
Peak (kW)	.32	.32
Load Factor	.68	.68

Load Type: ☒ AC ☐ DC

Annual Average (kWh/d): 5.18

Fig.4 A household load demand

E. Grid Extension

We need extension to connect the system to National Grid. The cost will increase or decrease with the increase or decrease of the distance between propose plant & National grid.

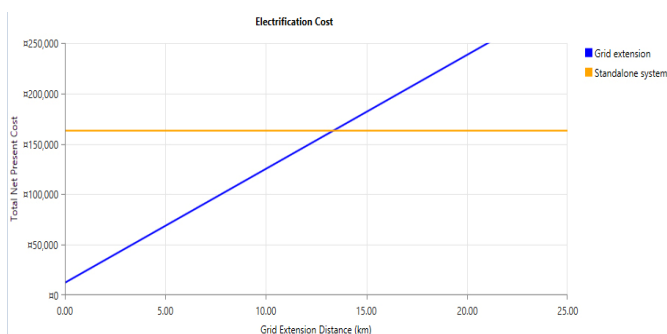


Fig.5 Grid Extension Curve

V. OPTIMIZATION RESULT

We use HOMER Pro software for simulation. We give input data of solar radiation, wind speed and load on the basis of HOMER software requirement. Fig.6 shows simulation result by HOMER Pro

Sensitivity Cases: Left Click on sensitivity case to see optimization cases.									
Architecture	Grid (kW)	EnergySection® (kW)	Dispatch	COE (\$)	NPC (\$)	Operating cost (\$)	Initial capital (\$)	Ren Frac (%)	System
PV (kW) 100.0 8 1	FB200-1600	999.999 50	CC	(\$0.040)	(\$190,469)	(\$16,602)	\$24,150	100	
Optimization Cases: Left Double Click on simulation to examine details.									
Architecture	Grid (kW)	EnergySection® (kW)	Dispatch	COE (\$)	NPC (\$)	Operating cost (\$)	Initial capital (\$)	Ren Frac (%)	System
PV (kW) 100.0 8 1	FB200-1600	999.999 50	CC	(\$0.040)	(\$190,469)	(\$16,602)	\$24,150	100	

Fig.6 Simulation by Homer Pro

Production			Consumption			Quantity		
kWh/yr	%		kWh/yr	%		kWh/yr	%	
Generic flat plate PV	186,036	45.17	AC Primary Load	1,890	100.0	Excess Electricity	409,672.0	99.5
Norvento nED 100	225,809	54.83	DC Primary Load	0	0.0	Unmet Electric Load	0.0	0.0
Total	411,844	100	Total	1,890	100.0	Capacity Shortage	0.0	0.0

Quantity	Value
Renewable Fraction	100.0
Max. Renew. Penetration	449,741.0

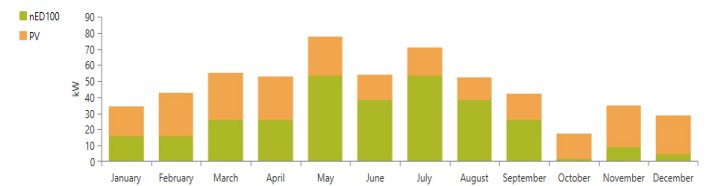


Fig.7 Electricity Production curve

The following result are obtained

Component	Production (kwh/yr)	Friction %
PV	186,036	45.17
Wind turbine	225,809	54.83
Total	411,844	100

Load	Consumption (kwh/yr)	Fraction %
AC	1,890	0.45
DC	0	0
Grid sell	409955	99.55

We connect the system to grid for commercial purpose

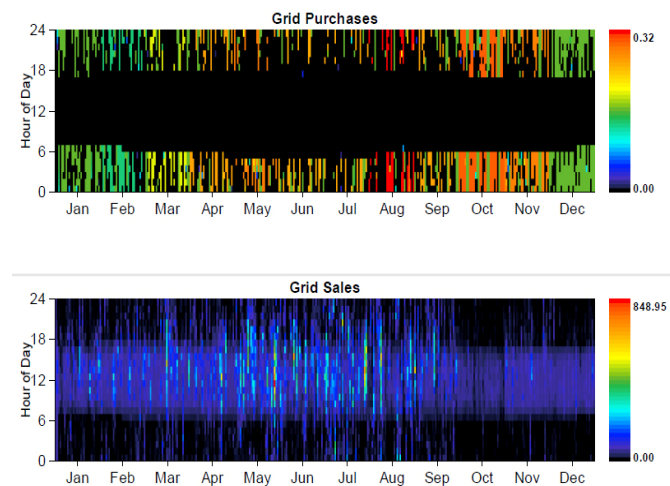


Fig.8 Grid Purchases & sales curve

VI. PROPOSED CONTROL SYSTEM DIAGRAM

Our main target is to supply electricity 50kw per hours. But we are failed to get 50kw per hour constantly from HOMER. So we control the system using programmable logic controller like as Siemens 200. We use two energy meter as sensor ,it's count energy then PLC device will control the system.

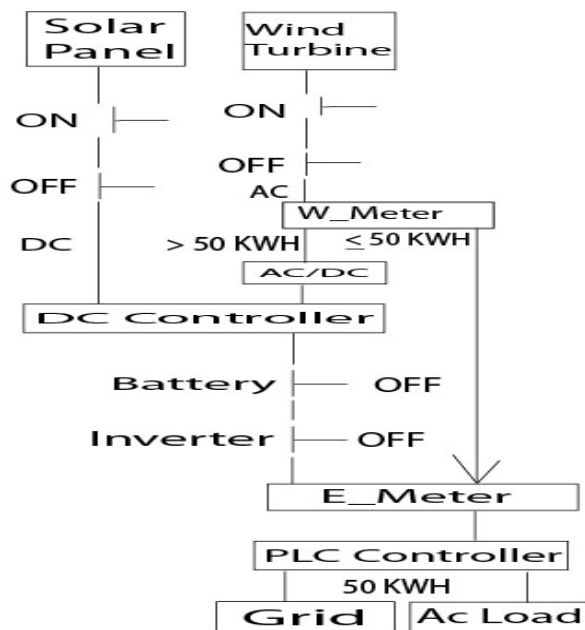


Fig.9 Proposed control diagram

VII. COST

currently 1\$ = 80 BDT

A. PV: Generic flat plate PV

Size	Quantity	Capital	Replacement	O&M(yr)
100KW	150w*667	40000\$	30000\$	100

B. Wind turbine

Size	Quantity	Capital	Replacement	O&M(yr)
10KW	8	80000\$	50000\$	1000

C. Battery

Size	Quantity	Capital	Replacement	O&M(yr)
1600kwh	1	80000\$	50000\$	100

D. Extra

Name	Size	Capital	Replacement	O&M(yr)
Converter	80kw	1000\$	500\$	none
Inverter	50kw	1000\$	500\$	none
Grid Extension	10km	1000\$		100\$
PLC	1pcs	500\$	300\$	None

we propose a hybrid plant for 25 year , battery will replace after 15-20 years. fig. shows the overall cost curve

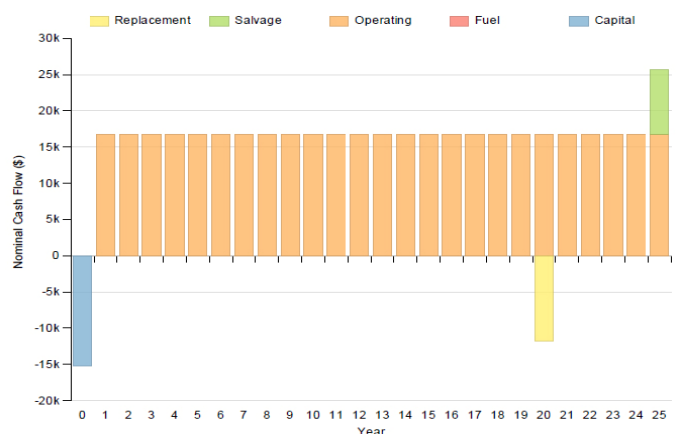


Fig.10 the overall cost curve

VIII. CONCLUIONS

This model is effective than existing conventional plant in any areas in Bangladesh. The hybrid system is not only reliable also profitable in business sense. It also helps to reduce our power crisis and provide clean energy. It does not contain any risk for production & environment. We also collected wind data at 30 feet but simulate at 20 m. If we setup the wind turbine at 20 meter, the production of electricity will increase.

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Integration of Smart Hybrid Uninterruptible Power Supply

Humdan Islam, Md Mujahidul Islam, Md Rezaul Hasan Rony

Abstract— In our country meeting the demand of electricity is a very big concern. The crisis of the electricity is mainly due to increasing of load, lack of proper maintenance and other factors. To worsen this situation further, our resources are limited as well. Moreover our electricity penetration in rural areas has not improved yet. The fact that only 10% of the rural households have electricity connection and there are some parts of Bangladesh which will not get the access of electricity connection from the national grid within next 30 years [1] does not help the cause either. To overcome this crisis many projects and research has been done particularly focusing on renewable energy resources. The project in this paper illustrates the effective ways of utilizing solar and hydro energy to provide uninterruptible power supply (UPS) scheme during load shedding period also it could be used as standalone system with a minor adjustment to generate electricity. The whole project is a miniature version which could be implemented on large scale.

Keywords— Hydro, Solar, GSM Module, Twin Charger Controller, Relay Switching, Voltage Regulator, Inverter, Microcontroller

I. INTRODUCTION

The main focus of our project is to utilize the renewable resources to provide uninterruptible power supply (UPS). In this project we concentrated particularly on solar and hydro energy. The energy generated from these sources would be used to charge batteries. Then the DC voltage would be converted to AC voltage using an inverter. Since we are intending to provide uninterruptible power, in order to do this the inverter would be coupled with the main supply in such a way that whenever the main supply fails the inverter would provide power to the load. Furthermore a GSM module will be integrated as well. This module will help to turn on or off the UPS manually using short message service (SMS) when the batteries are fully charged or discharged. Potential benefit of such project is that the source of energy is free because renewable resources are used. Another major advantage is that, since our project would employ three batteries, the likelihood of failing to produce continuous power is further reduced.

II. WORKING PRINCIPLE

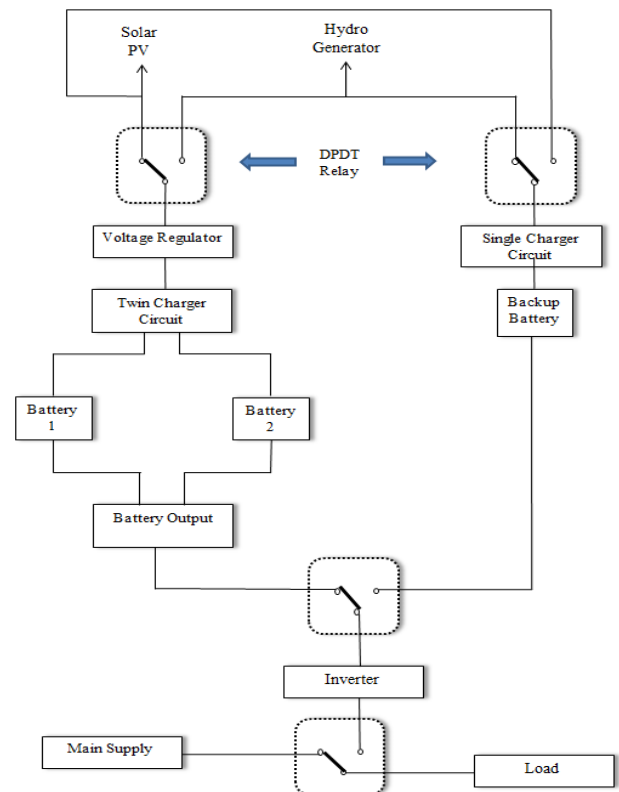


Figure 1: Block Diagram of the UPS

The block diagram above illustrates the overall project in detail. At the top, a solar panel of 18V and 10W as well as two micro hydro generator of 3.5W each [2] has been connected with a DPDT (Double Pole Double Throw) relay. The DPDT relay has a rating of 10Amp and 28Volt [3]. The voltage regulator circuit is connected with solar panel in normally close position (N/C) and the micro hydro generator is connected in normally open position (N/O) respectively. The output of the voltage regulator is connected with the input of twin charger circuit. The purpose of the voltage regulator is to fix the voltage at 14Volts since a 12Volts battery is used in this project.

On the other hand, the single charger circuit is connected with solar panel in N/O and in N/C with the micro hydro generator. The coil of the DPDT relay does not have any polarity [4]. This coil is energized by voltage generated from hydro generator. The purpose of the single charger circuit is to charge the backup battery (12Volt) which acts as tertiary protection (primary and secondary protection is provided by Battery 1 and Battery 2 respectively). Considering a situation where both solar panel and hydro generator is operational, the coil will be energized by the hydro generator hence voltage regulator will be connected

with hydro generator (N/O position) and single charger will be connected with solar panel (N/C position).

An inverter has been used here to convert DC to AC. The inverter is connected with the twin battery charger output in N/O position and in N/C position with the output of the single charger circuit. The SPDT (Single Pole Double Throw) relay has a rating of 10Amp and 12Volt [5]. This SPDT relay has been functioned in such a way that when Battery 1 & 2 reaches lower threshold, the inverter will be automatically connected to backup battery. To achieve this function of the SPDT relay, the coil energization of the relay was controlled by a comparator.

The output of the inverter is connected with the load using a SPDT relay (N/C position). Main supply is connected with the load in N/O position. The relay coil is energized by main supply. Suppose a situation arises when main supply is off due to load shedding or any other reason, the load will be automatically connected to the inverter output. The inverter will provide uninterruptible power as long as Battery 1, 2 and backup battery does not run out.

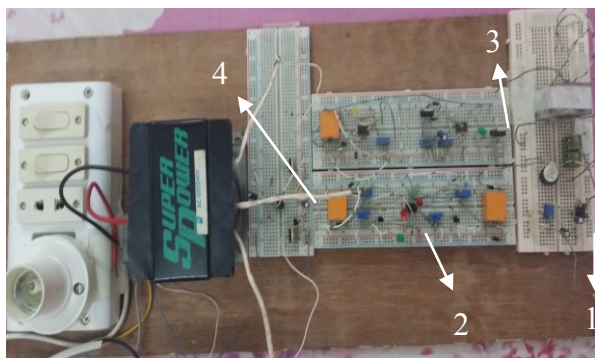


Figure 2: Overview of the UPS

- 1) Voltage Regulator
- 2) Twin Charger Circuit
- 3) Single Charger Circuit
- 4) Inverter



Figure 3: Mico Hydro Generator



Figure 4: 10W 18V Solar Panel

A. Twin charger circuit

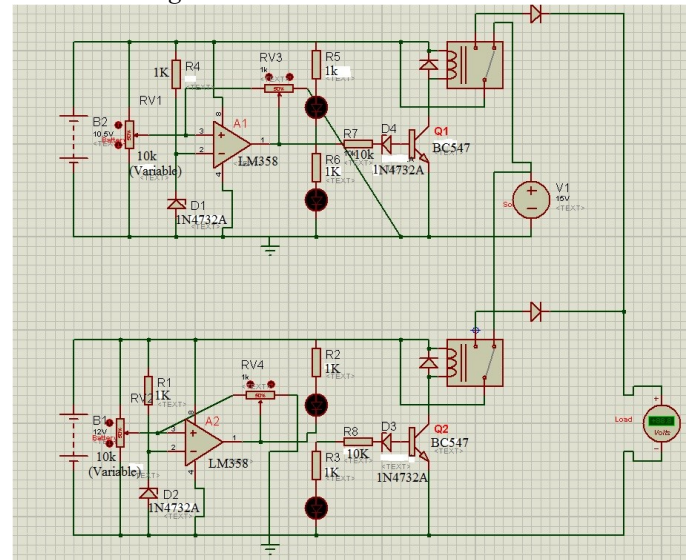


Figure 5: Circuit Diagram of Twin Charger Controller in Proteus Simulation Software

The twin charger controller consists of dual operational amplifier (opamp) IC LM358 etc. A variable resistor of 100kilo Ohm has been connected between inverting pin and output pin of each of the opamp. This variable resistor is used to set the upper cutoff [6].

Another variable resistor of 10 kilo Ohm is used to set lower cutoff. Green LED and Red LED indicates battery charged and discharged condition respectively. When any of the battery is discharged it will be connected to input source (N/C position). When any of the battery is fully charged then it will be connected with the input of the inverter (N/O position) [7]. If both the batteries are charged then ampere hour (AH) rating will be doubled. Both of these batteries are 12Volts 1.35A and 4.5AH.

B. Single charger circuit

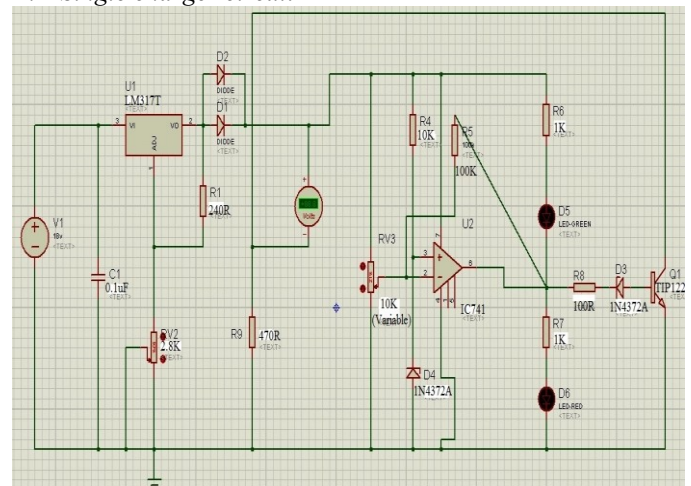


Figure 6: Circuit Diagram of Single Charger Controller in Proteus Simulation Software

The single charger controller consists of IC LM317, TIP122 and opamp 741. 2.2kilo Ohm resistor is used to set the threshold point (14V) [8].

C. Inverter

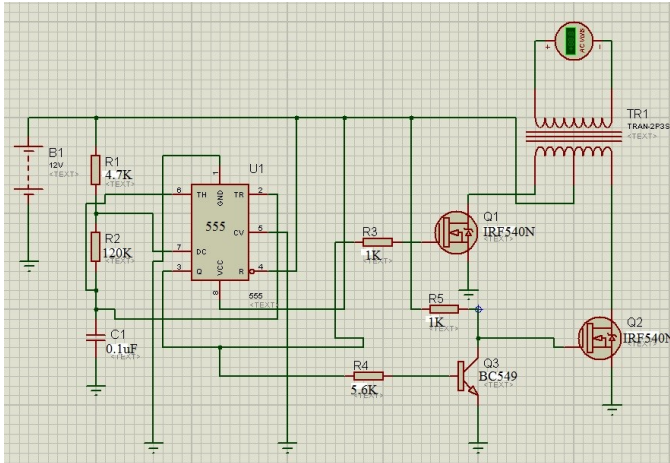


Figure 7: Circuit Diagram of Inverter in Proteus Simulation Software

The inverter in this project is a square wave inverter. The inverter circuit includes 10Amp step up (center tapped) 12V-220V transformer, Transistor BC549, IC555 and IRF540N etc. IRF540N is a silicon controlled rectifier (SCR). SCR is used for switching purpose at high voltage [9].

D. Implementation of GSM

One of the major feature of this project is the integration of GSM module. This module is coupled with relay shield and microcontroller. Advantage of using GSM module is that it can be used to switch between main supply and the UPS even if the UPS is placed in remote areas. Another advantage is that when the UPS battery is fully charged also main supply is available (e.g. no load shedding occurs), at that moment using the microcontroller and GSM module the load can be fed by UPS. This ensures that the pressure on the main supply is eased.

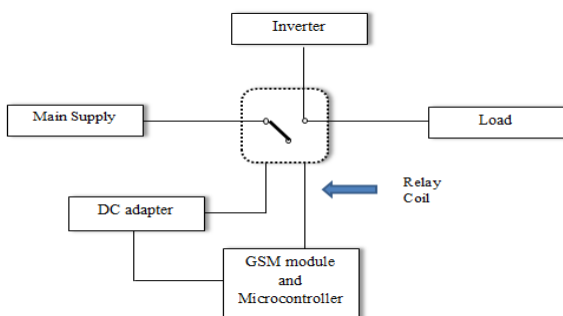


Figure 8: Block diagram of integration of GSM and Microcontroller into the UPS.

The principle function of the GSM module and microcontroller is to control the energization of the SPDT relay which connects load to the main supply or UPS. DC adapter is used energize the coil and hence to keep the relay in N/O position (i.e. load connected to main supply).

III. SIMULATION & CALCULATION

Voltage generated from micro hydro generator was recorded against flow speed of water. A water flow sensor was used in tandem with a microcontroller to check the water flow speed. Water flow sensor consists of a plastic valve body, a water rotor, and a hall-effect sensor. When fluid flows through the rotor, the rotor rolls. Its speed changes with different rate of flow. The hall-effect sensor outputs the corresponding pulse signal [10]. The voltage was recorded for a single hydro generator.

- Flow Rate = (Pulse frequency × 60)/7.5Q
 - ❖ Pulse Frequency is the pulse signal obtained from the output of the water flow sensor
 - ❖ 'Q' is the flow rate in L/min
- Input Power from source (Theoretically)= 3.5W×2 + 18.25W= 25.25W
- Transformer Output (Theoretically)= 10A×12V=120W
- The output current (hence the output power) varied when applying different types of load.

TABLE 1: WATER FLOW RATE VERSUS VOLTAGE

No.	Flow Speed (in Liter per hour)	Voltage (volt)
1	50	0
2	100	0
3	150	1.3
4	200	2.4
5	250	3.8
6	300	4.7
7	350	6.23
8	400	7.3
9	450	8.35
10	500	9.4
11	550	10.3
12	600	11.15
13	650	11.9
14	700	12.8
15	750	13.5
16	800	14
17	850	14.5

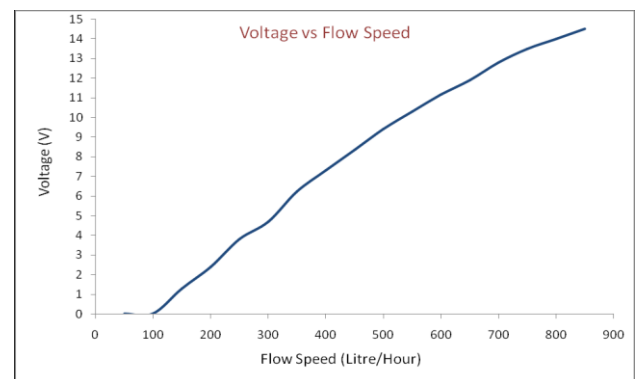


Figure 9: Graph of Voltage versus Flow Speed

IV. DEVICE SPECIFICATION

A. Micro Hydro Generator

- Water inlet/outlet: 1/2"
- Output Voltage: DC9.8~18.5V/no-load,
- DC 8.8 - 15V/0. 1K load Max power: 3.5W
- Output Current: 128 to 260 MA/0. 1 K load
- Water Pressure: 0.08 ~ 0.45MP
- Maximum Water Pressure: 0.55MP
- Maximum Pressure: 1.6MP
- Dimensions: 85.4 x 42.5 x 81.4mm

B. Water Flow Sensor

- Minimum Working Voltage: DC 4.5V
- Working Voltage: 5V~24V
- Flow Rate Range: 1~60L/min
- Load Capacity: ≤10mA(DC 5V)
- Operating Temperature: ≤80°C
- Water Pressure: ≤2.0MPa

C. Solar Panel

- Max. Power(Pm): 18.25W
- Open Circuit Voltage (Voc): 21.96V
- Max. Power Voltage (Vmp): 18.25V
- Max. Power Current (Imp): 0.55A
- Short Circuit Current (Isc): 0.59A
- Solar cell: Poly crystalline

D. Arduino Compatible Relay Shield

- Rated coil Voltage: 5V DC
- Operating Frequency: 50/60HZ
- Nominal Current: 10A at 125V AC, A at 10A at 28V DC

III. APPLICATION

A. Urban Use

In urban areas solar panel could be placed on the roof top and adjusting the tilt angle equal to latitude of the location and also facing it due south to gain maximum solar radiation [11].

On the other hand, micro hydro generator could be installed in association with the overhead water reservoir on the roof. Water flows constantly from reservoir and there is a continuous fluid force available which could be utilized to drive the generator. To be more precise the micro hydro generator should be connected with the reservoir in a close loop. This ensures that water takes a parallel path into the generator from the reservoir and then again returns to the main line.

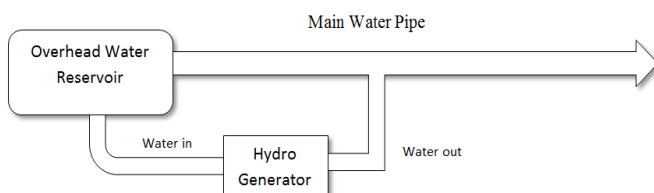


Figure 10: Installation of hydro generator with overhead water reservoir

B. Rural Use

The solar panel could be placed in similar manner used for urban use.

Since water pumps are frequently used in rural areas for irrigation purpose. There is a fluid force available when water is being pumped out. This force of water is being completely unutilized. Micro hydro generator could induce voltage using such force.

C. As Standalone system

By removing the circuitry involving connection of main supply, this UPS can be used as standalone system in remote areas.

V. COST CALCULATION

Transformer = 600 BDT

Battery= 3×280= 840 BDT

Micro Hydro Generator= 2×1100= 2200 BDT

Solar Panel= 1000 BDT

IC, BJT, MOSFET, Resistors, Capacitors etc. = 500 BDT

Relay Shield= 450 BDT

GSM Module=4000BDT

Breadboard=4×220= 880 BDT

Arduino Uno=800 BDT

Grand Total= 11,270 BDT

- ❖ All calculations were done in Bangladeshi currency
- ❖ Cost estimated here was a rough estimation (cost may slightly vary)

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Coordination Control of a Hybrid Ac-Dc Micro-grid at Different Generation and Load

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Abstract—The hybrid micro-grid is developing because of providing reliable electricity from renewable energy sources. This paper introduces hybrid AC/DC micro-grids which will help to develop distribution generation system and the interconnection between distribution generation and customer. Here, a hybrid smart micro-grid system is designed in MATLAB (SIMULINK) with some renewable energy sources (wind, PV system, fuel cell). This system will have the ability to deliver electrical energy on both micro grids variation of generation and load. Different coordination control systems are proposed for assuring stability under various generations and load conditions. The simulation results show proper power sharing among AC micro-grid, DC micro-grid and utility grid. AC and DC bus voltages remain steady both in autonomous and grid-tied mode.

Keywords—renewable energy sources; distribution generation; coordination control system; grid tied mode; autonomous mode.

I. INTRODUCTION

AC using Renewable energy is necessary because of reduction of fossil fuel and high price of oil. Also, increase environmental problem by products from conventional fossil fuel, need a new energy system. Development of renewable energy is an important advantage to fulfill electricity demand [1].

Micro-grid system usually optimizes operation (AC to DC, DC to AC). This helps to reduce conversion losses and coordinating with generation and load resources [2]. This micro-grid system is proposed with some renewable energy sources like wind, PV panel and fuel cell. This hybrid micro-grid system helps to fulfill the load demand. One very appealing technology for grid connected operation is tie line controls, which will regulate the active and reactive power flow between the Micro-grid and the utility grid at the point of interconnection. These controls essentially allow the Micro-grid to behave as an aggregated power entity that can be made dispatch able by the utility [2].

The conventional electric generation systems with fossil fuel have existed over 100 years for efficient conversion of power at different voltage level [3]. Recently, regional distribution networks with renewable sources directly made a connection with the substation and consumer [4]. The main advantage of using this kind of network is made connection with various AC and DC generation and load. To achieve this,

power electronics play an important role to coordinate control the system.

II. HYBRID MICRO-GRID

Interconnection of small distribution generation establishes a new concept “micro-grid”. Micro-grid is a system that have distribution energy resources, load, control system and ability to work on both grids tied mode and isolated mode [5]. The micro-grid can be connected to the electrical distribution network generates and distribute electricity to the customer. A hybrid micro-grid can be involved with different kind of sources like wind, PV cell, fuel cell, bio-mass etc. with energy storage batteries [6]. It has the ability to use local distributed generation and produced waste heat can be reused for coal fired generation. It has ability reused the by-products before it goes into the environment. For establishing micro-grid system, proper controlling, monitoring and management are needed.

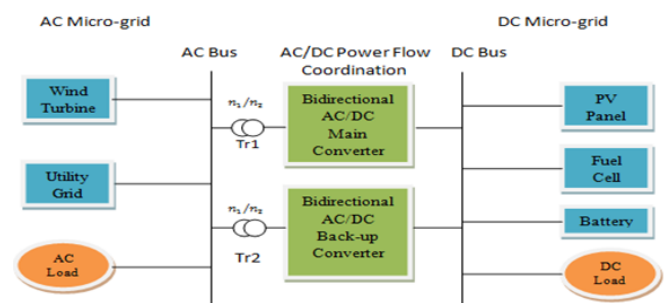


Fig. 1. Proposed hybrid AC/DC micro grid system

III. MODELING OF THE SYSTEM

A. Proposed smart micro-grid

The configuration of the proposed hybrid micro grid system shown in Fig. 1, where various kinds of AC and DC sources as well as loads are connected to the corresponding AC and DC networks. AC and DC sides are connected together through bidirectional converter. The AC bus of the hybrid grid is tied to the utility grid.

Fig. 2, is the Simulink (MATLAB) model of the hybrid micro grid, where boost converter is used to connect PV arrays and fuel cell to the DC bus. A DFIG wind generator is connected to AC bus in order to simulate AC sources. Various AC and DC loads are connected to the AC and DC buses for

simulating loads. PV modules are connected both in series and parallel. Again fuel cells are connected both in series and parallel. A wind generator system consists of doubly fed induction generator (DFIG) with AC/DC/AC converter. The AC and DC buses are coupled through bidirectional converter in order to exchange power.

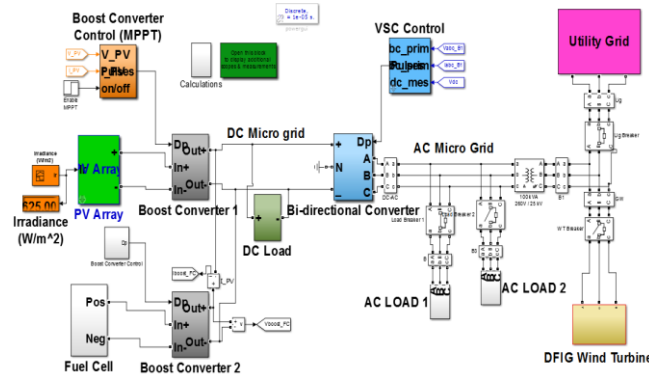


Fig. 2. Simulink (MATLAB) model of hybrid micro-grid

B. Operation of Grid

The hybrid micro grid performs its operations in two ways. In grid tied mode the bidirectional (VSC) converters provides stable DC bus voltage as well as synchronize the AC bus voltage with utility grid in this mode. It also exchanges power between AC and DC buses. When output power of DC sources is greater than DC loads this converter reacts like inverter and on this condition power flows from AC to DC side. When generation of total power is less than the total load at DC side, the converter injects power from AC to DC side. This converter helps to receive power from the utility grid incase the total power generation is less than the total load in the hybrid micro grid. Otherwise hybrid grid injects power to the utility grid. Battery role is not important as the power is balanced by utility grid [7].

In autonomous mode the battery can play a very important role for balance voltage stability [7]. In this thesis voltage stability has been ensured only by controlling loads rather than using battery. This means voltage will be stable only if total load is less than total generation.

C. PV Panel Model

PV the equation of the solar cell,

$$I_{pv} = I_{ph} - I_D = I_{ph} - I_0 \left(e^{\frac{q(v + IR_s)}{AKT_C}} - 1 \right) - (v + IR_s / R_{sh}) \quad (1)$$

The photon current,

$$I_{ph} = [I_{sec} + K_i(T_c - T_R) \cdot SI/1000] \quad (2)$$

Where, K is Boltzmann's gas constant (1.38×10^{-23} J/K), T_c is Absolute temperature of the cell (K), q is Electron charge (1.602×10^{-19} J/V), I_0 is Saturation current, v is Voltage of PV module, A is Ideal factor, T_R is Cell's reference temperature and SI is Solar Irradiance [3].

D. PEM Fuel Cell

Proton exchange membrane fuel cells, also known as polymer electrolyte membrane (PEM) fuel cell are used in this paper [8], [9].

E. Wind Turbine Generator Model

The DFIG wind turbine is used in this paper. The basic principles of DFIG are described in [3].

IV. Coordination Control of the System

A. Control of the Converters

The objective of control the boost converter of PV array is to achieve MPPT of PV array by regulating duty cycle of boost converter. Incremental Conductance (IC) method regulates the duty cycle of boost converter to track the MPPT of PV array. Since the PV panel output power is the product of voltage and current. MPP tracking is carried out by differentiating the Power with respect to voltage and setting it to be equal to zero to determine the expected operating point

$$\frac{dP_{pv}}{dV_{pv}} = I_{pv} + V_{pv} \frac{dI_{pv}}{dV_{pv}} \quad (3)$$

$$\frac{dI_{pv}}{dV_{pv}} = -\frac{I_{pv}}{V_{pv}} \quad [\text{at MPP}] \quad (4)$$

dI_{pv}/dV_{pv} is called maximum power identifier factor. Using this factor, IC proposed to track the MPP of the PV module [10].

The objectives of the bi-directional converter controller are to keep a stable dc micro-grid voltage for variable dc loads and to synchronize the ac micro-grid with utility grid. The control scheme for the bi-directional converter used in this paper is based on basic principles described in [11].

The control objective of boost converter for FC is to extract maximum power from FC. To achieve this a fixed duty cycle is selected by simulating and observing results for different duty cycle. This fixed duty cycle is 0.67.

One of control objectives of the back to back ac/dc/ac converter of DFIG is to regulate rotor side current to synchronize ac grid and the other objective is to achieve MPPT. In this paper direct torque control (DTC) scheme is used as the control method for the rotor side converter of DFIG. The DTC scheme is described in literature [12].

V. SIMULATION RESULT

Uncertain change in source and load conditions is created in this simulation. The grid tied mode as well as the autonomous mode are also simulated by disconnecting utility grid by circuit breakers. In grid tied mode power is balanced by utility grid. But at autonomous mode it is important to keep balance between loads and sources power. In this paper, power balance in autonomous mode is achieved by controlling loads. This simulation has been run for 10 sec., and system operates in autonomous mode from 2 sec. to 3 sec. and rest of the time it operates in grid tied mode.

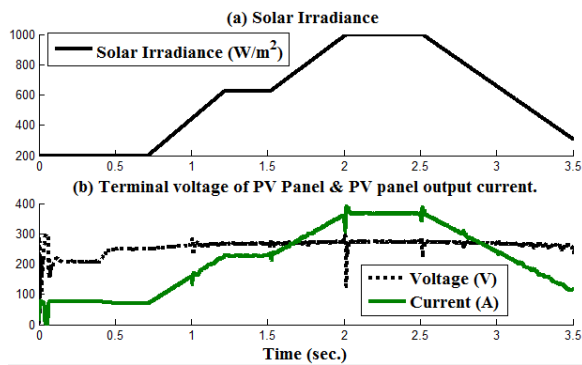


Fig.3. (a) Solar irradiance (b) Terminal voltage of PV panel and PV panel output current

In this paper Incremental Conductance(IC) algorithm is used for MPPT of PV based on the corresponding solar irradiation. The solar irradiance is shown in Fig. 3(a), and Fig. 3(b), shows PV panel terminal voltage and output current response. The terminal voltage is about 260 volts and it has some transient response from 0-0.4 sec. It can be seen that the output current varies to follow the solar irradiance and to achieve maximum power. The terminal voltage and output current of FC are shown in Fig. 4,

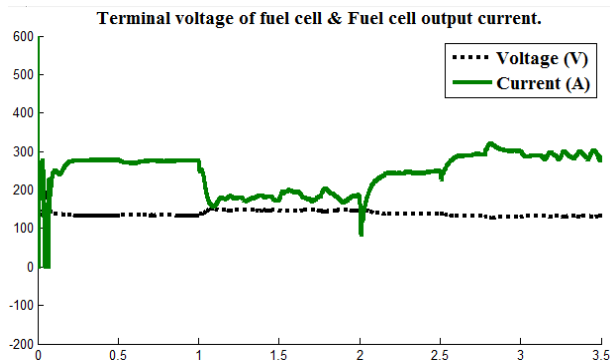


Fig.4. Fuel cell Terminal voltage and output current

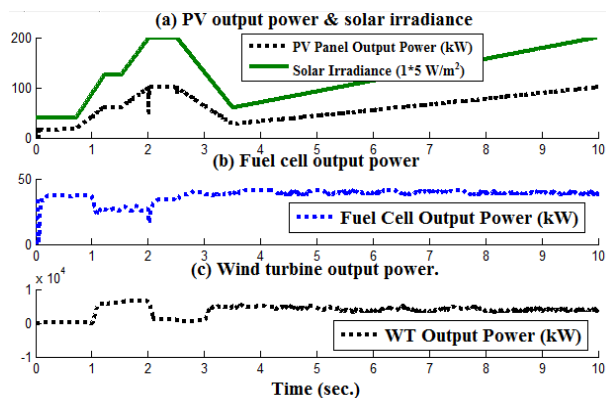


Fig.5. (a) Solar irradiance (radiation level times 1/5 for comparison) versus PV output power (b) Fuel cell output power (c) Wind turbine output power.

The solar irradiation (radiation level times 1/5 for comparison) and corresponding PV output power are shown in Fig. 5(a). The maximum output power of PV array is 100kW

for 1000 W/m^2 solar irradiation. Fig. 5(b), shows the FC output which is about 40-45 kW. Fig. 5(c), shows the WT output which is 4 MW. The WT was isolated from main system for 0 to 1 sec. That is why Fig. 5(c), shows zero output for 0 sec. to 1 sec. The main reason for isolation of WT is to verify that hybrid system can run with or without WT. Again WT output decreased to less than 700 kW from 2 to 3 sec because within this time utility grid was disconnected from systems so the useful WT power is equal total load minus sum of other generation. The isolation of hybrid grid from utility grid was made to verify the ability to operate in off grid mode. There are several transient responses in "Fig. 5," because of sudden change in loads and or system configuration. All these changes were made intentionally to verify the system stability.

Fig. 6(a), shows the DC generation versus load variation. Fig. 6(b), Shows power sharing between AC and DC micro-grid. In Fig. 6(a), at 2.3 sec. the dc generation is 144kW and dc load is 30 kW. At that time Fig. 6(b), shows the value +113, which represents that, at that time ac micro-grid receiving 113 kW power from dc micro grid. On the other hand at 0.6 sec. Fig. 6(a), shows the dc generation is 53kW and dc load is 185 kW. At the same time Fig. 6(b), shows the value -132, which represents that, at that time ac micro-grid injecting 132 kW power to dc micro grid.

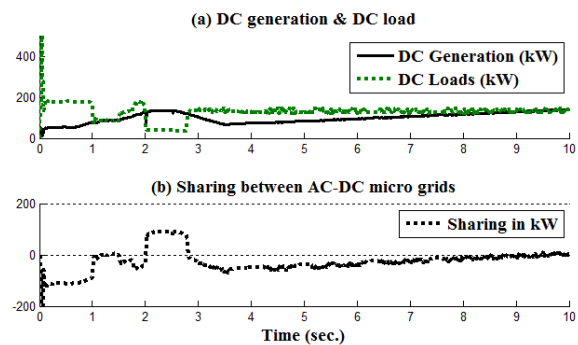


Fig.6. (a) DC generation versus DC load (b) Sharing between AC-DC micro grids.

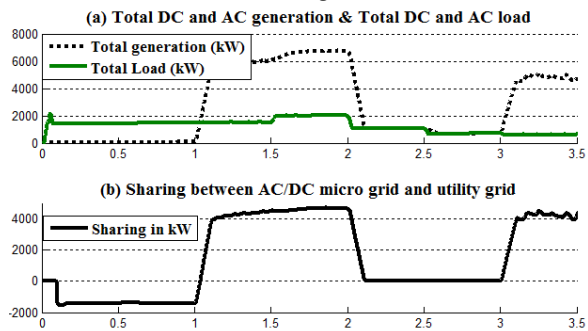


Fig.7. (a) Total DC and AC generation versus total DC and AC load (b) Sharing between AC/DC micro grid and utility grid.

Fig. 7(a), shows the total generation versus load variation. Fig. 7(b), Shows power sharing between hybrid grid and utility grid. From 2 sec. to 3 sec. the curve remains zero because at this time utility grid is disconnected from system and system is operating in autonomous mode. At autonomous mode there is no power exchange between utility and hybrid grid. In Fig. 7(a), at 0.6 sec. total generation is 53kW and total load is 1475

kW. At that time Fig. 7(b), shows the value -1422 kW which represents that, at that time hybrid grid is receiving 1422 kW power from utility grid. On the other hand, at 1.3 sec. Fig. 7(a), shows the total generation is 6000kW, total load is 1432 kW and Fig. 7(b), shows the value 4568 kW, which represents that, at that 4568 kW power is being shared from hybrid grid to the utility grid.

Fig. 8, shows the AC micro-grid RMS voltage which is about 200- 210 Volt and the DC bus voltage. DC micro-grid voltage which is about 400 Volt. Both AC and DC voltage have transient response because of load or system configuration change. But it can be seen that DC bus voltage is more sensitive than AC bus voltage.

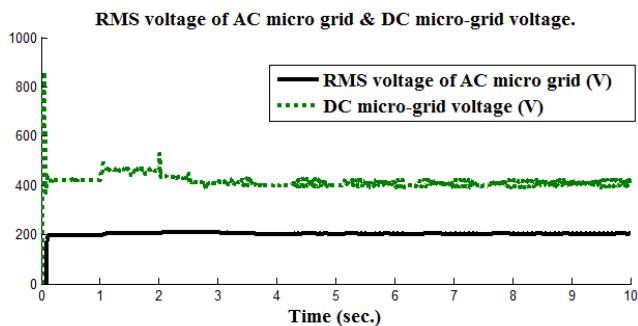


Fig.8. Transient response of DC bus voltage and AC bus RMS voltage.

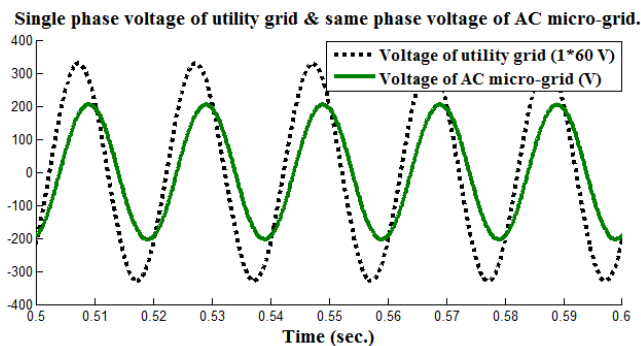


Fig.9. Single phase voltage of utility grid versus same voltage of AC micro-grid

Fig. 9, shows single voltage of utility grid and AC micro-grid. By comparing AC quantities it can be seen that AC micro-grid has a constant 13 degree phase difference from utility grid and both of them have same frequency of 50 Hz.

VI CONCLUSION

In this paper the hybrid grid system configuration is simulated in MATLAB environment. The goal is to accelerate realization of the main benefit to use renewable energy. The hybrid grid may be effective for small isolated industrial plants with PV systems, wind turbine and fuel cell as the major power supply. This system reduce back to back ac-dc-ac conversion lose and long distance transmission lose and increase energy efficiency. As this system is not totally dependent of utility grid, it also improves reliability.

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Development of an LCL Filter for a Residential Microgrid System

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Abstract— Energy crisis is becoming one of the major challenges in today's world due to increased energy demand and at the same time decreased energy resources. Energy from renewable sources can never be exhausted. They cause less emission and therefore stand out as a potentially viable source of clean and limitless energy. Residential Microgrid, which is a new concept, can operate as standalone or as a grid-connected system. Renewable energy sources, particularly the photovoltaics, are the major component of such systems. A single-phase inverter is the main control subsystem of a residential Microgrid. Elimination of generated harmonics from the inverter is an issue. In this paper, the development process of an LCL filter for a residential Microgrid inverter system is described that can filter out most of the harmonics generated from the switching circuit of the inverter.

Keywords— *Microgrid, LCL filter, pulse width modulation, rectifier, single-phase grid-tied inverter, smart grid.*

I. INTRODUCTION

The world in the past decade has witnessed growing interest in alternative sources of energy due to increased energy crisis worldwide. Distributed Energy Resources (DER) are attractive solutions for meeting the increased energy demand and reliably. Energy from renewable sources such as the sun, geothermal, biomass and wind can never be exhausted. They cause less emission and therefore stand out as a potentially viable source of clean and limitless energy. Distributed Generation (DG) is the major component of DER. DG consists of the following main components:

- i. Generating unit (e.g. PV, wind turbine);
- ii. Energy Storage System or ESS (e.g. battery, ultra-capacitors, flywheel);
- iii. Loads (e.g. electronic loads, plug-in hybrids, Microgrid, main grid); and
- iv. Power electronic interfaces (e.g. DC-DC converters, inverters, rectifiers).

Generally voltage drop along the feeder of a distribution system increases with the increase of demand. On the other hand the voltage along the feeder crosses the upper limit with

increased generation. When the generation is surplus then the excess energy can be stored in the energy storage devices. When the load demand is higher than the generation, the excess demand can be met by supplying energy from the energy storage devices.

The smart grid can allow a Microgrid system to optimize the use of intermittent renewable energy sources and enhance the reliability of regional power systems. A Microgrid is a locally islanded grid that can operate as a standalone or as a grid-connected system and powered by gas turbines or renewable energy. The feasibility of a Microgrid as a standalone system will be improved if it can generate enough power for the load it is designed for [1]. Solar photovoltaic along with storage devices for Microgrid systems can provide an inexpensive and sustainable means of power supply for Microgrid loads [2]. It is also possible to design a residential Microgrid rated in the kW range [2]. Figure 1 shows the block diagram of a residential Microgrid system. A single-phase inverter is the main control subsystem of a residential Microgrid. An inverter is a power electronic apparatus which converts DC power or direct current to AC power.

The generation of high quality AC energy from PV sources is very crucial, because an inverter can produce non-sinusoidal currents. Non-sinusoidal current can cause additional non-sinusoidal voltage drop across the line impedances and therefore increases the voltage distortions delivered to the load. Another problem of PV system is that the energy supplied by this source is not constant, and fluctuates according to the surrounding conditions (e.g. intensity of sun rays and weather condition). Disturbances in the supplied energy sources may cause a lot of damage and reduce the life span of the equipment. The inverter is controlled by a sophisticated control algorithm such as Sinusoidal Pulse Width Modulation (SPWM) technique which ensures nearly sinusoidal current generation. In order to ensure nearly sinusoidal current a filter is essential at the output of an inverter. A low pass filter is used between the inverter and the grid in order to limit the excessive current harmonics injected in the point of common coupling [3-4]. LCL filter topology is much more attractive

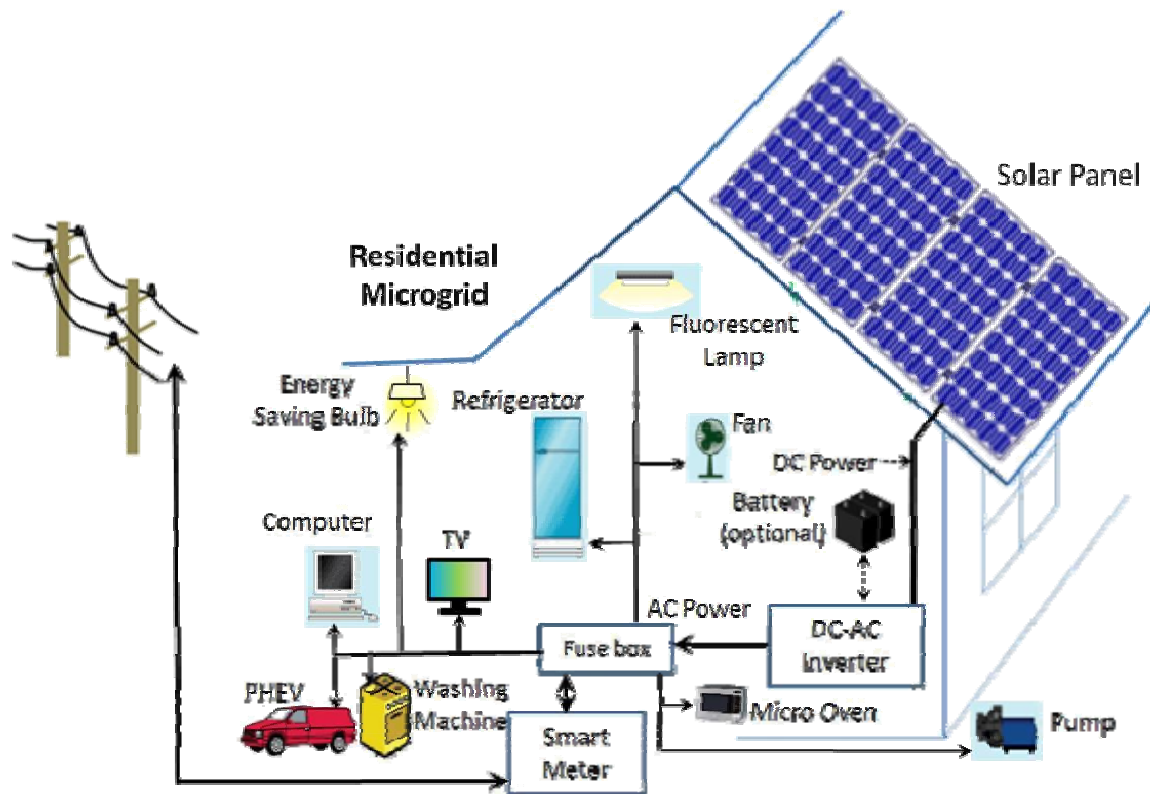


Figure 1. Block diagram of a residential Microgrid system.

than L type or LC type filters. LCL filter has excellent harmonic suppression capacity.

This paper describes the development process of a low-distortion, low-noise switching DC/AC inverter for a single phase residential Microgrid system. An LCL filter is implemented at the output of the inverter. The development process is validated through simulation and experimentation.

II. DEVELOPMENT OF THE INVERTER CIRCUIT

Figure 2 shows the block diagram of the single phase inverter for the residential Microgrid system. The inverter converts the DC from the PV or other sources (e.g. battery) by using a pair of MOSFET switches. By switching the two MOSFETs ON and OFF alternately, the current is made to flow in one direction for half of the cycle and in the opposite direction for another half cycle, producing an alternating magnetic flux in the transformer's core. As a result a corresponding AC voltage is induced in the transformer's secondary winding. MOSFETs are used as electronic switches to convert the DC into AC because they make the most efficient high-current switches. When the MOSFETs are OFF, they are virtually an open circuit. When they are ON, they behave like a short circuit (only a few milliohms). So very little power is wasted as heat within the MOSFET [5-6]. Energy storage device, which is optional in this case, can be also be connected with the Microgrid system. Energy storage devices also help in maintaining a constant frequency of the power system. For example, the battery can be

charged when the frequency crosses the upper limit and the battery can be discharged to the power system when the frequency of the system falls below the lower limit. Figure 3 shows a conceptual block diagram of the frequency control system through energy storage devices. We have followed the development process of the inverter with and without LCL filter. Figure 4 shows the schematic of the inverter without any filter at the output. We have done the simulation of the output voltage of the step-up transformer both in time domain and frequency domain. Figure 5 shows the output voltage in time domain and Figure 6 shows the output voltage in frequency domain. Both Figures 5 and 6 show that the output of the inverted and step-up voltage is not a pure sine wave and contains many harmonic signals. Therefore it is essential to incorporate a filter circuit at the output of the circuit. It will also be shown in section V that the output of the developed hardware without filter also exhibits non-sinusoidal waveform. Schematic shown in Figure 4 was drawn by using the software program PROTEUS.

III. REASON OF INCORPORATING LCL TYPE FILTER

It was mentioned in section I that a non-sinusoidal wave can cause severe disturbance in the power system. An appropriate filtering circuit is essential at the output of the inverter in order to reduce the generated harmonics due to semiconductor switching. There are several types of filters. First-order passive L-type filters can be used on the AC side of PWM inverters to attenuate the current ripples that result from inverter switching. This type of filter has a restriction in low-switching frequency applications of high-power PWM inverters due to the inevitable large filter size.

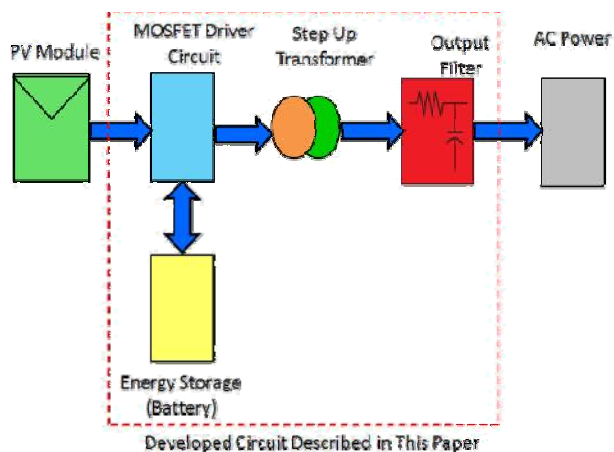


Figure 2. Block diagram of a single-phase inverter with PV and energy storage module.

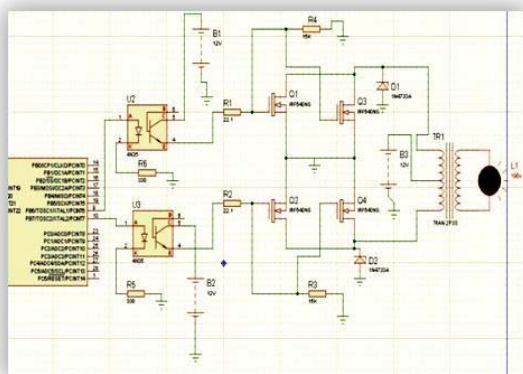


Figure 4. Schematic of the single phase inverter without LCL filter.

On the other side inductance greatly decreases the dynamics of the whole system. Second-order passive LC-type filters have been widely used on the AC terminals to attenuate the voltage ripple that results from inverter switching. The design of the filter is a compromise between the value of the capacitance and inductance. The higher capacitor has positive effects on the voltage quality. On the other hand higher inductance value is required to achieve demanded cut-off frequency of the filter. The resonant frequency of the filter becomes dependent on the grid impedance if the inverter with this kind of filter is connected to the supply grid. Therefore this type of filter is not also suitable at the inverter output.

Another option is the use of third order LCL filter at the inverter output. The attenuation of the LCL filter is 60 dB/decade for frequencies above the resonant frequency. It also provides better decoupling between the inverter and the grid impedance and lower current ripple across the grid inductor. Therefore we have used an LCL filter at the inverter output.

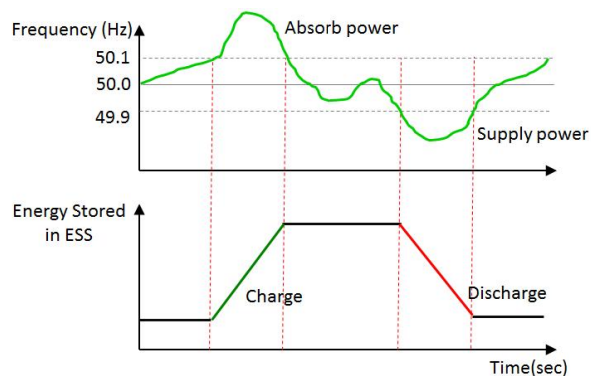


Figure 3. Charging and discharging of an energy storage system and frequency control.

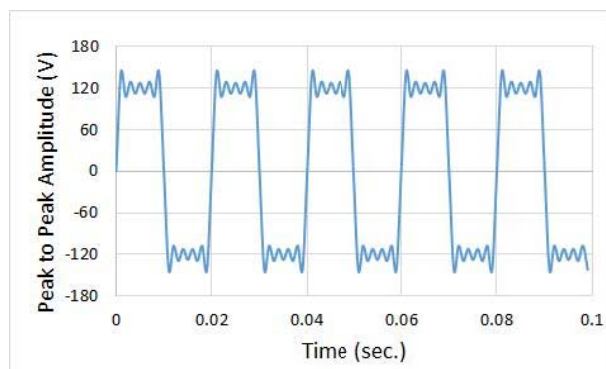


Figure 5. Output voltage of the step-up transformer without any filter at the output.

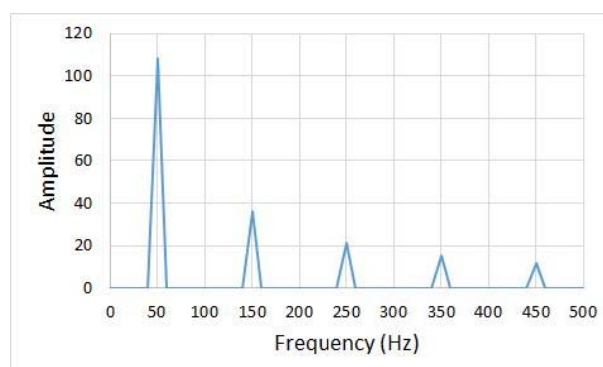


Figure 6. Frequency spectrum of the output voltage shown in Figure 5.

IV. DEVELOPMENT OF LCL TYPE FILTER

The LCL filter has good current ripple attenuation even with small inductance values. However it also can bring resonances and unstable states into the system. Therefore the filter must be designed precisely according to the required parameters of the specific inverter. The most important parameter of the filter is its cut-off frequency or resonant frequency. The cut-off frequency of the filter must be minimally one half of the switching frequency of the inverter in order to have enough attenuation in the range of the inverter's switching frequency. The cut-off frequency must have a sufficient distance from the grid frequency, too. The cut-off frequency of the LCL filter can be calculated as:

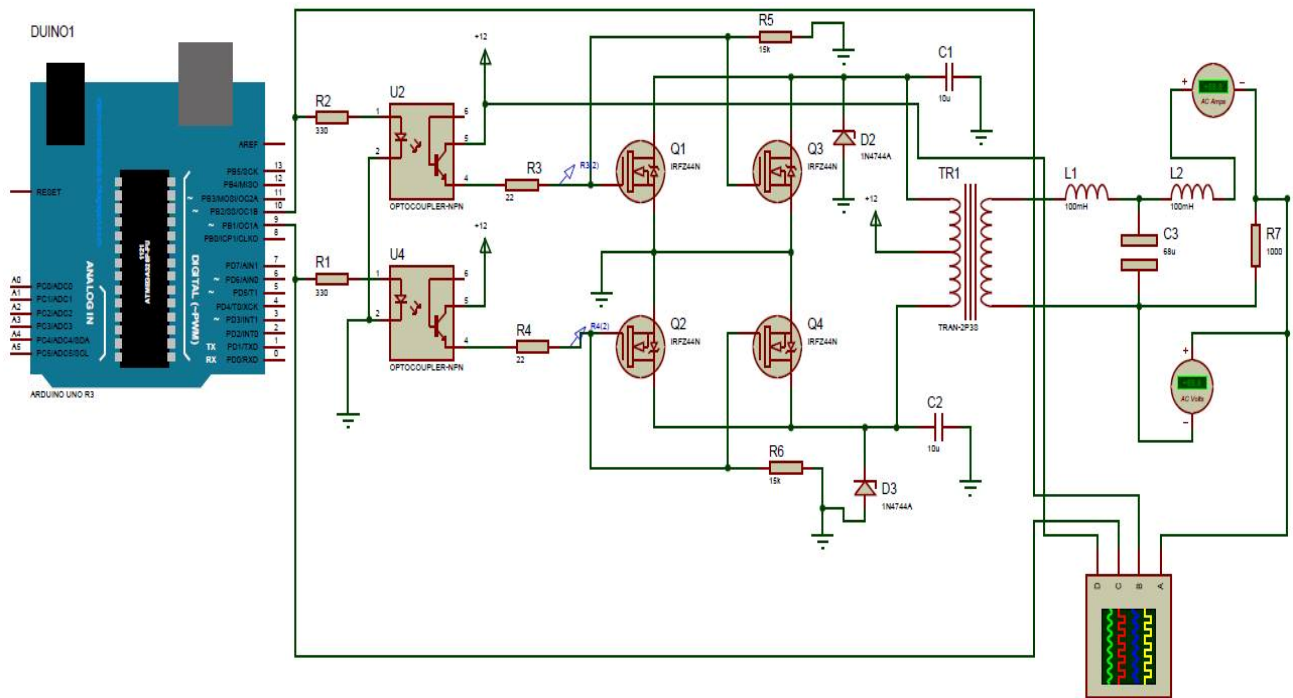


Figure 7. Schematic of single-phase inverter circuit with LCL filter.

$$f_{res} = \frac{1}{2\pi} \sqrt{\frac{L_1 + L_2}{L_1 L_2 C_3}} \dots \dots \dots (1) [7]$$

The parameter L_1 , L_2 , and C_3 are shown in Figure 7 which shows the schematic of the inverter with the added LCL filter. Figure 8 shows the simulated sine wave output of the inverter after the LCL filtering and without connecting any load. It can be seen from Figure 8 that the output is not a pure sinusoidal voltage yet and further improvement of the filter is needed which is a future work. Schematic shown in Figure 7 was drawn by using the software program PROTEUS.

Figure 8: Output voltage of the step-up transformer without LCL filtering.

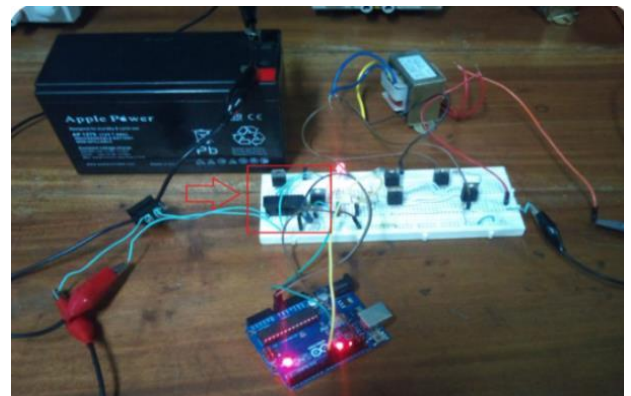


Figure 9 Experimental setup of the single phase inverter

Figure 9 shows the experimental setup of the inverter circuit. Parts and components used for the hardware setup are shown in Table 1 (Appendix). A DC battery (rating: 12 V/7.9 Amp) was used in the experiment as source of DC supply.

Figure 10 shows the observed output of the transformer through an oscilloscope without any added filter. The output voltage shows that it is not a pure sine wave, rather the shape is somewhat square in shape. A filter is needed to make the voltage output sinusoidal. On a second stage we added an LCL filter at the output of the inverter and observed the output wave shape.

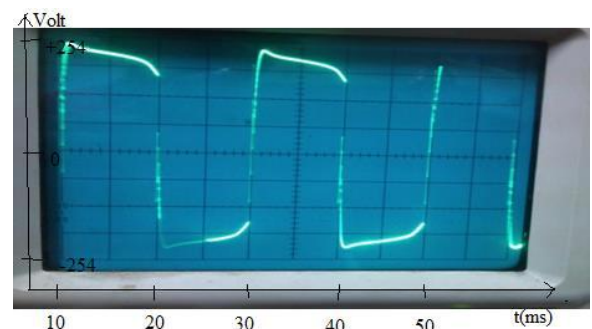
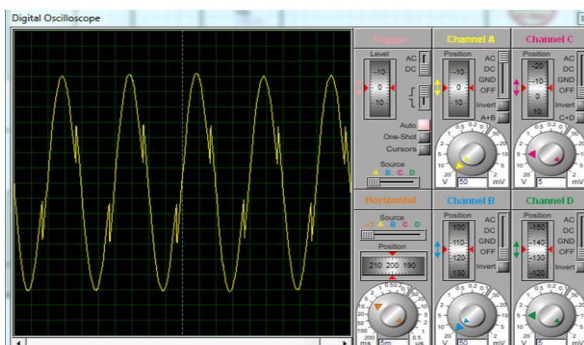


Figure 10. Experimental output voltage of the step-up transformer without any filter at the output of the inverter.

Figure 11 shows that the output shape of the wave has improved but it is not a pure sinusoidal voltage yet. It was mentioned in section IV that further improvement of the inverter is needed. We have used an off the shelf Arduino board for the purpose of controlling the inverter through pulse width modulation (PWM). The duty cycle was 50 % in the experiment.

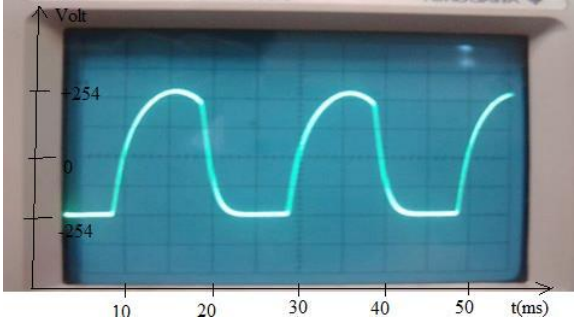


Figure 11: Experimental output voltage of the step-up transformer with an LCL filter at the output of the inverter.

VI. CONCLUSION

The development process of a single-phase inverter for a residential Microgrid system is presented in this paper. We have investigated the development process of the inverter both through simulation and experimentation. An efficient filter is needed at the output stage of the inverter for achieving a pure sinusoidal voltage. LCL filter is the most suitable for this kind of application. LCL filter has good current ripple attenuation characteristics with less inductance values. LCL filter has lower impedance and harmonics than other filter like L, LC etc. Filter must be designed precisely according to the parameters of the specific converter [7]. L type as well as LC type filters can bring resonances and unstable states into the system. LCL filter provides better decoupling between the inverter and the grid impedance and the LCL filter described in this paper better fits to residential Microgrid operation. The experimentation results show that the filter can eliminate the harmonics from the output of the developed inverter. Further improvement of the filter circuit is necessary for getting a pure sinusoidal output voltage which is a future work.

ACKNOWLEDGEMENT

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Appendix

Table 1. List of parts and components used for the development of single phase inverter.

Item	No.
Battery(12V 7.9A)	1
Fuse	1
Switch	1
Optocoupler(4N35)	2
MOSFET(IRF44ZN)	4
Arduino Uno R3	1
5V Regulated IC	1
12V Regulated IC	1
Transformer(12/220V)	1
Resistor(22 Ohms)	2
Resistor (330 Ohms)	2
Resistor (15K Ohms)	2
Capacitor(10u)	2
Diode	1

Grid Connected Hybrid Power System Design Using HOMER

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Abstract—This paper presents an economical expediency of grid connected hybrid (PV/Wind turbine) power system for rural area applications in the southern city of Bangladesh, Lobon Chora, Khulna. The analysis is carried out by investigating the potentials of wind and solar energy and collecting data from different sources. Hybrid Optimization Model for Electric Renewable (HOMER) software is used to analyze the available data and economic feasibility of the proposed hybrid power system. In this paper, two types of models named off-grid and on-grid are designed and optimized for comparison purposes. Also, a sensitivity analysis has been performed for each model in order to determine the effect of variations in grid energy cost on its total cost. The simulation results indicate that the proposed grid connected hybrid (PV/Wind turbine) power system is most suitable and cost competitive for the mentioned region.

Keywords—Grid; HOMER; hybrid power system; renewable energy.

I. INTRODUCTION

Renewable energy sources (RES) offer an efficient solution to the global warming and high fuel cost. Therefore, there is a growing interest in RES, particularly photovoltaic (PV) and wind energy [1], [2]. Recently, hybrid power systems using RE are becoming most popular due to their potential advantages [3].

Although wind and solar has several advantages, a solar or wind generator in a stand-alone system cannot supply the load continuously due to their intermittent nature [4]. As load demands are always changing with time, the changes in solar or wind energy generations do not always match with the time distribution of consumer's demand [5]. Therefore, there is a need of additional battery storage or other components for providing continuous power supply to the load. It has been investigated that a hybrid PV/Wind/battery system is a secure source of electricity [6], [7], [8]. However, due to the high cost of battery energy storage, a stand-alone system is very expensive [9]. Therefore, it is essential to find an effective solution to this problem. On the other hand, a grid connected hybrid power system with proper planning requires storage with less capacity which, in turn, reduces the cost.

Therefore, the main goal of this paper is to analyze the economic feasibility of grid connected hybrid (PV/Wind turbine) power system to meet the load requirement on a specific area. For a comparison purpose, an off-grid hybrid (PV/Wind/battery) power system is also designed. The rest of the paper is organized as follows. Section II describes the system model. Methodology is given in Section III. Simulation models are provided in Section IV and optimization results are evaluated in Section V. Finally, conclusions are drawn in Section VI.

II. SYSTEM DESCRIPTION

In this paper, both on and off-grid hybrid power system models have been designed using HOMER to evaluate and determine the cost of different models. HOMER simulation software needs some input data to evaluate the optimization results for different combinations which are described in the next section [4].

A. Load Profile

The considered area is dominated by irrigation pumps during peak demand. In this paper, the load profile considered average energy consumption by the proposed area and it is 654.73 kWh/d. Figure 1 shows the daily average load profile in which 17:00 to 23:00 h time period has been found as peak demand which determines the size of the system. Here, 101.32 kW is considered as peak load consumption. The scaled annual average (kWh/d) is assumed 654.73. Fig. 2 shows the monthly average load in the proposed region.

B. Wind Speed and Solar Radiation

The data of wind speed and solar radiation is obtained from NASA surface meteorology and solar energy database. Wind speed data obtained at 50 m above the surface of sea level for the location of Khulna, Bangladesh [10], as shown in Fig. 3. The figure shows that the wind speed ranges from 2.64 to 4.33 m/s. The highest wind speed occurs in June. Monthly average solar radiation data is shown in Fig. 4. As used in HOMER software, the latitude and longitude of Khulna district are 22.5° N, 89.5° E, respectively. The annual average solar radiation is estimated to be 4.56 kWh/m²/day [10].

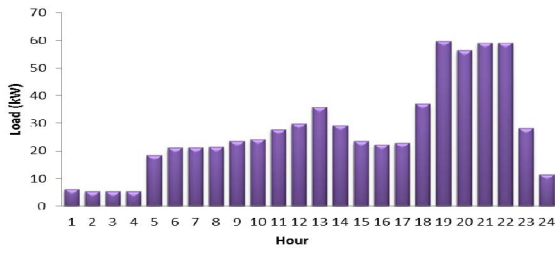


Fig. 1. Load profile (daily).

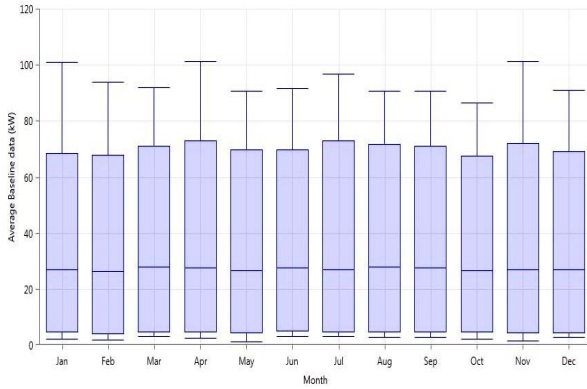


Fig. 2. Load for a complete year (monthly average).

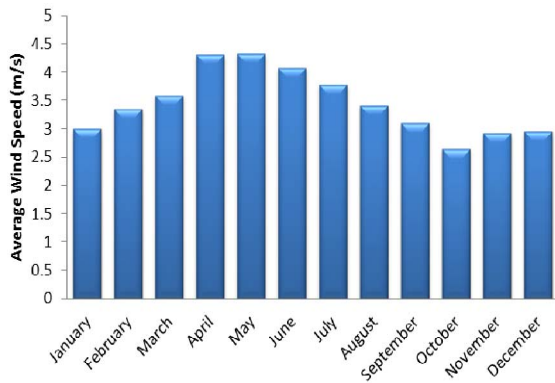


Fig. 3. Wind speed (monthly average).

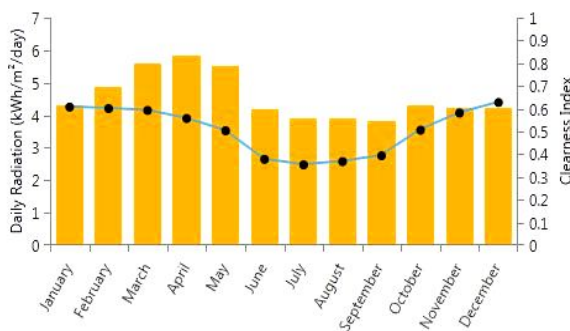


Fig. 4. Solar radiation and clearness index (monthly average).

III. METHODOLOGY

A. HOMER Software

National Renewable Energy Laboratory (NREL) in the United States (USA) developed HOMER software [10]. It is generally used for the design and analysis of hybrid power system. In this paper, electrical load, solar radiation and wind speed data, component details and costs are provided as input information to HOMER.

B. Cost Analysis Procedure by HOMER[10, 11, 12]

1) *Net present cost (NPC)*: NPC indicates the installation cost and the operating cost of the system throughout its lifetime which is calculated as follows [12]:

$$NPC = TAC / CRF(i, Rpr_j)$$

where, TAC, CRF, i , and Rpr_j are the total annualized cost (\$), capital recovery factor, interest rate in percentage, and project life time in year, respectively.

2) *Total annualized cost*: It is the sum of the annualized costs of every equipment of the power system including capital and operation and maintenance cost. It also includes replacement and fuel cost [12].

3) *Capital recovery factor*: It is a ratio which is used to calculate the present value of a series of equal annual cash flows [12].

$$CRF = \frac{i \times (1+i)^n}{(1+i)^{n-1}}$$

where, n and i represents the number of years and the annual real interest rate, respectively.

4) *Annual real interest rate*: It is a function of the nominal interest rate shown as [12]

$$i = \frac{i' - F}{1 + F}$$

where, i , i' and F is the real interest rate, nominal interest rate, and annual inflation rate, respectively.

5) *Cost of energy (COE)*: It is the average cost/kWh of useful electrical energy produced by the system. The COE is calculated as follows [12]:

$$COE = \frac{TAC}{L_{prim,AC} + L_{prim,DC}}$$

where, $L_{prim,AC}$ and $L_{prim,DC}$ are the AC primary load and the DC primary load, respectively.

IV. SIMULATION MODEL

The components are chosen from HOMER to perform simulation. Figure 5 shows the grid connected hybrid power system design using HOMER which consists of PV array, wind generator, converter, load, battery and grid connection. Figure 6 shows the same design but without grid connection.

To estimate the system performance under different situations, HOMER simulates the above two arrangements at the same area and same load based on different costs such as the estimated installation cost, operation and maintenance cost, replacement cost, interest, cost of energy.

The main components of the grid connected hybrid system are wind turbine, PV array, battery bank, and a power converter. For economic analysis, the following values are used.

A. Photovoltaic Array:

Name: Generic Flat Plate PV, manufactured by Generic. Its efficiency is about 13% and lifetime about 25 years. Initial installation cost of PV arrays is assumed as \$3000 per kW and operation and maintenance cost practically zero.

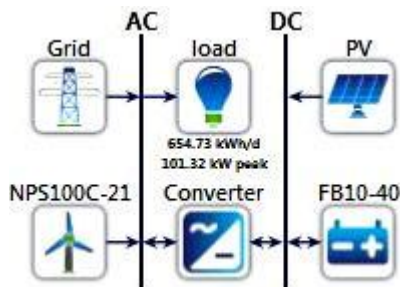


Fig. 5. The arrangement of hybrid power system (grid connected).

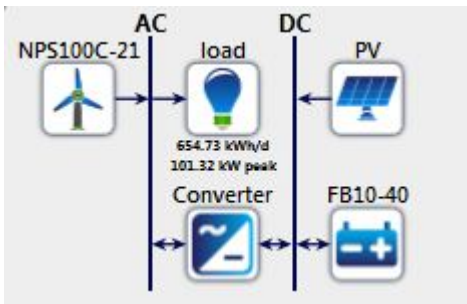


Fig. 6. The arrangement of hybrid power system (off-grid).

B. Wind Turbine:

For this study, three Northern Power NPS100C-21 100 kW wind turbines have been used [10]. Hub height is 23m, 30m, 37m, rotor 30.1m and its lifetime about 20 years. The capital and operation and maintenance costs are \$32,000 and \$300, respectively.

C. Power Converter:

Name: system converter, manufactured by Generic. Its efficiency is about 90% and lifetime about 20 years. In this paper, the capital cost for a 1 kW converter is \$300. Replacement cost is \$300 and operation and maintenance cost practically zero.

D. Battery Storage Bank:

The name of the battery bank is CELLCUBE®FB10-40, manufactured by Gildemeister, with a capacity 40 kWh, lifetime 876000 kWh. Its efficiency is about 64%. In grid-connected designs, a battery is only used during any fault on the grid to maintain the continuity of supply.

E. Grid:

In grid connected design, a grid is used as a backup power component or excess power absorber. The grid supplies power when there is not enough power from RES to meet the load demand and grid consumes power when excessive power is available.

Considering the availability of renewable energy sources in the considered area, wind and solar energy are used to design the hybrid power system. A battery energy storage is used to support the system in it's off grid mode.

V. OPTIMIZATION RESULTS

Figures 7 and 8 show the optimization results for the hybrid power model with and without grid connection, respectively. For grid connected system, as shown in Fig. 7, the minimum COE obtained from the result is \$0.0995. In this scenario, the percentage of renewable energy contribution is 73%. In the proposed grid connected model, an optimum number of renewable energy sources is activated and supplies electricity to the load. From Fig. 8, it can be clearly seen that off grid power system for the same load is more expensive (\$0.391) than grid connected system (\$0.0995). The NPC for on-grid and off-grid design is \$535,661 and \$1.21M, respectively. From Fig. 9 it can be seen that the total yearly production of the proposed hybrid model is 436,186 kWh/yr and consumption by the load is 238,967 kWh/yr. The rest of the production is sold to the grid. The monthly average electricity production from different units using the proposed model is shown in Fig. 10.

Architecture							Cost				
PV (kW)	NPS100C-21	FB10-40	Grid (kW)	Converter (kW)	COE (\$)	NPC (\$)	Operating cost (\$)	Initial capital (\$)	Ren Frac (%)		
150	3		999,999	180	\$0.0995	\$535,661	-\$4,977	\$600,000	73		
150	3	2	999,999	180	\$0.101	\$542,572	-\$4,906	\$606,000	73		
200			999,999	180	\$0.134	\$662,548	\$661.25	\$654,000	62		
200		2	999,999	180	\$0.135	\$669,536	\$737.62	\$660,000	62		

Fig. 7. Screenshot of simulation for finding optimal design (grid connected).

Architecture								Cost				
				PV (kW)	NPS100C-21	FB10-40	Converter (kW)	COE (\$)	NPC (\$)	Operating cost (\$)	Initial capital (\$)	Ren Frac (%)
				300	3	40	180	\$0.391	\$1.21M	\$2,773	\$1.17M	100

Fig. 8. Screenshot of simulation result of off grid power system (PV/Wind/Battery).

System Architecture:	Generic flat plate PV (150 kW)	Grid (999999 kW)	Total NPC:	\$535,661.00
	Northern Power NPS100C-21 (3) Cycle Charging		Levelized COE:	\$0.09953
	System Converter (180 kW)		Operating Cost:	(\$4,977.00)

Cost Summary	Cash Flow	Electrical	Renewable Penetration	Generic flat plate PV	Northern Power NPS100C-21	Grid	System Converter	Emissions			
Production		kWh/yr	%	Consumption		kWh/yr	%	Quantity	kWh/yr	%	
Generic flat plate PV		198,814	45.58	AC Primary Load		238,976	57.40	Excess Electricity		0.0	0.0
Northern Power NPS100C-21		124,625	28.57	DC Primary Load		0	0.00	Unmet Electric Load		0.0	0.0
Grid Purchases		112,747	25.85	Grid Sales		177,328	42.60	Capacity Shortage		0.0	0.0
Total		436,186	100.00	Total		416,305	100.00				

Quantity		Value
Renewable Fraction		72.9
Max. Renew. Penetration		111.1

Fig. 9. Snapshot of the production and consumption scenario.

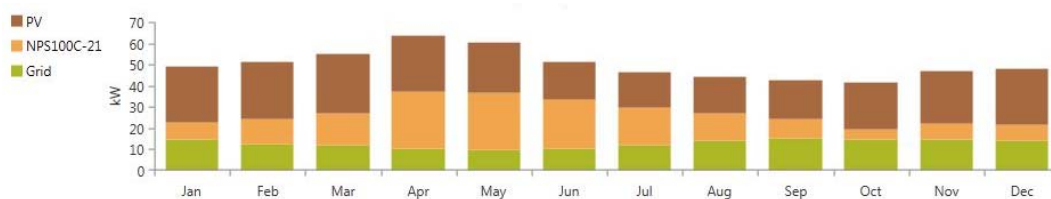


Fig. 10. Monthly average electricity production using the proposed model.

VI. CONCLUSION

This paper presents a comparative analysis between an off-grid and a grid connected hybrid power system in a specific area in Bangladesh. The optimization result shows that the grid connected hybrid (PV/wind) system is more efficient and economic compared to the traditional hybrid system (PV/Wind/Battery) for the same load. From the simulation result, it is also investigated that Net Present Cost of the proposed model is less than off-grid model. Although off-grid hybrid system uses 100% renewable energy but it needs an extra-large battery bank for storage of electricity. On the other hand, excess output power of off-grid system goes unused. A grid connected hybrid system does not require an extra battery bank in normal operating condition and excess power that is produced by renewable energy are added to the grid. Therefore, it can be said that the proposed grid connected hybrid power system is most suitable and cost effective as it offers several benefits.

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Solar PV- Hydrogen Fuel Cell System for Electrification of a Remote Village in Bangladesh

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Abstract—Main objective of this paper is to suggest an electricity generation system based on renewable energy sources to fulfill the electrical demand of remote villages of a developing country like Bangladesh. The suggested system is composed of photovoltaic array, fuel cell stack, electrolyzer and hydrogen storage tank. This system will replace the conventional method of storing energy by introducing hydrogen storage system. The system is known as Solar PV- Hydrogen Fuel cell system. The optimization and the simulation of the proposed system were performed with Hybrid Optimization of Multiple Energy Resources (HOMER) software. In order to determine the optimal combination of the system, different combinations of PV arrays, Fuel cell, Electrolyzer and Hydrogen storage tank were selected on the basis of the Net Present Cost (NPC).

Keywords— PV array, Fuel cell, Electrolyzer, Hydrogen, optimization.

I. INTRODUCTION

Electricity is the most essential ingredient to upgrade the socio-economic condition and to alleviate poverty of a country. The supply of electricity has a great impact on the national economy. Reliable and sufficient electricity supply has a great positive impact on our GDP and GDP is one of the key measures to understand the economy of a country. Thermal plants run by coal, oil & gas are the major conventional ways to generate electricity. In today's world, maximum amount of the global electrical demand is met by using the conventional fossil fuel such as petroleum, coal, and natural gas.

The electricity demand is increasing very rapidly, as a result the reserve of the fossil fuel under the Earth is also decreasing at the same rate and it is also estimated that within upcoming few decades the existing reserve of the fossil fuel will be used up if we continue to use the fossil fuel at this rate.

Bangladesh has achieved significant development in electrical energy sector in last few years. Bangladesh had achieved the capability of generating 10,000MW of electricity in 2013. Despite of all this developments in the energy sector, still a significant portion of the total population is still not having any electricity in their houses. The scenario is worst in remote villages (Chars) as the grid extension to those areas is very costly.

It's tough to think that in this modern age of technology many people of our country are not having the access to electricity. In this paper main focus was given on electrification of the Char [island] areas. Life of the villagers in remote villages is so miserable. The areas are completely inaccessible and can only be reached by boat and foot. The people living there are exposed to nature and due to the daily tides, the land gets flooded on a regular basis.

In this paper a comparatively new energy solution is suggested for those isolated villages and the proposed model will be the alternative to diesel generators and batteries. This paper mainly includes something new for our country [Bangladesh] i.e. fuel cell included in the stand alone power system. Stand-alone power system is an off grid power system that usually operates with renewable energy sources. Stand-alone power system is mostly used for remote areas.

II. COMPONENT DESCRIPTION

The suggested system is composed of a Fuel cell, Electrolyzer, Solar Photovoltaic arrays and Hydrogen storage tank. The main components of the system are introduced in brief in the following sections.

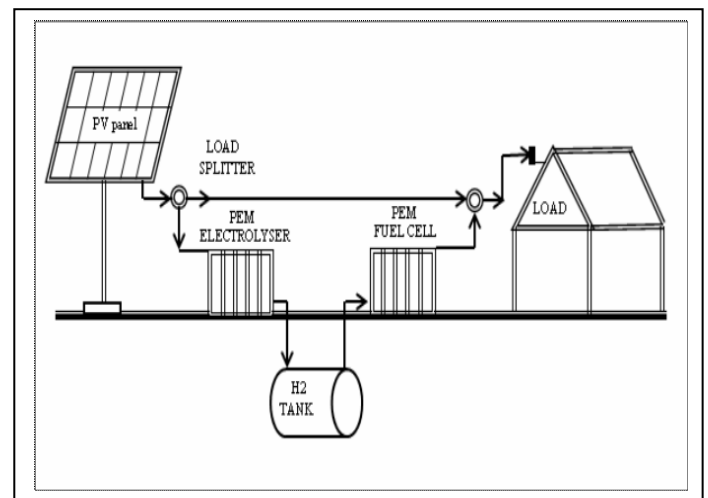


Fig. 1. Schematic diagram of the Solar PV-Fuel cell system showing its' main components. [1]

1. Fuel Cell

Fuel cell is defined as an electrochemical device that continuously converts the chemical energy of a fuel and oxidant into electrical energy and heat as long as the fuel and oxidant are supplied to the electrodes.

In a typical fuel cell, fuel is fed continuously to the anode (negative electrode) and an oxidant (often oxygen from air) is fed continuously to the cathode (positive electrode). The electrochemical reactions take place at the electrodes as fuel and oxidant are supplied, producing electrons. Electrons are produced at the anode and are consumed at the cathode. Electrons move from anode to cathode via an external circuit. As a result this movement of electrons produces electricity. Within the fuel cell, ions flow between anode and cathode. [2]

2. Electrolyzer

Electrolyzers are used widely in industry to synthesize various types of products, in a sense electrolyzers are reverse of fuel cells. When electric current is forced through an electrolyzer from an outside source, chemical reactions generate new products occurring at the electrodes[3]. An electrolyzer converts electrical energy into chemical energy which produces hydrogen through the process of electrolysis which is very well known electrochemical process for hydrogen production.

3. Solar Photovoltaic

Solar Photovoltaic (PV) modules are solid-state devices that convert sunlight directly into electricity without an intervening heat engine or rotating equipment. PV equipment has no moving parts, so as a result it requires minimal maintenance and has a long life. Photovoltaic systems can be built in virtually any size, ranging from milliwatt to megawatt, and the systems are modular, i.e., more panels can be easily added to increase the output.

III. WORKING PRINCIPLE OF THE SYSTEM

In the suggested system, Solar PV array is the main generator of electricity. When the sunlight will be available, Solar PV arrays will function and convert the sunlight falling on it into electricity. But in village areas, the demand is much less in daytime so there will be a mismatch in between demand and supply. Therefore, a reliable energy storage system will be required, and in this case the energy will be stored as hydrogen. This system will operate in three modes, which are explained below:

MODE 1:

The system will operate in this mode when the electricity generated by the Solar PV arrays will be exactly equal to the demand; as a result there will be no excess electricity. In this mode, the fuel cell and electrolyzer will not function.

MODE 2:

The system will operate in this mode when the electricity generated by the Solar PV arrays will be much more than the demand, so there will be significant amount of excess electricity. This excess electricity will be passed to the

electrolyzer to decompose water into hydrogen and oxygen. Then the produced hydrogen will be passed to the hydrogen storage tank to store for future use. In this mode, Solar PV arrays, electrolyzer and hydrogen will function. The system will operate in this mode in daytime when the sunlight is available in sufficient amount.

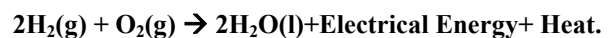
REACTION OCCURRING IN MODE 2:



MODE 3:

The system will operate in this mode mainly when the sky will be cloudy or in night time, basically when sufficient sunlight will not be available. In this mode the fuel cell will play its role. The hydrogen stored in the tank will be passed to the fuel cell and the fuel cell will use this hydrogen to generate electricity by recombining hydrogen and oxygen. In this mode, only fuel cell will function.

REACTION OCCURRING IN MODE 3:



IV. METHODOLOGY

In this section, the method is discussed that was followed for this paper. Optimization of the size and number of components in this system were required to study the feasibility of using this Solar PV-Fuel cell system for a remote island area in Bangladesh. There are many different possible sizes and hardware configurations to be determined. The size of the PV system, electrolyzer, hydrogen storage tank, fuel cell was determined to ensure the system would meet the required load at least possible cost.

The techno economic optimization of the proposed stand-alone system was performed with a software named Hybrid Optimization of Multiple Energy Resources, mostly known as HOMER. For our work, HOMER Legacy version 2.68 Beta was used.

A. Electrical Load of the system

The load demand data was estimated according to the probable daily use of a villager living in an isolated village. For a particular location, a year was divided in two parts on the basis of probable consumption, period of low consumption [November - February] and period of high consumption [March - October].

This division was done on the basis of the season, as in many cases electrical demand of a house depends on particular season. In this paper, 100 houses each with 5 CFL bulbs and 2 fans, school consisting of 3 rooms with 6 fans, bazar/shops with 5 CFL bulbs which will be in use from 6-9 pm and doctor's chamber/small clinic with 5 CFL bulbs and 2 fans are considered for electrical load calculation. The power rating of a electric fan is considered to be 60W and CFL bulbs - 23W and 11W. The annual average electrical load for the considered area is estimated to be 90.1 kWh/ day and the annual daily peak is 16.3 kW.

Solar radiation and clearness data for the site (Nijhum Dip) was obtained from the NASA Surface Meteorology and Solar Energy website [https://eosweb.larc.nasa.gov/sse/]. The annual average global radiation is 4.762 kWh/ m² / day and with an annual average clearness index of 0.510.

B. Component Costing

The capital cost of the PV panel is considered BDT 70/watt. This is the cost of the PV panels which was used in a project named “Solartech 5.1kW AC Solar Pump” at BADC, Tangail conducted by Electro Group. Life time of the PV panels is assumed to be 20 years.

The capital cost of the electrolyzer in the system is considered BDT 75,000/kW (\$1000/kW) and lifetime is assumed to be 15 years [4]. The capital cost of the fuel cell used for this paper is BDT 2,25,000/kW [5]. The cost of the hydrogen storage tank is taken BDT 106/kg and the lifetime is predicted to be 25 years[6].

V. SIMULATIONS & RESULTS

For simulation purpose, per unit capital cost of each component of the system, operation and maintenance cost, component sizes for consideration, lifetime, efficiency, slope for the PV panels, primary hourly load data, interest rate, project lifetime, dispatch strategy, generator control, different constraint, coordinates of the particular location [Nijhum Dip] was given as input in HOMER in order to perform the simulation.

To perform sensitivity analysis, for some particular case multiple numbers of input was given. In HOMER, the solution is a particular hybrid system configuration. As HOMER searches for the optimal hybrid system design, it typically evaluates hundreds or thousands of solutions. HOMER performs an hourly time series simulation of every solution and eliminates solutions that cannot meet the required load. In this paper, HOMER examined all of the 5,038,848 possible solutions by combining the components.

According to the results of the optimization process for the stand-alone PV-Fuel cell system, the optimal system compromises 70 kW PV panels, 15 kW PEM-type fuel cells, 40 kW electrolysis units and hydrogen storage tank with a capacity of storing 400 kg of compressed hydrogen. Load following control strategy is optimal in this case.

According to the optimization results, a DC configuration is economically more favorable, and the PV-Fuel cell system had a total annual electrical energy production of 131,083 kWh; 80% (105,059 kWh) of which was produced by the PV array and 20% (26,024 kWh) was produced by the fuel cell. The system had a total annual electric energy consumption of 122,064 kWh, 74% (90,628kWh) of which is consumed by the electrolysis units and 26% (31,436kWh) was consumed by the DC primary load.

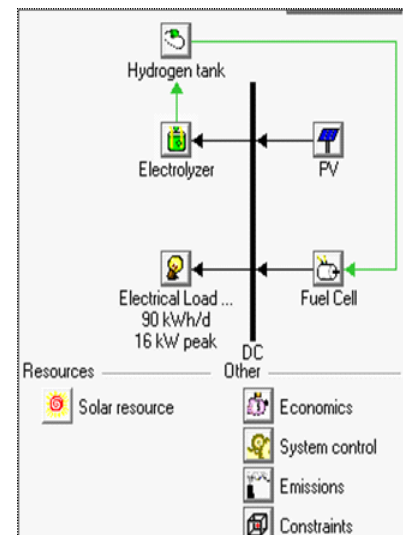


Fig. 2: Schematic of the PV-Fuel cell system.

The optimized cost of electricity [COE] of the system is 41.316 BDT/kWh. The system with battery bank only is much more expensive to set up and the COE is also more than the proposed system and on the other hand a system with the generator can provide electricity with less COE but the cost for operation and maintenance will be much higher and is highly dependent on the international market.

Sensitivity analysis results demonstrate that the capital cost of the components especially PV and Fuel cells significantly affect the final cost of energy produced by a stand-alone power system. According to the sensitivity analysis, if the capital cost of PV panels, Fuel cells, Electrolyzer and Hydrogen storage tank can be reduced to 50% of the present cost, then the COE of the system will be 25.011BDT/kWh (reduction of COE by 39.5%).

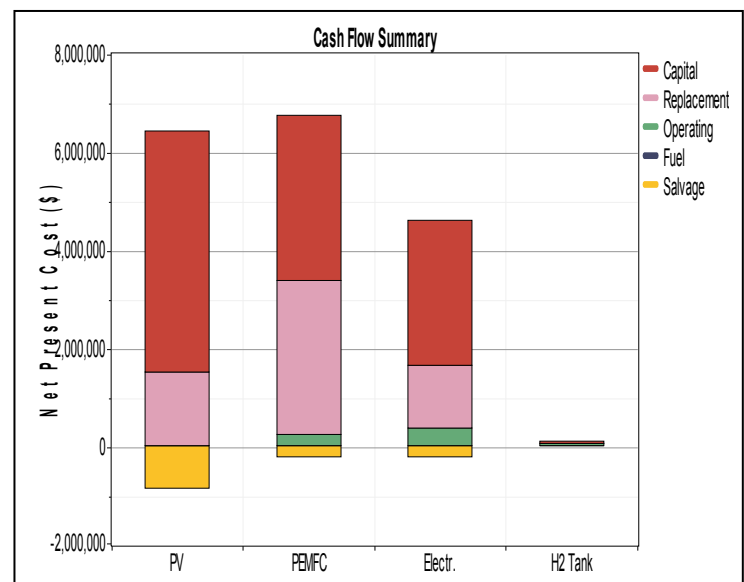


Fig. 3: Net Present Cost [NPC] by component type.

TABLE I. NET PRESENT COST OF THE COMPONENTS.

SOLAR PV:	BDT 571,574
FUEL CELL:	BDT 6,535,463
ELECTROLYZER:	BDT 4,402,299
HYDROGEN TANK:	BDT 93,533

TABLE II. COMPARISON BETWEEN SOLAR PV-FC & PV-BATTERY SYSTEM.

	SOLAR PV- FC	PV - BATTERY
COST OF ELECTRICITY	BDT 41.316/ KWh	BDT 44.858 /KWh
NET PRESENT COST	BDT 16,602,864 / YEAR	BDT 18,846,994 /YEAR

VI. CONCLUSION

Availability and reliability of the power source are the core aspects of energy security. The market for backup power system is predicted to increase in line with the increase in electricity grid instability in isolated areas. Fuel cell technology has entered this application sphere because it offers a number of advantages over incumbent systems. At present, this technology is used where its benefits, such as reliable and extended operation in remote areas, overcome the main disadvantage of high capital costs.

The main aim of this work was to propose a reliable electrical energy supply solution for the underprivileged isolated regions of Bangladesh. The installation of electrical system to supply electricity with an energy storage system strongly depends on the economic viability of the system. Therefore, the proposed system was analyzed both technically and economically. In this work, the simulation was carried out

in order to find out an optimized system with different optimized parameters. The simulation results indicate that the replacement of the conventional technologies, namely diesel generators and batteries by Hydrogen Fuel cell technology is technologically feasible, reduces emissions, noise and fossil fuel dependence and increases renewable energy penetration. The cost of energy produced by the proposed system is 41.316 BDT/kWh which may not be affordable for the poor people living in remote areas. Favorable government policies with subsidy and low interest rates may help the widespread deployment of PV-Fuel cell in remote areas.

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Organic Solar Cells: Historical Developments and Challenges

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Abstract— Organic Solar Cells (OSCs) represent a recent photovoltaic (PV) technology that uses organic semiconductor materials to convert sunlight into electric energy. OSCs have recently shown significant attention due to their technological advantages of flexibility, light weight, cost effective fabrication and significant improvement in the power conversion efficiency. But the efficiencies of OSCs are still lower than crystal or inorganic silicon solar cells. By studying the photovoltaic properties of organic solar cells it is possible to increase the cells efficiency. If the efficiency can be raised to half of the inorganic solar cell the production cost will be reduced to one-tenth. This paper describes the historical background, present situation, future prospectus and technological breakthrough of OSCs. Also explain the long-term sustainability advantages of OSCs with compared to inorganic photovoltaic technologies in terms of energy payback time and CO₂ emission.

Keywords—Photovoltaic; Organic; Inorganic; Silicon; Solar Cell;

I. INTRODUCTION

Organic solar cells are assured for low cost energy production, but if only further developments in their design and manufacture make them useful for real-world applications. A photovoltaic (PV) device converts light directly into electric current and voltage. Traditionally this is done by forming a diode from the junction of two regions in an inorganic semiconductor, one doped to be n-type and the other p-type. An n-type semiconductor has a small fraction of its atoms substituted with atoms one space over to the right on the periodic table like phosphorus which have five valence electrons. Similarly p-type has some atoms replaced by that one space to the left like boron which have three valence electrons. The dopants in the n-type have one more electron than all their neighbours and readily give it up, while the p-type dopants have one fewer electron and will accept electrons. When the electrons donated from the n-type region are accepted in the p-type region they form what's called the depletion region. Due to the net build up of charge there is a net electric field in this region. Semiconductors have "band gaps" of allowed energies. The highest band that would be filled at 0 K is called the valence band, while the next highest, separated by some band gap of forbidden energies, is called the conduction band. Because the conduction band is mostly empty, the electrons that are in it can move relatively easily.

When a photon with enough energy strikes an electron in a semiconducting material it is excited into the conduction band. At the junction of p- and n-type semiconductors the electron and resulting positive hole would then be accelerated apart due to the electric field. The result is an electric potential and current that is the output power.

Solar cells can now be made from single crystal, multi crystal, amorphous, and ribbon silicon. Other materials that have been widely studied but do not have large market shares include CdTe, GaAs, Ge, and copper indium gallium (di) selenide (CIGS). These other materials have mostly failed to be widely commercialized (outside of niche markets like space) because of their high cost. CdTe is an exception that has recently emerged as a competitive technology [1]. But even silicon based solar cells, currently the most cost efficient that can be produced on a commercial scale, still have a rather poor power to price ratio [2]. Solar will never be competitive for large scale energy production until the price per kWh is brought in line with those other sources. The cost of solar energy is instead driven by the cost of the initial capital investment. The "fuel" in this case is free. The cost then is determined by the manufacturing cost of making the panels, the installation cost, maintenance costs, and the lifetime over which those costs can be amortized. However, even if the price of these cells has been radically reduced, it is still too much expensive and photovoltaic energy account for less than 0.1% of the total world energy production [3]. Previous research estimates that only 0.1% of incident sun power can meet the world's current energy needs using solar cells with around 10% conversion efficiency.

Organic Photovoltaic's (OPV) are one strategy for bringing down the cost of solar energy. By replacing the crystalline semiconductor with some sort of organic molecule the manufacturing cost of solar cells might be brought down to the same level as for making sheets of plastic, because that is what they would be. Currently organic solar cells are significantly less efficient than most inorganic solar cells (1-11% vs. 15-40%), but it might still make sense to use them (organic solar cell) even the efficiency is half of the inorganic solar cell they could be produced one-tenth of production cost of electricity. Fig. 1 shows the different types of solar cells efficiency record from National Renewable Energy Laboratory (NREL) chart from the year of 1975 to 2015.

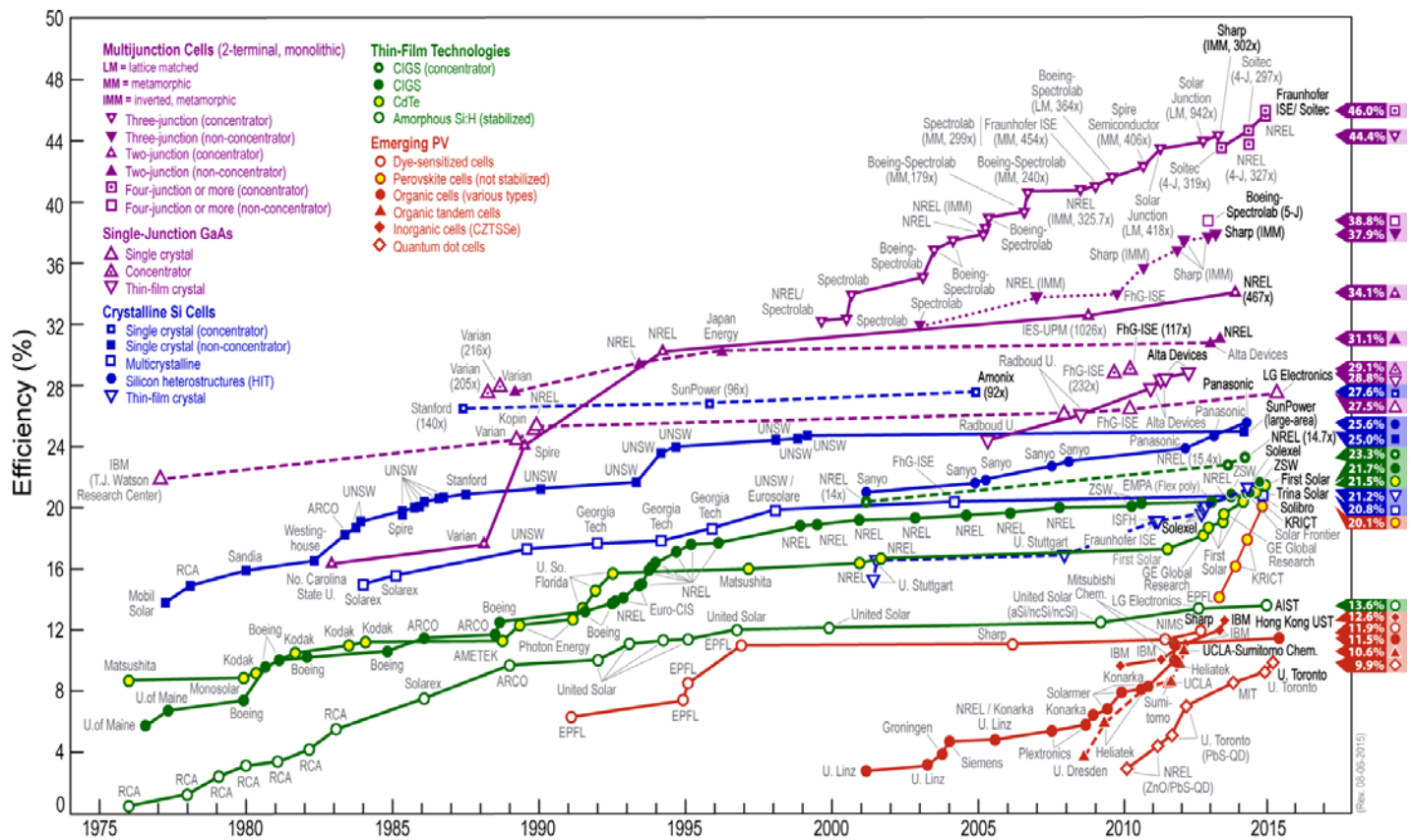


Fig. 1 NREL-Chart of power conversion efficiencies of best research solar cells from 1975 to 2015 for various photovoltaic technologies

II. ORGANIC SOLAR CELL

The external behaviour of organic photovoltaic is same as inorganic PVs, But the voltage and current are generated is quite different. The organic PV material is not crystalline, so there are not nice bands for the electrons and there isn't electric field to drive them. Because of that, when an excited electron is created by incident light it will quickly recombine with its hole unless something causes them to separate before recombination can happen. For that reason OPVs consist of two materials, one that electrons prefer and another that holes prefer. Once the electron and hole are separated into distinct materials, they can just diffuse apart due to their respective concentration gradients. Thus there are three roles that will need to be filled by the materials: absorption of light in the visible spectrum, a semiconducting material that will take the electrons (the acceptor), and a semiconducting material that will take the holes (the donor). The conducting materials must be semiconductors so that the cell can maintain an output voltage and not just produce photoconductivity. If the charge carriers are generated far from the electrodes, the materials will also need to be efficient charge conductors (long carrier lifetimes) so that the charges are not lost before collection. Many material combinations have been tried that fit these requirements. They can be classified into three primary categories: molecular, polymer and hybrid.

Now that the basic electronic role of the donor and acceptor materials has been established, consider their physical interaction. Because excitations will only separate into charge carriers at the interface of the two materials, it is desirable to maximize the interface surface area to volume ratio. This can be done by making very thin films of one material on top of the other, by mixing the two materials and forming what is known as a bulk heterojunction.

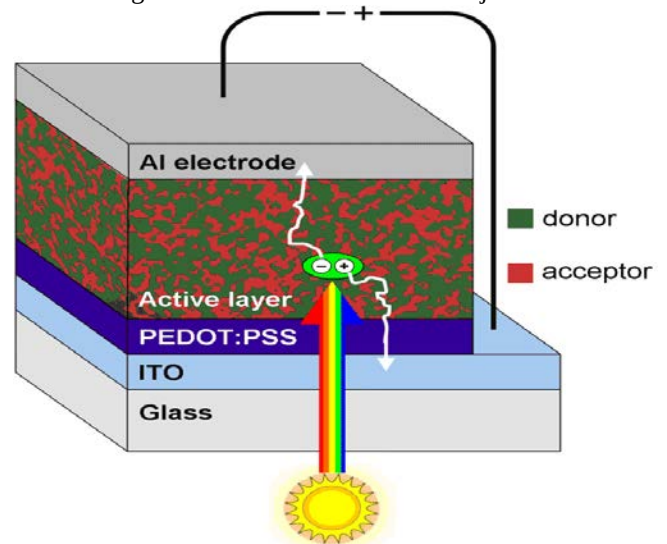


Fig. 2 Illustration of the basic structure of an organic solar cell

In the construction of a working solar cell the organic layer is just one of the necessary components. The device must also be designed to efficiently get light in and charge out. Fig. 2 show a basic schematic of OSCs [4]. On the side where light will come from there is a transparent conducting electrode. Indium tin oxide (ITO) is very commonly used for this purpose. There is also often a glass substrate that provides mechanical support. Anti-reflection coatings can help minimize losses due to non-absorption. On the other, dark side the electrode usually consists of a thin film of metal (such as Au or Al) that has been evaporated in to the device. Strategies have been developed for certain materials, such as adding buffer layers, to solve this problem.

III. TECHNICAL DEVELOPMENT OF OSCs

It is highlighted that organic chemistry has a strong development since 1965. The most attractive features of chemistry are that they discovered the single molecule or new group of molecules might hold the possibility of changing the world. That is the mission of the Holy Grail of all searchers involved in the field of OSCs. The first observation of photoconductivity in organic compound was reported in 1906 by Pochettino [5]. Organic semi-conductors are less expensive than inorganic semi-conductors. During 1970 to 1980 decades many works have been done in OSCs field, however very low efficiencies were achieved because of low concentration and mobility of free charge carriers.

Last ten years a new and strong interest has been devoted to OSCs. A new interest is motivated by two recent developments in the organic semi-conductor field. First the quantum efficiency of the electron transfer from an excited polymer to C_{60} is very high and fast [6], it is promising for charge carrier separation in PV cells and secondly the development of efficient organic displays based on organic light emitting devices (LED). Those displays are now developed using low cost technology like OLED can be tested for PV solar cells realization [7]. Plastic OSCs are solid and consist either of two organic layers or a homogeneous mixture of two organic materials. One of them is organic dye or a semiconducting polymer - donates the electrons and other component serves as the electron acceptor. Generally, a fullerene is used for this. After the charges have been separated, the holes are transported to the anode and the electrons are transported to the cathode, to supply a direct current for the load.

Few years ago, Shaheen and co-workers have reported solar cells 2.5% efficiency [8]. Also, $CuPc/C_{60}$ bi-layer cell, with an excitation-blocking layer, yields 3.6% efficiency. Recently more than 5% efficiency has been announced for optimised structures in both devices. Single junction solar cells made-up with materials such as $PTB7/PC_{71}BM$ have been reported with efficiencies as high as 9.2% [9]. Higher OPV efficiencies can be achieved by fabricating Tandem solar cells consist of multiple active layers with different band gaps. This device, high energy photons are absorbed by the first layer, which has the largest band gap, and lower energy photons pass through to the second absorbing layer for capture by a material for a lower band gap. The current published and certified record for an OPV cell is held by Mitsubishi Chemical at 10.7% for a tandem device [10], even higher efficiencies 12% have been reported [11].

Recent notable development in OSCs is the beginning of high-performance small molecule absorbers/ donors. Conjugated polymers have controlled the field for the last decade, but challenges associated with polymer synthesis such as batch-to-batch variation in performance, have led to a search for small molecules that can competitor polymers in devices [12]. So many cases, small molecules are not suitably soluble to be processed in solution, and vapour deposition processes are naturally more expensive and therefore less attractive. While vapour-deposited small molecule OSCs have made substantial progress recently [13], particular excitement has been generated by reports of solution-possible small molecule OSCs [14].

A. Energy payback time

OSCs have an exceptional sustainability performance compared to other PV technologies. An important life cycle metric for evaluating the energy production technology is the time required to produce the energy invested throughout its life cycle. It is called the energy payback time (EPBT). For a PV module, the EPBT is the time required for the PV module to generate the equivalent amount of energy that originally was used to manufacture the PV module [15]. In the current scenario, OPV modules have an efficiency of 1% and a lifetime of two years, for mid-term efficiency 10% and life time 10 years while in the long term scenario, the efficiency increases to 15% and the lifetime is as long as 20 years [16]. EPBT for silicon, CdTe, and OPV modules are shown in Fig. 3.

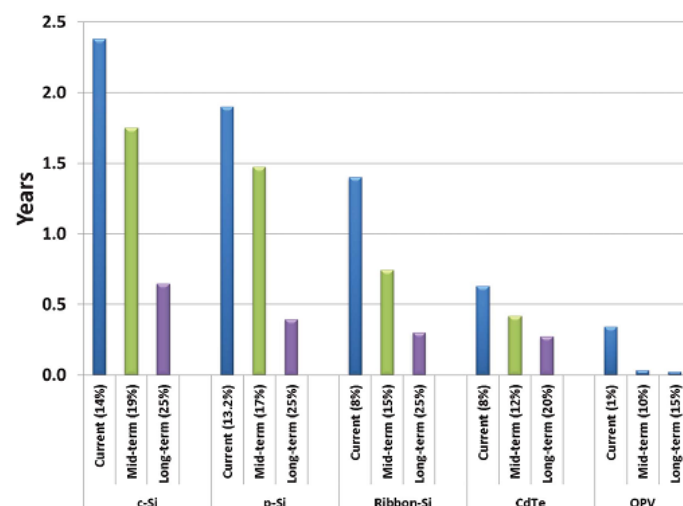


Fig. 3 EPBT for silicon, CdTe, and OPV modules for current, mid-term and long-term are shown [17]

B. Green house gas emissions

Greenhouse gas (GHG) emissions of a device life cycle indicate its environmental impact on the global climate. In the current scenario, the GHG emission factor of OPVs is higher than those of silicon and CdTe PVs. The CO_2 emission factors of all PV modules will decrease drastically from the current scenario to the long-term future. In particular, OPV emissions become the smallest in the long-term future scenario Fig. 4. The significant decrease is mainly due to the increase of conversion efficiency and operating lifetime. The study does expose the significant environmental benefits of OPV technologies in the long-term future.

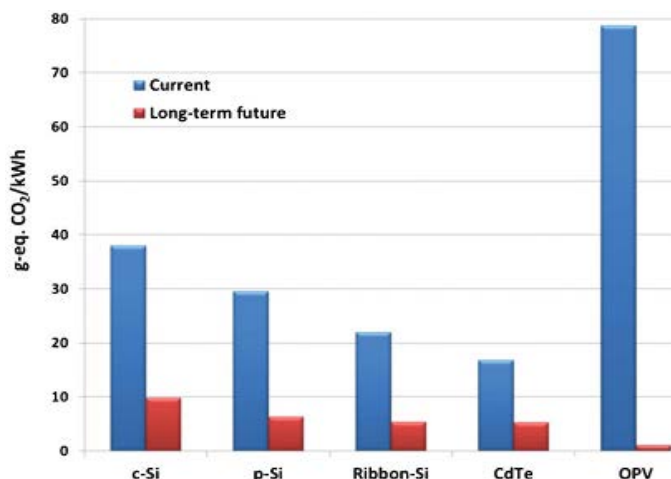


Fig.4. Carbon emissions for silicon, CdTe, and OPV modules for current and long-term are shown [17].

C. Power conversion Efficiency

The crucial obstacle to increase the efficiency of OSCs is the range of spectral absorption of light. The power conversion efficiency (PCE) has exceeds 9% for single-junction PSCs [18-21] and 11% for tandem-junction cells [22]. However, to promote the OSCs in the practical applications, to enhancements in the efficiency are urgently required for single-junction OSCs. For high-performance single-junction OSCs, the film thickness of the photoactive layers is around 100–200 nm to minimize the charge recombination loss for charge extraction due to the low carrier mobility. The overall light absorption of PSCs in such a thickness scale is inevitably limited between 60% and 80%. To avoid this problem, the optical light has capture solar radiation more effectively into an ultrathin photoactive layer with efficient charge extraction and thus to improve the light-harvesting efficiency. The deterministic aperiodic nanostructures (DANs) have suggested a route to controlling the light harvesting in organic optoelectronics with the uncomplicated fabrication procedures. The single-junction PSCs were constructed with a configuration of ITO/ZnO/PTB7-Th: PC₇₁ BM/MoO_x/Al with and without the incorporation of DANs.

$$J_{ph} = J_L - J_D,$$

Where J_L and J_D are the current densities under illumination and in dark, and

$$V_{eff} = V_0 - V$$

Where V is the applied voltage and V_0 is the compensation voltage at which $J_{ph} = 0$.

The J_{ph} increases linearly with the voltage at a low value of V_{eff} (<0.1 V).

And tends to saturate at a sufficiently high value of V_{eff} (>1 V).

The higher J_{ph} usually means a higher charge extraction efficiency, which leads a relative higher fill factor (FF). In this case, the superior $J_{ph} - V_{eff}$ characteristics from the patterned PSC clearly identify the higher charge extraction property in the patterned PSCs due to the optimized contact between the ZnO film and the photoactive layer. In summary, high-performance single-junction PSCs with efficiency over 10% have been demonstrated by

implementing the DAN patterning for broadband self-enhanced light absorption with optimum charge extraction [23]. We expect this result will promote the development of high-efficiency PSCs with new device structures.

IV. CONCLUSIONS

With current achievements, OSCs are placed a nice applications that because of their advantages. The efficiencies and lifetimes continue to improve a new opportunities will emerge. In the relatively near future, OSCs will achieve a strong role in third-world, off-grid uses. In the long term, climate change and other factors demand TWs of energy be supplied by the sun, and the future of OSCs will have unique advantages for very large-scale power generation because of their scalability as well as their remarkably low cost, low energy payback time and carbon emissions, making them a most important competitor in the world energy marketplace.

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DESIGN AND IMPLEMENTATION OF A PIEZOELECTRIC ENERGY HARVESTING SYSTEM USING A SUPER CAPACITOR

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Abstract — a method of harvesting vibrational energy with a piezoelectric element using a step-down DC-DC converter suggested by previous references is investigated for application based circuits. Based on the experimental data the current found from the piezoelectric material was very low so a super capacitor was introduced to increase the output current and also to store the harvested piezoelectric energy. Results based on vibrations picked up from the blades of a small DC fan show that vibrational energy can be harvested and the design of the prototype was successful.

Keywords — DC-DC step down converter, energy harvesting, piezoelectric element, super capacitor.

I. INTRODUCTION

Nowadays, wireless electronic devices have become a part of our everyday life. These devices are powered by remote electrical power supplies which are still not efficient when it comes to duration for which they can power these devices. This problem has spurred an interest into developing an alternative renewable energy source which can provide these devices with additional power hence increasing the efficiency of the devices.

Piezoelectric energy harvesting is an innovative way through which vibrational energy can be converted to electrical energy by mechanical excitation of the piezoelectric material. The drawback of piezoelectric material is that they do not provide a constant voltage and also a very low current. Thus this kind of material are said to be source of micro energy harvesting which can only be used by low energy consuming devices for an ultra low power solution [1].

In this paper a previously designed circuit is investigated which was able to harvest energy from piezoelectric material for any degree of mechanical excitation. Experimental results show a relationship between the charging time of the energy storage device and the number of piezoelectric elements. Based on those results, an array design is proposed from which a significant amount of energy can be harvested and stored at the least amount of time.

II. THEORY

The system of harvesting energy from piezoelectric material can be represented using the block diagram shown in Fig. 1.

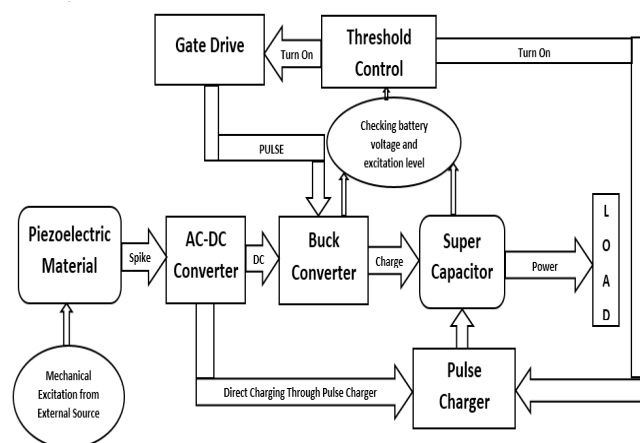


Fig. 1: Principle of Operation

Additional sub circuits are required for an optimized operation and the principles of operation of those sub-circuits will also be discussed in the subsequent sections.

As shown in Fig. 2, for the sake of simulation the electrical characteristics of a vibrating piezoelectric element can be modeled as a sinusoidal current source. Therefore voltage from the piezoelectric element must first be rectified using a full bridge rectifier. Moreover when the piezoelectric element is mechanically excited it produces a high voltage spike which charges up the smoothing capacitor to a very high voltage very quickly but the amount of charge stored in this capacitor is very low and hence the current produced is also low.

After the energy was harvested from the piezoelectric material, the harvested energy was used to charge up a generic capacitor before investigating with a super capacitor. From the experimental setup, the charging time for each type of capacitor was measured and then energy stored and current generated were both calculated.

As the charging of the capacitors was taking a very long time, it was decided that the capacitors should be charged to a predetermined voltage level.

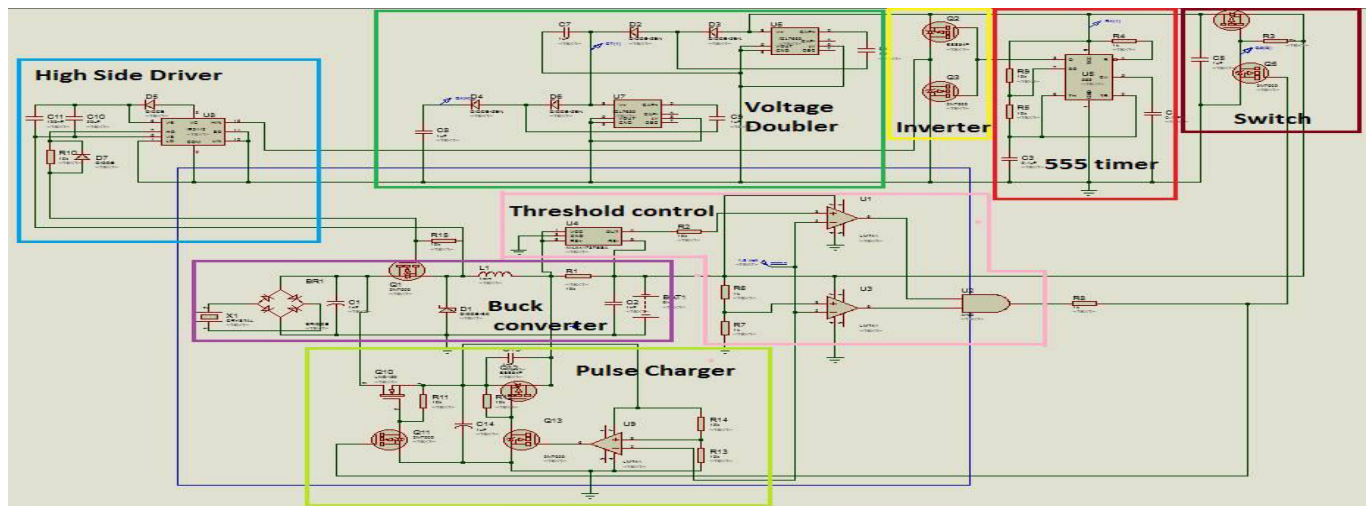


Fig. 2: Simulation Layout of Piezoelectricity Harvesting Circuit

The following parameters were then calculated using experimental results as described below:

- Time (t) to charge a capacitor up to predetermined voltage determined from experimental setup.
- Total charge stored in the capacitor calculate using:

$$Q = CV \quad (1)$$
- Total Energy stored in capacitor calculate using:

$$E = QV \quad (2)$$
- Harvesting Power was confirmed to be proportional to Energy per second.
- Output Voltages were measured across the AC-DC and DC-DC step down converter from the experimental setup and subsequent current values were also calculated.

III. CIRCUIT DESCRIPTION

The simulated schematic shown in Fig. 2 shows the sub-circuits: Duty cycle generator, voltage doubler, high side driver, threshold control and the pulse charger, along with main harvesting circuit (bridge rectifier, and the buck converter).

The buck converter consists of an N-channel MOSFET (2N7000), an inductor (10mH), a capacitor (2200uF) and a diode. 2N7000 is a fast switching MOSFET which is ideal for DC to DC step down conversion. The switching frequency of this MOSFET is provided by the duty cycle generator which consists of a 555 Timer (operating in astable mode) and a CMOS inverter. The duty cycle for the circuit is designed to be very small. Therefore for a small duty cycle (5%), a much higher voltage is needed to drive the gate of the MOSFET. The high side driver is a special purpose IC whose output only depends on the pulse or the duty cycle. It can produce a similar output waveform from an input with different voltage levels. The IR2110 IC is

used here for this purpose. But the biasing voltage of this IC is high (around 12V) and the voltage of the main storage device (at the load) of the main circuit is around 3V. Therefore a voltage doubler circuit is used to drive the IC. The voltage doubler consists of two ICL7660 ICs. The pulse-charger circuit was used at lower mechanical excitation levels to harvest energy by methods of direct charging. The pulse-charger is actually a secondary DC-DC step-down circuit consisting of a depletion-mode n-channel MOSFET (LND150), and an enhancement-mode p-channel MOSFET. The gate drive circuit and the pulse charger circuit operate depending upon the excitation level of the piezoelectric material and the voltage level of the battery. Threshold control circuit checks the excitation level of piezoelectric material and the voltage level of the storage device and determines whether energy would be harvested via the gate drive circuit or the pulse charger circuit.

IV. CIRCUIT IMPLEMENTATION

In the experimental setup the piezoelectric element is mechanically excited using a small DC fan as shown in Fig.3. The mechanical excitation on the piezoelectric element was varied by varying the speed of the DC fan. Energy was harvested using the designed circuit and was stored in capacitors at the output of the DC-DC step down circuit.

Results were recorded based on different scenarios that were implemented in the experimental setup:

- Scenario 1: Harvesting energy using the step down converter. And the harvested energy was stored in a generic electrolytic capacitor.
- Scenario 2: Harvesting energy by direct charging. And the harvested energy was stored in a generic electrolytic capacitor.

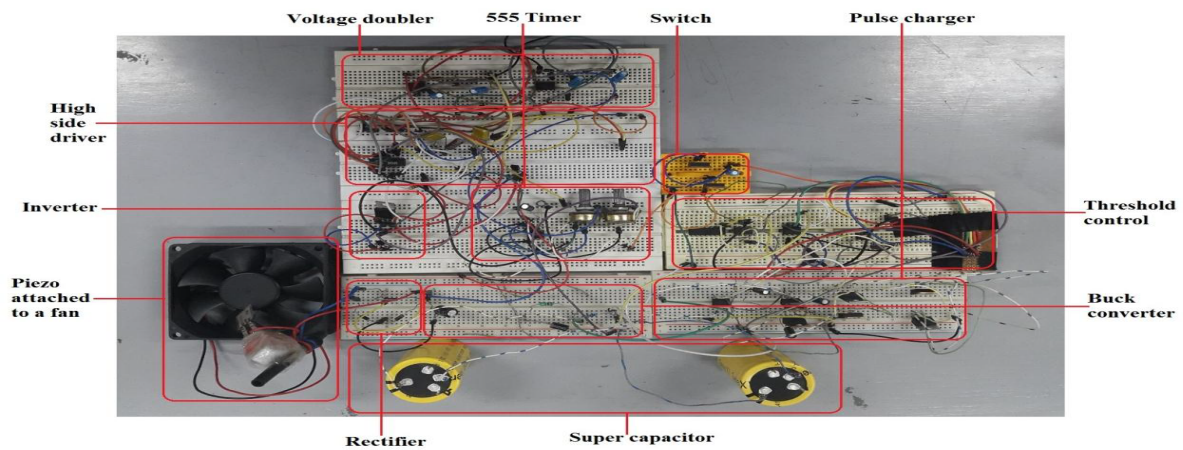


Fig. 3: Experimental setup of Piezoelectricity Harvesting Circuit

- Scenario 3: Energy harvesting by connecting the piezoelectric elements in both series and again in parallel. And the harvested energy was stored in a generic electrolytic capacitor.
- Scenario 4: Energy harvesting by connecting the piezoelectric elements in parallel. And the harvested energy was stored in a super capacitor.

The purpose of using the super capacitor was that when generic capacitors are used as the storage device for the harvested energy, the current provided by them is very low (in the μA range). The super capacitor used in this experiment (PowerStor XV3560) was 400F (ref. datasheet) and was thus able to provide a significantly higher amount of current (in the mA range) which can be used to power up external devices.

V. DATA ANALYSIS

The experimental results for power and current were recorded for the specific scenarios. Results obtained from the experimental setup were repeated at least for three times for each scenario and the most recurring values were selected and presented in this paper.

From table I and table II we can see that as the number of piezoelectric element connected in series is increasing, the charging time of the capacitor decreases and the calculated power and current increases. Furthermore, it can be seen that more energy is harvested when the harvesting circuit was used compared to the direct charging circuit (via pulse charger).

TABLE I. CALCULATED POWER AND CURRENT VALUES BASED ON EXPERIMENTAL DATA FOR DIFFERENT NO. OF PIEZOELECTRIC ELEMENTS (CONNECTED IN SERIES) AND USING THE MAIN HARVESTING CIRCUIT

No. of Piezo-electric element	Charging Time, t (Sec)	Energy, E (mJ)	Power, P (μW)	Current, I (μAmp)
1	126	5.778	45.857	9.17
2	55		105.055	21.01
3	30		192.500	38.52

TABLE II. CALCULATED POWER AND CURRENT VALUES BASED ON EXPERIMENTAL DATA FOR DIFFERENT NO. OF PIEZOELECTRIC ELEMENTS (CONNECTED IN SERIES) AND BY DIRECT CHARGING

No. of Piezo-electric element	Charging Time, t (Sec)	Energy, E (mJ)	Power, P (μW)	Current, I (μAmp)
1	81	3.38	41.780	2.19
2	40		84.600	4.43
3	25		135.36	5.43

TABLE III. CALCULATED POWER AND CURRENT VALUES BASED ON EXPERIMENTAL DATA FOR DIFFERENT NO. OF PIEZOELECTRIC ELEMENTS (CONNECTED IN PARALLEL) AND BY DIRECT CHARGING

No. of Piezo-electric element	Charging Time, t (Sec)	Energy, E (mJ)	Power, P (μW)	Current, I (μAmp)
1	81	3.384	41.780	2.19
2	16		211.50	11.08
3	12		282.00	14.78

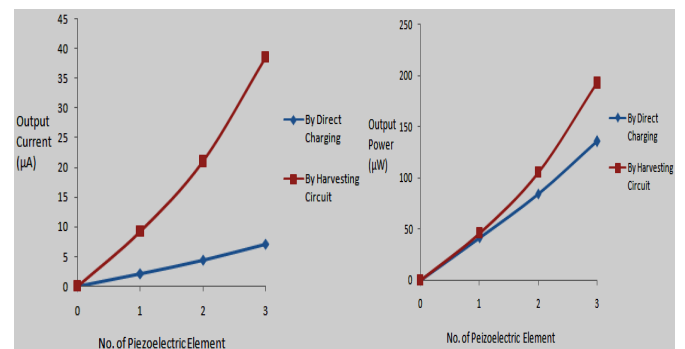


Fig. 4: Comparison between determined current and power from piezoelectric electric element with and without harvesting circuit

From Fig. 4 we can conclude that when the designed energy harvesting is used, the current across the load increases by approximately 5 times and the power stored increases by approximately 1.4 times of what it would have been possible without the harvesting circuit.

From Fig. 5 we can conclude that connecting piezoelectric materials in parallel gives us a both a higher current (approx. 2.5 times) and power (approx. 2.5 times) compared to connecting in series. Therefore for real life applications, connecting piezoelectric elements in parallel will ensure a better performance when harvesting energy.

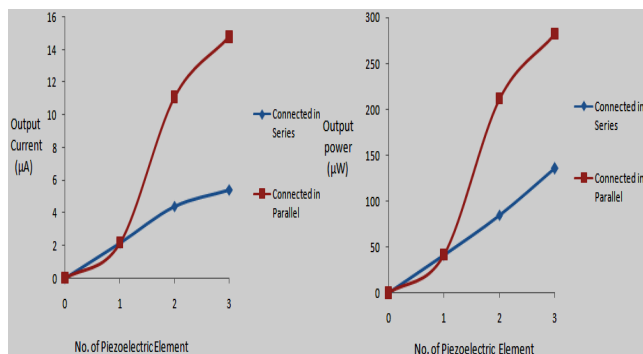


Fig. 5: Comparison between determined current and power from piezoelectric electric element connected in series and parallel

When the generic capacitors were replaced with super capacitors it was observed that the super capacitors were taking too much time to charge but the current that is provided by it is much higher than that provided by a generic capacitor. Therefore to find experimental values for the super capacitor, first outcomes were predicted based on previous results and theoretical values. Then experimentally, the change in voltage of the super capacitor was recorded for a fixed amount of time (2 hours) and using that data the total charging time of the super capacitor was calculated and compared with the theoretical and predicted results. In 2 hours the super capacitor was charged up to 3.73 mV (experimentally). Therefore to charge the capacitor to 2.7 V (maximum voltage level of the super capacitor) the time required was calculated to be 1447.7 hours.

VI. FURTHER DEVELOPMENTS

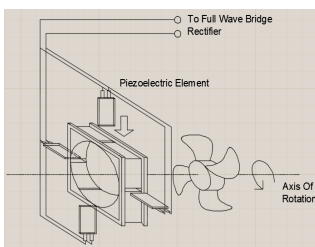


Fig. 6(a): Isometric View Of The Proposed Array

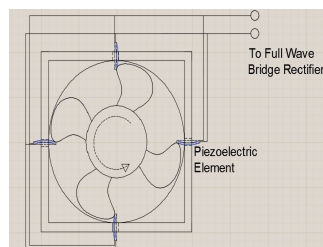


Fig. 6(b): Orthographic View Of The Proposed Array

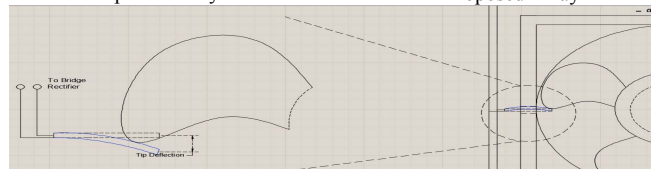


Fig. 6(c): Tip Of The Piezoelectric Element Being Deflected By The Blade Of The Fan Causing It To Produce A Voltage

From the experimental results it was concluded that the designed circuit was able to harvest energy when the tip of

the piezoelectric element was deflected by the DC fan blade. But the time to harvest a significant amount of energy was still very high. To overcome this problem industrial grade piezoelectric element such as PIC153 [2] could be used in the future which can provide up to 600nC of charge for 10 mm deflection (ref. datasheet). Further the piezoelectric element used in this experiment (LDT0-028K) could only provide 10 to 12 nC of charge for 10 mm deflection (ref. datasheet). So if the industrial grade piezoelectric element were used, the charging time of the super capacitor could be reduced from 1447.7 hours to approximately 24 hours. The charging time can be further reduced significantly by increasing the number of piezoelectric material in an array. Further, an array of piezoelectric elements was designed (as shown in Fig. 6(a-b)) so that they can be placed in specially designed slots of a DC fan so that maximum charge is produced when the tip of the piezoelectric element hits the fan's blade (as shown in Fig. 6(c)).

VII. CONCLUSION

In this paper the main objective was to present an approach to harvesting electrical energy from mechanically excited piezoelectric elements and based on the experimental results the data showed similar trends with previously established results[3]. Calculations based on experimental results showed that for best results a large array of piezoelectric elements connected in parallel along with super capacitor is necessary. This will produce considerable harvested energy which will then be able to drive an external load and also power itself up, thus making the system autonomous. Also, it was seen that the designed system was cost effective as the building cost of the prototype circuit was approximately BDT 1,000 but price will increase with respect to the type of piezoelectric material used. Further the system was able to harvest energy for varying degree of mechanical excitation which shows that the system can be used in real life applications where vibrational energy is abundant. The designed system along with the proposed modifications will be able to harvest enough energy to provide power to low power devices such as wireless sensors or store the harvested energy in power banks to charge up cell phones.

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A Study of Starting Methods for an Induction Motor Using an Arbitrary Waveform Generator

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Abstract—In this paper the use of an arbitrary waveform generator to generate desired signals in real-time is illustrated by simulating an induction motor operation along with various starting techniques. An off-line digital simulation of an induction motor is carried out on MATLAB/SIMULINK. The resulting sample-data files pertaining to signals such as electromagnetic torque and rotor speed are used in the arbitrary waveform generator to trace waveforms on an oscilloscope in real-time. This arbitrary waveform generator is developed on the Real-Time Application Interface (RTAI) enabled-Linux platform. The paper presents results of some case studies carried out by employing different starting techniques such as direct on-line and star-delta starting methods. Such a tool will be of great help in enhancing the class room learning.

Keywords—Real-Time Application Interface, Arbitrary Waveform Generator, Induction motor starting.

I. INTRODUCTION

In electrical industries, one of the most important constituents of electrical loads is the induction motor. Even for residential and commercial use, and in various day-to-day operations, the induction motors are widely employed. Thus, an increased usage of induction motors has made the study of various models and starting methods employed for induction motors, a very important aspect of the power system studies [1]. In understanding the model of an induction machine, it is worth noting the following aspects of its characteristics which differ from those of a synchronous machine [2]:

- The rotor has a symmetrical structure. This makes the d-and q-axis equivalent circuits identical.
- The rotor speed is not fixed, but varies with load. This has an impact on the selection of a d-q reference frame for modeling.
- There is no excitation source applied to the rotor windings. Consequently, the dynamics of the rotor circuits are solely determined by slip.

Having modeled an induction motor, it is interesting and essential to understand its starting performances when different starting methods are employed. When very large motors are started directly using a direct on-line (DOL) starters, they cause voltage dips on the supply lines due to large starting current surges. To limit the starting current surge, induction motors are generally started at a reduced voltage and then a full supply voltage is connected when they

run up to near rated speed [3]-[6]. In this paper, a digital simulation of an induction motor is carried out by employing different starting methods such as direct on-line and star-delta starting. The off-line simulated waveforms such as phase-currents, torque developed and speed of the rotor are generated as real-time signals using an arbitrary waveform generator developed in the Power Systems lab, Electrical Engineering department, National Institute of Technology Karnataka, India [7]. Such a facility is very useful to test and design various control and protection equipment in a cost effective and realistic environment. The arbitrary waveform generator has been built using the real-time application interface (RTAI) on Linux platform [8]-[11].

The paper is organized as follows: Section II presents a detailed modeling process of the induction motor using MATLAB/SIMULINK. Section III describes the features of the arbitrary waveform generator. Section IV presents a comparison of induction motor waveforms for different starting methods using oscilloscope captures.

II. MODELING OF INDUCTION MOTOR USING MATLAB

The induction motor considered for modeling has the following specifications:

3-phase, 2.4 kW, 460 V, 60 Hz, 1750 rpm, 4-pole, induction motor. $R_s = 1.77 \Omega$, $R_r = 1.34 \Omega$, $x_{ls} = 5.25 \Omega$, $x_{lr} = 4.57 \Omega$, $x_m = 139 \Omega$, $J = 0.025 \text{ kg-m}^2$. Assume full-load efficiency of 90% and full-load power factor=0.9 lagging.

Here, it is to be noted that synchronous reference frame (SRF) has been used for the simulation. The various subsystems along with the associated equations are described below:

1) *Park Transformation*: The equations used in this subsystem are:

$$v_{0dq} = [P] v_{abc} \quad (1)$$

where

$$[P] = \sqrt{2/3} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ \cos(\theta) & \cos(\theta - 2\pi/3) & \cos(\theta + 2\pi/3) \\ \sin(\theta) & \sin(\theta - 2\pi/3) & \sin(\theta + 2\pi/3) \end{bmatrix}$$

with $\theta = \omega_s t$. The corresponding blocks are shown in Fig. 1. A balanced 3-phase sinusoidal source voltage is assumed with per-phase peak voltage equal to $\sqrt{\frac{2}{3}}460$ V. An inverse Park transformation may be used as follows to compute i_{abcs} from i_{0dqs}

$$i_{abcs} = [P]^T i_{0dqs} \quad (2)$$

with $i_0 = 0$ for a balanced system. This in turn can be used to compute 3-phase real power as P_{vi} .

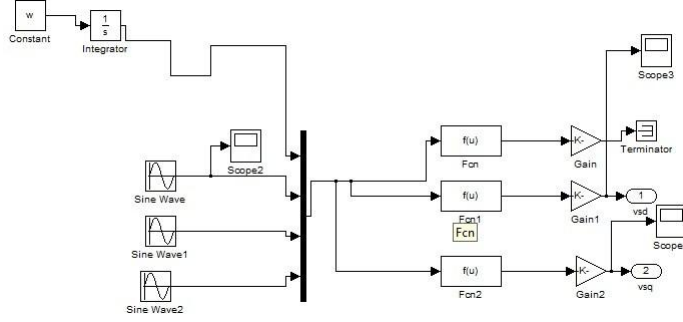


Fig. 1. Park Transformation Subsystem.

2) **Stator Winding Subsystem:** The state-equations used for the stator d- and q-axes windings are:

$$\begin{aligned} \frac{d\psi_{sd}}{dt} &= v_{sd} - R_s i_{sd} - \omega \psi_{sq} \\ \frac{d\psi_{sq}}{dt} &= v_{sq} - R_s i_{sq} + \omega \psi_{sd} \end{aligned} \quad (3)$$

Set $\omega = \omega_s = 2\pi 60$ electrical rad/s for SRF-based simulation. The respective blocks are depicted in Fig. 2.

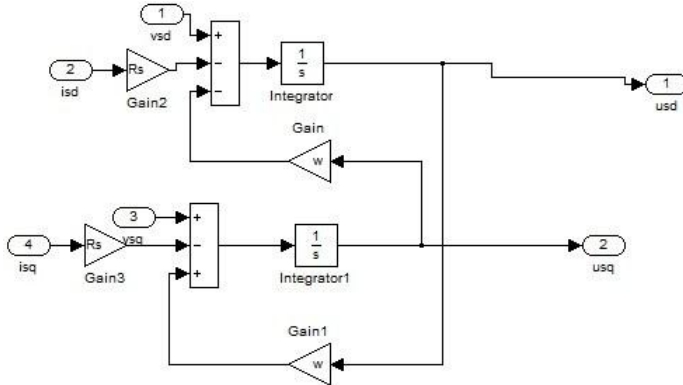


Fig. 2. Stator winding subsystem.

3) **Rotor Winding Subsystem:** The state-equations for the rotor circuits are:

$$\begin{aligned} \frac{d\psi_{rd}}{dt} &= v_{rd} - R_r i_{rd} - (\omega - \omega_r) \psi_{rq} \\ \frac{d\psi_{rq}}{dt} &= v_{rq} - R_r i_{rq} + (\omega - \omega_r) \psi_{rd} \end{aligned} \quad (5)$$

(6)

In the above equation v_{rd} and v_{rq} are set equal to zero to simulate a cage rotor. Set $\omega = \omega_s = 2\pi 60$ electrical rad/s for SRF-based simulation. Further, ω_r represents the rotor speed in electrical rad/s and is obtained using (8) as $\omega_r = \frac{P}{2} \omega_{rm}$, where $P = 4$ (poles). The corresponding blocks are shown in Fig. 3.

4) **Current Equations Subsystem:** The following expression is used to compute the machine currents from the flux-linkages obtained by solving the differential equations:

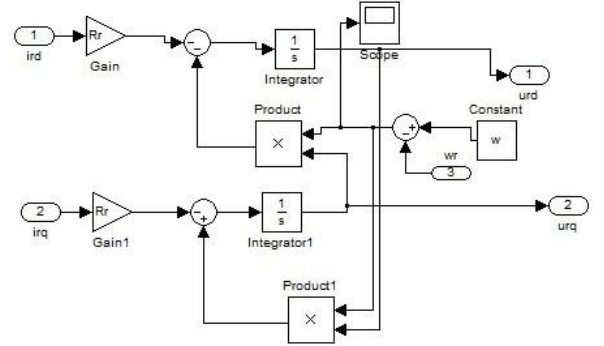


Fig. 3. Rotor winding subsystem.

$$\begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{rd} \\ i_{rq} \end{bmatrix} = \begin{bmatrix} L_{ss} & 0 & L_m & 0 \\ 0 & L_{ss} & 0 & L_m \\ L_m & 0 & L_{rr} & 0 \\ 0 & L_m & 0 & L_{rr} \end{bmatrix}^{-1} \begin{bmatrix} \psi_{sd} \\ \psi_{sq} \\ \psi_{rd} \\ \psi_{rq} \end{bmatrix} \quad (7)$$

All inductances are obtained using the reactances specified. Further, note that $L_{ss} = L_{ls} + L_m$ and $L_{rr} = L_{lr} + L_m$.

5) **Mechanical Equations Subsystem:** The state-equation used is as follows:

$$J \frac{d\omega_{rm}}{dt} = T_e - T_L \quad (8)$$

where the develop torque in Nm is

$$T_e = (\psi_{rd} i_{rq} - \psi_{rq} i_{rd}) \frac{P}{2} \quad (9)$$

and ω_{rm} is mechanical rotor speed in mechanical rad/s. J is the moment of inertia of the motor and load. T_L is the load torque and is assumed to be zero (i.e., free acceleration).

III. REAL TIME APPLICATION INTERFACE AND ARBITRARY WAVEFORM GENERATOR

Real Time Application Interface extension to Linux operating system makes it possible to perform hard real-time controls using a commercial PC. Normally, hard real-time control tasks are accomplished by employing dedicated processing units, for example, using digital signal processors and micro controllers having guaranteed interrupt latency. RTAI allows users to write real-time application in Linux environment. It provides guaranteed hard real-time scheduling, retaining all the features and services of Linux operating system. It adds a small real time kernel below the standard Linux kernel and treats the Linux kernel as a low priority real-time task. RTAI provides a large section of inter process communication and other real time services. Communication between the real-time tasks and outside world is achieved with the help of a multifunctional peripheral component interconnect (PCI) card [12].

The arbitrary waveform generation arrangement comprises of a PC which has the Linux operating system installed with the RTAI and an external PCI interface card. Two DAC interfacing cards are used for observing the output waveform on an oscilloscope. Each DAC interfacing card consists of a voltage level-shifter to convert the unipolar DAC output to bipolar. An active low-pass filter is cascaded at the output of the level-shifter to produce a smooth analog signal. The two DACs on the PCI card receive data from the RTAI module running in the PC at a frequency which is set by the tick period [7] of the real-time task. This tick-period is decided based on the sampling frequency of the signal to be generated. In most of the experiments performed, the sampling frequency used is 80 samples per cycle for a 60 Hz signal.

A. Process Flow for Arbitrary Wave Generation

- 1) From the off-line simulation on MATLAB, the data files are generated. These files are renamed as results1.dat and results2.dat and are saved in an appropriate folder location.
- 2) Now a program file opens both the files and converts the float data into character format which is readable by the RTAI-based kernel module.
- 3) The output data files after conversion will be named as newresults1.dat and newresults2.dat.
- 4) The RTAI-based kernel module file skt.c takes the above two files as input and inserts a kernel module that initiates a real-time task and generates two waveforms at the PCI card pins DAC0 and DAC1.
- 5) At the output of the DAC-interfacing cards an oscilloscope is connected to show the generated waveforms.
- 6) To verify the waveforms, using the data files results1.dat and results2.dat are plotted in MATLAB.

IV. CASE STUDIES FOR DIFFERENT STARTING METHODS

The following starting methods for the induction motor have been studied:

- 1) Direct on-line (DOL) starting method.
- 2) Star-delta starting method.

A. DOL Starting Method.

In this case important waveforms like torque, speed, stator currents, rotor currents, input and output powers, stator and rotor D-axis winding and Q-axis winding currents, stator and rotor fluxes are plotted and are compared with respect to oscilloscope captures. In the following lines only two waveforms such as torque and speed are presented.

Figs. 4 and 5 show the torque and speed waveforms for the DOL starting method. In the oscilloscope figure, the yellow waveform denotes the torque, whereas, the pink waveform depicts the speed signal.

B. Star-delta Starting Method.

For star-delta starting method, changes are made in the SIMULINK programs so that at the beginning of the simulation the supply voltage is set to $\frac{460}{\sqrt{3}}$ V (phase value) and when the speed reaches approximately 80% of the rated speed (which is around 300 rad/s) the voltage is switched to the line value. This switching instant corresponds to 0.56 s from the start of the simulation. The MATLAB plots are shown in Fig. 6.

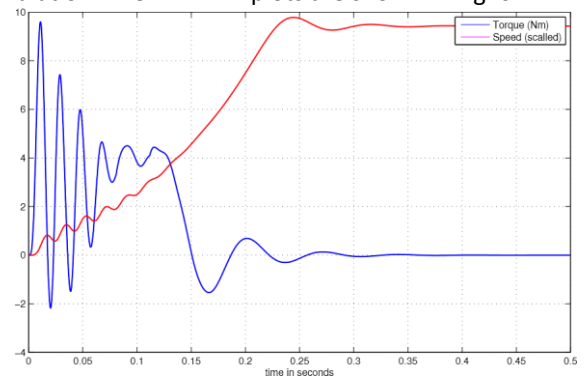


Fig. 4. MATLAB plot of torque and speed signals (Direct on-line).

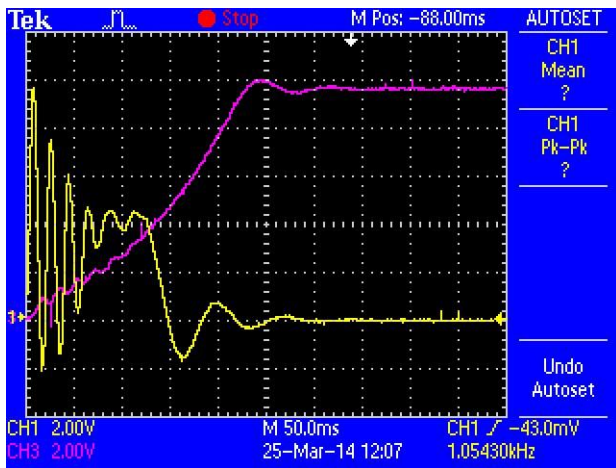


Fig. 5. Oscilloscope capture of torque and speed signals (Direct on-line).

In Fig. 7 the torque signal for the star-delta and the DOL starting methods is compared. Here, the yellow waveform represents the torque waveform for the DOL starting method and the pink trace denotes the torque signal for the star-delta starting method.

In Fig. 8 the speed signal for the star-delta and the DOL starting methods is compared. Here, the yellow waveform represents the speed waveform for the DOL starting method and the pink plot denotes the speed signal for the star-delta starting method. From the figure it can be seen that in case of the DOL starting the motor reaches the rated speed (on no load) with a run-up time of approximately 0.3 s, whereas with the star-delta starting, the run-up time is higher (0.7 s) due to the delayed application of the rated voltage.

V. CONCLUSION

In this paper off-line simulation of an induction machine with different starting methods namely direct on-line and star delta starting, is carried out. The obtained data files pertaining to various signals are re-played in real-time using an arbitrary waveform generator developed in the lab. The oscilloscope captures presented in the paper for different starting methods for the induction motor illustrate that such a tool really

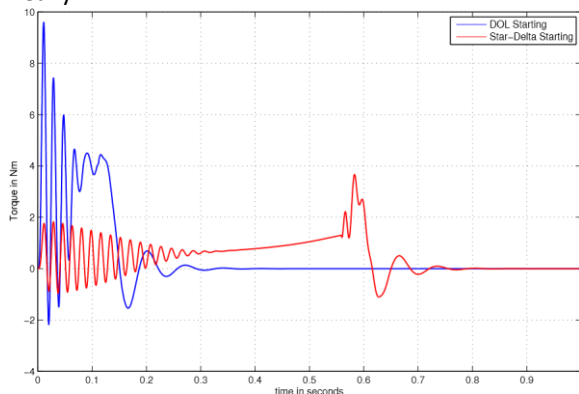


Fig. 6. MATLAB plots for the DOL and the start-delta starting methods torque signal.

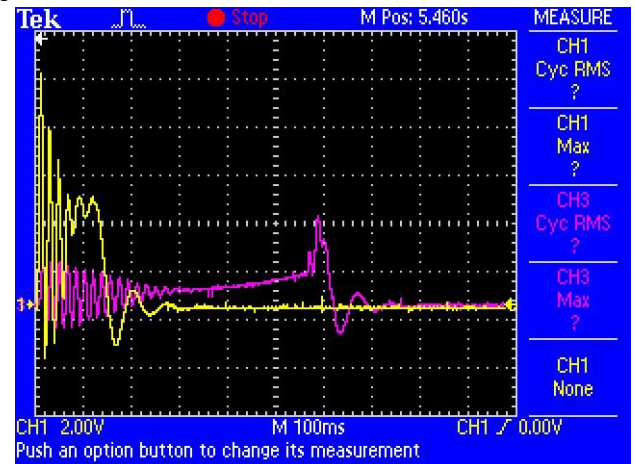


Fig. 7. Oscilloscope capture for the DOL and the start-delta starting methods - torque signal.

provides a new way to understand the theoretical concepts, and greatly augments the class room teaching-learning process.

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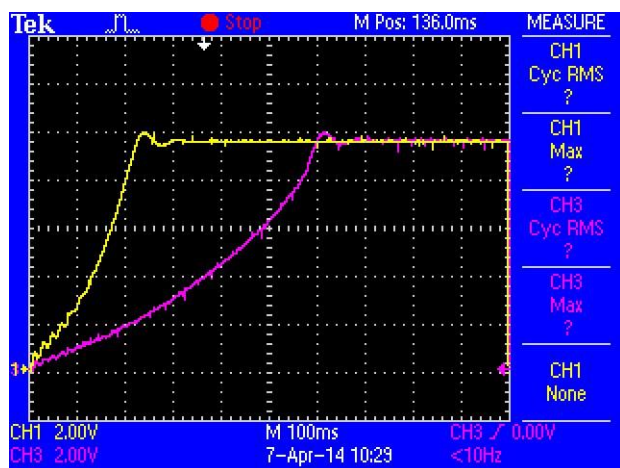


Fig. 8. Oscilloscope capture for the DOL and the start-delta starting methods - speed signal.

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Power Generation for Auto Street Light Using PZT

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Abstract—Energy consumption is an indicator of development of the universe. Modern world needs a huge amount of electrical power to meet up the current demand whereas the resources of conventional energy sources are diminishing steadily in response of vast consumption of energy. So, alternate sources of energy are required not only to fill up the large gap between demand and supply of electricity but also they should be clean, environment friendly and sustainable. This research work is based on the piezoelectric effect in which certain materials have the ability to build up an electrical charge from having pressure and strain applied to them. This research paper focused on the applications of auto street light for transportation facilities using that energy. When the roads or the tires of vehicle are engineered with piezoelectric technology, the energy produced by the pressure of moving vehicles is captured by piezo sensors and converted it into electrical charge by Lead Zirconium Titanate as a Piezoelectric Transducer (PZT), then stored the energy and used as an energy generation source. This energy source can be used for auto street lighting as a source of roadside power generation unit. In this research work, generated voltage and current depend on the system configuration of PZT, stress on the PZT, displacement of PZT from its original state and the way of connection of PZT plates.

Keywords—Energy consumption, sustainable, piezoelectric effect, Lead Zirconium Titanate, street light.

I. INTRODUCTION

The capacity of conventional energy sources are decreasing while the demand of energy is increasing every moment to fulfil the rapid development of the world. Fossil fuels are a finite resource as it takes burning 1 year while duration of making takes 15,000 years [1]. Moreover, power generation from fossil fuel emits carbon dioxide (CO₂) which causes smoke and responsible for global warming. Consequently, the earth temperature is increasing and the CO₂ creates earth thermal blanket which causes radiated heat from the earth cannot pass to the atmosphere. So, it is burning question to harvest renewable energy sources that can reduce the CO₂ emission and meet the future energy crisis. Generating electricity from renewable sources of energy rather than fossil fuels offer significant public health benefits since positive environmental impact awareness to replace the conventional energy sources by renewable energy. Renewable energy can also improve the energy security of the present world.

Piezoelectric effect is shown by specific materials having the ability to build up an electrical charge from pressure and strain applied to them. Many researchers are trying to develop system to generate power from different piezoelectric materials. Yun et al. showed the possibility of piezoelectric power generation from lead free material Lithium Niobate (LiNbO₃) crystalline [2]. Piezoelectric Barium Titanate (BaTiO₃) thin film has been applied to convert mechanical energy to electrical energy by Park et. al [3]. A system was

studied about frequency conversion in Potassium Niobate (KNbO₃) by Hohlaet .al [4]. Dakual et. al designed a system to produce piezoelectric power generation using ZnO [5] and Gupta et. al proposed electricity generation due to vibration of moving vehicles using piezoelectric effect [6]. Researchers are currently trying to develop automatic street light system using non conventional sources of energy. A control system designed for auto street light using LDR by Sudhakar et. al [7]. Rohaida Husin et. al showed automatic street lighting system based on low cost microcontroller [8].

In this research work, the implemented system is efficient since the provision to improve the efficiency. By increasing the number of PZT plates, efficiency and power generation both are increased that represent the flexibility of the implemented system. By keeping continuous input of pressed and released, efficiency can be controlled.

II. METHODOLOGY

The piezoelectric effect is a special material property that exists in many single crystalline materials [9]. Quartz, Rochelle salt, topaz, tourmaline, sugar cane, Berlinitite (AlPO₄), bone, tendon, silk, enamel, dentin, Barium Titanate (BaTiO₃), Lead Titanate (PbTiO₃), Potassium Niobate (KNbO₃), Lithium Niobate (LiNbO₃), Lead Zirconium Titanate (PZT) etc. are examples of such crystalline structure. PZT is well known energy converting source because it has high electro mechanical coupling coefficient [10]. In this research work, PZT is used to generate electricity for auto street light.

During the movement of a vehicle on road, it losses energy to the road surface in the form of vibration and sound etc. due to the transfer of its weight on the road surface and hits the wind. This energy can be tapped and converted into the usable form such as in electrical energy done by PZT.

III. SYSTEM DESIGN

When vehicle pressurizes on the piezoelectric plates, reasonable voltage is produced as proportional to the displacement of piezo crystal materials. The AC signal with high frequency is converted into DC for application purpose. The increase in number of piezoelectric plates causes higher generated voltage. The charging circuit charges a 12V battery where the inverter circuit converts this DC voltage to AC voltage. Dark sensing circuit uses to sense the night and switch to inverter so that street light lit on. Fig. 1 shows the whole system is represented by a block diagram.

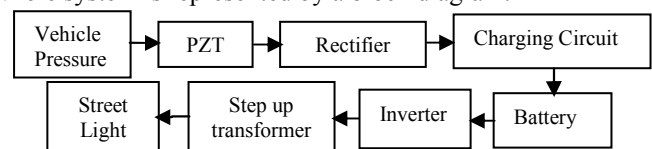


Fig. 1: Block Diagram

IV. SYSTEM CONSTRUCTION

In this research work, the following components are used to generate power for auto street light

- Piezoelectric plates
- Rectifier
- Frame
- Charging circuit
- Dark sensing and inverter circuit
- Step up transformer
- Street light

A. Piezoelectric Plate

It is the main part of the design and when it is subjected to press and release, it produces positive and negative voltage respectively whereas the holding period does not generate any voltage. Fig. 2 shows PZT plate having positive and negative terminal.

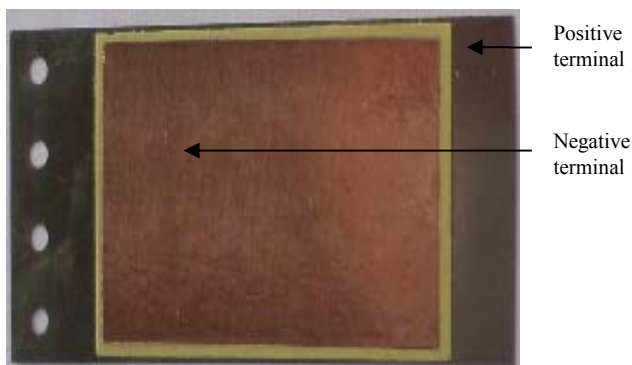


Fig. 2: Structure of PZT Plate

B. Rectifier

A rectifier converts AC to DC in only one direction and full wave rectifier and schottky diode are used in this study.

C. Frame

A spring with high spring constant is placed in the frame to bring the PZT at its original position. Fig. 3 shows the design of frame used in this work.

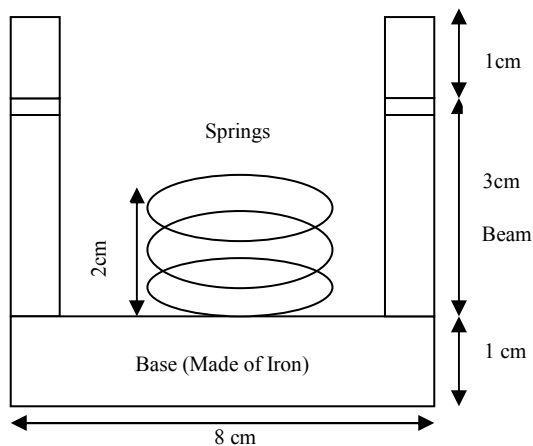


Fig. 3: Design of Frame

The PZT is covered and insulated with wool and placed in the frame. There are clips, nuts and bolts are proper sized to lock PZT and jutes or rubbers to work as insulator. This PZT plate is needed to lock with the road or tire of the vehicle. Fig.4 shows the mechanical set up of the design.



(a)



(b)

Fig. 4: Mechanical Setup (a) Top view (b) Front view

D. Charging Circuit

The generated energy through the piezoelectric effect is not sufficient enough for directly powering most electronic devices. Fig. 5 shows the construction PZT rectifier circuit and it is used as a unit block to charge the battery. It consists of a full wave rectifier and the battery intended to be charged. In this design, schottky diode (1N5819) is used which has threshold voltage smaller than that of a p-n junction diode.

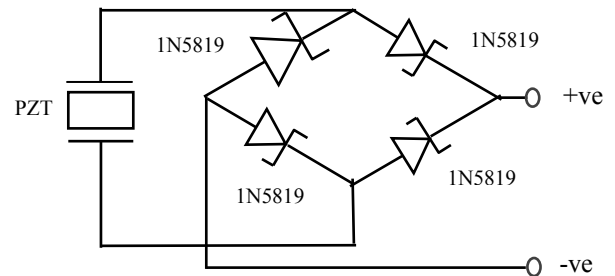


Fig. 5: Unit block of PZT rectifier circuit

PZT set produces energy for every pressure on it and this PZT AC signal is converted into DC signal after rectification. The charging circuit for 10 PZT plates is shown in Fig. 6. Capacitor is used to store DC voltage in battery.

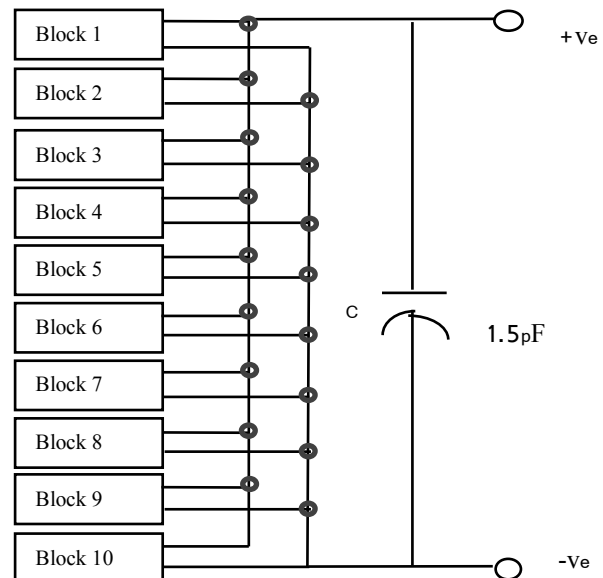


Fig. 6: Charging Circuit (for 10 PZT)

E. Dark Sensing and Inverter Circuit

LDR is a light sensing detector and a comparator is used to detect the dark while another comparator input is reference voltage. Relay is used to switch on the inverter circuit which is self-commutated voltage source square wave type. IC 555 is used as the clock pulse of the flip-flop and it is in astable mode. RV2 is used to control the frequency ranging from 10 Hz to 100 Hz. R3 & C1 both create the time for switching and R1 & R2 provide a voltage divider for the flip-flop circuit's clock. IC 4013 and D-type flip-flop is used in this work. Fig.7 shows the dark sensing and inverter circuit.

F. Step Up Transformer

Step up transformer is used to make the voltage to 250 V AC from 15V AC that is converted from 12 V DC by flip flop.

G. Street Light

A streetlight is a raised source of light on the edge of a road or walkway. Modern lamps may also have light-sensitive photocells that activate automatically when light is or is not needed: dusk, dawn or the onset of dark weather. Generally, fluorescent lamps are used for street light in Bangladesh. These lights require 220 V AC to lit on.

V. SYSTEM IMPLEMENTATION

Design of the system is implemented in Chittagong Collegiate School (CCS) situated in Ice Factory Road, Chittagong-4000, Bangladesh on 15 July 2015. Fig. 8 shows the implemented system to generate power for auto street light.



Figure 8: System Implementation

VI. EXPERIMENTAL DATA ANALYSIS

A. Mathematical Expression

In this system, efficiency is expressed by

$$\eta = \frac{V \times I}{m \times g \times l} \times 100\% \quad (1)$$

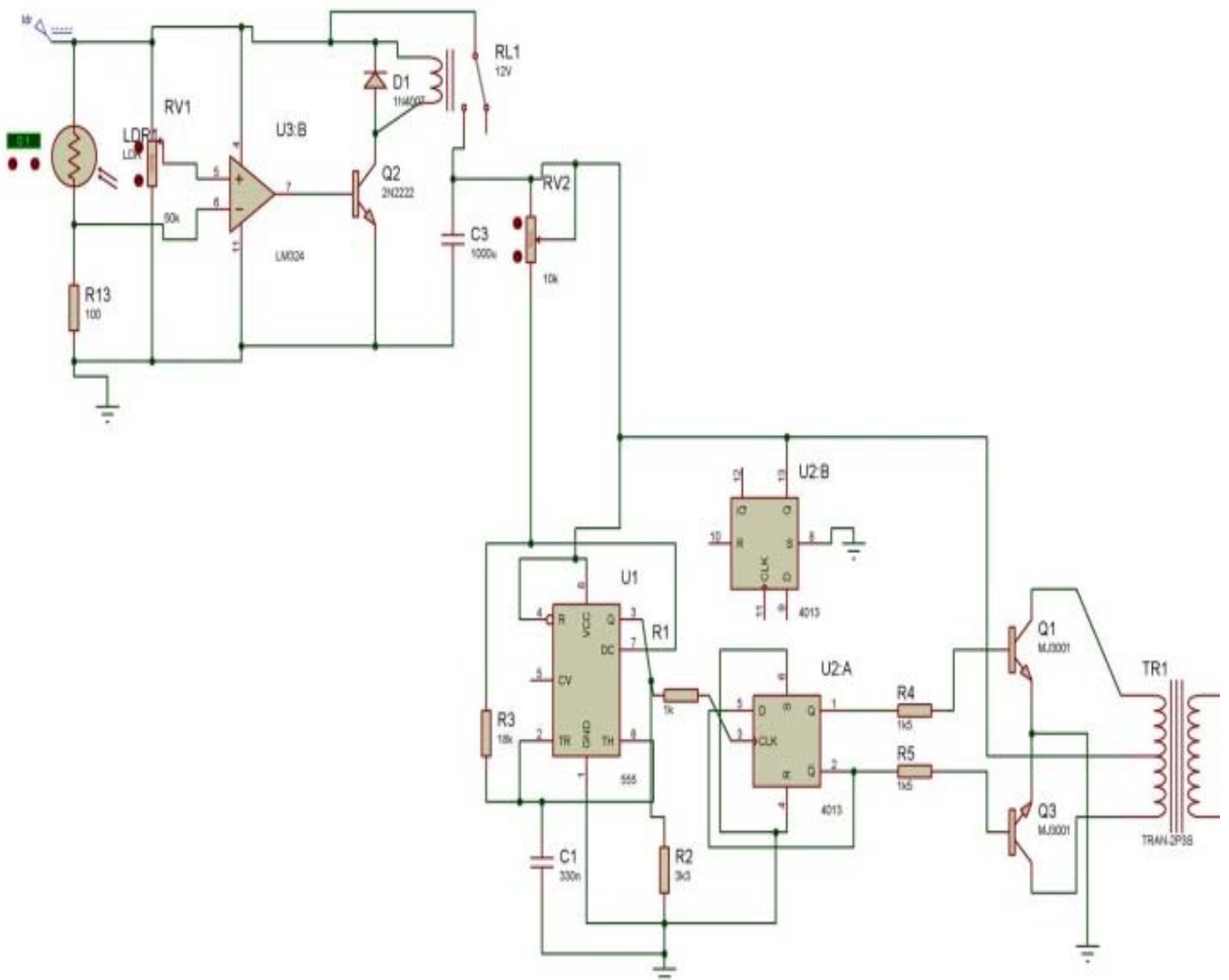


Fig. 7: Dark Sensing & Inverter Circuit

where, V is the generated voltage of the system in volt
 I is the current through the system in ampere
 M is the mass of the vehicle in Kg
 g is the gravitational acceleration in m/s^2
and l is the displaced length of PZT/time

1st case: For 6 PZT plates, Efficiency calculation is

$$\eta = \frac{21.9 \times 0.00022}{60 \times 9.8 \times 0.0002} \times 100\% = 4.1\%$$

2nd case: For 8 PZT plates, Efficiency calculation is

$$\eta = \frac{27.4 \times 0.00025}{60 \times 9.8 \times 0.0002} \times 100\% = 5.82\%$$

3rd case: For 10 PZT plates, Efficiency calculation is

$$\eta = \frac{29.6 \times 0.00028}{60 \times 9.8 \times 0.0002} \times 100\% = 7.05\%$$

From the above calculations, it is seen that efficiency is increased with the increased number of PZT plates.

B. Charging Time

If 3448 PZT plates are used for a 4.5Ah 12 V battery, the charging time is

$$t = \frac{Ah}{I} \quad (2)$$

Where, t is the time in hour

Ah is the ampere-hour rating of the battery

I is the generated current for 3448 PZT plates

Charging time of the battery, $t = \frac{4.5}{0.5} = 9$ hours.

VII. RESULT

During experiment, generated voltage and current are increased with the increase of the number of PZT plates. The summarized form of this research work is shown in Table I.

TABLE I: RELATION OF VOLTAGE & CURRENT WITH NUMBER OF PIEZOELECTRIC PLATES

Observation No.	No. of PZT plates	Voltage (Volt)	Current (mA)
1	6	21.9	0.22
2	8	27.4	0.25
3	10	29.6	0.28

In this research work, the relation between generated voltage and number of PZT plates is shown in Fig. 9.

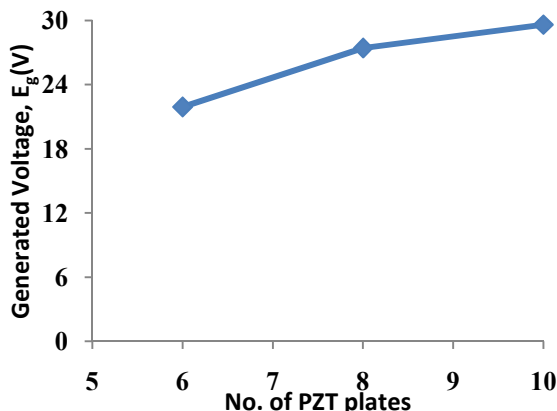


Figure 9: Relation between generated voltage and number of PZT plate

VIII. ADVANTAGES

The following advantages can get using this technology

- No manual work necessary during generation
- Simplicity
- No fuel transportation problem
- Cost effective
- Easy to maintain
- Easy to control
- Reduces labour cost
- Low power consumption
- Energy available all the year round

IX. CONCLUSIONS

The implemented system is a resource of renewable energy and power generation using this technology is not only environmental friendly but also clean, cost effective and safe. Sustainable development is possible by adopting this technology. It clearly tackles two problems that world is facing today, saving of energy and disposal of incandescent lamps can be tackled very efficiently by the system.

ACKNOWLEDGMENT

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Electricity Generation from Sewage Sludge: A Highly Prospective Renewable Energy for Power Problem Solution in Bangladesh

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Abstract—In many countries power is being generated from various biogas or other renewable resources. One of the very common sources of biogas is sewage sludge which is generally a threat to the environment. In Bangladesh this threat could be turned to a blessing by using the treated outcomes of sewage sludge into generating electricity. In Dhaka city everyday a large amount of sewage sludge is being produced, which is approximately 1,841,980 kg/day. If this huge amount of sludge is being perfectly entreated, then we could get a huge amount of biogas and energy from it. In this paper the generated electricity in the aspect of Dhaka city from the sewage sludge has been calculated which could be approximately 58.12 MW per day basis. The average electricity demand of Dhaka city in 2013 was 520.18 MW/day. But the supply power was 483.48 MW/day on an average. From the sludge per monthly total generated power could reach as high as 1.74 GW approximately. So the generated power from the sewage sludge will not only be an upright contribution to the national grid, but also reduce our environmental threat as well.

Keywords—*Sewage Sludge; Biogas; Electricity Generation; Sedimentation Tanks; Lagoons; DESCO; Dhaka WASA.*

I. INTRODUCTION

Through recent years the Economist Intelligence Unit (EIU) has rated the Dhaka city as the second last livable city in the world by considering its pollution, cleanliness and other criterion [1]. The steps taken to decrease pollution growth are not working adequately. However, to keep clean environment the Government has set up a sewage sludge treatment plant near Dhaka city which mainly takes the sludge in and treats it out.

The treatment process depends on the characteristics of the sludge. The amount of sludge can be reduced to as little as a third of its initial volume. Unlike, wastewater treatment plants (WWTPs) sludge are recycled in influent way. Various methods including mechanical, biological, chemical pre-treatment and pre-treatment of sludge dewater ability to improve the digestibility can be accomplished. The most common treatment options are aerobic digestion, anaerobic digestion and composting. After completing the treatment the sludge raw material can be used to produce biogas. Mainly the biogas can be used for electricity generation which further depends on rotor structure, generator size, generation facilities

as well as the techniques. Different types of process vary the efficiency of the power production and also the availability matters to generate power, based on specific zone.

In this paper some emerging technologies and concepts are discussed to procure electricity from sewage sludge. Moreover, deep contrasts are shown with graphical analysis in case of current power generation with the additional generated power from sludge material. This paper is organized as follows: Section II describes the review of sludge characteristics. Treatment of sewage sludge is described in Section III. Section IV extensively represents the power generation from sewage sludge. Section V provides Power Contribution Analysis. Future works and concluding remarks are drawn in section VI and VII.

II. REVIEW OF SLUDGE POWER AND CHARACTERISTICS

After the research and development of sewage power technology many countries around the world are looking forward to utilize this promising source of renewable energy. In 2009, Northumbrian Water, one UK water company used Advanced Anaerobic Digestion in its wastewater treatment process to generate biogas and used it in an on-site CHP unit which generated 4.7 MWe from four on-site CHP engines. Annually the advanced digestion facility had an annual output of 37 GWh [2].

In 2010, researchers at the University of Nevada, US conducted an experiment to transform sewage sludge to biosolids at a Reno-based treatment facility to electrical power which yielded positive results. The process involves drying the sludge and gasifying it, thus using the gas for electricity production. The project could process 20 pounds of sludge per hour. The plant's entire electrical power could potentially be supplied on-site by the process, say researchers working on the project. They concluded that with the biosolids produced in the US, as much as 1,500 MW of renewable energy could be produced at wastewater treatment plants (WWTPs), enough to power nearly 1.2 million US homes [3]. The researchers also estimated, which would be further refined through the research, showed that a full-scale system could potentially generate 25,000 kilowatt-hours per day to help power the local reclamation facility [4]. Enerzea Power Solution, Bangalore, India, a Government approved clean Energy Power Company is also working on sewage power technology.

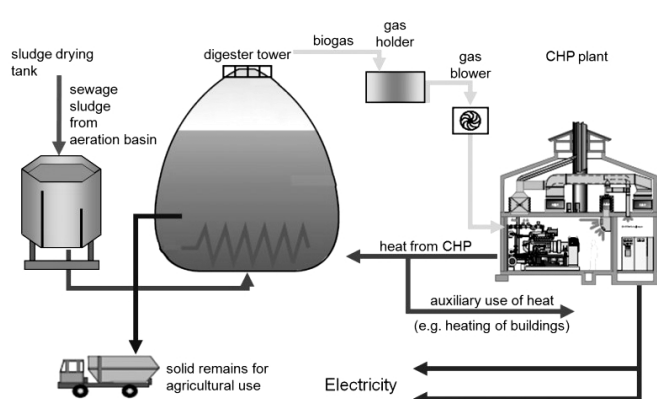


Fig. 1. Sludge power generation process of Western Branch Diesel, Portsmouth, Virginia, United States.

Fig. 1 shows an existing process of sewage power generation system by Western Branch Diesel, US. From above overviews of the worldwide approach towards sludge power, it is inevitable that the large amount of sewage sludge collected per day in Dhaka city can also be turned into electricity. To make a proper use of it the characteristics of sludge should be known first. The sludge contains considerable amounts of organic matter, pathogens and chemical contaminant which should be properly handled to get rid of health hazard [5]. So, different zones of Dhaka city carry different types and quantities of sludge.

A. Basic Classification

Basically sludge can be primarily classified in two categories, i.e. Class A and Class B. Class A type, also known as exceptional quality sludge. This kind of sludge is typically dried and pasteurized. Class B includes only that sludge's which are not included in Class A and they are usually undigested and volatile [6].

B. Primary Sludge

Sludge from the primary settling tank is called primary sludge which is usually gray and slimy and in most cases has an extremely offensive odor. Primary sludge can be readily digested under suitable conditions of operations [7].

C. Activated Sludge

When waste-activated sludge are taken from a wastewater treatment plant the oil and grease in the influent are first removed then recycled back to the aeration basin. The activated sludge sonication tests are conducted with the help of a cell-breaker [8].

D. Tertiary Sludge

It is derived from a tertiary treatment process and compromises the fractional of secondary sludge that remains in the effluent from the secondary that is removed by the tertiary treatment filters and contains no more than 1.0% dry solids [7].

E. Humus Sludge

Humus sludge is the product of settlement of effluent from biological filters. Mainly bacterial and fungal material are

sloughed from the filter media. The typical humus sludge contains 0.5-2.0 % dry solids [7].

F. Surplus Activated Sludge

To maintain an optimum level of mixed liquor suspended solids a portion of the sludge is removed at regular intervals and it is known as surplus activated sludge. Surplus activated sludge rarely contains more than 1.0% dry solids [7].

III. TREATMENT OF SEWAGE SLUDGE

A. Thermal Treatment

Bacteria are destroyed by heat sterilization temperature in the sludge, which is raised to a level. The most commonly used method for thermal pasteurization of primary and excess sludge in waste water treatment plants is done with a temperature greater than 65°C to 100°C [9].

B. Chemical Treatment

The sludge hygienisation is controlled by maintaining the pH value of 12 which can be achieved by the calcium chemicals. Hygienisation is usually controlled by measuring the physical and chemical properties of the waste water. Chemical dosing equipment and a storage silo mixing unit, together with the sludge and chemicals needed for operations and maintenance treatment is relatively easy [9].

C. Biological Treatment

By stabilizing the organic waste composting (3-6 days at 60°C) and humus, an aerobic bacterial decomposition process is carried out [6]. Anaerobic digestion (AD) facilities are sometimes called biogas plants. AD stabilizes the sludge to reduce the number of pathogens and from fossil fuels to clean, renewable alternative to continuous and biogas (64% CH₄, 35% CO₂-) production. Biogas is a clean fossil fuel, persistent and renewable option [10].

D. Combustion and Wet Oxidation

Wet thermal oxidation and wet oxidation process include sections performed at a water temperature of 150-330°C and atmospheric pressure of 1-22 MPa by using pure oxygen. So the process is occurred in two ways, the first of which is below 374°C and a pressure of 10 MPa in the second sub-critical condition is occurred, the situation in the supercritical 374°C and a pressure of 21.8 MPa. The main output is more than 95% of mineral material and low-molecular organic substances in the sludge which contain less than 3% [10].

E. Gasification Process

A combustible gas containing combustible materials gasification of non-feed stocks is a thermal process whereby air is converted. The first stage is a thermal drying to remove the remaining water. Exploiting the dry solid content output, 85-93%, depending on the type of gasifies. Thus the sludge is ready for gas, i.e. Sub-stoichiometries oxygen input with the 'incineration' [11].

IV. POWER GENERATION FROM SEWAGE SLUDGE

A. Selecting of Generator

The salient pole induction generator would be a good choice because the turbo alternators can run at a very high speed. It consists of a smooth, solid forged steel cylinder, having a number of slots milled out at intervals along the outer periphery for field coils. These rotors are designed for turbo generators running at 3600rpm. The prime mover of salient pole rotor is hayed with a turbine or a combustion engine where the non-salient pole rotor is generally connected with gas or steam turbine.

B. Model for Electricity Generation

After completing the treatments the extracted sludge raw materials can produce biofuels. By taking some purification steps the biogas is supplied to the power plant where it uses the biogas to rotate a turbine and electricity is generated from the rotation of the turbine. Thus, it produced electricity which is further supplied to the national grid. The block diagram Fig. 2 shows all steps from collection of sewage sludge to the electricity generation.

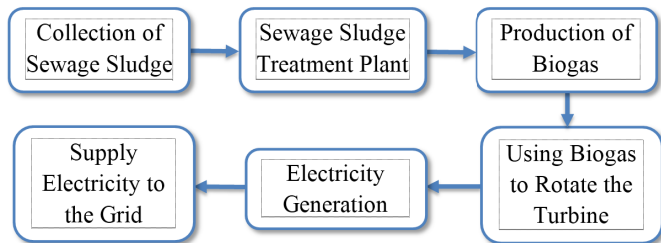


Fig. 2. Block Diagram of the Model for Electricity Generation.

C. Generated Power Calculation Procedure

Table I. shows the percentage of sewage sludge collection corresponding to sedimentation tanks from different locations in Dhaka city which have been collected from Dhaka WASA (Water Supply & Sewerage Authority) office. The calculation procedures of electricity generation from sewage sludge are also shown in step by step.

TABLE I. THE SEWAGE PERCENTAGE IN DIFFERENT AREAS (NEARBY DRAIN AND NEARBY SEWER)

Zone (Respondents)	Nearby drain (nos.)	Nearby drain (%tank)	Nearby sewer (Nos)	Nearby sewer (%tank)
Uttra	5	38.46	5	38.46
Tejgaon	1	11.11	6	66.67
Shyampur	6	18.75	3	9.38
Hazaribagh	17	23.94	8	11.27
Mirpur	5	27.78	63	350.00
Jatrabari	14	31.11	59	131.11
Demra	1	1.00	2	2.00
kotwali	6	6.67	3	3.33
Gulshan	25	15.15	0	0.00
Badda	24	24.24	4	4.04
Dhanmondi	8	9.76	5	6.1
Total		207.97		622.38

So the Total percentage = (Nearby drain + Nearby sewer)

$$= 830.33\% \text{ sewage sludge/tank}$$

Calculation from Sedimentation Tanks

(Data are collected from Pagla Sewage Treatment Plant, Dhaka)

(A Lagoon + B Lagoon) = 8 (Need 8 days to fill-up)

The average density (D) of water sludge solids is 1400 kg/m³ and the sedimentation tank has the following measurements.

Diameter (d) = 35m,

Radius (r) = 17.5m,

Height (h) = 3.5m

Sedimentation Tanks volume (STV) $v = \pi r^2 h = 3367.40 \text{ m}^3$

4 tanks of STV = $4 \times 3367.40 \text{ m}^3$

= 13469.61 m³/8day

(Sedimentation tanks are filled up in 8 days)

= 1683.70 m³/per day

Calculation from Lagoons

Lagoon A types: $(10.8 \times 8 \times 2.5) \text{ m}^3 = 216 \text{ m}^3$

Lagoon B types: $(7.6 \times 8 \times 2.5) \text{ m}^3 = 152 \text{ m}^3$

Total Volume of Lagoons = $368 \times 8 \text{ m}^3 = 2944 \text{ m}^3$

So, Per day = $\frac{2944}{8} \text{ m}^3 = 368 \text{ m}^3$

Sludge volume (v) = (STV - Volume of Lagoon) = 1315.7 m³

We know, Mass(m) = $(1315.7 \times 1400) \text{ kg} = 1841980 \text{ kg}$

Weight = $1841980 \text{ kg} \times 9.81 \text{ ms}^{-2} = 18069823.8 \text{ N}$

Power Generation in daily, monthly and yearly basis

Electrical power generation potential,

$$= \frac{0.00116 \times \text{Wweight} \times \text{NCV}(\text{net calorific value})}{24}$$

$$= \frac{0.00116 \times 18069823.8 \times 2000}{24}$$

$$= 1.75 \text{ MW only for (100\%)}$$

But we get 830.33% for one sedimentation tank

4 sedimentation tank, we get = $(830.33 \times 4) = 3321.32\%$

Total generated power = $\frac{1.75 \text{ MW} \times 3321.32}{100} = 58.12 \text{ MW (Daily)}$

[1.7436 GW (Monthly), 20.9232 GW (Yearly)]

For (Watt / kg) calculation,

Watt per kg in daily = $\frac{1.75 \text{ MW}}{1841980 \text{ kg}} = 0.95 \frac{\text{watt}}{\text{kg}}$ (daily)

Hence, $28.5 \frac{\text{watt}}{\text{kg}}$ (monthly), and $342 \frac{\text{watt}}{\text{kg}}$ (yearly)

V. POWER CONTRIBUTION ANALYSIS

A. Sewage Sludge in Different Areas

Table II. shows the amount of daily power that can be generated from different sectors in Dhaka city. In total the

generated power is 58.12 MW/day from drains and sewers located Dhaka city.

TABLE II. POWER GENERATION PER DAILY BASIS

Zone (Respondents)	Nearby Drain (nos.)	Nearby Drain (MW/Day)	Nearby Sewer (No's)	Nearby Sewer (MW/Day)
Uttra	5	2.7	5	2.7
Tejgaon	1	0.78	6	4.66
Shyampur	6	1.31	3	.66
Hazaribagh	17	1.67	8	0.79
Mirpur	5	0.94	63	24.5
Jatrabari	14	2.18	59	9.18
Demra	1	0.07	2	0.14
Kotwali	6	0.47	3	0.23
Gulshan	25	1.06	0	0
Badda	24	1.69	4	0.28
Dhanmondi	8	0.68	5	0.43

The pie chart in Fig. 3. and Fig. 4. show the percentage of electricity generation per daily basis based on Table II.

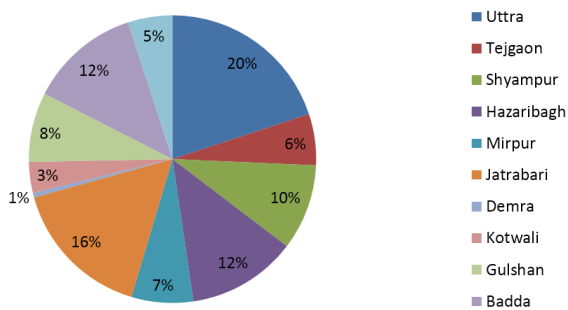


Fig. 3. Percentage of power generation per day (Nearby drain).

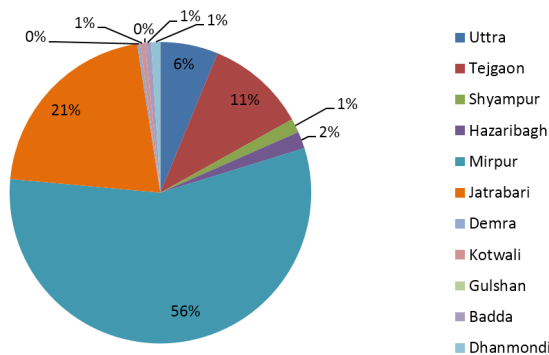


Fig. 4. Percentage of power generation per day (Nearby sewer).

B. Graphical Analysis Comparisons

The following Fig. 5, Fig. 6, Fig. 7, and Fig. 8 show the monthly demand power, supply power, and additional power which can be aided from sewage sludge electricity generation. The core data for demand power and supply power are collected from Dhaka Electric Supply Company Ltd (DESCO) office. The additional power (calculated power from sewage sludge electricity generation) is compared with those collected demand and supply power. All curves are plotted in OriginLab.

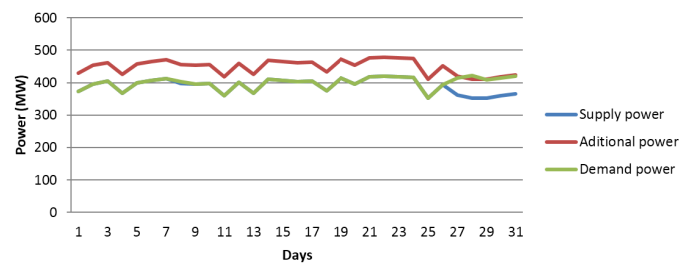


Fig. 5. Power demand curves for January, 2013.

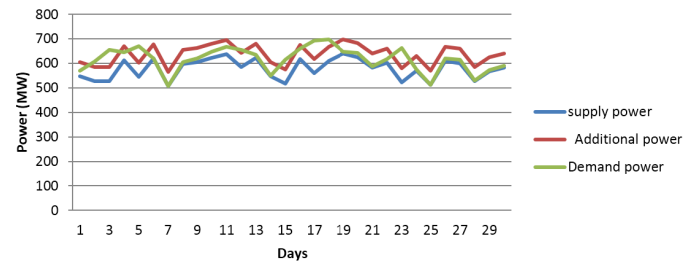


Fig. 6. Power demand curves for June, 2013.

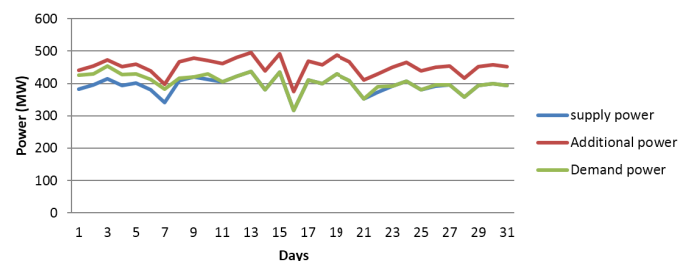


Fig. 7. Power demand curves for December, 2013.

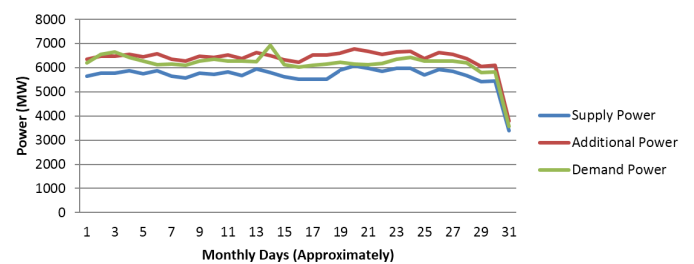


Fig. 8. Power demand curves for all monthly days, 2013.

January 2013 Here the observed average supply power is 390.35MW and the average demand power is 399.74MW. With the additional power (sewage sludge electricity generation) the total average supply power is 448.47 MW.

June 2013 Here the observed average supply power is 578.33MW and the average demand power is 616.17MW. With the additional power (sewage sludge electricity generation) the total average supply power is 636.45 MW.

December 2013 Here the observed average supply power is 394.29 MW and the average demand power is 403.93 MW. With the additional power (sewage sludge electricity generation) the total average supply power is 452.41 MW.

In 2013 the observed total supply power is 175,999MW and average supply power is 5677.39MW. And the total demand is 190,934MW and average demand power is

6159.16MW. By adding the additional power (sewage sludge electricity generation) the total additional power is 197,812.8MW and average supply power is 6381.058MW. So from all above calculations and plots it can be easily understood that the gap between demand power and supply power can be met from this highly prospective renewable energy concept: Electricity generation from sewage sludge in Bangladesh.

VI. FUTURE WORKS

Beside the discharge of domestic wastewater to the drainage system and the lack of suitable facilities for collection, transport, treatment and disposal of septic tank sludge, the main environment issue in Dhaka City includes the lack of treatment of industrial effluents discharged to the environment and the subsequent effects on the receiving waters or ground waters. The Dhaka East-Dasherbandi sewage catchment situated as 23°45.9 North and 90°27.8 East [12] comprises an area of approximately 72 square km, including parts of Badda, Khilgaon, Gulshan, Cantonment, Kafrul and Tejgaon, The River Balu flows along the eastern border of this catchment shown in Fig. 9.



Fig. 9. Dhaka East-Dasherbandi sewerage catchment boundary line.

At present (EST. 2011) the population is 1.4 million of this area and it is projected that the population of the catchment area will increase to 3.6 million by 2035. So if a power plant could be built in this site where power can be generated from sewage sludge which will ultimately contribute in the national power grid. So the daily increment of sewage sludge will become the resource of electricity generation which will stand as one of a very important backup power plant in the future in respect of Dhaka city.

VII. CONCLUSION

For multiple utilization of energy, such as clean, renewable energy efficient and environmentally friendly way to convert the waste activated sludge is currently developing a considerable concern. Sludge is being treated to protect the environment but proper outcomes are not yet accomplished in Bangladesh. Hence, emerging techniques has been used in this

paper to generate electricity from sewage sludge. The overall result indicates a significant amount of additional power that can be added to the national grid if necessary attempts are taken by governing bodies. The electricity generation from sludge material will not only solve the national power crisis but also make the city more clean and livable.

ACKNOWLEDGMENT

The authors acknowledge the authorities of Dhaka Electric Supply Company Ltd (DESCO) for giving permission to access the certified data. The Authors are also grateful to the staffs of Pagla Sewage Treatment Plant, Dhaka and Dhaka WASA (Water Supply & Sewerage Authority) for their assistance in collecting the valuable data and other information.

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Design and Performance Analysis of a Biomedical Implantable Patch Antenna

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Abstract—Implantable Medical Devices (IMDs) consist of miniature implantable antenna that is embodied inside and its bidirectional communication is monitored through external monitoring device. This paper presents an implantable miniaturized antenna working in the Medical Implant Communication Service (MICS) band and being carried out in human body phantom models representing implantation inside the arm which helped to design the proposed antenna. Furthermore, different characteristics of the antenna such as S parameter, return loss and directivity were contemplated using FR4 substrate material in the same band.

Index Terms—Implantable Medical Devices (IMDs), Biomedical Telemetry, Medical Implant Communication Service (MICS) band, Human phantom model

I. INTRODUCTION

Biomedical engineering helped to upgrade the health care and quality of life. IMDs (Implantable medical device), as a part of wireless telemetry, play a vital role in modern advanced healthcare system. With integrated implantable antenna bidirectional telemetry is monitored with the help of external monitoring equipment and it also plays a vital role in variety of diagnostic and therapeutic functions [1].

Implantable antennas have been characterized inside both tissue equivalent liquids and mimicking gels or in animal models. Before there were no globally accepted frequency band dedicated to biotelemetry for implantable medical device, until recently the ITU-R Recommendation SA.1346, which dedicated the use of 402.0-405.0 MHz frequency band for Medical Implant Communication Service (MICS) [2]. Due to its availability worldwide, workable in low-power and low-cost circuits, higher data rate of transmission with relatively low noise portion of the spectrum and smaller antenna designs, our main focus is on the MICS band in this paper [3]. Moreover the 433.1-434.8 MHz, 868-868.6 MHz, 902.8-928 MHz, and 2400- 2500 MHz Industrial, Scientific, and Medical (ISM) bands are also suggested for implantable medical device biotelemetry in some countries [6]. However, focus is on the MICS band because of its advantages of being feasible and acceptably propagating through human tissue [1]. In addition to the antenna size being small (small gain and narrow bandwidth), more challenges are introduced due to loss and complex structures of the human body in which these antenna's radiate. Due to their small sizes, uniform structures and inaccurate equivalent dielectric material, simplified body models offers many uncertainties into the design of implantable antennas [6]. Patch designs are currently receiving considerable attention for implantable antennas, because they are highly flexible in design, shape, and conformability thus allowing for relatively easy miniaturization and integration into the shape of the implantable medical device [1]. For optimum performance, the antenna is designed considering the environment where the antenna is implemented, which in this case dielectric properties of skin and fat in arm [4]. A good radiation performance and

miniaturization are the main two requirements that should be fulfilled for the implanted system [5].

II. ANTENNA DESIGN STRATEGY

To design an implantable antenna several methods and strategies have been proposed. These are mainly dictated by the fact that the antenna should work inside human body instead of free space [8]. In this paper a MICS patch antenna is designed in a free space environment and further it is implanted inside a three layered human phantom model. In this paper, a patch antenna at 404 MHz for implantable purpose has been designed.

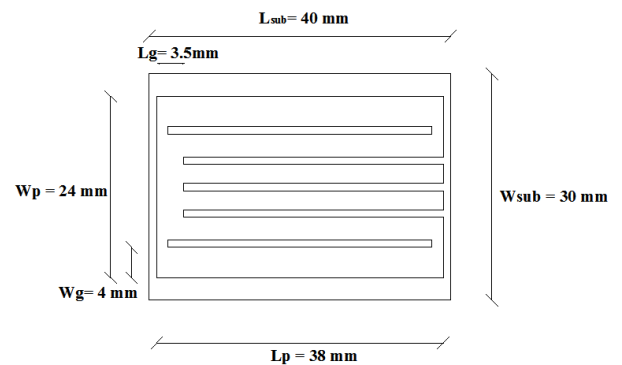


Fig.1. Geometry of the designed antenna

To design an implantable antenna miniaturization and biocompatibility plays a role factor. The antenna is designed with respect to the miniaturization and biocompatibility factor. Implantable antennas must be biocompatible in order to preserve patient safety and prevent rejection of the implant [1]. Because of the conductive characteristic of human tissue it can short circuit the implantable antenna if it directly contact with the metallization. Biocompatibility and prevention of undesirable short-circuits are especially crucial in the case of antennas that are intended for long-term implantation. The thickness of the biocompatible insulation layer is an important factor in antenna design [7].

III. SIMULATIONS AND RESULTS

The characteristic of the antenna are affected by the depth of the muscle inside which the antenna should be located and the electrical properties such as: dielectric constant and conductivity of the skin, fat and muscle.

A. Antenna Performance in Free Space

The antenna is designed in free space and the result of the proposed antenna is observed there. In free space the antenna resonance at 0.9133 GHz which is also a proposed band for the Medical Implant Communication Service (MICS).

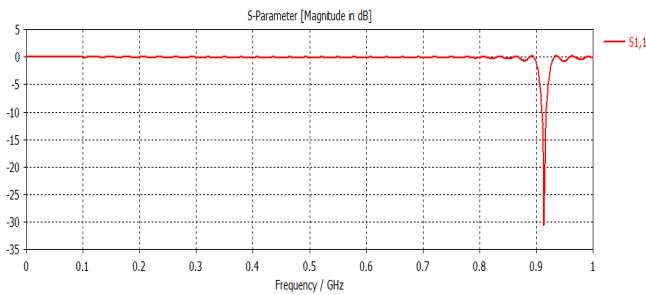


Fig.2. S parameter of the designed antenna in free space

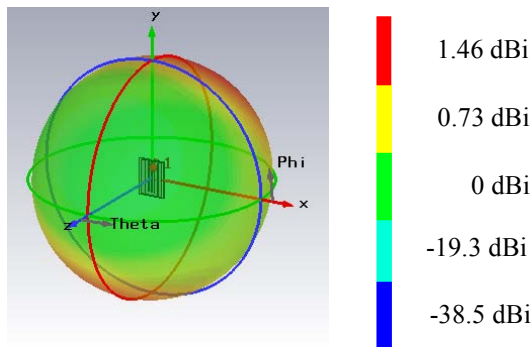


Fig.3. 3D view of the far field region of the designed antenna in free space

It can be seen from fig.2 that the designed antenna has a return loss of -30.1 dB.

$$\text{Return loss (dB)} = 10 \log_{10} (P_i/P_r) \quad (1)$$

Equation (1) shows the return loss of the antenna where P_i is the incident power and P_r is the reflected power.

From fig.3 it can be observed that, the designed antenna has a directivity of 1.460 dBi with a total radiation efficiency of -16.47 dBi.

B. Antenna Performance Inside Human Phantom Models

For designing an implantable antenna, it is important to observe the performance of the designed antenna and obtain the desired resonant frequency in an environment that is alike a human body. For that, the performance of the antenna is observed inside a human body model by using two phantom layered models with the FR4 as substrate.

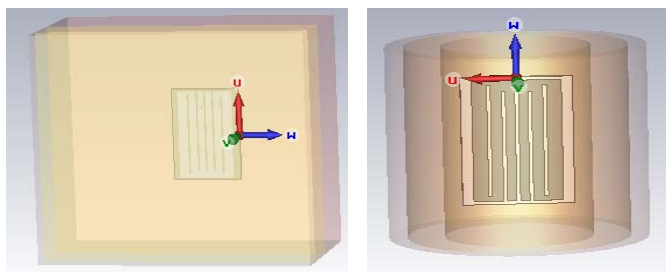


Fig.4. Rectangular and Cylindrical model for the designed antenna

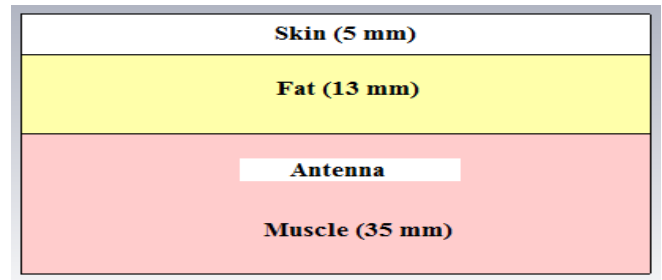


Fig.5. Layer dimension for simulation model

For simulation the dielectric properties of skin, fat and muscle have been used [9].

TABLE I DIELECTRIC PROPERTIES OF DIFFERENT LAYERS

Layer	Rho (Kg/m ³)	Thermal Conductivity (w/k/m)	Heat Capacity (Kj/k/Kg)	Thermal Diffusivity (m ² /s)
Muscle	1041	.53	3.546	1.4358e-007
Fat	910	.201	2.5	8.8516e-008
Skin	1100	.293	3.5	7.6104e-008

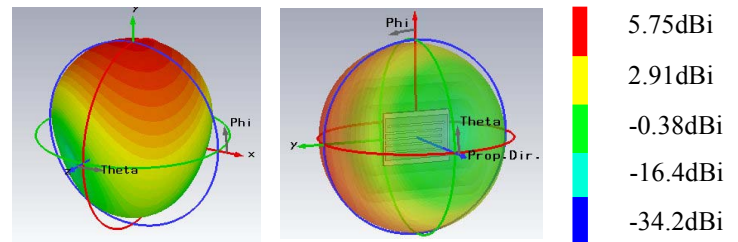


Fig.6. Far field region for two different phantom models

From fig.6 it can be seen that, the 3D view of far field region of the designed antenna for two different human phantom models. The designed antenna inside both block layered and cylindrical layered models is directive.

TABLE II ANTENNA PERFORMANCE FOR TWO DIFFERENT PHANTOM MODELS

Model Type	Resonant Frequency (GHz)	Return Loss (dB)	Directivity (dBi)
Block Layered	0.404	-12.1	6.291
Cylindrical Layered	0.404	-12.5	3.778

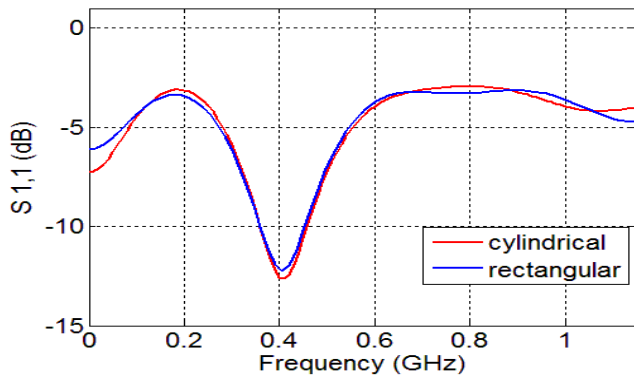


Fig.7. S parameter of the designed antenna in two human phantom models.

As it can be seen from table II that the directivity of the block layered model and cylindrical layered model is respectively 6.291 dBi and 3.778 dBi.

From fig.7 the observation is that, both the phantom models is operating at the resonant frequency of 0.404 GHz with a better return loss.

Three different substrate materials like FR4 (lossy), Bakelite (lossy) and Rogers RO3003 (lossy) are being used for designing to get better performance of the designed antenna [1].

TABLE III ANTENNA PERFORMANCE FOR DIFFERENT SUBSTRATE MATERIALS

Materials	Resonant Frequency (GHz)	Return Loss (dB)	Directivity (dBi)	Total Radiation Efficiency (dBi)
FR4	0.404	-12.1	6.291	-45.04
Bakelite	0.404	-12.1	6.066	-44.91
Rogers RO3003	0.404	-12.1	5.773	-44.13

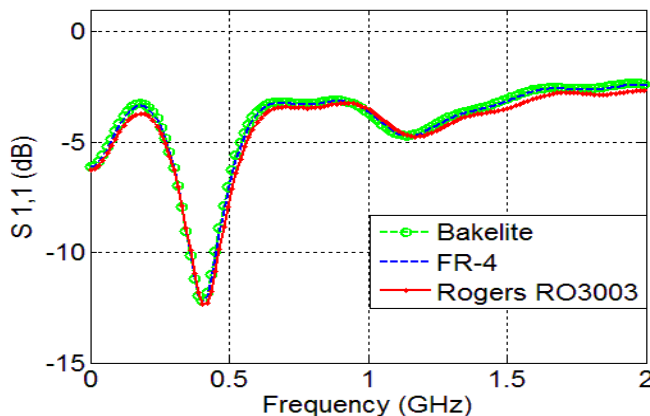


Fig.8. S parameter of the designed antenna for three different substrate materials

It can be seen from table III and fig.8 that, in comparison with the other substrate materials FR4 is better in return loss, directivity and total radiation efficiency.

TABLE IV ANTENNA PERFORMANCE FOR DIFFERENT PATCH AND GROUND MATERIALS

Materials	Resonant Frequency (GHz)	Return Loss (dB)	Loss	Directivity (dBi)
Silver	0.404	-12.1		6.291
Aluminum	0.404	-12.0		6.290
Copper	0.404	-12.1		6.291

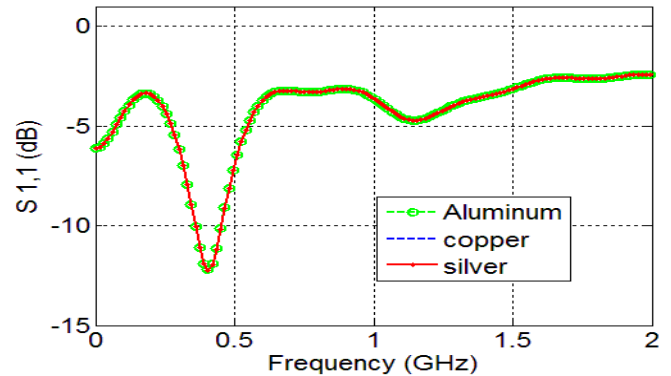


Fig.9. S parameter of the designed antenna for three different patch and ground materials

From table IV and fig.9 we can see that for different types of patch and ground materials, both the return loss and the directivity are quite close to each other.

IV. CONCLUSION

The paper proposes a design of a miniaturized implantable patch antenna for biomedical application. The antenna has a size of 40*30*1.525 mm, which is small enough for being implanted. Miniaturization and biocompatibility are the two most important factors as the proposed antenna is implantable. So for miniaturization several techniques along with a shorting pin between the patch and the ground was introduced. For making the antenna biocompatible the antenna was covered with an insulator layer. Considering all these factors and with the help of few formulas finally an antenna was designed which has a resonant frequency at 404 MHz that falls in the Medical Implant Communication System(MICS) band inside human phantom model with a bandwidth of 92.7 MHz. The method used in designing the antenna was transient analysis. The antenna was designed using FR4 substrate and using silver as patch and ground materials. The performance of the antenna was observed in two types of human phantom models. The radiation pattern inside human phantom models is almost same. So it can be concluded that the performance of the antenna will not change inside the arm of human body. The volume of the antenna was kept constant for all types of material. The

designed antenna is smaller in size and the performance of the antenna is better.

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Design of a Miniaturized UWB Ingestible Antenna for Wireless Capsule Endoscopy

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Abstract—In recent years, non-invasive Wireless Capsule Endoscope (WCE) has been explored extensively for the diagnosis of GI tract diseases. However the application of WCE is limited due to capsule size, propagating efficiency of the antenna, reducing effects on the human body and the data rate of the telemetry system. In this paper a miniature Ultra-wideband (UWB) antenna is designed for WCE by exploiting several miniaturization techniques such as using high-permittivity dielectric substrate and lengthening of the current flow path on the patch surface. The volume of the optimized antenna is only 81 mm^3 . The performances of the antenna are investigated inside a homogeneous human muscle layer. The proposed antenna resonates at 8.36 GHz with a wide -10 dB bandwidth of 500 MHz. Maximum power of 21.115 mW can be set as input to the proposed antenna in order to comply with the IEEE C95.1-1999 safety standards.

Index Terms—Antenna, UWB technology, Capsule Endoscopy

I. INTRODUCTION

With the tremendous advancement of communications and VLSI technology, Wireless Capsule Endoscope (WCE) shows a great promise for inspecting the human gastrointestinal (GI) tract for diseases in a non-invasive way [1]. The WCE is an ingestible capsule that can able to capture images during its journey through the GI tract including the stomach, large bowel or colon and part of the small bowel after being swallowed. The capsule is composed of several key parts such as image sensors, lighting, control unit, wireless communication unit, power source, and mechanical actuator as shown in Fig. 1. For real-time data transfer from and to the capsule, antennas with wide-bandwidth capabilities are required. Also miniaturization of the antenna is vital to save precious space of WCE capsule. In addition, the antenna is to design with omnidirectional radiation pattern for reliable transmission of data regardless of the orientation and location of the capsule or receiver [2].

Currently, companies such as Olympus, IntroMedic, Given-Imaging develop and manufacture widely popular WCE capsules. However, one of the major drawback of those WCE is low data rate [1]. Recently many antennas are proposed for WCE in different frequency band. Different spiral and cylindrical helical antennas [3], [4], [5], [6] have been proposed in Medical Implant Communication Services (MICS) band 402-405 MHz. Also, conformal meandered antennas have been presented in different frequency bands [7], [8], [9], [10]. However, for both MICS band (402-405 MHz) and WMTS

band (1395-1400 MHz), WCE can able to transfer signals with poor data rates.

The FCC rule for Ultra-wideband (UWB) transmitter is that, at any point in time, the radiator has a fractional bandwidth equal to or greater than 0.2 or has a bandwidth equal to or greater than 500 MHz irrespective of the fractional bandwidth [11]. UWB antennas has emerged for wireless implanted devices due to the introduction of WBAN (Wireless Body Area Network) standard IEEE802.15. A UWB antenna in the frequency band (3.1-4.5 GHz) is proposed for the medical capsule endoscope application [12]. However, due to the complex geometrical shapes of the antenna and the feeding arrangements, it might be hard to develop WCE commercially with that antenna.

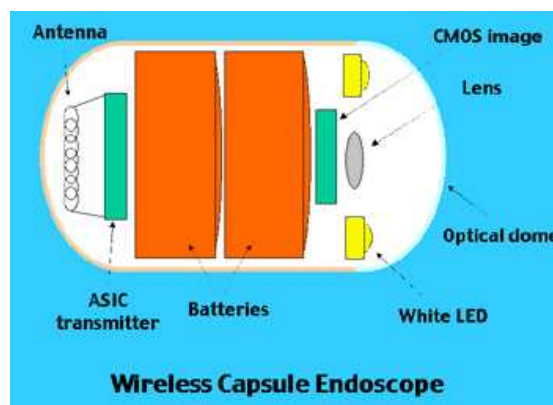


Fig. 1. The capsule incorporates the imaging system at one end, the UHF telemetry system at the other, and batteries in between [13].

In this paper, a small size planar slotted patched antenna is designed and optimized to operate inside a human body at UWB frequency band. Several miniaturization techniques such as using high-permittivity dielectric substrate and lengthening of the current flow path on the patch surface are exploited for optimized the antenna. At the same time, omnidirectional radiation pattern is required for receiving unit irrespective of the capsule location inside the body. The finite integration technique (FIT) in Computer Simulation Technology's (CST's) Microwave Studio is used in the simulation.

The paper is organized as follows. Section II describes the human body model used in this research work. Antenna geometry is given in section III. In section IV, the performances of the antenna are evaluated and those are compared with the other reported antennas for WCE applications. Concluding remarks are given in section V.

II. HUMAN BODY MODEL

A single layer tissue model has been chosen because it is simple to model, and exhibits much faster computational times compared to more complex tissue models (e.g., anatomical tissue models). The study [14] shows that this model provides a fairly accurate result for any proof of concept. More complete body model will be studied in the future work.

A cubic homogeneous model of the human muscle is incorporated for the performance analysis of the antenna. This muscle tissue has relative permittivity, $\epsilon_r = 45.007$, conductivity, $\sigma = 8.2903$ S/m, and loss tangent, $\delta = 0.3962$ at 8.36 GHz [15]. The dimensions of the muscle are $40\text{mm} \times 40\text{mm} \times 40\text{mm}$ as shown in Figure 2. The antenna is placed in the center of the muscle box. The muscle equivalent material, namely, *muscle* in CST Microwave studio is customized with above mention parameters to build the muscle box.

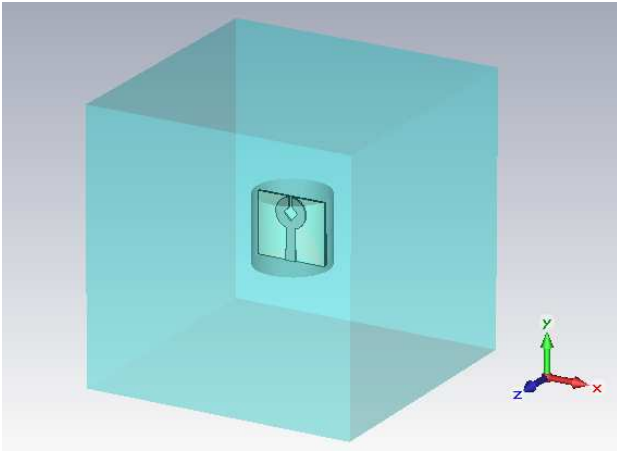


Fig. 2. A single-layer homogeneous muscle box with the antenna inserted in the middle of the box

III. ANTENNA DESIGN

The main motivation in designing the antenna is to reduce the antenna size so that the capsule can easily be swallowed and move in human body. The proposed UWB antenna as shown in Fig. 3 operates at UWB frequency range. The antenna is designed on Roger R03010 substrate which is generally used in implantable patch antennas for biomedical telemetry. The specifications of the substrate are as follows, relative permittivity, $\epsilon_r = 10.2$, and loss tangent, $\tan\delta = 0.0023$. This high-permittivity dielectric assists in antenna miniaturization because of the shortening of the effective wavelength.

The antenna incorporates a ground plane under the substrate which can be used as the reference ground of all electronic devices that are integrated inside the capsule. The antenna is covered with teflon ($\tan\delta = 0.001$, and relative permittivity $\epsilon_r = 2.25$). Teflon is commonly used bio-compatible materials for the ingestible capsule to preserve patient safety. The antenna is fed by the wave-guide port with 50Ω matching.

In Table I, sizes of different antenna parts of the proposed antenna are shown. The circular patch of the antenna is fed by a 50 ohm microstrip feed line with a width of 1.6 mm. The antenna is fabricated on a $10 \times 9\text{mm}^2$ substrate with thickness of 0.25 mm. The designed antenna has a circular disc with a radius of 2.25 mm which is fed by a 50 Ohm microstrip feed

TABLE I
GEOMETRICAL DIMENSIONS OF THE OPTIMIZED ANTENNA

Parameters	Value (mm)
Length of substrate	9.0
Width of substrate and ground	10.0
Thickness of substrate	0.25
Length of ground	4.5
Thickness of ground	0.4
Length of ground notch	2.5
Width of ground notch	0.5
Radius of circular patch (CP)	2.25
Thickness of the patch	0.25
Length and width of square slot in CP	1.5
Width of rectangular slot in CP	0.5
Width of feedline	1.6

line whose width is 1.6 mm. The radius of the patch, a , is calculated using the equation 1 [16].

$$a = \frac{1.8412v}{2\pi f_r \sqrt{\epsilon_r}} \quad (1)$$

where v is the speed of light in the medium, f_r is the resonance frequency, and ϵ_r is the relative permittivity of the substrate.

The main circular patch and the feed is connected through a matching section with a width of 1.3 mm. The antenna is designed by cutting a small $1.5 \times 1.5\text{mm}^2$ slot along with a rectangular portion of $2.25 \times 0.5\text{mm}^2$ vertically on the circular patch. Due to the lengthening the current flow path on the patch surface, a more compact antenna size is obtained [17]. A rectangular notch on the top of the ground is introduced to improve the impedance matching of the antenna [18]. Therefore, the volume of the optimized antenna is only 81mm^3 ($10\text{mm} \times 9\text{mm} \times 0.9\text{mm}$). All the parameters except the radius of the patch of the optimized antenna are obtained using CST Microwave Studio.

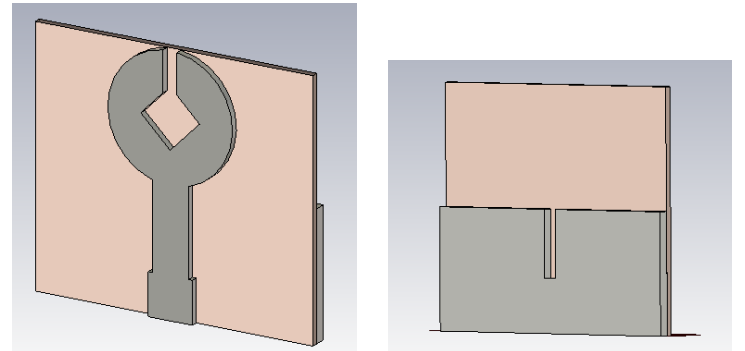


Fig. 3. The proposed antenna: front view (left) back view(right)

IV. PERFORMANCE OF THE PROPOSED UWB ANTENNA

A. S-parameter analysis of the antenna

The reflection co-efficient ($|S_{11}|$) frequency response of the proposed antenna inside the single-layer model shown in

Fig. 4. It can be shown from the figure that the antenna resonates at 8.3572 GHz with a reflection coefficient magnitude $|S_{11}|$ of -22.97 dB and a wide -10 dB-bandwidth of 501.8 MHz from 8.1423 GHz to 8.6438 GHz where VSWR is less than 2 shown in Fig. 5. Therefore the antenna can be used in a UWB system.

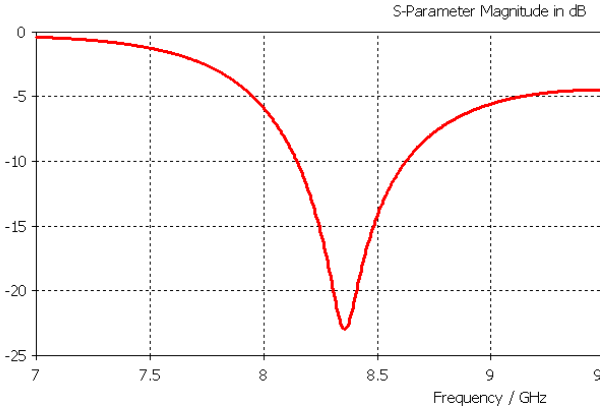


Fig. 4. Reflection co-efficient, S_{11} , frequency response of the proposed optimized antenna inside the human muscle

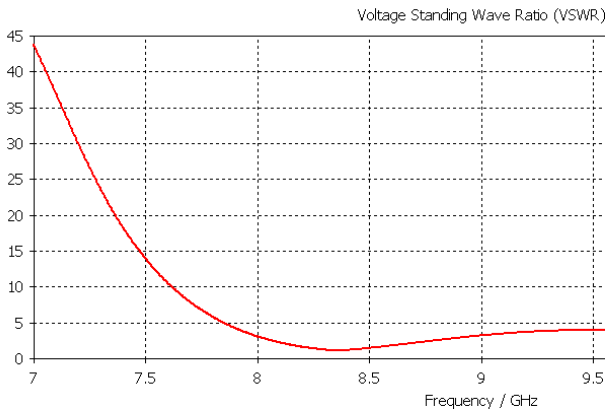


Fig. 5. VSWR inside the human muscle

B. Radiation pattern analysis

The far field radiation patterns are shown in Fig. 6. From the 3D radiation pattern in Fig. 6 (right), the maximum far field gain value is found equal to -26.6 dB. From the 2D radiation pattern in Fig. 6 (left), main lobe magnitude is 0.6 dBi, main lobe direction is 36 degree, angular width (3dB) is 21.7 degree and side lobe level is -1.2 dB. Both the figures indicate that the proposed antenna pattern is omnidirectional. This is one important requirement for ingestible antennas as the direction of the transmitted signal should be independent of the transmitter's position. The radiation efficiency and the total efficiency of the antenna in free space are 98.24% and 95.27% respectively.

C. Currents, Electrical and Magnetic fields distribution

To illustrate the operation of the proposed antenna, surface current distribution is presented in Fig. 7. The current is oriented at the feeding point, which is placed at the end of circular patch. It shows the effective current path which is responsible for the antenna operation is the one between the feeding point and the circular patch. The magnetic field

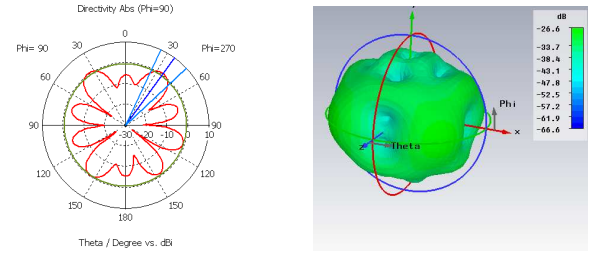


Fig. 6. Far-field gain radiation pattern in 2D (left) and 3D (right) inside the human muscle at 8.3572

distribution and the electric field distribution are shown in Fig. 8.

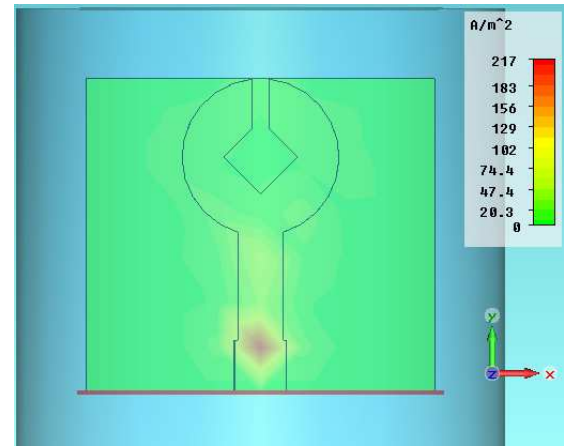


Fig. 7. Simulated surface current at the resonance frequency of 8.3572 GHz)

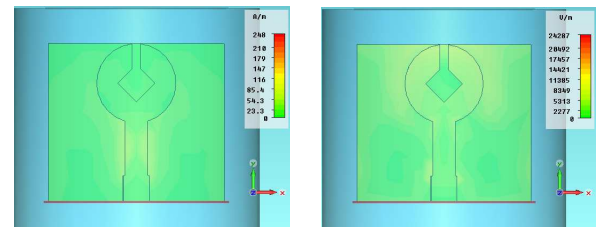


Fig. 8. Simulated magnetic field distribution (left) and electric field distribution (right) at the resonance frequency of 8.3572 GHz

D. SAR analysis

Specific Absorption Ratio (SAR) is a measure for electromagnetic energy absorbed by biological tissue mass when exposed to radiating device. IEEE C95.1-1999 standard limits the SAR averaged over any 1g of tissue in the shape of a cube to 1.6 W/Kg. For the proposed antenna, SAR is calculated by CST MW studio and is equal to 75.7736 W/Kg averaged over 1g of tissue for input power of 1W. To comply with IEEE safety standards, the maximum allowed net input power for the proposed antenna is limited to 21.115 mW.

E. Comparison of selected antennas

The performance of the proposed antenna is compared with other different research works mentioned in Section I. The performance parameters include shape, bandwidth, operating frequency band, and volume of the antennas. The proposed antenna shows higher bandwidth compare to all the antennas except one mentioned in [12]. However, comparing with

[12], the proposed antenna is simple in design. Moreover, our proposed antenna is the smallest one.

TABLE II
COMPARISON OF SELECTED ANTENNAS

Reference	Antenna shape	Bandwidth	Volume(mm ³)
[6]	spiral	75 MHz (460 MHz-535 MHz)	346
[8]	conformal	420 MHz (330 MHz-750 MHz)	436
[9]	conformal	260 MHz (370 MHz-630 MHz)	2057
[10]	conformal	785 MHz (215 MHz-1 GHz)	950
[12]	conformal	1.7 GHz (3.1 GHz-4.8 GHz)	134
this work	patch	500 MHz (8.14 GHz-8.64 GHz)	81

V. CONCLUSIONS

A small, slotted circular patched type antenna surrounded by a bio-compatible layer is designed and proposed for WCE. Several techniques are applied to minimize the occupied volume of the designed antenna (10mm×9mm×0.9mm). The antenna is designed to resonate at 8.3572 GHz with a reflection coefficient magnitude $|S_{11}|$ of -22.97 dB and a wide -10 dB-bandwidth of 501.8 MHz. To comply with the IEEE C95.1-1999 standards, the maximum allowable net input power is 21.115 mW.

Because of its UWB and small size, this antenna can be used with very small capsules for transferring data from inside body at high data rates. In addition to that, the omnidirectional radiation pattern of this antenna guarantees data reception regardless of the orientation of the capsule inside the digestive tract.

To manufacture a prototype antenna and to test the antenna will be the future work. Also the video transmission measurement system will be set-up to investigate the data transmission rate and the signal quality of WCE systems.

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Effect of Pump Characteristic (H-Q) Curves on the Efficiency of Total Artificial Heart

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Abstract—Nowadays, heart failure of human being is a very common occurrence all over the world. Total Artificial Heart (TAH) is an outstanding achievement of biomedical engineering where the patient's natural heart is removed and replaced with two mechanical pumps that provide perfusion to the systemic and pulmonary circulations. This paper focuses on the effect of pump characteristic (H-Q) curves on the efficiency of TAH. To show the effect of pump parameters, a simplified model of human cardiovascular system (CVS) is developed with an electrical circuit. A numerical simulation is conducted in order to find out the different physiological pressures and pump flows in response to change of parameters related to pump performance curves. Finally, the effect of H-Q curves on the efficiency of TAH is analyzed.

Keywords—Total Artificial Heart; pump characteristic; H-Q curves; human cardiovascular system; efficiency.

I. INTRODUCTION

Heart failure is a common feature of the present world. Day by day, patients with heart failure, a physical condition that often cannot be treated with surgical therapies or medicines, are increasing in the hospitals. In case of most of the patients those suffer with heart failure, the only option for treatment is heart transplantation via an organ donor or by artificial means. But, at present, the doctors look for artificial heart therapies to the patients due to lack of donor hearts for increasing demand. In case of artificial heart therapies, the significant feature is the use of mechanical circulatory devices and these mechanical circulatory devices are designated as either Total Artificial Hearts (TAH) or Ventricular Assist Devices (VADs) [1]. TAH (Total Artificial Heart) is an outstanding achievement of biomedical engineering where the patient's natural heart is removed and replaced with two mechanical pumps that provide perfusion to the systemic and pulmonary circulations. The pumps that can be used in TAH are typically classified into two categories, pulsatile pumps and continuous flow pumps [2]–[3]. Between these two types of pumps, continuous flow pumps increase the mechanical reliability, improve patient survival and thus reduce the maintenance, as well as the device size with respect to pulsatile pumps [4]. Beside this, use of continuous flow devices provide fixed support throughout the cardiac cycle that must be limited in

order to prevent suction and ventricular collapse. Consequently, continuous flow pumps do not unload the ventricle to the same extent as compared to previous generation positive displacement pumps [5]–[6]. There are different types of continuous flow pumps that can be differentiated by the relationship between pump head pressure and pump flow (H-Q) curve. In case of centrifugal pump, typically it shows a shallow H-Q curve which means large changes in flow for comparatively small changes in head pressure [7]. But, in case of axial flow pump, it has a steeper H-Q curve: small changes in flow for comparatively greater changes in head pressure and mixed flow pumps can have H-Q curves with characteristic that fall somewhere in between centrifugal and axial flow pumps.

As mechanical pump is one of the most important part of the TAH, so pump characteristic plays an important role to the cardiac output of the TAH and physical condition of human body. That's why the effect of pump characteristic (H-Q) curves on the efficiency of TAH is being investigated. In case of pump performance curves, the change of flow (Q) with respect to pressure head (H) depends on the slope (m) of the H-Q curves and zero-flow pressure of the pump (H_0).

II. METHODS AND MODEL DEVELOPMENT

A. Electrical Circuit for Human Cardiovascular System

To show the effect of pump parameters, a simplified model of human cardiovascular system (CVS) is developed with an electrical circuit which is shown in Fig. 1. The model consists of mechanical pumps (left and right pumps), the systemic and pulmonary circulation, cannulas that make the hydraulic connections between the pump and the cardiovascular system etc. Here, resistors represent the viscous property of blood flow and capacitors model the elastic property or compliance of the vessel wall. There are four capacitive elements included in the model such as C_{pi} , C_{po} , C_{si} , and C_{so} which represents pulmonary arterial compliance, pulmonary venous compliance, systemic arterial compliance and systemic venous compliance respectively. The numerical values for all compliances are listed in the Table I. To simplify the model of the cardiovascular system, the leakage flow and bronchiolar circulations were neglected. Beside this, the continuous flow left & right pump are modeled with two identical linear pump

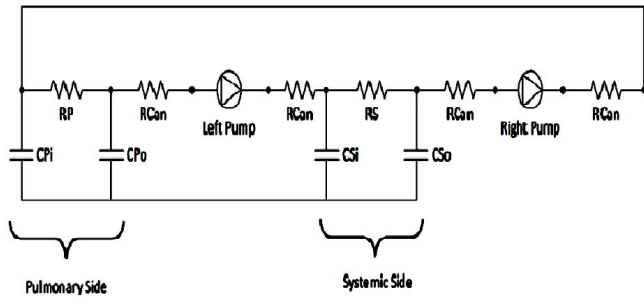


Fig. 1. Electrical circuit model of cardiovascular system (CVS).

performance curves:

$$H_l = m_l Q_L + H_{ol} \quad (1)$$

$$H_r = m_r Q_R + H_{or} \quad (2)$$

where H_l & H_r denote the left & right pump head pressures; Q_L & Q_R denote the left & right pump flows; m_l & m_r represent the slopes of the left & right pump characteristic curves and H_{ol} & H_{or} denote the zero-flow pressures of left & right pump. Also, pump head pressure is the difference between aortic pressure (P_{ao}) and left ventricular pressure (PLV). From above equations, the amount of flow $Q(t)$ and by integrating $Q(t)$ over the cardiac cycle, mean flow can be obtained.

The model parameters are listed in the following table:

TABLE I. PARAMETERS OF THE MODEL OF CVS

Parameters	Symbols	Value
Systemic Vascular Resistance (SVR)	R_s	0.85 mmHg/ml/s [8]
Pulmonary Vascular Resistance (PVR)	R_p	0.09 mmHg/ml/s [8]
Pulmonary Arterial Compliance	C_{pi}	3 ml/mmHg
Pulmonary Venous Compliance	C_{po}	15 ml/mmHg
Systemic Arterial Compliance	C_{si}	1.5 ml/mmHg
Systemic Venous Compliance	C_{so}	115 ml/mmHg [9]

B. Simulation Protocol

The model was implemented using the Simulink toolbox in MATLAB (The Mathworks, Inc., Natick, MA, USA). In order to investigate the effect of pump parameters on the efficiency of Total Artificial Heart (TAH), the Simulink model was simulated using different parameters such as slope of left pump characteristic curve (m_l), slope of right pump characteristic curve (m_r), zero-flow pressures of left & right pump (H_{ol} & H_{or}) etc. Every parameter was changed many times and respective physiological pressures such as mean pulmonary arterial pressure (P_{pi}), mean left atrial pressure (P_{po}), mean aortic pressure or mean arterial pressure (P_{si}), mean right atrial pressure (P_{so}) etc. and left & right pump flows (Q_L & Q_R) were recorded. The slope values were chosen to cover the range of H-Q slopes in pumps that are used clinically ($m = -5$ to -10 mmHg/lpm). The Heart Mate II axial flow pump (Thoratec Co., Pleasanton, CA) has an average H-Q slope of $m = -10 \pm 1$ mmHg/lpm [10]. The Heart Ware HVAD (Heart Ware international Inc., Framingham, MA) and Terumo Dura Heart (Terumo Heart Inc., Ann Arbor, MI) centrifugal flow pumps have average H-Q slopes of $m = -$

7 ± 1 mmHg/lpm and $m = -5 \pm 1$ mmHg/lpm, respectively [11]–[12].

III. SIMULATION AND OPTIMIZATION OF PARAMETERS

A. Simulation Results

A numerical simulation was conducted in order to find out the different physiological pressures and pump flows in response to change of parameters related to pump performance curves. Fig. 2 illustrates the changing patterns of different physiological pressures such as P_{pi} , P_{po} , P_{si} , P_{so} etc. and left & right pump flows (Q_L & Q_R) with the change of slope of left pump characteristic curve (m_l).

If the value of m_l is increased along the negative axis, then it is observed from simulation result that left & right pump flows (Q_L & Q_R) decrease gradually. But left pump flow & right pump flow remain same for all values of m_l . Beside this, with the increase of m_l , P_{pi} & P_{po} increase consecutively and P_{si} & P_{so} decrease linearly.

Fig. 3 demonstrates the effects of the variation of slope of right pump characteristic curve (m_r) on the different physiological pressures such as P_{pi} , P_{po} , P_{si} , P_{so} etc. and left & right pump flows (Q_L & Q_R). With the increase of m_r along the negative axis, mean pulmonary arterial pressure (P_{pi}) & mean left atrial pressure (P_{po}) decrease gradually and become more and more negative which may cause suction problem. On the other hand, mean right atrial pressure (P_{so}) increases with the increase of m_r along the negative x-axis. Beside this, with the increase of slope of right pump characteristic curve (m_r), mean aortic pressure (P_{si}) and left & right pump flows (Q_L & Q_R) decrease linearly.

Fig. 4 depicts the changing manner of various physiological pressures such as P_{pi} , P_{po} , P_{si} , P_{so} etc. and left & right pump flows (Q_L & Q_R) with the change of zero-flow pressure of left pump (H_{ol}). It shows that if the value of H_{ol} is decreased from 115 mmHg to 66mmHg, then the value of P_{pi} increases from 10.98 mmHg to 14.72 mmHg consecutively and the value of P_{po} increases from 1.2 mmHg to 8.36 mmHg

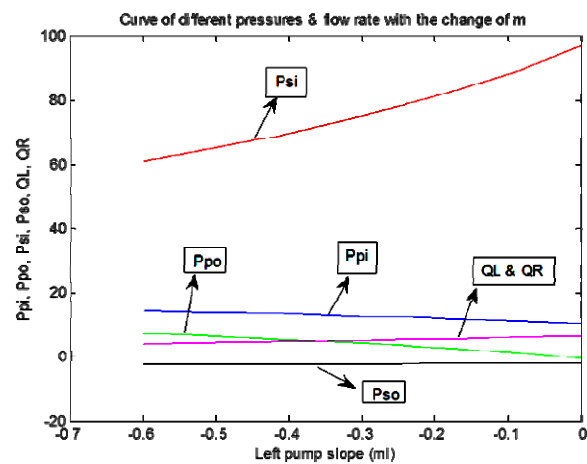


Fig. 2. m_l vs. P_{pi} , P_{po} , P_{si} , P_{so} , Q_L , Q_R .

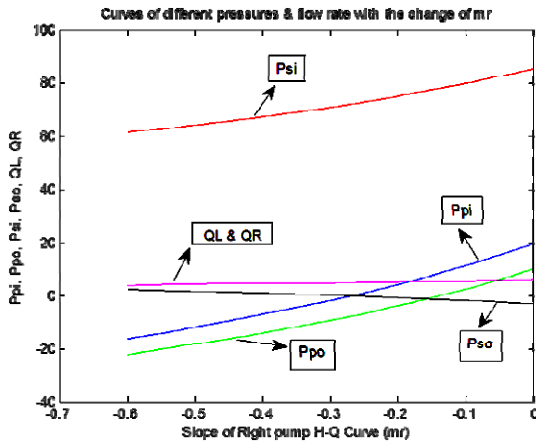


Fig. 3. mr vs. Ppi, Ppo, Psi, Pso, QL, QR.

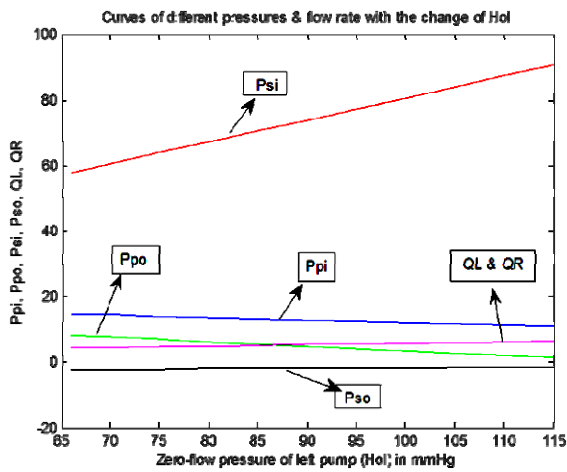


Fig. 4. Hol vs. Ppi, Ppo, Psi, Pso, QL, QR.

linearly. But the value of Pso is going to be more and more negative. On the other hand, mean aortic pressure (Psi) and left & right pump flows (QL & QR) decrease from 90.8 mmHg to 57.83 mmHg and 6.52 L/min to 4.24 L/min respectively with the decrease of Hol. From Fig. 5, it can be seen that if zero-flow pressure of right pump (Hor) decreases, then the value of Ppi, Ppo, Psi and left & right pump flows decrease linearly; but Pso increases linearly.

B. Effect Analysis

These numerical studies have successfully demonstrated the effects of pump characteristic curves on the efficiency of TAH. The above results explain how physiological pressures and pump flows vary with the variation of slopes of pump characteristic curves and zero-flow pressures of pumps. If analyzed, it is obvious from the result that sometimes physiological pressures are going beyond the range of human body and sometimes pump flow is decreasing to a low value which will be quite harmful to the human body. If Pso and Ppo

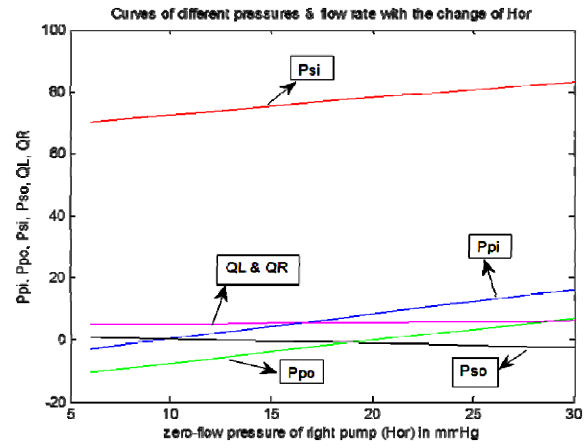


Fig. 5. Hor vs. Ppi, Ppo, Psi, Pso, QL, QR.

increase beyond their normal range, then edema may occur. Also, Psi is responsible for hypertension. So, it would be necessary to choose some specific values of slopes of H-Q curves and zero-flow pressures of pumps which will give suitable pressures and suitable flow for human body.

C. Proposed Optimization

From the previous study, it is known that the normal range of mean aortic pressure (Pao or Psi), mean pulmonary arterial pressure (Ppa or Ppi) and mean left atrial pressure (Pla or Ppo) are 70 – 110 mmHg [13], 9 – 18 mmHg and 4 – 12 mmHg [14] respectively. Based on this range, the result obtained from simulation can be compared. From the simulation results, it is clear that for some values of slope, the physiological pressures are going to be negative which may create suction problem in human body. Also, it is found that, for some values of zero-flow pressures of pumps and slopes of pump characteristic curves, pump flows decrease beyond the normal cardiac output. Considering above factors the result has been optimized. The optimized values for slope of left pump characteristic curve (ml), slope of right pump characteristic curve (mr), zero-flow pressure of left pump (Hol) and zero-flow pressure of right pump (Hor) are -3.5 mmHg/lpm, -1.5 mmHg/lpm, 100 mmHg and 25 mmHg respectively. Using the optimized values in the Simulink model, the values obtained for Ppi, Ppo, Psi, Pso, QL and QR are 12.13 mmHg, 3.4 mmHg, 80.7 mmHg, -1.811 mmHg, 5.8 L/min and 5.8 L/min respectively. These physiological pressures and pump flows reflect the normal cardiovascular pressures and cardiac output. Fig. 6 shows the H-Q curves for both left and right pump.

IV. CONCLUSIONS

The effect of pump characteristic (H-Q) curves on the efficiency of Total Artificial Heart (TAH) has been analyzed. In order to do this a simplified model of human cardiovascular system (CVS) was developed using electrical circuit. Then numerical simulation was conducted in order to observe the changes in different physiological pressures and pump flows with respect to the changes in various parameters of pump characteristic curves. These result showed the effect of pump characteristic curves on the efficiency of TAH. Based on these

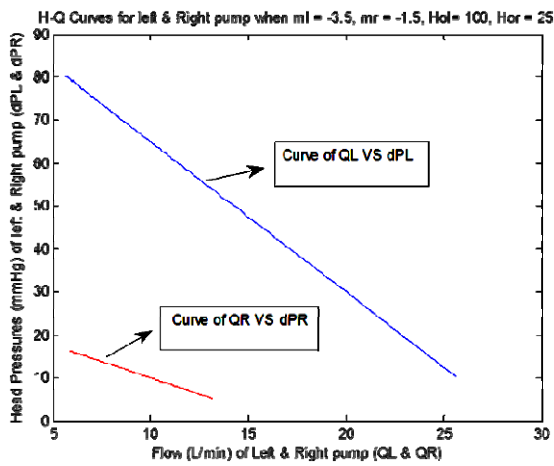


Fig. 6. Optimized H-Q curves for left and right pumps.

result, optimized values for parameters of pump characteristic curve have been proposed. A numerical simulation taking those optimized values into account gave the normal cardiovascular pressures and cardiac output, as expected.

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DESIGN AND PERFORMANCE MEASUREMENT OF AN IN-BODY IMPLANTABLE MINIATURIZED SLOT DIPOLE RECTANGULAR PATCH ANTENNA FOR BIOMEDICAL APPLICATIONS

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Abstract— An in-body implantable miniaturized slot dipole rectangular patch antenna operating in the Industrial, Scientific, and Medical (ISM) band (2.4–2.4835 GHz) for biomedical applications is presented in this paper. Primary reason behind choosing ISM band is its flexibility towards larger bandwidth with smaller size of antenna. The main specialty of this antenna is its tiny dimension & bendability which makes it practically implantable for in-body biomedical applications. Actual size of the antenna is (8.20x12.5x0.045) mm. For making it suitable for implementation, the antenna was immersed inside a human tissue model where proper electrical properties of the human tissue muscle model were maintained. All the parameter calculations and simulation results were suitable to support the Industrial, Scientific, and Medical (ISM) band in both bent and planar conditions. This antenna can be used for real time wireless monitoring of blood pressure, temperature, glucose level, and other in-body applications that functions in the ISM band. To make the designed antenna suitable for implantation, it is embedded in a silicon block which acts like a superstrate. Finally, the safety measurements were taken into account by stimulating the Specific Absorption Rate (SAR) distribution of the antenna to make it perfectly biocompatible.

Index Terms—ISM Band, Bent Antenna, Implantable Antenna, Human Tissue Model, Specification Absorption Rate (SAR).

INTRODUCTION

I. Implantable medical devices are very much popular in now a days for their flexibility towards a wireless & hassle free communication between patients body & monitoring unit. A total system circuitry works to produce responsive data from different parts of the body & an antenna on that system transmits the data to an external receiver [1]. Biomedical sensors have also put a well remark on this field of wireless biomedical communication field. But sensors has always a limitation with the availability of interstitial fluid that surrounds it & helps it operate successfully [2]. Wireless communication has earned that level of reliability where now a days, different kind of artificial body parts like cochlear implants, brain and cardiac pacemakers [3-5], nerve signal recorders [6], artificial eyes etc. are implanted & monitored successfully. As this sensitive devices need very small but efficient implantable antennas for transmitting & receiving data, this is where the main motive of this research work comes in.

II. When it comes to designing a biocompatible antenna, there are certain bands which are allocated for this purpose. One of the most common type amongst biomedical antennas are microstrip or planar-inverted F antennas (PIFA), operating in the Medical Implant Communications Service (MICS) band which ranges from 402–405 MHz But the main problem with this kind antennas is the bigger wavelength due to its lower frequency range [7-8]. There are also some PIFA antennas in MICS band, which is low in size but it has limitation in lower

fractional bandwidth & lower reflection coefficient [9]. The antenna was designed to operate at Industrial, Scientific, and Medical (ISM) band because of its wide range of bandwidth which ensures higher bit rate as well as smaller wavelength which perfectly matches the biocompatible antenna criteria's [10-11]. Previous research work in this sector has some limitations like use of biocompatible material to embed the antenna, simulation of antenna performance inside human tissue model, bendability etc. This paper covers all these criteria's for performance measurement as well as minimizing the size of the antenna by keeping all the performance satisfactory & maintaining FCC guidelines [12].

III. In this paper, a modified model of antenna [13] is proposed & developed which will be operating in 2.45 GHz ISM Band recommended by the European Radio-communications Committee (ERC) for ultra-low-power active medical implants [14]. Proposed model of antenna is rectangular patch shaped Co-Planar Waveguide (CPW) fed Slot Dipole antenna. Substrate of the antenna is of Polyimide whereas material of the patch is copper. For making the antenna biocompatible, we are embedding the antenna into Silicon layers. To make the antenna more versatile antenna performance will be observed inside human tissue model where it will prove it's in body application capability. For keeping the safety issues for biocompatible antennas in high priority, the Specific Absorption Rate (SAR) was measured with an input power of 2 mW following IEEE C95.3 standard [15] & compared the results with the ICNIRP [16].

MODELING OF THE ANTENNA

A flexible slot dipole antenna was designed at first in free space using the CST Microwave Studio. Reference was taken from an existing slot dipole antenna for the initial design from [13]. Designed model of the antenna with parameters is shown below Fig. 1-

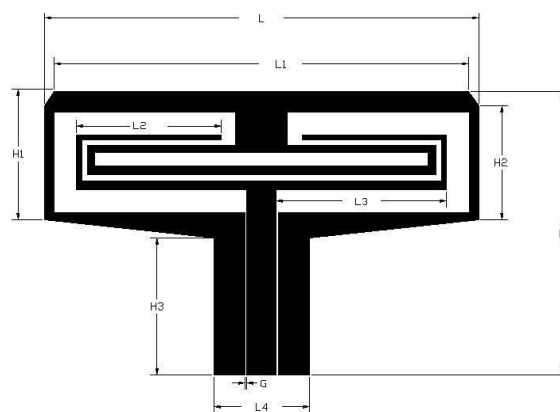


Fig. 1. Model of the antenna

All the parameter values are shown below in the table-

TABLE I. Antenna Size Parameters

Parameter	Size (mm)
L1	11.9325
L2	4.1500
L3	4.900
L4	2.7500
H1	3.7651
H2	3.2972
H3	3.9525
G	0.05
L	12.500

After characterization of the antenna inside the human tissue model, the result was checked and by putting importance on the S11 parameter, the antenna's shape was redesigned and re-optimized so that it covers the ISM band at a resonant frequency of 2.45 GHz. Upon obtaining the desired frequency, current density across the surface of the antenna was observed. Slots were cut where the current density was minimum. This method of iteration led to a minimum return loss of -38.42 dB as seen from the S11 parameter plot.

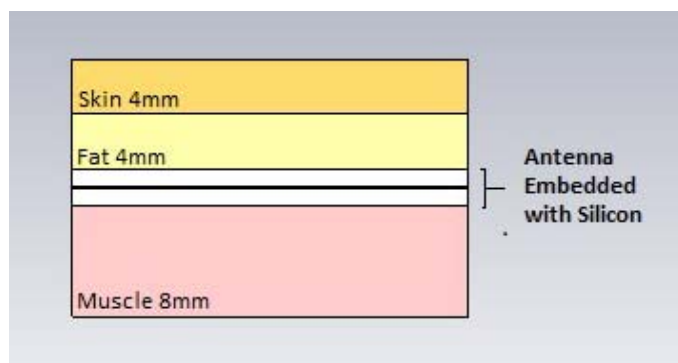


Fig. 2. Three Layer Geometry Model

Frequency (GHz)	ϵ_r	σ S/m
2.0	53.29	1.45
2.4	52.79	1.70
2.45	52.72	1.73
2.5	52.66	1.77
3.0	52.05	2.14

TABLE II. Dielectric Values of Human Muscle Tissue at Different Frequencies

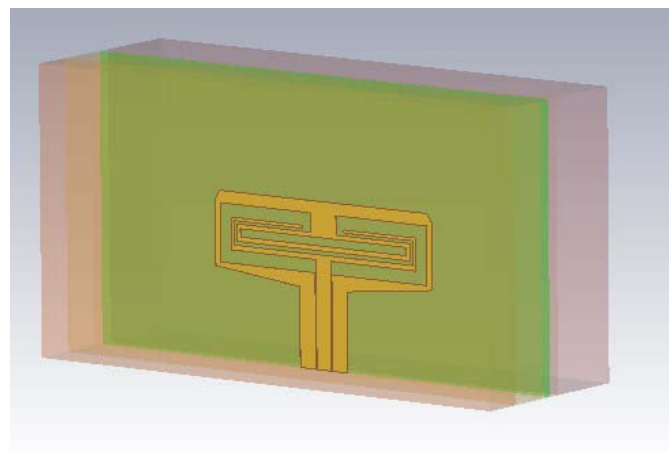


Fig. 3. Antenna inside Human Phantom Model

For simulations inside the human body model, three layer geometry blocked phantom model was used as shown in Fig. 2 and Fig. 3. The antenna was immersed in a silicon block and then was embedded between the fat and muscle layer. The characteristic of the antenna is affected by the dielectric constant, conductivity of the skin, fat and muscle. The depth of the muscle inside which the antenna was located was taken into account to get the desired resonance frequency.

ANALYSIS OF ANTENNA CHARACTERISTICS

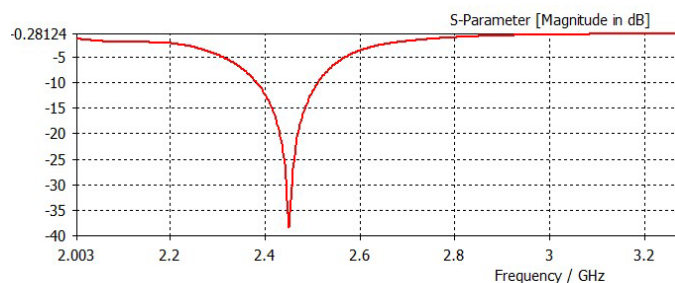


Fig. 4. S11 Parameter of the antenna inside Human Phantom model

Fig. 4 shows that the resonant frequency is at exactly 2.45 GHz which lies in the desired ISM band. The S11 parameter also known as the reflection co-efficient shows how much power is radiated or reflected from the antenna. The return loss of the antenna as seen from the graph is -38.42 dB which means the loss of the reflected back signal is very less compared to the radiated signal. This means the antenna shows very little loss in the signal radiated. The bandwidth of the antenna can also be calculated by drawing a straight line at -10dB. From the graph the bandwidth is found to be 127.9 MHz (2.511 GHz-2.3831 GHz).

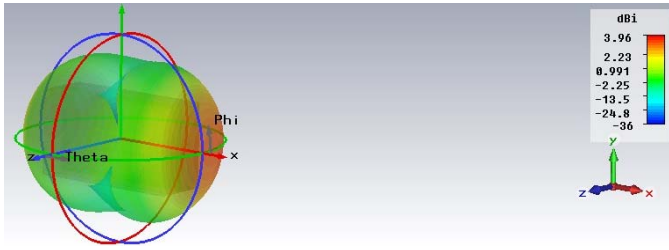


Fig. 5. 3D view of Far-Field Region of the antenna inside Human Phantom Model

It can be seen from Fig. 5 that the designed antenna inside the Normal Human Phantom model is directive with a directivity of 3.964 dBi and the total radiation efficiency of it is -22.45 dB.

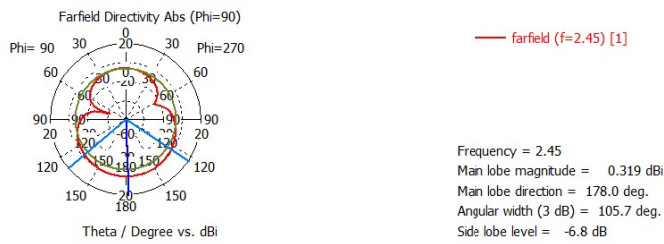


Fig. 6. Polar view of Far-Field Region of the antenna inside Human Phantom Model

From the Fig. 6 it can be observed that the main lobe magnitude is 0.314 dBi and it is centered at 179.0 degrees with 49.3 degrees angular width.

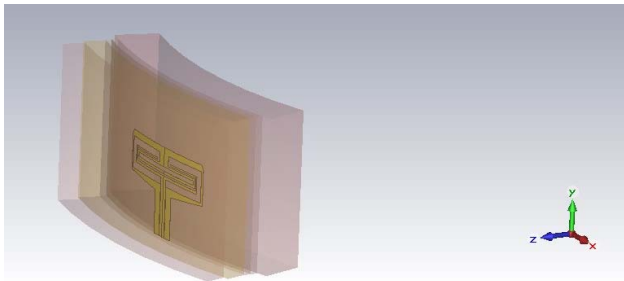


Fig. 7. Slot Dipole Rectangular Antenna Undergoing Bent Test

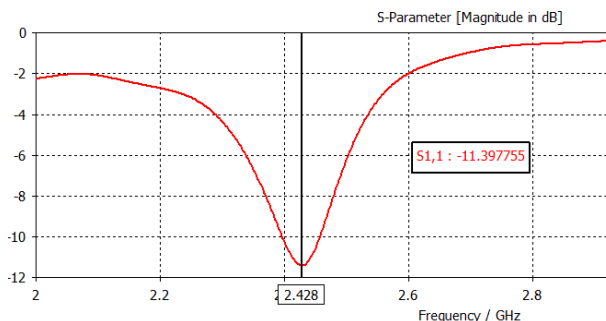


Fig. 8. S Parameter of Bent Antenna inside Human Phantom Model

As the antenna is designed for implantable purposes it must be flexible meaning the antenna must bend along with the surface of body where it will be implanted. The performance of the antenna hence also must be made conformal to the curved parts of the body. Thereto, the antenna was bent around the x-axis using a cylinder of 90mm in diameter as shown in Fig. 7. Basic S11 parameter of the antenna on the bent position of the antenna is shown on Fig. 8.

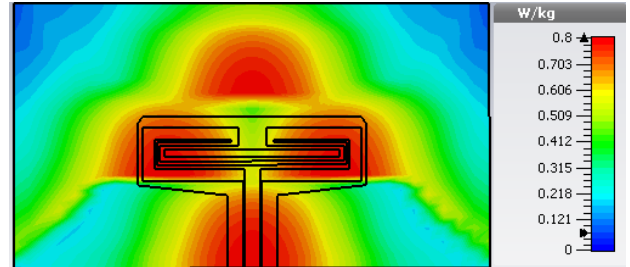


Fig. 9. SAR distribution for an input power of 2mW at x-y plane

TABLE III. Calculated SAR in Different Planes with maximum Input Power

SAR values at 2.45Ghz	y-z plane	x-z plane	x-y plane
C95.3-2002, 1-g averaged SAR(W/kg) for an input power of 2mW	0.8389	0.8166	0.8

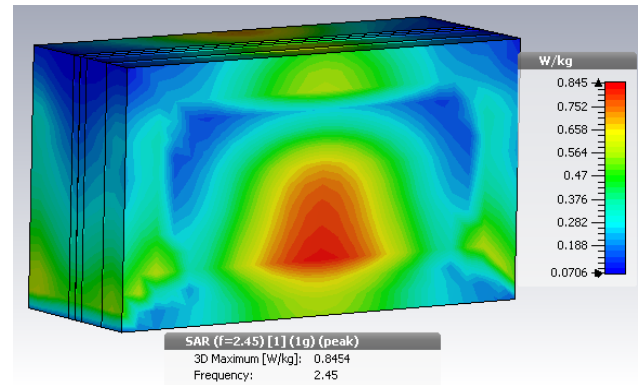


Fig. 9. 3D SAR distribution inside the human phantom model.

Since the antenna is designed for implantable purposes, the SAR needs to be evaluated mainly for safety concerns. The SAR is measured in the same three layer geometry model block around the antenna, as seen in Fig. 3. The standards for SAR calculation was regulated by IEEE C95.3-2002 standard, where the 1-g averaged SAR should not exceed 1.6W/kg as given by FCC and ICNIRP guidelines. The SAR was measured at the resonant frequency of 2.45 GHz when an input power of 2mW was provided. Fig.10 shows the SAR distribution when z=0 at x-y plane. The maximum SAR value

is seen to be at 0.8W/kg. The peak values of SAR in y-z plane and x-z plane was also evaluated inside the three layer geometry model and summarized as in TABLE III. We can also observe the peak SAR is distributed in the center region as seen by the bright red spot. The maximum 1-g averaged SAR measured inside the 3D model peaked to 0.8454W/kg for an input power of 2mW. This was well below the mark to fulfill the ICNIRP and FCC guidelines.

TABLE IV. Antenna performances for three types of materials in Human Phantom Model

Materials	Resonant Frequency (GHz)	Return Loss (dB)	Directivity (dBi)
Copper	2.45	-38.42	3.964
Aluminum	2.45	-35.9	3.964
Gold	2.45	-37.2	3.964

TABLE V. Antenna performances for three types of materials in Human Phantom Model

Materials	Resonant Frequency (GHz)	Return Loss (dB)	Directivity (dBi)	Total Radiation Efficiency (dB)
Polyimide	2.45	-38.4	3.964	-22.45
FR4	2.45	-38.8	3.964	-22.45
Rogers RO3003	2.45	-37.9	3.964	-22.45

The antenna was checked by using different patch materials and substrate to observe the performance in different conditions to get better efficiency and minimize the return loss. TABLE IV & TABLE V shows the data analysis with different material for patch & substrate of the antenna which proves the versatility of this antenna.

CONCLUSION

A flexible slot dipole antenna with a rectangular patch was designed in this paper for different biomedical applications. After designing the antenna it was immersed in a human muscle tissue model for testing its biocompatibility & observing its behavior in different measurement parameters. Antenna performance with different substrates & patch materials was observed for increasing the diversity of the proposed antenna. Far-field radiation pattern & SAR calculation showed the directivity & safety measurement of the antenna. Some primordial points regarding this antenna are-

- Design of a slot dipole CPW fed rectangular patch antenna compatible for implementing into human body.
- Measurement & analysis of antenna gain, reflection coefficient, E-field, Far-field & SAR calculation in both planar & bent state of the antenna.
- Performance measurement of the antenna by putting it into human muscle tissue model.
- Comparing antenna behavior by changing different substrates & patch materials to find out the best suitable material and substrate for making the antenna biocompatible.

By analyzing the results, a large bandwidth was observed both in planar & bent state. Resonant frequency of the antenna was 2.45 GHz which is exactly same with ISM band & the return loss was measured at -38.42 dB which was also very good to support the criteria's of the in body biocompatible antenna. Observed far-field radiation had a gain at -42.41 dBi with a $\theta = -73^\circ$ at $\phi = 90^\circ$.

The directivity of the radiation pattern of the antenna was also satisfactory. SAR of the antenna was measured at 1-g averaged SAR with an input power of 2 mW following IEEE C95.3 standard technique & the results were satisfactory according to FCC & ICNIRP. Future continuation of this work should cover the integration of this antenna with a complete biomedical system or device. Practical performance measurement of the antenna inside a human tissue mimicking gel should also be done. Finally, by analyzing all the results & for the miniaturized size of the antenna, conclusion can be made that the proposed antenna is practically implantable & ready to give a good performance in biomedical applications.

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Performance Analysis of SSVEP Based Wireless Brain Computer Interface for Wet and Dry Electrode

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Abstract— A Brain-computer Interface (BCI) is a communication pathway to provide ease to the users for interacting with the outside surroundings after translating brain signals into machine commands. The modern Steady-state Visual Evoked Potential (SSVEP) based Electroencephalographic (EEG) signals has become the most sophisticated methodology for a BCI paradigms. So, the perfection of SSVEP signal make the perfection of the BCI paradigm. The use of gel based wet electrode for the extraction of EEG signal is too much noisy and unpredictable for long time measurement which degrades the quality of SSVEP signal in a consequence degrades the performance of modern BCI paradigm. In our research, we are trying to solve this degradation of the quality of SSVEP signal. To accomplish this goal, a typical wireless BCIs using dry electrode is proposed for long term application without sacrificing Information Transfer Rate (ITR), Signal to Noise Ratio (SNR). After extracting SSVEP signal using dry electrode, Analog to Digital Conversion (ADC) is proceeded for the wireless transmission for remote BCI paradigms. Finally, after receiving this signal any BCI paradigms can be operated with high degree of accuracy.

Keywords—Brain-computer Interface (BCI); Steady-state Visual Evoked Potential (SSVEP); Electroencephalographic (EEG); Information Transfer Rate (ITR); Signal to Noise Ratio (SNR); BCI Paradigms.

I. INTRODUCTION

Biomedical engineering (BME) is the engineering in which the designs and principles of engineering are applied to the medicine and biology for the purpose of better healthcare (e.g. therapeutic or diagnostic). In other sense, the collaboration between the engineering science and medical science is known as the biomedical engineering. The purpose of this field is to make the gap closed between engineering and medicine. This is an emerged field compared to others. Most of the works of biomedical engineering are research and development. There are several applications of biomedical engineering such as development of biocompatible prostheses, various diagnostic and therapeutic medical devices including clinical equipment to micro-implants and common imaging equipment such as MRIs and EEGs. On the other side, Brain-computer Interface (BCI) is a direct pathway of communication between the brain and an external device i.e. computer [1], [2]. A typical pictorial representation of BCI and its compliments is shown in Fig. 1. It translates the signals from brain activities into machine codes or commands. The signals that are achieved from the brain activity are the electroencephalography (EEG) signals [3]-[5].

These electroencephalogram (EEG) signals reflect the electrical activity of the brain [6]. These can provide high temporal resolution in milliseconds that directly reflects the dynamics of the generating cell assemblies [6], [7]. The acquisition of these signals by BCIs can be divided into two different categories: invasive BCIs and noninvasive BCIs [2], [8].

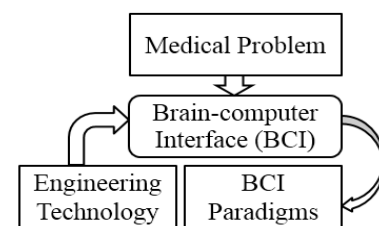


Fig. 1. Pictorial representation of Brain-computer Interface (BCI)

An invasive BCI is implanted directly into the grey matter of the brain to obtain the highest quality of brain activity signals or to send external signals into the brain [9], [10]. However, invasive BCIs are dependent on surgical techniques and are also potentially risky because of the interaction between the device and brain tissues when they are used in the long term. Brain-computer interfaces (BCIs) based on noninvasive scalp electroencephalography (EEG) have become an increasingly active research area [11], [12]. These are another major BCI research direction. These noninvasive devices are worn on the outside of the head and they are easily removable [2], [11]. There are some difficulties with the noninvasive methods. As the conductive gel desiccates or the glue loses its adhesion, there is an increase in the contact impedance between the electrode and scalp. This causes a large reduction in the Signal to Noise Ratio (SNR). Shorts can also occur between neighboring recording sites due to smearing of the conductive gel or due to sweat [5]. This procedure is also time consuming, uncomfortable, and even painful for participants since skin preparation usually involves abrasion of the outer skin layer [2], [4]. The main obstacles of wet electrode based EEG measurement has following limitations-

1. The recording of signal may take more time
2. Sometimes painful to the subjects
3. May be unsuitable for all kinds of participants
4. May be involved the abrasion of the outer skin layer

These problems of wet electrode based EEG measurement can be minimized by using dry electrodes [5], [10].

II. PROPOSED METHODOLOGY

The proposed methodology is described below in a single block diagram:

In this block the overall proposed methodology is shown in Fig. 2, which composed of a visual stimulator for visually stimulate the brain. Subject is visually stimulate in order to extract SSVEP based EEG for BCI paradigms. Then EEG data is transmitted and received without wire for some BCI paradigms. In Fig. 3, actual dry electrode based signal extraction and transmission is shown. The signal from the dry electrode is amplified using instrumentation amplifier after that this signal is filtered for the removing of the unwanted components from the SSVEP signal for better performance of BCI paradigm. Prior to wireless transmission there required an analog to digital conversion (ADC) which also shown in this block.

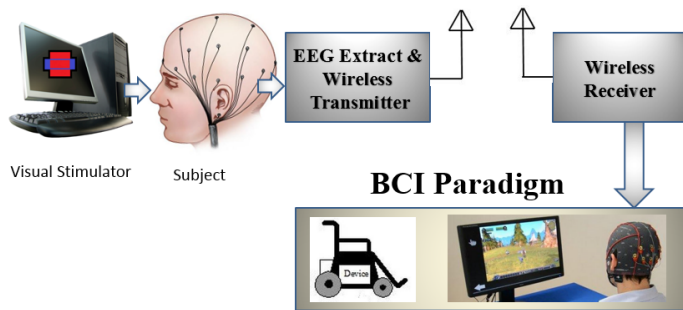


Fig. 2. Proposed block diagram for wireless BCI paradigms

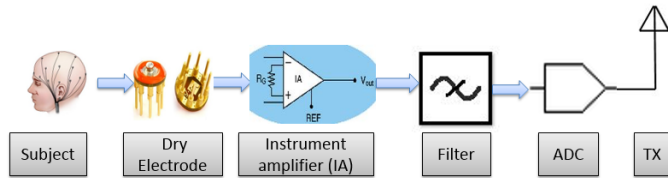


Fig. 3. Actual EEG extraction and wireless transmission.

III. DESIGN OF PROPOSED SYSTEM

The overall design of each block shown in Fig. 2 and Fig 3. are described in various separate sections as below:

A. Electrode Placement on scalp

EEG signal is the most sophisticated signal that have very low magnitude typically in the μV range. This negligibly small voltage can be measured very accurately using scalp activity of the brain. Gel based wet or dry electrode is used to extract the brain activity or to measure the EEG signal. The position of the electrodes (Wet or Dry) is shown in Fig. 4.



Fig. 4. A typical placement of dry electrode along scalp.

The skin of a human being is the largest organ of the body which contain about 20 square feet area. Skin has three layers which are mention below-

1. The outermost layer of skin is known as the epidermis which provides a waterproof barrier.
2. The next layer known as the dermis which beneath the epidermis, contains tough connective tissue, hair follicles, and sweat glands.
3. The deeper subcutaneous tissue which is made of fat and connective tissue.

The equivalent circuit of skin anatomy with a dry electrode is shown in Fig. 5.

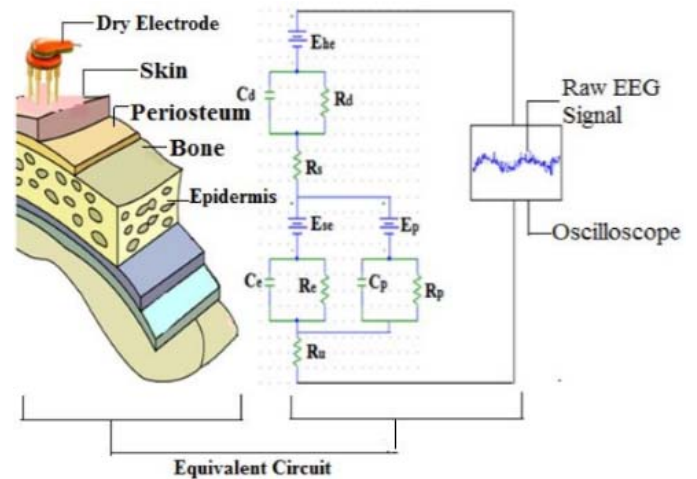


Fig. 5. Skin anatomy and equivalent circuit of dry electrode

B. Visual Stimulator

To stimulate the brain visually, a stimulator is designed using JAVA script. A typical screen shot of this stimulator is shown in the Fig. 6. The characteristics of extracted signal depends on the configuration of laptop used for the visual stimulation. The configuration used for our stimulation was Intel (R) core (TM) 2 Duo CPU T6600 @2.20 GHz, 4 GB RAM (Installed memory) and 32-bit OS. The refresh rate of the display was 60 Hz.

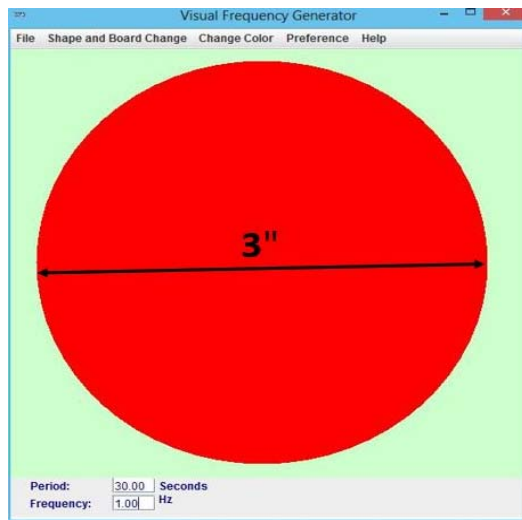


Fig. 6. A typical red colored circular stimulator with diameter 3".

C. EEG Acquisition

Extraction of EEG signal from the subjects is partially depends on the subject condition which means either subject is relax or tensed. Subject should be seated in front of visual stimulator as shown in Fig. 4. After visually stimulate the subject desired EEG is extracted using Fig. 7. The electrode position and corresponding brain anatomy is shown in Figure 8. The neural activity is of brain is measured using MP-36 whose representation along with brain is shown in Fig. 8.

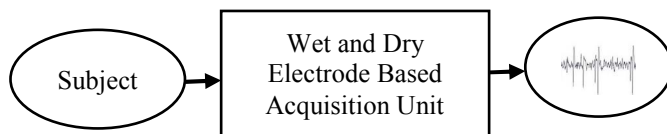


Fig. 7. EEG measurement using dry and wet electrode

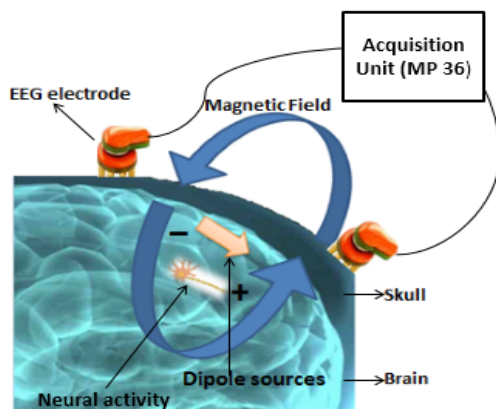


Fig. 8. Electrode placement and EEG measurement using MP-36

D. Wireless Transmitter and Receiver

Infrared light is electromagnetic radiation with longer wavelengths than those of visible light, extending from the nominal red edge of the visible spectrum at 700 nanometers

(nm) to 1 mm. This range of wavelengths corresponds to a frequency range of approximately 430 THz down to 300 GHz. Most of the thermal radiation emitted by objects near room temperature is infrared. The proposed IR transmitter and receiver circuit is shown in Fig. 9.

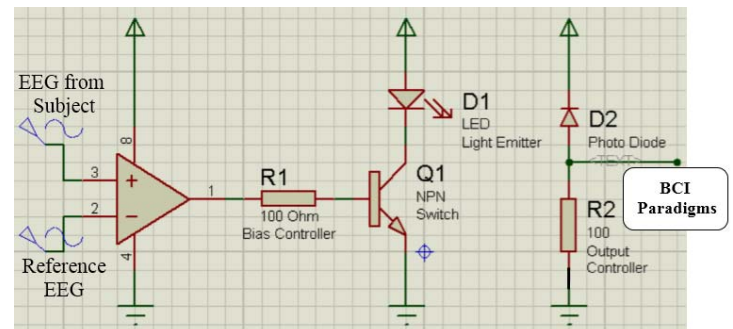


Fig. 9. Proposed model for wireless transmission of EEG data

IV. SIMULATION RESULTS

In our research, two important comparative performances are analyzed for dry and wet electrode during EEG signal recording. One is the randomness and other is the noise level. Both of these performances are analyzed with respect to the EEG data recording time.

In statistics, randomness is the measure of lack of predictability or desired pattern in any event occurring in a definite time. By definition, with the increase of randomness performance is decreased. The randomness analysis for wet and dry electrode is shown in Fig. 10. The red and black colored dot lines indicate randomness for wet and dry electrode respectively. From this analysis it can be seen that with the starting of recording EEG signal randomness is much lower for wet electrode than dry electrode. But with the increase of recording time the rate of randomness for dry electrode is reduced comparatively with respect to wet electrode. In Fig. 10 it is shown that at 15 seconds randomness is equal for both electrode and then it is lower for dry than wet electrode.

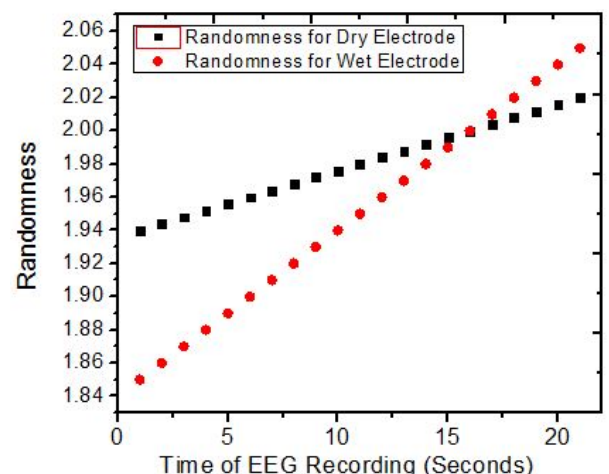


Fig. 10. Randomness analysis of wet and dry electrode

The noise level analysis for wet and dry electrode is shown in Fig. 11. In this figure the red and black colored bars indicate the noise level for wet and dry electrode respectively. This figure shows that the noise level is less for wet electrode than dry electrode with the starting time of recording signal. But the slope of noise level is more for wet electrode than dry electrode with the increasing of time. Therefore the performance of signal measurement also decreases for using wet electrode with the increase of recording time. Basically it is true that for using gel in wet electrode the conductivity between skin and electrode increase as well as performance is better than dry electrode. The abrasive paste and electrolytic gel using with wet electrode are sticky products which make the scalp and hair dirty. With the increase of SSVEP based EEG signal recording time from the scalp of human being the conductivity of gel may be decreased because when the gel starts to dry, the transductive properties of gel starts to disappear. For this reason, it is appreciable to use wet electrode for short time EEG measurement. But for long time measurement of SSVEP based EEG signal, using dry electrode is more effective than using wet electrode after some time of EEG recording.

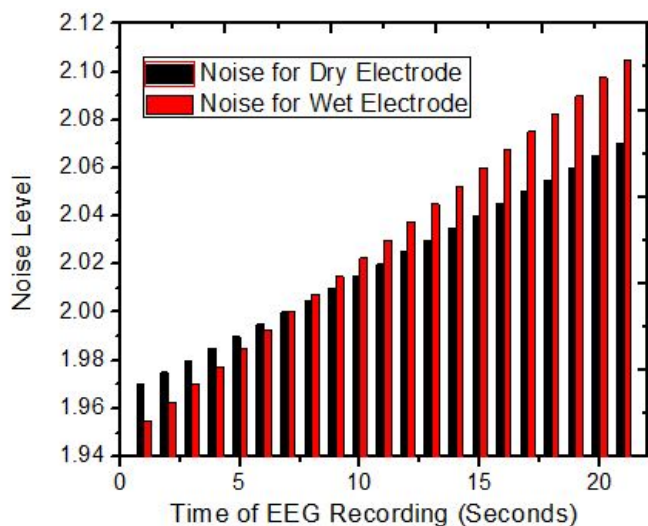


Fig. 11. Noise level analysis of wet and dry electrode

V. CONCLUSION AND FUTURE WORKS

For the smooth and perfect operation of Brain-computer Interface (BCI) paradigms, perfect acquisition system of SSVEP based EEG signal is required first of all which does not necessitate bald head of the performers. In this paper, this perfection of recording SSVEP based EEG signal is analyzed on the basis of comparative performance analysis for wet and dry electrode. From this comparative performance analysis, it can be concluded that though wet electrode is useful for short-term measurement of EEG signal from the scalp but for long-

term measurement dry electrode is more useful for perfect measurement of SSVEP based EEG signal. For effective wireless transmission of signal IR transmitter and receiver circuit is designed which has better noise immunity characteristics. By using our designed system effective measurement of SSVEP based EEG signal can be obtained which also can be used for operation of different BCI paradigms with proper degree of performance as well as this system can be easily applicable to different clinical applications where long-term measurement is required with higher degree of accuracy.

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A Simulation Study on Viscosity Change Effects in Ultrasound Based Carotid Atherosclerosis Diagnosis

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Abstract— Common cardiovascular diseases like heart attack and stroke are the major cause of mortality and disability in recent decades. Increased blood viscosity is an independent risk factor for atherosclerotic cardiovascular diseases. Elevated blood viscosity is associated with atherosclerosis and its growth. The knowledge of the arterial hemodynamics resulted due to atherosclerosis with different level of blood viscosity can clarify the relationship between artery stenosis and symptoms, and, finally, the risk of heart attack and stroke. Hence proficient imaging of blood flow has great clinical importance to understand the origin and evolution of disease. Therefore, investigation and detection of viscosity change and its effect might be helpful for better diagnosis of carotid atherosclerosis. In this simulation work, hemo-disturbances are studied in a carotid artery finite element model (FEM) with atherosclerosis under different viscosities consideration using computational fluid dynamics (CFD). The obtained FEM data are used to reconstruct color flow image of blood velocities using ultrasound radio-frequency (RF) signals to study atherosclerosis detecting the viscosity changes. The findings of this study suggest that ultrasound based blood flow image can detect the viscosity change effect related to atherosclerosis growth. We think the findings will be helpful to develop effective ultrasound imaging technique for atherosclerosis diagnosis.

Keywords— *blood viscosity; carotid atherosclerosis; finite element model; computational fluid dynamics; ultrasound imaging.*

I. INTRODUCTION

Carotid atherosclerosis is a major risk factor of sickness and mortality in patients with coronary heart disease. In atherosclerosis, plaques are deposited inside the artery wall which consists of fat, cholesterol, calcium and other substances of blood. The plaques are hardened gradually and narrow the flow channel of artery and hence, it disturbs the flow of blood to different parts of our body. The location most frequently organ that affected by atherosclerosis is the carotid bifurcation. Atherosclerosis in carotid artery impedes the flow of oxygen-rich blood into our brain and can cause brain stroke.

Recent studies show that, blood viscosity is increased in human body with carotid atherosclerosis. Blood viscosity is the measure of thickness and stickiness of blood. Viscosity is defined by the capacity of blood to flow through the vessels. It is an important bio-physiological factor that specifies the friction of the blood creates against the vessel surface and the

difficulties felt by the heart to pump blood in order to deliver oxygen to organs and tissues. Blood is a non-Newtonian fluid and its viscosity changes with its velocity. Usually blood at diastole is 5 to 20 times as viscous as the same blood at systole [1]. Therefore, study on the effects of blood viscosity along with the dynamics of blood flow in the circulatory system can explain why the atherosclerotic plaques are found only in particular areas inside the body.

Blood viscosity has importance as higher viscosity of blood creates lower shear stress in fat deposited areas. Therefore, the blood coagulates and spins around, initiating eddies and turbulence during the intervals of heart beats. This turbulence flow gradually damages the cells of the artery vessels. As a result, plaque tends to deposit densely in the low shear regions of the arteries where atherosclerosis grows up. This in turns causes the blood flow impeded further which then leading to further turbulence and further growth of plaque. Once the plaque growth reaches at an acute level, it explores a major probability of cardiovascular diseases like heart attacks and brain stroke [2].

Blood viscosity also affects blood pressure. Elevated blood pressure is a major risk factor for developing atherosclerosis and becomes a stronger risk factor after age 45 [3]. Elevated blood viscosity increases peripheral resistance. Blood pressure must increase to maintain cardiac output when there increased resistance due to hyper-viscous blood. Increased blood pressure then damages arteries further.

A proper understanding of the flow behavior and effective imaging of flow can help us in diagnosing during primary stage and give us proper direction to understanding of atherosclerosis and development of cardiovascular disease. In several studies, researcher suggested blood flow model using the fluid dynamics concept to analyze the mechanical behavior of the blood vessels. These models are useful to investigate the vessels condition and flow information using image or measurement data directly or inversely [4-8]. Some studies used the CFD data to construct the ultrasound based velocity images using pulsed-wave (PW) Doppler technique or color flow imaging (CFI) for blood flow analysis [9]. But, still there are scopes to improve the image quality of the flow which is important for effective diagnosis of atherosclerosis. We think, incorporation of viscosity related information can enhance the image quality of ultrasound based blood velocity images.

In this work, we plan to investigate viscosity change effect of blood flow in carotid bifurcation with atherosclerosis using Computational fluid dynamics (CFD) concept. Using CFD data, ultrasound based multiple color flow images of blood velocities are planned to simulate in order to detect and analyze the usefulness of viscosity related contrast in the diagnosis of carotid atherosclerosis and its growth. We focus on carotid bifurcation artery due to complex behavior of local flow of blood, their susceptibility to atherosclerosis and the effect of blood viscosity on atherosclerosis because the location is easily accessible with ultrasound.

The paper is organized as follows. Section II describes the methodological parts of this study emphasizing on CFD and ultrasound based color flow image formation. The simulation results and images are shown and discussed in section III. Finally, conclusions are drawn.

II. METHODS

Atherosclerosis in carotid artery can be studied using computational fluid dynamics concept. Investigation on increasing blood viscosity due to atherosclerosis and its effect in the carotid artery is our major consideration. A better understanding can be achieved through computer simulation of the blood flow model of stenosis in carotid artery bifurcation. Using a combination of computational fluid dynamics and finite element analysis techniques allows one to study the physics involved with atherosclerosis.

The blood in carotid artery is assumed to be incompressible, laminar and non-Newtonian fluid governed by Navier-stokes equations. The continuity equation for an incompressible medium is given below:

$$\nabla \cdot u = 0 \quad (1)$$

and Navier-stokes equation is expressed as [10]:

$$\rho \left(\frac{\partial u}{\partial t} + u \cdot \nabla u \right) = -\nabla p + \nabla \cdot \tau + \rho f^b \quad (2)$$

where u is the velocity vector, ρ is the density, p is the pressure, τ is the deviatoric viscous stress tensor and f^b is the body forces per unit mass accounting for gravitation effect.

In this study, we apply finite element method to solve the blood flow model. Computational fluid dynamics has recently been applied to vascular flows, such as in the study of the flow dynamics in the stenosed carotid bifurcation [11]. We have used CFD to solve differential equations of fluid flows numerically which is an effective technique for computing blood flow data.

The scattering centers (red blood cells) are taken from the FEM data to make the synthetic phantom for ultrasound imaging. The position of the scatterers for multiple times instant is necessary for velocity estimation. The flow estimation can be done from the shifting information of the scatterers using ultrasound RF signals. For ultrasound RF signals generation, we plan to use the widely accepted ultrasound image simulation software Field II [12-14]. Field II allows modeling arbitrary probes and implementing realistic scanning sequences to generate RF signals for simulating ultrasound images using the model data.

Color flow imaging (CFI) is now found useful in assessing blood flow in many clinical conditions. The color flow images reflect the projected velocity components in the direction of ultrasound beam propagation. A rate of change of phase can be interpreted as a time or frequency shift and the velocity of the target can be calculated from these information [15-17]. The time shift can be calculated using auto-correlation or cross-correlation of ultrasound echo signals. We have followed the standard auto-correlation based algorithm to estimate the color flow map using Field II ultrasound image simulator [17].

III. RESULTS AND DISCUSSIONS

We have made a 2D FEM model of stenosed carotid artery bifurcation as our major emphasis on viscosity effect. We consider the blood as non-Newtonian fluid and laminar flow for the model. Figure 1 shows the stenosed carotid artery bifurcation model with common carotid artery (CCA), internal carotid artery (ICA) and external carotid artery (ECA). As our emphasis is on viscosity effect, we consider steady state analysis. A fixed density of 1060 kg/m³ and variable viscosity of 3.5 mPas, 7 mPas and 10 mPas has been considered for the analysis. The model is meshed into 17524 elements as shown in Fig. 2. The artery wall is assumed as rigid. The inlet boundary condition has been set considering a parabolic profile with a maximum velocity of 0.5 m/s. The outlet boundary is defined as a pressure outlet with initial pressure of 0.0 Pa. All wall boundary layers were defined as no-slip walls, resulting in a velocity of 0.0 m/s along the wall surface. The simulations have done using MATLAB [18].

In ultrasound simulation image, the velocity profile calculated using the autocorrelation algorithm. The CFD data based map of the blood velocity in the carotid can be an good reference to compare the ultrasound simulated images to investigate the viscosity effects.

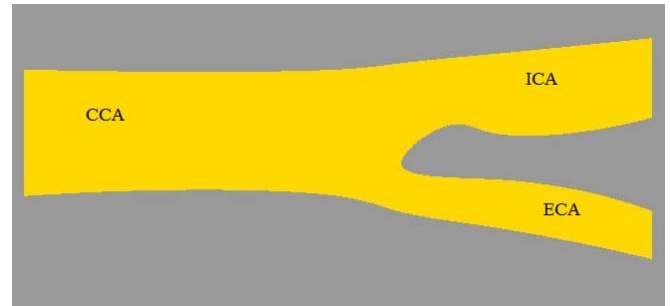


Fig. 1. Stenosed carotid artery model with bifurcation

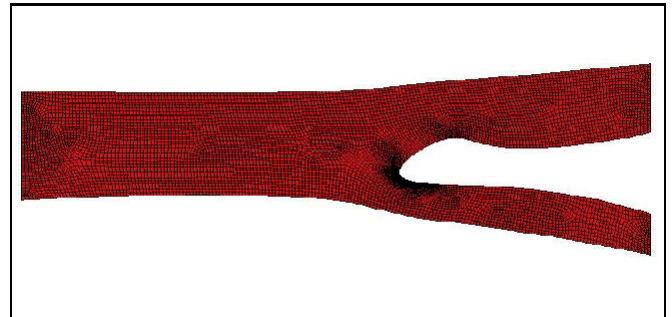


Fig. 2. Mesh diagram of the stenosed carotid artery model

For ultrasound wave propagation simulation and radiofrequency (RF) signal generation, we have used Field II program due to its accurate modeling of transducer properties, beam forming, wave propagation and interactions. We set a linear probe with 5MHz center frequency and 100MHz sampling rate was modeled. 256 RF lines were simulated using 196 transducer elements of which 64 active elements firing with focal length to cover the phantom. Hanning apodization is used to transmit and receive aperture. We have considered 10 time steps to generate RF signals for each frame using CFD data of 10 time steps. Velocity map frames are calculated using the phase shift estimation from autocorrelation of corresponding RF signal envelope of different frames.

CFD simulation images for the 3 different blood viscosities are shown in Fig. 3. Ultrasound based color flow images are shown for the 3 different blood viscosities in Fig. 4. Though inlet velocity was 0.5 m/s, the maximum velocity reaches on about 2.5 m/s due to atherosclerosis stenosis. These images can be compared with corresponding CFD simulation images. In the ultrasound color flow images, a parabolic velocity inlet similar to reference CFD simulated velocity map have been noticed with some distortions.

In the top CFD simulation image frame, where blood viscosity is 3.5 mPas (normal), the blood velocity is normal in the CCA and high velocity gradients are observed at the outer wall of the ICA and ECA near the apex-induced separation due to atherosclerotic stenosis in the carotid bifurcated region that narrow the artery. The estimated velocity fields in the ultrasound color flow images show covenant with the CFD results with a little fluctuation.

In the middle CFD simulation image frame, where blood viscosity is 7 mPas, the resistance of blood increases for increasing viscosity. As a result, blood velocity decreases in the CCA and arise complex flow pattern in the bifurcation region that is the cause of further growing of carotid atherosclerosis. The blood velocity also decreases in the ICA and ECA near the apex-induced separation. Insufficient blood might supply into the brain that is the indication of stroke. The estimated ultrasound color flow image shows slight fluctuation in compared to the CFD simulation image.

In the bottom CFD simulation image frame, where blood viscosity is 10 mPas, the resistance of blood highly increases for further increase of viscosity. As a result, blood velocity more decreases in the CCA and a large zone of turbulence flow is present in the bifurcation which tends to grow carotid atherosclerosis and narrow the artery. The blood velocity more decreases in the ICA and ECA near the apex-induced separation and extremely less blood supplied that increases the risk of stroke. The estimated flow map in the ultrasound flow images show a covenant with the CFD results also.

The hemodynamical disturbances in atherosclerotic carotid bifurcations for different level of blood viscosities are shown complex flow pattern during high degree of blood viscosity and significantly different from each other mainly due to different degrees of viscosities. From both CFD and ultrasound images, it is observed that the velocity images are different for different viscosity. As the viscosity increase with atherosclerosis growth, effective estimation of these velocity differences using ultrasound technique might prove additional information to enhance the quality of color flow image.

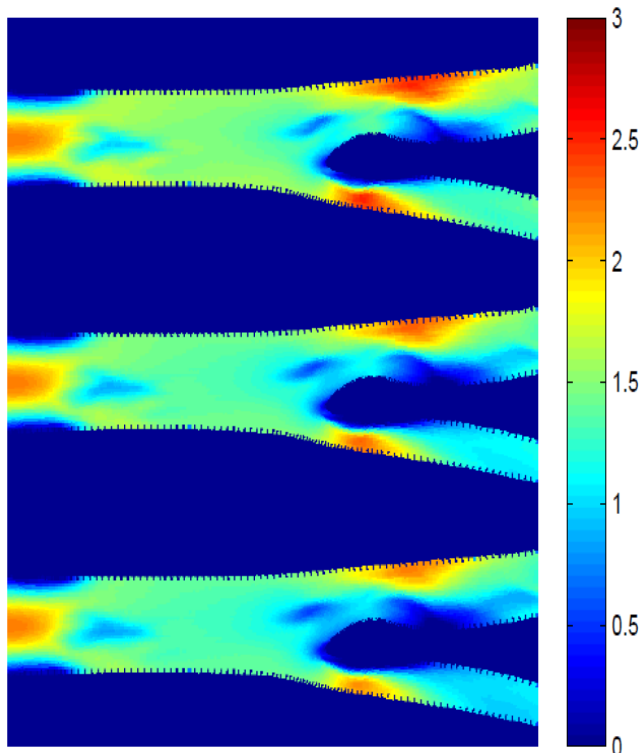


Fig. 3. CFD simulation image of stenosed carotid artery for blood viscosity 3.5 mPas (top), 7 mPas (middle), and 10 mPas (bottom) respectively.

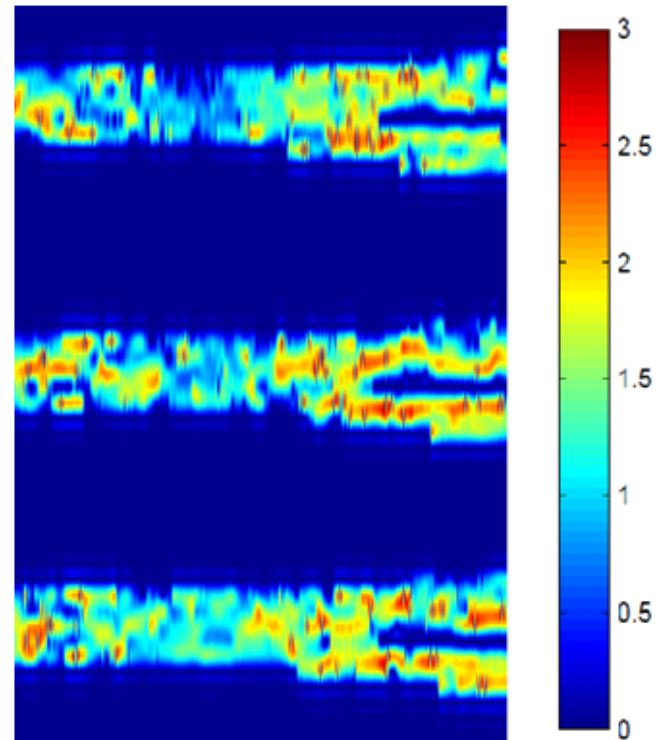


Fig. 4. Ultrasound simulation image for blood viscosity 3.5 mPas (top), 7 mPas (middle), and 10 mPas (bottom) respectively.

IV. CONCLUSIONS

In this work, the flow characteristics in carotid atherosclerosis bifurcations for different degree of blood viscosities are investigated using CFD and ultrasound simulated color flow image. We compared ultrasound simulated flow characteristics with CFD flow behavior and see that a good agreement between them. It was observed that the complex flow pattern differs from each other due to different degree of blood viscosities. A patient with carotid atherosclerosis increase blood viscosity and this increasing nature of viscosities further growing carotid atherosclerosis and arising more complex flow pattern. Using the information of the complex flow and velocity difference due to viscosity change, better quality color flow images can be constructed. Incorporation of viscosity related changes might be helpful to diagnosis the present conditions and predict the future progressions of the disease to minimize the risk of stroke.

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Development of Low Cost Wireless ECG Data Acquisition System

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Abstract—Cardiac malfunctioning is the leading cause of mortality today. Electrocardiogram (ECG) plays a vital role for diagnosis of cardiac disease. The frequency information from the ECG signal is highly useful in clinical applications such as sudden cardiac arrest prediction, cardiovascular disease detection etc. This paper presents the development of a low cost, wireless ECG data acquisition system for clinical purpose and analysis. In this system, Arduino Uno is used and the ECG signal is transferred to the laptop by Bluetooth serial communication. By using a Bluetooth dongle, this system can be used in desktop computer. An application is developed by processing, which visualizes and stores the ECG signal in real time. This application is built for making this system usable for clinical purpose. This application stores the ECG signal in text file, which can be used in MATLAB for analysis. This system can be used in Windows, Linux, or Mac OS. This system is portable, battery powered and it is capable of acquiring up to 22 hours of continuous electrocardiogram data.

Keywords— ECG; Processing; Arduino; Powerline interference

I. INTRODUCTION

An Electrocardiogram (ECG) is a recording of the electrical activity of heart generated by the cells of the heart, throughout the cardiac cycle (both normal and abnormal) [1]. ECG signals are usually recorded at the surface of the body and it is a bipolar low-frequency weak signal. Normally the frequency range of a normal ECG signal is 0.05-100Hz [2]. Its amplitude ranges from 0.1mV to 3mV. ECG signal plays a vital role in diagnosis especially for heart related diseases and disorders. By ECG interpretation many cardiac abnormalities can be detected including insufficient blood flow, death of heart muscle due to coronary thrombosis and enlargement of heart muscle [3].

Our research aims to develop a low cost ECG data acquisition system. In our previous paper we have developed an ECG monitoring system using smartphone [4]. In this research, we have modified the system for using it as an ECG data acquisition system which is suitable for both laptop and desktop computer. This paper presents a prototype ECG data acquisition system, which is portable, battery powered and it includes wireless facility for safety concern and reducing noise interference. It gives wireless connectivity up to 9 meter. As this system is battery powered, it can be used during load shedding and on those remote areas where electricity have not

reached. In this system, Processing is used for developing an application, which visualizes and stores the ECG signal in real time. It stores the ECG signal in text file, which can be taken in MATLAB for analysis and disease diagnosis. Processing is an open source programming language. For this, it can be used for clinical purpose as a single channel ECG monitor. This system consists three electrodes. By this system ECG signal can be taken from either any limb leads (Lead I, II and II) or any augmented limb leads (aVR, aVL, and aVF).

II. SYSTEM ARCHITECHURE

This system includes instrumentation amplifier AD620, operational amplifier LM741 and CA3130, Bluetooth module HC-06, Arduino Uno and filter. Arduino Uno is a low power and highly efficient microcontroller board based on the ATmega328 microcontroller.

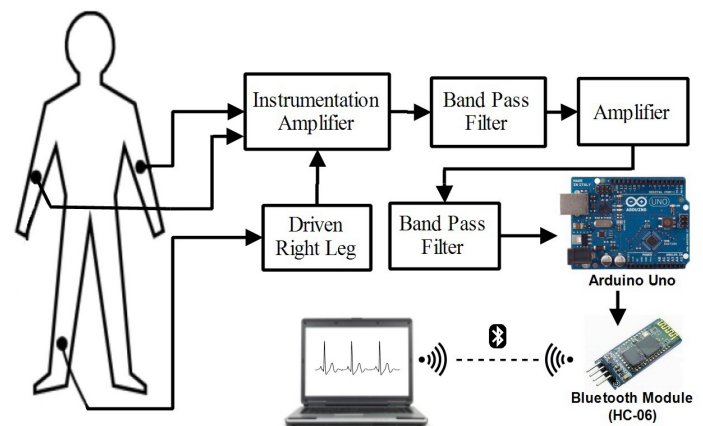


Fig. 1. System Architecture of the ECG data acquisition system.

III. PROPOSED METHODOLOGY

This system consists of three units: signal conditioning unit, power supply unit, and a microcontroller unit.

A. Signal Conditioning Unit

Silver chloride electrodes are used in his system for sensing ECG signal. A commercially available disposable electrode named Bio Protech T716 was used which is shown in Fig. 3 [5]. ECG signal is transferred to the instrumentation amplifier

input, by using the patient cable. Figure 2 shows the circuit diagram of this system. A commercially available instrumentation amplifier AD620 is used in this system. It is a low cost, high accuracy instrumentation amplifier, having CMRR greater than 100dB to nearly 1 kHz [6]. The important features of AD620 are low power consumption, gain range 1 to 10,000, wide power supply range (± 2.3 V to ± 18 V), higher performance than three op-amp IA designs, low noise. Total gain of 587 is achieved by tow stage amplification. Two stage amplification is done to avoid the risk of amplifier saturation. It generally occurs due to amplification of electrode offset voltages [7]. To set the gain of AD620, a resistor R_g is used. The gain is selected to be 11.5 by setting R_g to 4.7k Ω for first stage amplifier. The gain expression is given in Eq. (1).

$$\text{Gain} = \frac{49.4 \text{ k}\Omega}{R_g} + 1 \quad (1)$$

AD620 has high common mode rejection ratio of 100dB for gain of 11.5 [6]. Output voltage of the instrumentation amplifier is given by Eq. (2).

$$V_o = \text{Gain} \times (V_1 - V_2) + \frac{1}{2} \times A_c \times (V_1 + V_2) \quad (2)$$

Where, A_c is the common mode gain and V_1, V_2 are the voltage obtained from left electrode and right electrode. According to common mode rejection ratio (CMRR) we can write as,

$$\begin{aligned} \text{CMRR}(\log) &= 20 \log \frac{\text{Gain}}{A_c} \\ \Rightarrow 100 &= 20 \log \frac{\text{Gain}}{A_c} \\ \Rightarrow 100000 &= \frac{11.5}{A_c} \\ \Rightarrow A_c &\cong 0.00015 \end{aligned}$$

For second stage amplification CA3130 op-amp was used as a non-inverting amplifier and voltage gain of 51 was achieved in second stage amplification.

Generally, the ECG signal is superimposed by a large common mode voltage. The sources of common mode voltage are generally 50 Hz or 60 Hz line frequency noise and electrode DC offset potential [8,9]. A Driven Right Leg Circuit is used in this system. It eliminates the common mode signals. It inverts and amplifies the average common mode signal back into the patient's right leg, which cancels line frequency noise and improves the common mode rejection ratio [10].

For filtering, first a band pass filter is implemented by using a high pass filter of 0.03 Hz cut off frequency and a low pass filter of 159 Hz cut off frequency in cascade. It filters the amplified ECG signal from AD620 instrumentation amplifier. Then the filtered ECG signal is again amplified by CA3130 op-amp. After amplification ECG signal is again filtered by a band pass filter having same cut off frequency. All those filters used in this system are passive first order filter. General expression of the cut off frequency of the filter is given in Eq. (3) [11].

$$F_c = \frac{1}{2 \times \pi \times R \times C} \quad (3)$$

Arduino UNO board contains a 6 channel 10-bit analog to digital converter (ADC). It's internal ADC of works between 0V to +5V [12]. It returns a linear value from 0 to 1023 corresponding to 0V and +5V respectively [12-13]. A virtual ground of 3.3V is used as reference of AD620 instrumentation amplifier. This makes the bipolar ECG signal to Unipolar.

B. Power Supply Unit

In this system tow 9V disposable Lithium battery is used for making a dual supply of ± 5 V. For safety and short circuit protection, a voltage regulator LM7805 is used. In this system all equipment's get power from battery and ECG signal is transmitted wirelessly. It is done to avoid power line interference (50Hz) and other noise interference.

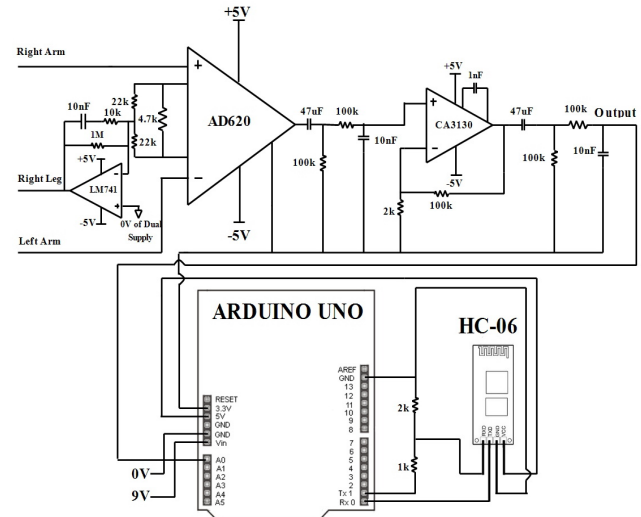


Fig. 2. Circuit diagram of the system.

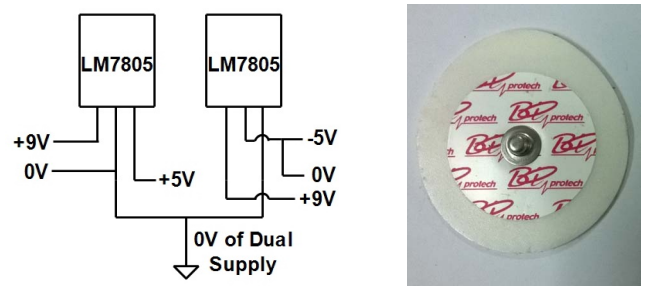


Fig. 3. (a) Designed dual power supply for the system. (b) Bio Protech T716 ECG electrode.

C. Microcontroller Unit

Arduino Uno board consist a microcontroller named ATmega328. For serial transmission baud rate is taken as 9600 bps and sampling rate of 320 samples per second is obtained. Sampling rate can be increased further by changing the baud rate or core clock speed [14]. We found sampling rate of 3400 samples per second at baud rate of 115200 bps and 16 MHz core clock speed. A Bluetooth module HC-06 does signal transmission between Arduino UNO and laptop. For desktop computer a Bluetooth dongle in needed for signal transmission. For isolating subject body from AC mains, ECG signal is transferred to laptop via Bluetooth serial communication. It

also reduces noise and power line interference by making the signal conditioning circuit isolated from AC mains.

IV. SOFTWARE IMPLEMENTATION

In software structure of this system there are two main parts, they are device layer and application layer. The device layer is to drive hardware and the application layer is for the user for storing and visualizing the ECG signal in real time. For programming the microcontroller, integrated development environment provided by the Arduino platform is used. Arduino Uno transfers the quantized value of ECG signal to Laptop via Bluetooth serial communication by using the Bluetooth module. For visualizing and storing the ECG signal an application is developed by Processing. Processing is an open source programming language and integrated development environment (IDE). It builds on the Java language, but uses a simplified syntax and graphics programming model and available for Linux, Mac and Windows operating system [15]. Serial library of processing is used for receiving the ECG data. ECG signal is visualized using line () function [15]. For sorting println() function of ECG signal is used , which creates a txt file and saves the values received from Arduino ADC. Processing IDE of version 2.21 is used for building the application. The main purpose of building this application is to make this system usable for clinical purpose.

V. EXPERIMENTAL MEASUREMENTS AND RESULTS

Figure 5 shows the prototype ECG acquisition system and experimental measurement performed by us. The ECG signal is visualized in laptop running on Windows 8.1 operating system having Intel B820 processor (Dual core 1.7 GHz, 2MB catch memory) and 4GB ram. In figure 5 every red horizontal line represents 1V of total 5V. As a virtual ground of 3.3V is used, ECG signal is visualized above 3V.

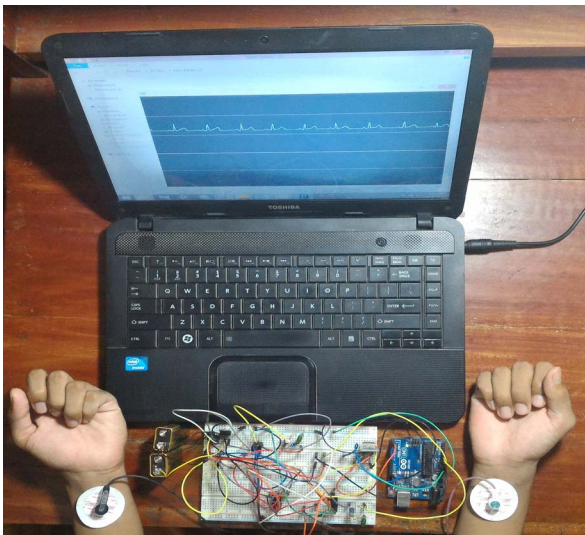


Fig. 4. Experimental measurements performed by us.

For analysis, MATLAB R2014a is used and the data of the ECG signal stored by the application is imported in MATLAB. The ECG signal is taken from a boy of 23 years old in LEAD I. Figure. 6 shows the ECG signal without filtering in MATLAB.

Figure. 7 shows the frequency spectrum of the ECG signal. In this figure, magnitude for 50Hz is 0.002224 and that is for common-mode voltage produced by the ac mains of 50 Hz frequency with line voltage 220V [16]. Due to this the ECG signal is contaminated by power line interference (50Hz).

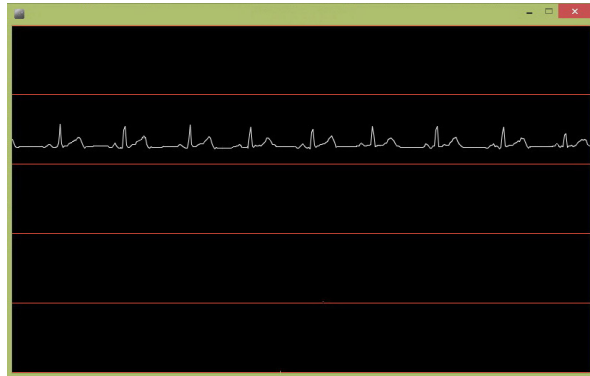


Fig. 5. Output of the application.

In this system signal conditioning and signal transmitting circuit is isolated from ac mains but there exist a small amount of power line interference. This is because of electric field coupling from the power line to the ECG leads or the power line to the patient and magnetic induction. Electric field coupling generally occurs when the mains supply or mains cable between the wall socket and the ECG equipment or the patient came into close proximity with an electric field. Magnetic field induction can occur by mobile phones, electrical equipment, nearby florescent lights [7,16].

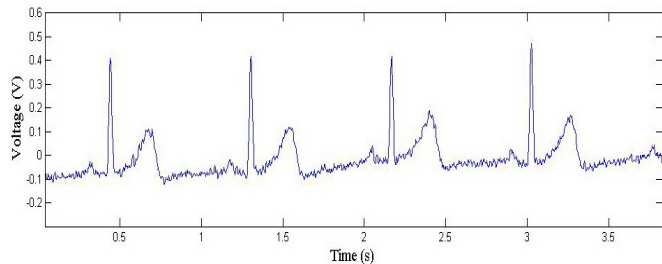


Fig. 6. ECG signal without filtering in MATLAB.

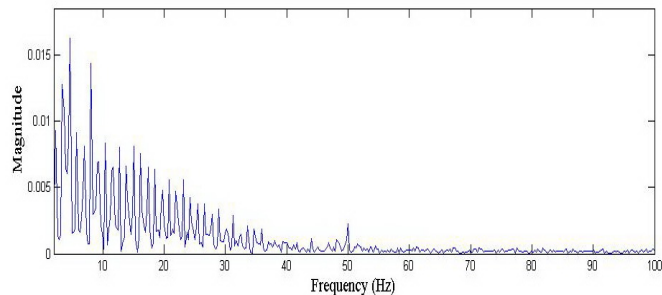


Fig. 7. Frequency spectrum of the ECG signal shown in figure 6.

Figure 8 shows the filtered ECG signal and figure 9 shows its corresponding power spectral density. For filtering a 4th order butterworth band pass filter having lower and higher cutoff frequency of 0.033Hz and 120Hz and a 4th order elliptic

band stop filter of 50Hz was implemented. Sampling frequency was 320 samples per second.

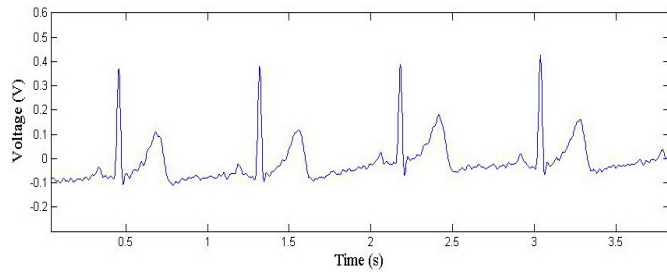


Fig. 8. ECG signal after filtering.

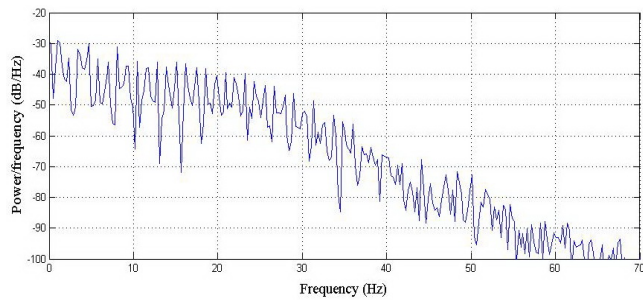


Fig. 9. Power spectral density of the filtered ECG signal.

TABLE I. shows different ECG parameters normal range and all those intervals measured by MATLAB are found in normal range [17].

TABLE I. MEASURED ECG PARAMETERS AND THEIR NORMAL RANGE

Parameter	Measured	Normal Range	Unit
Heart Rate	69	60-100	Beats per minute
RR Interval	0.864	0.6-1.0	S
QT Interval	0.375	0.36-0.44	S
PR Interval	0.121	0.12-0.20	S
QRS Interval	0.077	0.06-0.1	S

TABLE II. COMPONENT LIST FOR THIS SYSTEM

Component Name	Quantity	Unit Price (USD)	Total Cost (USD)
Arduino UNO	1	9	9
HC-06	1	6	6
Bluetooth Dongle	1	2	2
AD 620	1	6	6
Resistor	13	0.01	0.13
Capacitor	6	0.02	0.12
LM 7805	2	0.15	0.3
CA3130	1	0.25	0.25
LM 741	1	0.15	0.15
ECG Electrodes	3	0.09	0.27
9V Battery	2	0.5	1
Total Cost of the system (USD)			25.22

Table II shows the list of the component used for this system [18]. Total cost of this system is USD 25.22.

VI. CONCLUSION

The main aim of this research is designing and building a low cost, wireless ECG data acquisition system has been fulfilled. As this system is wireless and battery powered, noise interference is reduced and patient safety is ensured. This system is simple and effective for continuous ECG data acquisition. The cost of this system is USD 25.22, which makes this system highly inexpensive and ideal for under developed and developing countries. This system can be used as single channel ECG monitoring system with a laptop or desktop computer for clinical purpose and ECG data acquisition system for analysis. The component used for this system consumes very low power and the developed hardware for this system takes maximum current of 65mA. We have tested this system and found that it is capable of acquiring up to 22 hours of continuous electrocardiogram data. In this system maximum sampling rate of 3400 samples per second is obtained and during operation about 30% of the CPU was utilized by the application. In future, acquisition and monitoring of multichannel ECG and acquisition of EOG, EMG signals can be added to this system.

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Feasibility Assessment of Wind Energy Prospect In Bangladesh And A Proposal of 2MW Wind Power Plant At ‘Parky Saikat, Chittagong’

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Abstract— At present, a higher number of researches are being directed on alternative energy sources to supply power for modern life of today and the future. As the conventional sources of energy are diminishing day by day, one of the most promising sources is wind energy which can provide electrical power using wind turbines. The increase in the use of this type of energy requires greater consideration of wind speed, installation and especially the per unit cost of energy produced by wind turbines. This paper will discuss the feasibility of wind energy prospect in Bangladesh considering the wind speed at the particular representative coastal locations and also the wind turbine which can be used commercially. We also propose a 2MW wind power plant at ‘Parky Saikat, Chittagong’. For which the system and cost calculation is carried out by the ‘HOMER Energy’ simulation software.

Keywords—Wind Energy; Wind turbine; Wind Speed.

I. INTRODUCTION

In modern world, human life is directly dependent on electricity. All of our activities and standard of life is heavily controlled by energy especially electric energy. But energy crisis in Bangladesh is becoming acute day by day. The major part of the electricity comes from conventional power sources such as Coal, Gas, Oil, Diesel etc. As a consequence of the electricity shortage and the continuous emitting of large amount of carbon which is responsible for so called Green House Effect causes natural disasters. The continuation of this crisis arise the possibility of industrial losses for which the economic condition of our country may fall behind. This effect may cause the abrupt shutdown of major industries and commercial buildings that occurs the severe unemployment problem. In Bangladesh presently, 62% of the total population (including renewable energy) has access to electricity and per capita generation is 321 KWh, which is very low compared to other developing countries [1]. Due to electricity the entire development of the country is hindering. As a developing country Bangladesh use energy for its needs, but there is significant gap between available power and demand. For accelerating its development, Bangladesh need to fill up the shortage energy. Along with the government many private

companies and non-government organization are working in Bangladesh power sector. At present Bangladesh government has been emphasizing heavily on power sector to mitigate the energy crisis. Currently it ranks third in the world in the production of wind power energy with wind energy accounting for 70% of the total energy supplied by nonconventional energy sources. This is more so important as energy demand is like to increase by three times between 2005 to 2030[2].

II. BANGLADESH’S WIND ENERGY SCENARIO

Bangladesh is situated between 20°34' and 26°38' N latitudes 88°01' and 92°41' E longitudes with 160 million people living on 1,47,570 sq. km area. It has 710km coastal line which lays a potential prospect for harnessing wind energy but is yet to take advantage of wind energy to generate electricity. In the recent past, some initiatives have been taken for extracting wind energy [3]. At present there are only two wind power plants functioning in Bangladesh.

TABLE I. PRESENT WIND POWER PLANTS IN BANGLADESH

No	Wind Power Plants in Bangladesh		
	Name	Installed Capacity	Present Status
1	Feni Wind Turbine Power Plant	0.99 MW	Running
2	Kutubdia Wind Power Plant	1 MW	Running

III. WIND ENERGY RESOURCES IN BANGLADESH

In Bangladesh, winds are available mainly during the Monsoon and around one to two months before and after the Monsoon. During the months starting from late October to the middle of February, wind is either remaining calm or too low to be any use by a traditional wind turbine. Except the mentioned period of four months, if a wind turbine properly designed and located, can supply enough energy to be marketable [3].

There are some prospective location in Bangladesh such as Feni Muhuri Dam, Kuakata, Kutubdia, Cox's Bazar, Chorfession, Sitakunda, Parky Saikat, Shahporir Deep where wind speeds in monsoon season are enough to harness wind energy with proper wind turbine design.

TABLE II. AVERAGE WIND SPEED AT DIFFERENT HEIGHTS AT DIFFERENT LOCATIONS

No	Average Wind Speed of Different Location		
	Location	Height(m)	Average Wind Speed (m/s)
1	Cox's Bazar	25m	3.792
2	Chorfession	25m	4.433
3	Chittagong	25m	4.367
4	Kuakata	20m	3.135
5	Kuakata	30m	4.146
6	Kutubdia	20m	3.642
7	Sitakunda	20m	3.015
8	Sitakunda	30m	3.554

Table II shows the average wind speed of eight different locations at different heights [4]. It is noticeable that average wind speed for all locations is above cut in speed which usually ranges between 3 to 4 m/s.

TABLE III. AVERAGE WIND SPEED FOR TWO LOCATION AT 18.3METER HEIGHT

No	Average Wind Speed of Two Locations		
	Locations	Height(m)	Annual Average Wind Speed (m/s)
1	Shahporir Deep, Teknaf	18.3	4.66
2	Parky Saikat, Chittagong	18.3	3.67

Table III shows the average wind speed at 18.3m height for Shahporir Deep and Parky Saikat which was recorded from Nov-10 to Oct-11 and Feb-12 to Jan-13 respectively [4]. At these locations annual average wind speed is above the cut in speed.

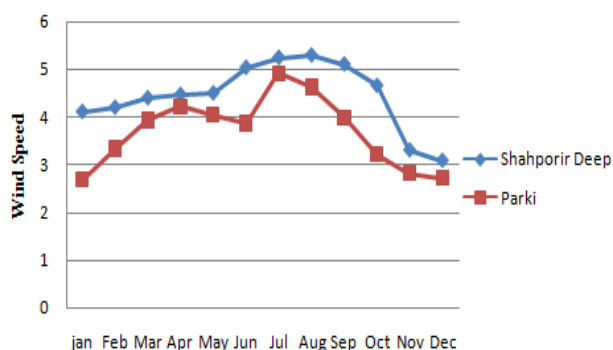


Fig. 1. Month Vs Wind Speed

Fig. 1 shows the month Vs wind speed at Shahporir Deep and Parky Saikat, Chittagong. It is clear that the wind speed is

higher during the monsoon seasons and low in the winter season.

Table IV shows the monthly and annual average wind speed at 50m height of four locations namely Muhuri Dam at Feni; Mognaghat at Cox's Bazaar; Parky Saikat at Chittagong and Kuakata at Patuakhali [3]. From the Table IV it is clear that annual average wind speed of these four sites is more than 6.5 m/s. It is an internationally accepted thumb rule that a site having the wind speed above the 6.0 m/s or higher is feasible for harnessing wind electricity with commercial viability [5].

TABLE IV. MONTHLY AVERAGE WIND SPEED AT 50M HEIGHT

Month	Monthly Average Wind Speed of Four Different Locations			
	Muhuri Dam, Feni (m/s)	Mognaghat, Cox's Bazar (m/s)	Parky Saikat, Chittagong (m/s)	Kuakata, Patuakhali (m/s)
Jan	5.10	5.30	4.90	4.6
Feb	5.30	4.80	5.10	5.50
Mar	7.0	7.30	7.60	7.70
Apr	7.70	7.90	7.80	8.30
May	8.10	8.20	8.20	7.90
Jun	7.20	8.20	7.60	6.90
Jul	7.40	8.40	8.10	7.70
Aug	6.80	7.70	7.40	7.50
Sep	6.70	7.10	6.90	6.90
Oct	6.20	6.80	6.40	6.30
Nov	5.60	5.90	5.60	5.50
Dec	4.90	5.40	5.10	4.80
Annual Avg. Speed (m/s)	6.50	6.90	6.73	6.63

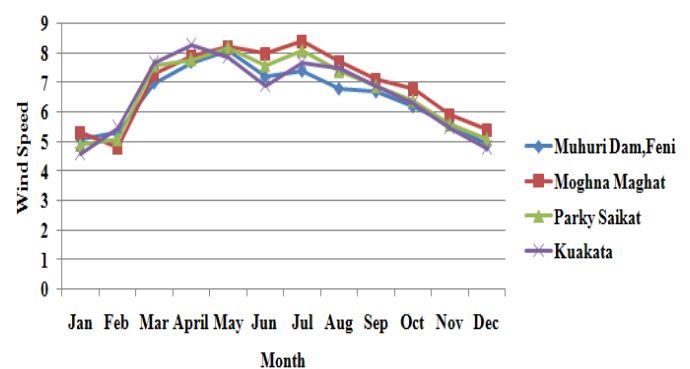


Fig. 2. Wind Speed Vs Month

In Fig. 2 shows the Wind speed Vs Month data at different location where Mognaghat shows the highest speed in the middle of July month.

IV. POWER LAW METHOD

A formula was proposed first by Hellman in 1916 [6]. It uses a single parameter α that depends on the stability of the atmosphere.

$$V_2 = V_1(h_2/h_1)^\alpha$$

Where,

h_1 = Measured height (m)

V_1 = Measured wind speed at h_1 height (m/s)

h_2 = Height of the wind speed estimation (m)

V_2 = Estimate wind speed at h_2 height (m/s)

α = Exponent of Hellman (1/7 for lakes, ocean)

TABLE V. AVERAGE WIND SPEED OF THREE DIFFERENT LOCATIONS AT DIFFERENT HEIGHT

No	Average Wind Speed (m/s) of Three Locations			
	Height	Parky Saikat, Chittagong	Kuakata, Patuakhali	Kutubdia, Cox's Bazar
1	25m	4.30m	4.50m	4.17m
2	50m	6.73m	6.63m	5.97m
3	75m	7.16m	7.02m	5.90m

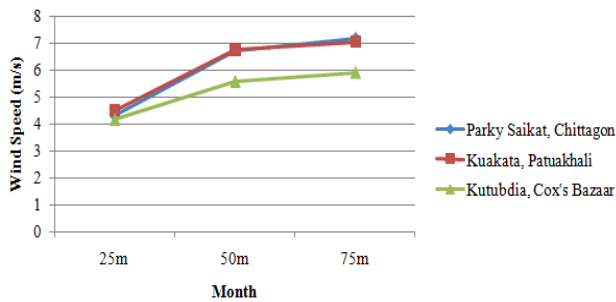


Fig. 3. Wind Speed Vs Height of Three Different Locations

Comparing the wind speed of Table III to Table IV and Table V, It is obvious that wind speed increases with the height. From the above data analysis it can be proofed that harnessing wind energy in Bangladesh is feasible.

V. COMMERCIALY USED WIND TURBINE

A wind turbine is a device that converts kinetic energy from the wind into electricity. There are mainly two kinds of wind turbine. They are horizontal axis wind turbine (HAWT) and vertical axis wind turbine (VAWT).

TABLE VI. DIFFERENCES BETWEEN HAWT AND VAWT

No	Differences Between VAWT And HAWT	
	VAWT	HAWT
1	VAWT is Omni directional.	HAWT is affected by wind direction.
2	YAW mechanism not required.	YAW mechanism required.
3	No power calculation for VAWT.	Power production by HAWT can be calculated.
4	It has a bit complex power	It smoothes transfer of the

No	Differences Between VAWT And HAWT	
	VAWT	HAWT
	transfer due to its Omni-directional characteristic.	power from the rotor to the gear box.
5	Used in small applications such as homes, office building roof.	Used in commercial wind power plant.

Mainly HAWT is used in commercial wind power plants worldwide between two types of turbines. The reasons of utility-scale wind farms are set up where wind is relatively stable. So catching the wind from any direction isn't a particular advantage. VAWT blades have rarely at an optimal angle to the wind or in clean air. So they can never be as efficient as a tri blade HAWT and won't generate more electricity. Most importantly for commercial power plants, it is must to calculate the electricity production which is only possible by HAWT. At present HAWT has the capacity factor 35%-47% which is increasing day by day. Many research are ongoing to improve its capacity.

VI. THEORY

Theoretical extracted (max.) power from wind is:

$$P = C_p \cdot \frac{1}{2} \cdot \rho \cdot A \cdot V^3 \quad (1)$$

Where,

P = Potential power

C_p = Power coefficient

ρ = Air density

A = Swept area of rotor

V = Wind speed

C_p is the percentage of power that can be extracted from the wind and take into account losses due to Betz law and those assigned to the internal mechanisms within the turbine. According to Betz's law, no turbine can capture more than 16/27 (59.3%) of the kinetic energy in the wind. The factor 16/27 (0.593) is known as Betz's coefficient. Practical utility-scale wind turbines achieve at peak 75% to 80% of the Betz limit[1].

VII. PROPOSED WIND POWR PLANT WITH COST ANALYSIS

Based upon the data analysis and turbine study, we have proposed a 2MW grid tied wind turbine power plant at 'Parky, Chittagong' and we simulated the system by 'HOMER Energy'. Our proposed plant mainly supplies the produced electricity to the grid from where it will be supplied to the families. We have used the SRC31 250/50 kW wind turbine model made by NEPC INDIA in our simulation[7].

We have assumed a load scheme for about 3000 families near 'Parky Beach' each family having the average demand about 600watts and max demand in peak hour can be 1800 kWh. Our system is grid tied as said earlier and system can take max 1000kW from grid in case of shortage. To fulfill the max demand our system required total 8 wind turbines which hub

height is 50m. Fig. 4 shows the seasonal load profile of our proposed load for 300 families and Fig. 5 shows the average monthly wind speed respectively.

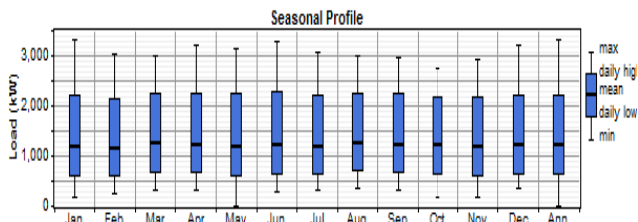


Fig. 4. Seasonal Load Profile

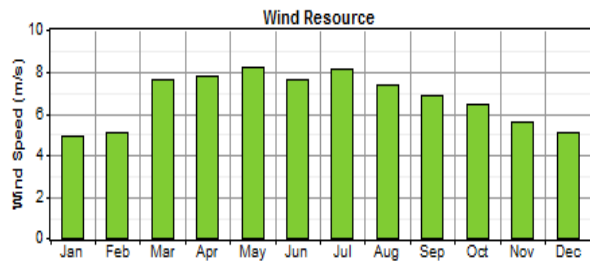


Fig. 5. Monthly Average Wind Speed at 50m Height at Parky Beach

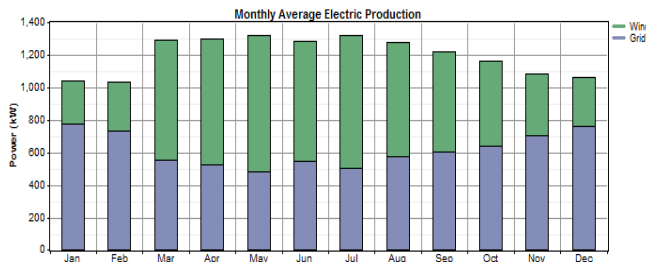


Fig. 6. Monthly Average Electricity Production

Fig. 6 shows the contribution from wind and grid in monthly average electricity production. The wind turbine reflects the contribution (renewable fraction) towards the electricity generation is 49% (5,128,367 kwh/year). Rest of the electricity production about 51% comes from the grid.

TABLE VII. CAPACITY SHORTAGE

No	Capacity Shortage		
	Quantity	KWh/yr	Percentage(%)
1	Excess Electricity	54,439	0.5
2	Unmet	1,345,179	12.4
3	Capacity Shortage	2,137,939	19.7

Table VII shows the amount of unserved loads. Capacity shortage comprises the both unserved loads and unserved operating reserve. In case of wind power plant it is usually take the value of maximum capacity shortage is 30% which is also taken from our simulation in HOMER energy. We find the capacity shortage is about 20% from our simulation, which is considerably a good quantity.

TABLE VIII. TOTAL COST SUMMARY

No	Cost Summary			
	Component	Capital (\$)	O & M (\$)	Total
1	NEPC 50/250 KW	2,240,000	51,615	2,291,614
2	GRID	100,000	2,458,919	2,558,919
3	Other System (Fixed Cost)	50,000	0	50,000
5	System	2,390,000	2,510,534	4,900,532

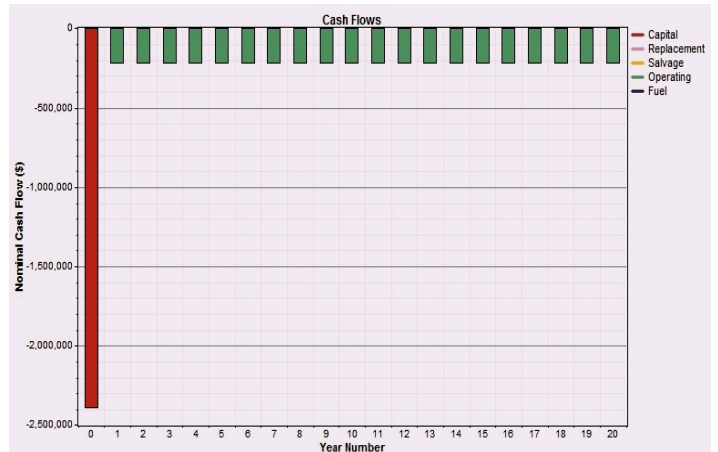


Fig. 7. 20 Years' Cash Flow

Table VIII shows the total cost summary for our proposed plant and in Fig. 7 shows the cash flow for the system's life time i.e. 20 years. So our proposed system is feasible at that location in compare with its speed and cost.

VIII. CONCLUSION

In this paper we concluded that harnessing wind energy in Bangladesh is feasible at medium and large scale via wind speed analysis at different locations at different height. Also we found that HAWT is better than VAWT for commercial viability. Hence we proposed a wind turbine power plant and analyzed a cash flow at 'Parky Saikat, Chittagong'.

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A Feasibility Study of Mini hydroelectric Power Plant at Sahasradhara Waterfall, Sitakunda, Bangladesh

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Abstract— Electricity production from hydropower has been, and still is today, the first renewable source used to generate electricity. This paper discusses the scope of developing a mini hydropower station at Sahasradhara waterfall in the Sitakunda upazila in the district of Chittagong, Bangladesh. In this research work, we are mainly concerned with providing an estimation of the output power to establish a mini hydroelectric power plant at Sahasradhara waterfall. The float method is used to measure flow rate and output power. From our survey it is found that Sahasradhara waterfall would be sustainable for a mini hydroelectric power plant with its head and available flow rate.

Keywords—Mini hydro; power generation; renewable energy; rural electrification; Bangladesh.

I. INTRODUCTION

Renewable energy is generally defined as energy that comes from resources which are naturally replenished on a human timescale such as sunlight, wind, rain, tides, waves and geothermal heat [1]. Renewable energy replaces conventional fuels in four distinct areas: electricity generation, hot water space heating, motor fuels, and rural (off-grid) energy services [2]. Rapid deployment of renewable energy and energy efficiency is resulting in significant energy security, climate change mitigation, and economic benefits [3]. At the national level, at least 30 nations around the world already have renewable energy contributing more than 20 percent of energy supply. National renewable energy markets are projected to continue to grow strongly in the coming decade and beyond [4]. Renewable energy provides 21.7% of electricity generation worldwide as of 2013 [5].

Hydroelectricity is the most widely used form of renewable energy, accounting for 16 percent of global electricity generation- 3,427 terawatt-hours of electricity production in 2010 [6] and is expected to increase about 3.1% each year for the next 25 years. International institutions such as the World Bank view hydropower as a means for economic development without adding substantial amounts of carbon to the atmosphere [7]. Small hydro is the development of hydroelectric power on a scale serving a small community or industrial plant. The definition of a small hydro project varies but a generating capacity of up to 10 megawatts (MW) is generally accepted as the upper limit of what can be termed as small hydro. Small hydro can be further subdivided into mini hydro, usually defined by power generation less than 1,000 kilowatts (kW), and micro hydro which is less than 100 kW.

II. BACKGROUND

The demand for electricity is on the rise in Bangladesh due to growing population and increasing economic activities. But the country cannot conform to the pace of increasing energy demand; therefore there is now a lack of adequate power generation capability. She is facing energy crisis in suburban and coastal areas, especially in hilly areas. The per capita energy consumption in the country is 252 kWh, which is very low compared to other developing countries [8]. The energy production in Bangladesh is mainly dependent on fossil fuels but the resources of fossil fuels are limited and its usage result in global warming [9]. Therefore to encounter the growing energy demand, we have to focus on renewable energy resources and these resources are considered as a promising solution for improving the quality of life of people [10].

A large hydroelectric plant was built in the Kaptai region using the Karnaphuli River during the 1960s, and was Bangladesh's only hydroelectric power plant for a long time. It is capable of producing a total of 230 MW of electricity [11]. Apart from Karnaphuli power station, Kaptai, two other prospective sites for hydropower generation at Sangu (140MW) and Matamuhuri River (75MW) are identified by Bangladesh Power Development Board (BPDB). Currently, there is a micro hydropower plant, with a power generation capacity of 50 kW, at Barkal area of Rangamati district implemented by BPDB [11].

III. MINI HYDROPOWER GENERATION

Flowing water creates energy that can be captured and turned into electricity which is called hydroelectric power or hydropower. There are many sizes of hydropower plant. Table I outlines the categories used to define hydropower plant according to output power.

TABLE I. CLASSIFICATION OF HYDROPOWER PLANT ACCORDING TO SIZE [12],[13]

Pico Hydro	From a few hundred watts up to 5 kW
Micro Hydro	Capacity up to 100 kW
Mini Hydro	Above 100kW but below 1MW
Small Hydro	Capacity above 1-15 MW; usually feeding into a grid
Medium Hydro	15 – 100 MW; usually feeding a grid
Large Hydro	More than 100 MW; usually feeding into a large electricity grid

A. Shahasradhara Waterfal in Chittagong District

Sitakunda is an upazila, or administrative unit, in the Chittagong district of Bangladesh. Sitakunda upazila occupies an area of 483.97 square kilometers (186.86 square miles) [14] which includes 61.61 square kilometers (23.79 square miles) of forest [15]. Part of Sitakunda is covered by the low hill ranges, while the rest is in the Bengal flood plain [16]. There are two waterfalls in the hills: Sahasradhara and Suptadhara [17]. Considering water flow, locality orientation, approachability, and security, we have chosen the Sahasradhara waterfall for study. The abridged portrayal of the site is shown in Table II.

TABLE II. SUMMARIZED DESCRIPTION OF SHAHASRADHARA WATERFALL

Location	Sitakunda Ecopark, Sitakunda, Chittagong
Number of waterfall in the area	Two
Villages in the vicinity	5 villages within 6km ² range of waterfall
Population	Low population density, total 150 houses
Similar existing project in the vicinity	None

B. Flow Measurement

All hydroelectric power generation depends on falling water. This makes hydropower extremely site dependent. First of all, a sufficient and dependable stream flow is required. Secondly, a reasonable height of fall of water, called the head, is necessary for power generation.

Basically, the cross section of an unobstructed area of the stream is measured and a weighted float such as a bottle weighted with pebbles is timed as it floats down a 30.48 meter (100 feet) course. The float method is illustrated in Fig. 1.

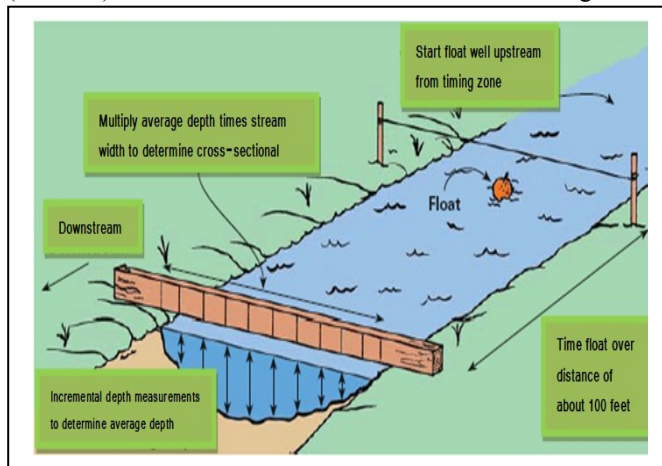


Figure 1. Flow measurement using float method.

C. Head Measurement

The gross head is the vertical distance that the water falls through in giving up its potential energy. The head measurement technique is depicted in Fig. 2.



Figure 2. Head measurement

D. Power Calculations

A hydropower resource can be evaluated by its available power. Power is a function of the hydraulic head and rate of fluid flow. The power available in falling water can be calculated from the flow rate, density of water, the height of fall, and the local acceleration due to gravity. We have measured the flow rate of water using float method. In SI units, the power is:

$$P = (Q) \cdot (H) \cdot (\rho) \cdot (g) \cdot (\eta) \quad (1)$$

Where P is power generation (kW), Q is stream flow (m³/s), H is the effective head (m), ρ is the density of water (kg/m³), g is acceleration due to gravity (m/s²), and η is a system efficiency factor [18].

Again,

$$Q = 0.83 \cdot (b \cdot d) \cdot (30.48/t) \text{ m}^3/\text{sec} \quad (2)$$

Where b is stream width in meter, d is average stream depth in meter and t is time in second for float to drift 30.48 meter (100 feet).

IV. DATA COLLECTION AND ANALYSIS

The climate in Bangladesh maintain three major seasons: summer, rainy season and winter. The data is accumulated over a period of time which covers the three seasons. Tables III, IV and V depict the collected data for these three seasons in the observed period.

TABLE III. DATA FOR SUMMER SEASON

Observation No.	Time	Duration, t (sec)	Average Stream Depth, d (m)	Head, H (m)	Stream Width, b (m)	Discharge, Q (m ³ /sec)	Power For 50% η (kW)
1	10.30am	1.74	0.076	31	3.048	3.37	512.43
2	11.00am	1.80	0.076	31	3.048	3.26	495.70
3	11.30am	1.82	0.076	31	3.048	3.22	489.62
4	12.00pm	1.77	0.076	31	3.048	3.31	503.30
5	12.30pm	1.79	0.076	31	3.048	3.27	497.22
6	13.00pm	1.75	0.076	31	3.048	3.35	509.38
7	13.30pm	1.81	0.076	31	3.048	3.24	492.66
8	14.00pm	1.83	0.076	31	3.048	3.20	486.58
9	14.30pm	1.78	0.076	31	3.048	3.29	500.26
10	15.00pm	1.76	0.076	31	3.048	3.33	506.34

TABLE IV. DATA FOR RAINY SEASON

Observation No.	Time	Duration, t (sec)	Average Stream Depth, d (m)	Head, H (m)	Stream Width, b (m)	Discharge, Q (m ³ /sec)	Power For 50% η (kW)
1	10.30am	1.25	0.082	31	3.201	5.31	807.41
2	11.00am	1.31	0.082	31	3.201	5.07	770.92
3	11.30am	1.30	0.082	31	3.201	5.11	777.00
4	12.00pm	1.33	0.082	31	3.201	4.99	758.75
5	12.30pm	1.29	0.082	31	3.201	5.15	783.08
6	13.00pm	1.27	0.082	31	3.201	5.23	795.25
7	13.30pm	1.32	0.082	31	3.201	5.03	764.84
8	14.00pm	1.34	0.082	31	3.201	4.96	754.19
9	14.30pm	1.28	0.082	31	3.201	5.19	789.17
10	15.00pm	1.26	0.082	31	3.201	5.27	801.33

TABLE V. DATA FOR WINTER SEASON

Observation No.	Time	Duration, t (sec)	Average Stream Depth, d (m)	Head, H (m)	Stream Width, b (m)	Discharge, Q (m ³ /sec)	Power For 50% η (kW)
1	10.30am	1.72	0.073	31	2.972	3.19	485.06
2	11.00am	1.75	0.073	31	2.972	3.14	477.45
3	11.30am	1.69	0.073	31	2.972	3.25	494.18
4	12.00pm	1.67	0.073	31	2.972	3.29	500.26
5	12.30pm	1.64	0.073	31	2.972	3.35	509.38
6	13.00pm	1.67	0.073	31	2.972	3.29	500.26
7	13.30pm	1.66	0.073	31	2.972	3.31	503.30
8	14.00pm	1.68	0.073	31	2.972	3.27	497.22
9	14.30pm	1.74	0.073	31	2.972	3.15	478.97
10	15.00pm	1.71	0.073	31	2.972	3.21	488.09

Fig. 3 indicates the measured flow rate of the waterfall for three different seasons throughout a year. The water flow is almost the same in a day, but it varies from season to season. The water flow rises from summer and reaches its top in the

rainy season. Then it starts to decrease again and goes down to the minimum value in winter. It is evident from the figure that the lowest flow rate is 3.14 m³/sec in winter season and the highest is 5.31 m³/sec in rainy season.

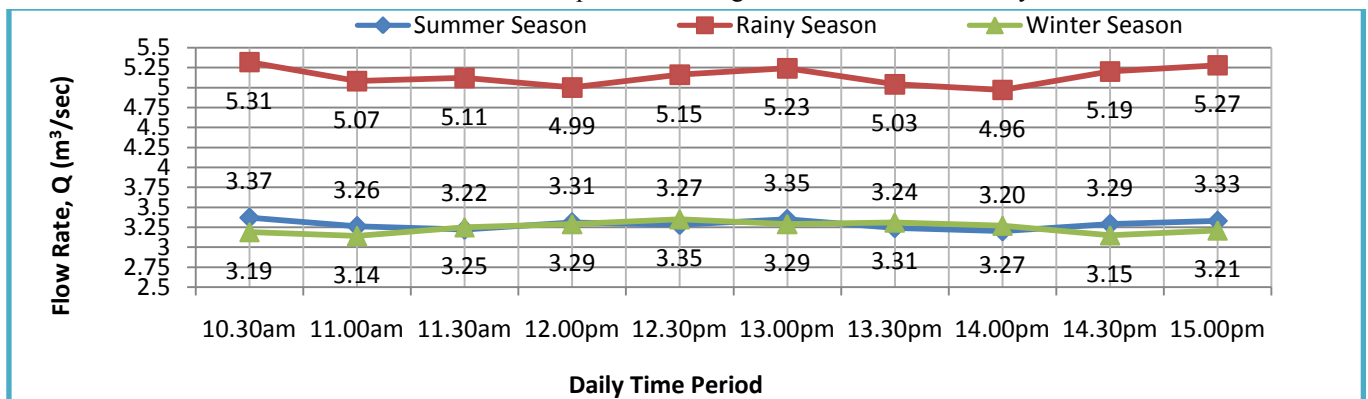


Figure 3. Flow rate of water in three different seasons.

Fig. 4 illustrates the relation between estimated power generation capability and the flow of water for the three seasons. It is apparent from the figure that the generated power is almost proportional to the flow rate. As observed from the figure, maximum power is generated in the rainy season which is about 807.41W. It is due to the abundance of water in this

season. But as the water level declines in summer and winter seasons, the total power generation decreases noticeably and reaches to 41% of its highest value. However, the calculated powers for summer and winter seasons are very close to each other.

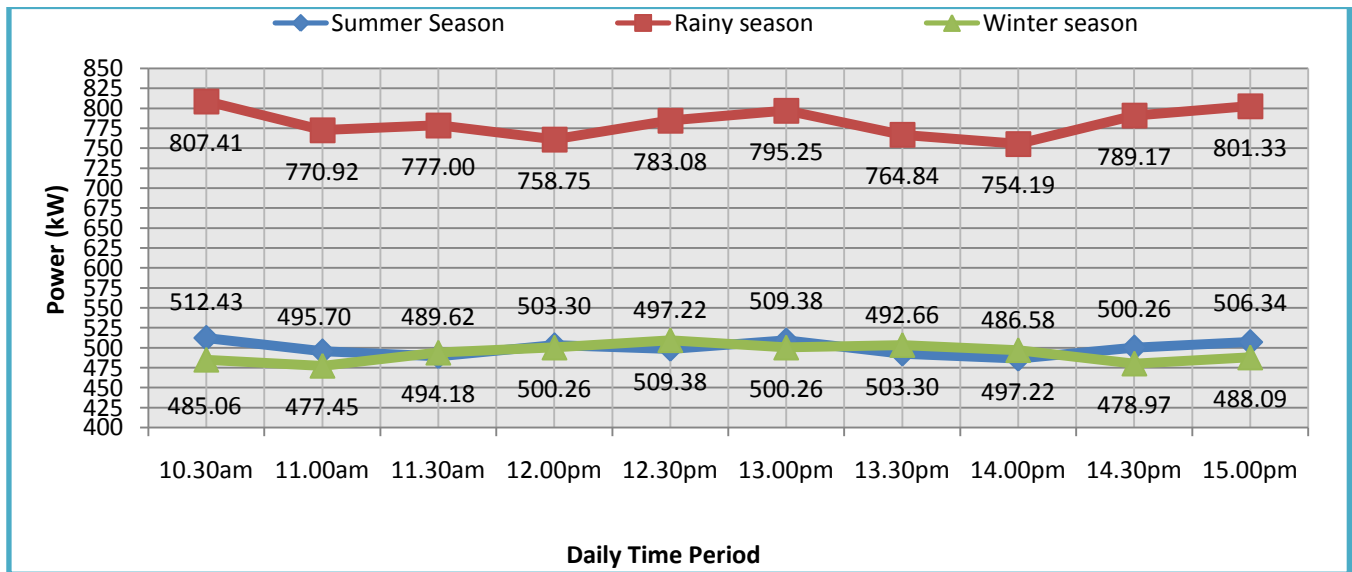


Figure 4. Power generation in three different seasons.

V. CONCLUSION

Hydroelectric power plant is a reliable, economical, long living and eco-friendly way to produce electricity. In this research work, a broad study has been carried out on the feasibility of mini hydroelectric power plant at Sahasradhara waterfall. From our study, we found that Sahasradhara waterfall should be capable of being sustained for mini hydroelectric power plant given its available head and flow rate throughout the year. A cross-flow turbine or a waterwheel will be the best option for the production of electricity with the available head and flow rate in that area. Moreover, if a mini hydroelectric power plant is installed in that area, it will ameliorate the living standards of the local people through new employment for maintenance of the plant, electricity for efficacious cultivation and blooming of ecotourism.

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Designing and Implementation of Small Scale Multi Stage Wind Turbine

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Abstract—This paper presents recommendations for the design and construction of a small multi-stage wind turbine power generation. It addresses the following topics and issues: turbine design, gearbox tooth choosing, generator choosing, implementation for both single and multi-stage wind turbine, data analysis by giving various wind speed. Firstly, a single stage wind turbine was designed and constructed. Then, a multi-stage wind turbine was made. The performance of both turbines was observed and shown in graphs. The results demonstrate that a multi-stage wind turbine can be more effective in terms of high wind speed and cost.

Keywords— bearing; blade; gear; rotor; shaft

I. INTRODUCTION

Renewable energy sources, such a wind energy, deliver power with minimal impact on the environment. Wind energy has great potential to lessen our dependence on traditional resources like oil, gas and coal. One of the drawbacks of wind turbines is that they don't generate anything like as much energy as a conventional power plant. The safe and high performance of these machines is possible only through technological progress in control systems, electronics and their integration its laws of mechanics that govern the behavior of such machines. Energy shortage is a real crisis in developing world. Such deficiency is increasing as fuel price is growing up. By addressing these issues, a lot of work has carried out on the wind energy as a source during the past 30 years [1] [2].

In order to effectively realize future goals for wind energy, the efficiency of wind turbines must increase beyond existing technology. One direct method for achieving increased efficiency is by improving the individual power generation characteristics of horizontal axis wind turbines [3]. The potential for additional improvement by traditional approaches is diminishing rapidly however.

The orientation of the shaft and rotational axis determines the first classification of the wind turbine. A turbine with a shaft mounted horizontally parallel to the ground is known as a horizontal axis wind turbine or (HAWT) [4] [5]. The overall objective was to use more blades in different sets in single wind turbine system [6]. This initiative includes fully instrumented HAWT system. Performance analysis was conducted for two set of wind turbine models. This paper focuses on various system components and their effects on overall output.

II. THE SYSTEM COMPONENTS

Two different wind turbines were designed and implemented. They are single stage and multi-stage wind turbine systems.

A. Single Stage Wind Turbine

In this project, at first a single stage wind turbine was designed and implemented. Fig. 1 shows the basic structure of single stage wind turbine.

A 12V DC generator was chosen for this system. DC generators are more efficient in power conversion. A wind turbine rotates at various speed in different wind speed. So using of DC generator in wind turbine is widely accepted [7] [8]. The DC generator's working voltage was 12V; No load RPM (before gearbox) was 5000 rpm; No load current at 12V was 0.23A, Full load current rating was 3.6A. The generator weighed 270gm.

Although wind speed is the most important factor for rotation, the blades are the main components for converting wind energy into mechanical rotation energy [9] [10]. For this reason the blade design is very important for efficiency.

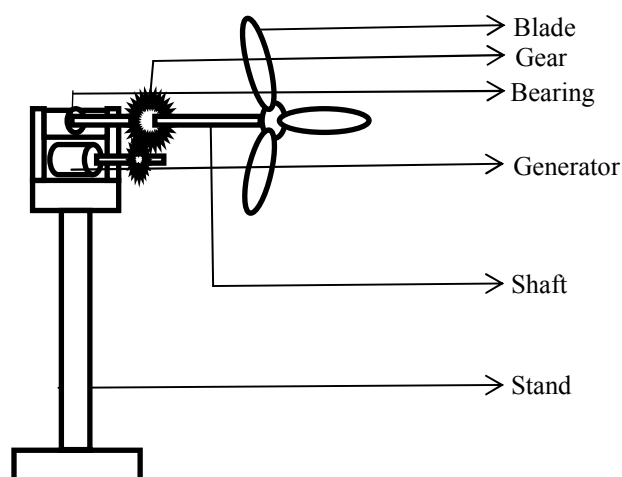


Fig. 1. Schematic diagram of single stage wind turbine.



Fig. 2. Implemented single stage wind turbine.

Different types of blades can be used in this system. As the system was small, a blade set consisting of 3 blades was used. The length of each blade was measured as 15cm and their pitch was 120°. The total blade set weighed 100gm.



Fig. 3. Implemented multi-stage wind turbine.

The coverage area of the blade set was measured as 3113.097 cm². The blades were made of PVC. For transferring power to DC generator a metal shaft was used. The shaft was 20cm long and had a radius of 3mm. Metal shaft was chosen to ensure that it does not bend to carry the weight of blade.

For attaching the shaft with the frame ball bearing was used. The purpose of a ball bearing is to reduce rotational friction and support radial and axial loads. It achieves this by using at least two races to contain the balls and transmit the loads through the balls. Ball bearing can tolerate some misalignment of the linear and outer races. The inner radius of ball bearing was 3mm and the outer radius was 6mm.

To increase the rotational speed of rotor, a gear system was introduced. The system was made of plastic to keep the weight as less as possible. Gears are machine elements used to transmit rotary motion between two shafts, normally with a constant ratio [11]. The pinion is the smallest gear and the larger gear is called the gear wheel. The teeth of gear were involute [12]. Triangle teeth were not used to reduce stress during rotation. The numbers of teeth in primary and secondary gears are 24 and 41 respectively. The resultant gear ratio was 1:1.7.1.

Wind turbines need to be installed higher from ground level. Due to this a wood tower was constructed. In order to guarantee the stability of a wind turbine a pile or flat foundation is used, depending on the consistency of the underlying ground. The foundation anchors the wind turbine to the ground. The tower height is 5ft. A clamp was used for attaching the generator to the frame and a hole for attaching the bearing. The wind turbine was completed by installing the components and connecting it to a load. A 12V battery was used to store the energy. Fig. 2 shows the completed single stage.

B. Multi-stage Wind Turbine

The main difference between a single stage and a multi-stage wind turbine is the number of blade sets. To achieve this a second blade set having same parameters and material of the first one was introduced in the new system. Fig. 3 illustrates the double stage wind turbine.

The position of multi-stage wind blades on a single tower needs to be accurate. If there is any imbalance, the shaft will be tilted causing a gear teeth mismatch. So the weights on two side of the tower need to be equal. The position of shaft on tower was calculated by using the following equation.

$$bw_1 + \left(x * \frac{sw}{l}\right) = (l - x) * \frac{sw}{l} + bw_2 + gw \quad (1)$$

Where bw_1 and bw_2 are the weights of blade set 1 and blade set 2 respectively. sw and l denote the shaft weight and length respectively. The weight of gear system is expressed by gw .

III. PERFORMANCE ANALYSIS

The performances of both turbines were evaluated. An electric fan was used to create desired wind speed in a close room. Then fan speed was regulated so as the wind speed. Different wind speed was measured by using an anemometer by regulating the fan from 50cm distance.

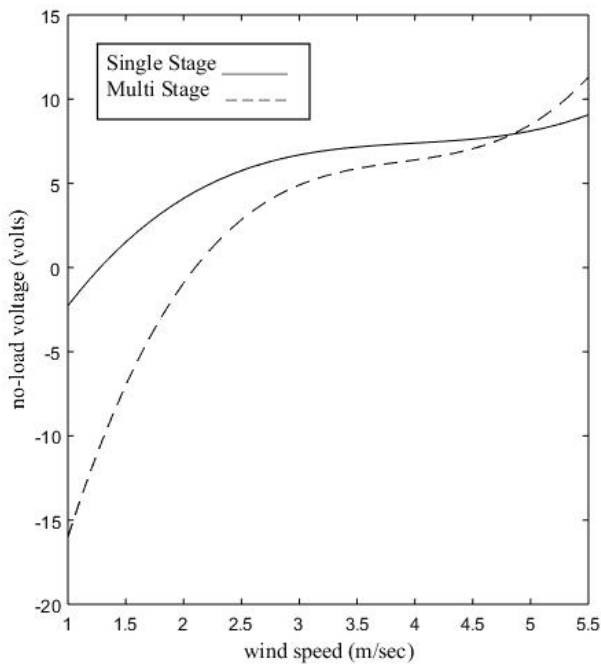


Fig. 4. Wind speed vs. no-load voltage

Then wind flow from fans was provided to the turbines at a fixed 90° angle from the same distance of 50cm. Two types of data were collected at different wind speeds. At first the system was kept in a no-load condition. During this condition, the no-load voltage was collected at different wind speeds. After that a simple electrical circuit consisting of $10k\Omega$ resistor and 1H inductor was connected as a load. As DC power is generated by the system, the DC currents and voltages at loaded condition at various wind speeds were recorded. Then the resultant output power was calculated.

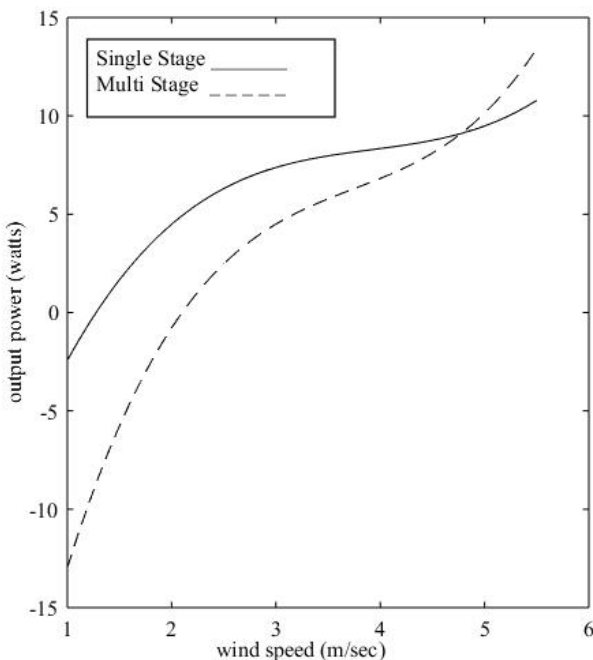


Fig. 5. Wind speed vs. output power

The same procedure was followed for both single stage and multi-stage system. It is to be noted that creating uniform wind speed in a closed room is quite difficult. In case of multi-stage system where we had two set of blades two uniform fans were used to supply wind to two blade sets. Two fans were set exactly 50 cm away at a same angle from two blade sets to ensure accurate wind velocity.

The wind speed generated from a single fan tends to reduce after certain distance. This is a limitation of using a fan for producing same wind speed at different distances. This is why two similar fans were used to supply equal wind speed from equal distance for two fans. It is also to be noted that output tends to change when wind vector is changed. It is found that maximum wind speed is achieved when wind vector is 90° (tangent) to the blade.

The collected data were used to create plots using MATLAB software. A third degree polynomial was used to fit the curves to data points. Fig. 4 shows the no-load voltage at different wind speed for two systems. This graph clearly shows at lower wind speed the no-load voltage for single stage system is better. In fact no voltage was recorded before 2.1m/sec wind speed for multi-stage system. This has occurred due to weight of two blades is more than a single one. So multi-stage system requires more torque to start. Once started, as wind speed is increased more than 4.6 m/sec the multi-stage system shows prospects of producing more voltage. Two blades catch more wind and can rotate the rotor at a higher speed. As a result we get more voltage.

Fig. 5 illustrates the output DC power at various wind speed for two systems. Here the output power of multi-stage system becomes better than single stage one when more than 4.8m/sec wind speed is applied. Although the output power in these systems seems to be low, but this is because of small blade size. A larger set of blades can definitely produce more power.

In addition, multi-stage wind turbine produced a more stable voltage output when wind speed was changed rapidly. This certainly proves that due to high inertia, multi-stage turbines can maintain voltage stability better than single stage one.

IV. CONCLUSION

Installing a multi-stage wind turbine can be a viable option at high wind speed areas. Installation cost is the main cost of any wind turbine system. Multi-stage turbines can be installed within the same area as single stage one requires. So it can save a lot of cost where land availability is limited. This system may not be suitable for locations like Bangladesh where wind speed is lower than 5m/sec but it can certainly work in areas where higher wind speed is available.

The main reason to design this system was to compare between single and multi-stage wind turbine. After a successful analysis, it was observed that a multi-stage wind turbine performs better than a single-stage wind turbine at higher wind speed in terms of voltage, power and stability.

The system implemented was a small one. But this prototype can work as a sample and can be helpful to design a

bigger one. Besides, small scale power generation can be a vital solution to the remote locality where people are deprived of regular electricity supply and can utilize this type of system to charge their small appliances like mobile, laptop, lights etc. by not spending a lot of money.

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Optimization of Unit Commitment Problem for Wind-Thermal Generation using Fuzzy Optimization Technique

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Abstract— The day ahead unit commitment (UC) problem is one of the biggest challenges of power system operation as it requires not only the demand of fulfilling the system constraints but also in most economical way. Adding uncertainty of renewable energy sources make it more difficult to handle. Because of the uncertain nature of Wind Energy Generator (WEG), some risk factors has been added in the solution process. So, solving UC with plants like WEG has become a vital topic of research in present days. A strategy of solving such problems is presented in this paper with having the objective of minimizing the cost and handling risks introduced due to WEGs. As a specialized formulation technique of uncertain optimization problem, Fuzzy optimization technique has been chosen and the problem has been formulated in Mixed Integer Linear Programming (MILP) technique.

Index Terms— Unit Commitment, Thermal units, Wind Energy Units, Fuzzy optimization Technique.

I. INTRODUCTION

Unit Commitment Problem (UCP) deals with determining the optimal on/off states of generating units over a certain time period. The generating units in operation must meet the demand of that period. All the running generating units need to satisfy their physical constraints such as minimum up/down time, minimum/maximum production capacity, ramp rate etc. The term ‘optimal’ signifies that the combination of online generating units will keep the total operating cost as low as possible while satisfying the generator physical constraints and system constraints. It is critical to derive minimum cost combination of generating units since cost of generation is a vital part of the total operating cost of the power system [1-2]. The solution of UCP involves determining all the feasible combination of ON/OFF states of the generating units and then solving economic dispatch for optimum generation of each running generating unit [3]. Since the starting and shutting down of thermal generating units may requires several hours, the unit commitment is done day ahead or week ahead to provide sufficient spinning reserve at real time [4].

Although there are methods for wind power forecasting, still there exists an average error about 10% in day ahead forecasting of wind power [4]. In case of committing thermal generating unit the emergency reserve is maintained to cover the loss of largest generating unit. But high penetration of wind

power to the system increases the emergency reserve requirement to ensure the reliability of supply. Besides maintaining the upward spinning reserve in traditional UC problems, there is also strong demand for enough downward spinning reserve to accommodate the sudden increase of wind power generation and thus, avoid the forced shutdown of thermal units, particularly at low loads times [4-5]. So the additional constraints due to the incorporation of wind power make the UCP difficult to achieve the optimal solution.

The literature is rich of techniques for the solution of unit commitment problem. UC problem can be solved by Dynamic programming (DP) [6] technique. Although dynamic programming provides flexibility [7], [8], it makes the problem complex and increases computation time with the increase of the dimension of the problem. The Lagrangian relaxation (LR) method [9], [10] can provide faster optimal solution if the dimension of the problem is small. The branch-and-bound (BB) method [11] approximates nonlinear cost function by linear one though having the demerit of increasing the execution time exponentially. A newly developed algorithm namely Direct Search Method (DSM) [7] is used to speed up the process of solution which also ease the evaluation of possible states for different generating units. Different artificial intelligence techniques like genetic algorithm operated improved binary particle swarm optimization (PSO) algorithm [7] has been used to solve UC problem for thermal units integrated with wind and solar energy systems.

We consider a hybrid system consisting of three thermal generating units and one wind power generation unit. Different constraints involving UC will be taken into account like power balance, max-min limit of plants, minimum up/down time, ramp rate, spinning reserve, uncertainty of wind power generation etc. As, the problem will be a Mixed Integer (MI) one with both discrete and continuous variables, we will have to be careful about formulating the model. Mixed Integer Linear Programming (MILP) technique has widely been used in formulating the UC problems because of its high computational efficiency. There are some optimizers available which can deal with the MI problems like CPLEX, GUROBI, MOSEK, XPRESS etc. We used CPLEX as an optimization tool to solve our UCP. Finally, to take care of the uncertainty of wind power properly, we used fuzzy optimization technique

as it creates flexibility in modeling and makes MILP based solution more practical.

II. MODELING

A. Mathematical model

Objective Function:

Minimizing the total Unit Commitment (UC) schedule cost can be provided by :

$$C = \sum_t \sum_n \sum_m (A_{nm} * U_{nm} + B_{nm} * P_{nm}) + \sum_t \sum_n (S_{cost} * S_{tn}) \quad (1)$$

where,

t = Time in hour

n = Number of generator

m = Segment number

B_{nm} , A_{nm} = Coefficient of the cost function

U_{nm} = On/Off status of the generators

P_{nm} = Output power of the generators

S_{cost} = Startup cost

S_{tn} = Status (1/0) which indicates the startup of a generator.

Constraints:

i) Power balance equation:

$$\sum_n \sum_m \sum_t P_{nm} + P_{wind} = P_{demand} + SR \quad (2)$$

where,

Spinning Reserve (SR) = $SR = 0.2 * \sum_n \sum_m \sum_t P_{nm}$

P_{demand} = load requirement

P_{wind} = generated power from wind plant

ii) Thermal limits limit of the thermal units:

$$U_{nm} * \underline{P}_{nm} \leq P_{nm} \leq U_{nm} * \overline{P}_{nm} \quad (3)$$

where,

$\overline{P}_{nm}$ = Maximum limit of the thermal generators

$\underline{P}_{nm}$ = Minimum limit of the thermal generators

iii) Segment switching rule:

$$\text{For thermal units: } 0 \leq \sum_m U_{nm} \leq 1 \quad (4)$$

$$\text{For wind unit: } \sum_m U_{nm} = 1 \quad (5)$$

iv) Ramp Rate

$$\text{Ramp up rate: } U_{tn} * P_{t+1} - U_{tn} * P_t \leq R_{up} \quad (6)$$

$$\text{Ramp down rate: } U_{tn} * P_t - U_{tn} * P_{t+1} \leq R_{down} \quad (7)$$

where,

P_{t+1} = power generation in (t+1)th hour

P_t = power generation in tth hour

R_{up} = ramp up rate

R_{down} = ramp down rate

v) Minimum up time:

$$\text{If } u(i, t) = 1 \text{ and } t_i^{up} < t_i^{up, min} \text{ then } u(i, t+1) = 1 \quad (8)$$

where, $t_i^{up, min}$ = minimum up time.

vi) Minimum down time:

$$\text{If } u(i, t) = 0 \text{ and } t_i^{down} < t_i^{down, min} \text{ then } u(i, t+1) = 0 \quad (9)$$

where, $t_i^{down, min}$ = minimum down time.

B. Modeling of S_{tn}

Evaluating the value of S_{tn} :

$$S_{tn} = \max (U_{tn} - U_{t-1, n}, 0)$$

TABLE I: DECISION MAKING OF STARTUP COST

U_{tn}	$U_{t-1, n}$	$U_{tn} - U_{t-1, n}$	S_{tn}
0	0	0	0
0	1	-1	0
1	0	1	1
1	1	0	0

Table I shows that only the non-zero value of S_{tn} indicates that the start-up cost should be added as it reflects the situation where the unit goes to ON state from OFF state.

C. Fuzzy Model

In this work we have formulated a mixed integer problem. It may be advantageous to allow some relaxation to mixed integer problem to reach to optimal solution [11]. This form of relaxation can be provided by modeling problem as a fuzzy model. Basic concept of fuzzy optimization is presented below. Fuzzy set theory is an effective mathematical tool to solve problems containing fuzziness or uncertainties.

Fuzzy optimization requires conversion of crisp set of objectives and constraints into membership function of fuzzy set. The membership function interprets the original objectives functions and constraints in the fuzzy platform. The value of the each membership function lies in the range between 0 and 1.

Suppose, $\tilde{A}_i(x_i)$ and $\tilde{B}_i(x_i)$ are the two fuzzy sets where $\mu_{A_i}(x_i)$ and $\mu_{B_i}(x_i)$ are their membership function respectively and x_i is the decision variable.

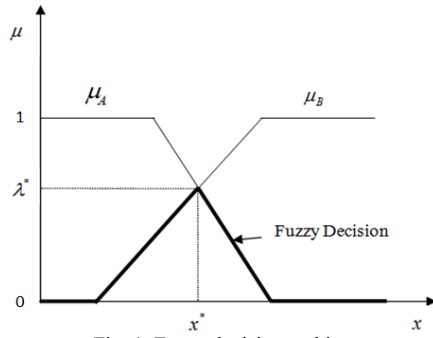


Fig. 1. Fuzzy decision making

$$\begin{aligned}\tilde{A}_i &= \{x_i, \mu_{A_i}(x_i)\} & x_{A_{\min}} \leq x \leq x_{A_{\max}} \\ \mu_{B_i}(x_i) &= \frac{x_{A_{\max}} - x}{x_{A_{\max}} - x_{A_{\min}}} & 0 \leq \mu_{A_i} \leq 1 \\ \tilde{B}_i &= \{x_i, \mu_{B_i}(x_i)\} & x_{B_{\min}} \leq x \leq x_{B_{\max}} \\ \mu_{B_i}(x_i) &= \frac{x_{B_{\max}} - x}{x_{B_{\max}} - x_{B_{\min}}} & 0 \leq \mu_{B_i} \leq 1 \\ \lambda &= \min\{\mu_{A_i}(x_i), \mu_{B_i}(x_i)\} & \text{where, } \lambda \text{ is the} \\ & & \text{confluence/intersection of } \mu_{A_i}(x_i) \text{ and } \mu_{B_i}(x_i)\end{aligned}$$

Maximum value of λ denotes the optimum solution where $\lambda = \lambda^*$ and $x = x^*$

Fuzzy optimization of the problem under consideration
Maximize λ

1. Complete Fuzzy Model of Objective function:

Our objective is to minimize total cost which is given by

$$C(x) = \sum_t \sum_n \sum_m (A_{nm} * U_{nm} + B_{nm} * P_{nm} * U_{nm}) + \sum_t \sum_n (S_{cost} * S_m) \quad (10)$$

Fuzzy set of total cost

$$\begin{aligned}\tilde{C} &= \{[C(x), \mu_C(x)], C_{\min} \leq C(x) \leq C_{\max}\}; \\ 0 &\leq \mu_C \leq 1\end{aligned}$$

Membership function of fuzzy set $\tilde{C}$

$$\mu_C(x) = \frac{C_{\max} - C(x)}{C_{\max} - C_{\min}}$$

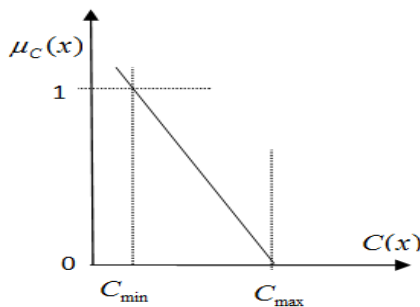


Fig. 2. Fuzzy model of cost minimization objective

2. Fuzzy Model of EENS minimization objective:

Power output from wind generator depends on the hourly wind speed which is highly intermittent in nature. One popular trick to minimize the uncertainty of wind power is to calculate the expected energy not served (EENS) by the wind generator in a particular hour [11]. The EENS for a wind generator is computed from its hourly wind power output and associated cumulative probability and is shown in equation (11).

$$EENS = E(x) = \sum_t \sum_{n \in WG} \sum_m \{(1 - F_{nm}) * (\prod_{r=1}^{WG} F_{tr, \max}) * P_{nm}\} \quad (11)$$

where, F_{nm} = cumulative probability of wind generation corresponding to wind speed

$$F_{tr, \max} = \max(F_{nm}) \forall m$$

WG = Number of wind generator

A high value of EENS indicates the high deviation of wind power output from the forecasted one and corresponds to lower value of cumulative probability values. The minimization of EENS value takes the high cumulative probability values of wind generator which results in lower wind power output. In order to obtain the minimum value the EENS function has been transformed to fuzzy objective function.

$$\tilde{E} = \{[E(x), \mu_E(x)], E_{\min} \leq E(x) \leq E_{\max}\}$$

$$\mu_E(x) = \frac{E_{\max} - E(x)}{E_{\max} - E_{\min}}$$

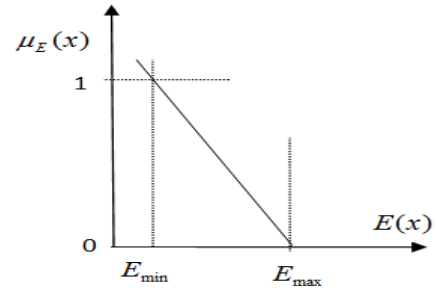


Fig. 3. Fuzzy model of EENS minimization objective

III. IMPLEMENTATION AND RESULTS

This section will give the details about the implementation strategy followed and the results for the test case. Five thermal generators and one wind generators are taken in to consideration in formulating the UC problem. Details specification (max-min limit, up time, down time, ramp up rate, ramp down rate, startup cost etc) of the thermal plants is presented in Table IV. The same table also contains the coefficient values of the cost functions for each of the five thermal units.

Forecasted wind energies are shown in table V. To incorporate the uncertainty of wind energy due to the fluctuations in wind speed, a probability amount of 60% has

been considered. Wind generator has been considered as always ON state.

The UC fuzzy optimization problem has been formulated in MATLAB with the help of a tool namely YALMIP [12] and CPLEX (IBM) has been used as solver. The branch and bound algorithm was effectively used there in order to meet the requirement of convergence. UC problem gives the optimal solution for next 24 time intervals (hours).

Spinning reserve of 20% of the forecasted load has been considered in the problem and is shown in table III. As the wind unit has been considered ON for all the hours, optimal on/off scheduling has only been made for thermal units which is shown in table I and figure V. All the detail results for thermal units are presented in table II which clearly reflects that the combined generation of the plants in each hour meets the required load demand with wind energy, reserve energy and not served wind energy taken in to consideration. Because of the condition of minimum down time of 5 hours, unit 4 while staying in OFF state at 12th hour, remains there for at least 5 hours.

As the main requirement of using Fuzzy optimization, the value of EENS has been minimized in each hour which reflects that the wind plant will be able to provide the forecasted power to be delivered by that. This reduces the uncertainty of using WEG as well. After the end of 24 hour, cumulative amount of energy which has not been served is only 0.36082MW which clearly reflects the success of optimization process. In this simulation, the value of maximum generation cost has been determined in deterministic way and it was 283660\$ and minimum cost was assumed 100\$. Total generating cost obtained from simulation is 256319.7159\$ which is reasonably lower than the deterministic value of cost that indicates the success of using Fuzzy in the given UC problem solving.

As described earlier in the theory portion, the value of fuzzy satisfaction factor away from 0 reflects the accuracy of solving the optimization problem. This particular UC problem gives the value of lambda (λ) closer to 1. This is a clear evidence of the success of using Fuzzy in minimizing the uncertainty due to WEG.

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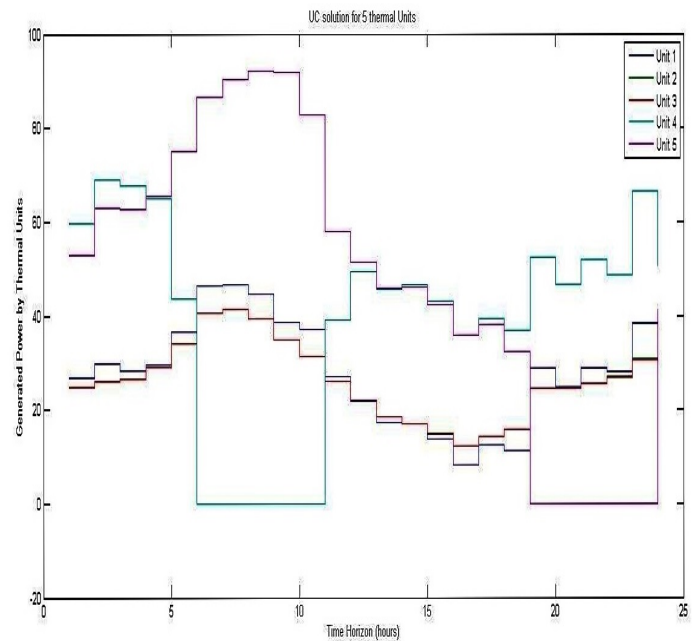


Fig. 4. Graphical representation of scheduling pattern of thermal units

TABLE II: SCHEDULING PATTERN OF THERMAL UNITS

Hours	Unit 1	Unit 2	Unit 3	Unit 4	Unit 5
1	1	1	1	1	1
2	1	1	1	1	1
3	1	1	1	1	1
4	1	1	1	1	1
5	1	1	0	1	1
6	1	1	1	1	1
7	1	1	1	1	1
8	1	1	1	1	1
9	1	1	1	1	1
10	1	1	1	1	1
11	1	1	1	1	1
12	1	1	0	0	1
13	1	1	1	0	1
14	1	0	0	0	1
15	1	1	1	0	1
16	1	1	1	0	1
17	1	1	1	0	1
18	1	1	1	0	1
19	1	1	1	0	1
20	1	1	1	0	1
21	1	1	1	0	1
22	1	1	1	0	1
23	1	1	1	1	1
24	1	1	1	1	1

TABLE III: HOURLY GENERATION OF THERMAL UNITS AND HOURLY EENS OF WIND GENERATOR

Hours	EENS (MW)	Power Generation (MW)				
		Unit 1	Unit 2	Unit 3	Unit 4	Unit 5
1	0.0399	3.7071	8.6739	8.7014	73.1214	94.5961
2	0.0129	5.6807	10.1795	12.0369	86.0474	99.9584
3	0.0118	4.1433	9.1220	10.1816	88.1068	100.4782
4	0.0092	5.2504	11.1583	11.8148	89.2171	100.7086
5	0.0126	15.7979	15.3758	0	91.8597	100.7091
6	0.0195	5.5244	11.9277	12.0597	87.6385	97.1692
7	0.0130	6.7515	11.6154	16.6357	89.0778	95.9226
8	0.0132	6.6470	11.8234	16.1400	85.5942	95.1086
9	0.0178	5.5136	9.8169	11.2154	79.2163	94.8055
10	0.0145	3.5412	9.1078	8.5842	65.9307	95.2107
11	0.0166	5.4999	16.6388	7.9479	48.7834	97.5466
12	0.0138	39.7915	21.6130	0	0	105.4492
13	0.0144	17.0130	14.3848	12.4016	0	102.1150
14	0.0131	36.8468	0	0	0	107.2162
15	0.0196	6.7504	10.3973	14.2456	0	97.6564
16	0.0124	1.6700	5.7288	6.6350	0	90.3786
17	0.0160	4.5354	8.7678	11.7266	0	93.3862
18	0.0092	2.9866	7.1921	8.9980	0	92.8325
19	0.0157	6.3392	12.4955	16.1543	0	95.2066
20	0.0113	4.0314	8.8199	12.2019	0	95.3581
21	0.0158	5.7293	12.2075	16.3652	0	97.4338
22	0.0128	5.7260	11.7827	15.4889	0	97.4852
23	0.0117	3.3929	8.3309	9.0668	49.7226	95.8585
24	0.0141	3.4691	8.4218	8.3378	68.4393	94.3560

IV. CONCLUSION

This work presents a solution technique of unit commitment problem using fuzzy MILP formulation. Presence of a wind generating plant along with the thermal units made the problem more challenging than normal. All the unit and system constraints were met during operation. Start-up cost is added with normal running cost by applying particular mathematical logics.

From the usual objective of minimizing the cost, a fuzzy model of objective was generated where attempt of reducing the uncertainty of wind unit was also incorporated. The solution using a MILP solver is presented where reasonable value of fuzzy membership function represents the success of the work. Test results represented also reflects the efficiency and robustness of this method.

APPENDIX

The system data used in this work has been provided in Tables IV-V

TABLE IV: THERMAL GENERATOR SPECIFICATIONS

Unit	P_{min} (MW)	P_{max} (MW)	Minimum Up Time (Hrs)	Minimum Down Time (Hrs)	Ramp Up Rate (MW)	Ramp Down Rate (MW)	Initial Condition	Coefficients of Cost Function		
								a	b	c
1	0.01	50	1	1	50	50	0	0	80	0
2	5	45	1	1	15	15	0	85.509	70.85831	0.18819
3	5	45	1	1	15	15	0	82.342	68.23392	0.18122
4	25	100	5	5	50	50	0	32.9875	64.415	0.0416
5	25	100	5	5	50	50	0	32.9875	57.915	0.0416

TABLE V: HOURLY FORECASTED LOAD, WIND POWER AND SPINNING RESERVE

Hours	1	2	3	4	5	6	7	8	9	10	11	12
Load Forecast (MW)	173	185	196	204	209	210	209	204	196	185	173	160
Wind Power Forecast (MW)	18.84	8.11	23.18	26.66	27.07	37.7	30.81	29.5	34.65	39.64	31.2	25.16
Spinning Reserve (MW)	34.6	37	39.2	40.8	41.8	42	41.8	40.8	39.2	37	34.6	32
Hours	13	14	15	16	17	18	19	20	21	22	23	24
Load Forecast (MW)	148	135	125	117	112	110	112	117	125	135	148	160
Wind Power Forecast (MW)	31.7	17.95	20.97	36	16	20	4.22	20	18.28	31.53	11.24	8.99
Spinning Reserve (MW)	29.6	27	25	23.4	22.4	22	22.4	23.4	25	27	29.6	32

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Design of a Single Chip Digital Weighing Machine Using ATmega32 Microcontroller Architecture

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Abstract – Now-a-days, the sales centres are using Digital Weighing Machines (DWM) for precision measurement of weight, acquisition of rate from keypad and computation of cost. The weight, rate and cost are shown on 7-segment bright display devices. The implementation of a DWM can be carried out using various types of hardware topologies and accordingly the cost varies. The FPGA based DWM is the most competitive one; but, the technology is beyond affordability of the small business houses due to very high initial capital investment for the design and fabrication of the FPGA chip. It is possible to realize a DWM of moderate cost using an ATmega32 microcontroller, which provides almost a single chip design particularly owing to its high-power (20mA) 32-IO lines, which can directly drive 7-segment display devices, ADC, and Keypad. In this paper, we have presented the design methodology of an ATmega32 based DWM, which has been prototyped and tested using RMCKIT.

Keywords – DWM, ATmega32, FPGA, microcontroller, load cell, HX710B serial ADC.

I. INTRODUCTION

With the advent of low-cost microcontrollers, high resolution serial ADC [1] and precision load cell [2], the mechanical weighing scales are being replaced very rapidly by electronic digital weighing machines (DWM). A scale just measures the weight of the goods; but, a machine measures weight, accepts rate from keypad, and computes the cost.

There are many brands of DWM in the market, which have been designed using FPGAs. The user manual just covers the operational procedures of the DWM; there is no information on the technical design aspects. Little information are available in the literatures except [3, 4], which have discussed designs using 89S52 and 8086 architectures with academic interest only. This paper has presented a practically realizable design methodology of a DWM using ATmega32 microcontroller (MCU) with a view to cover both the academic and practical interests. Fig. 1 depicts the hypothetical pictorial view of an ATmega32 based DWM.

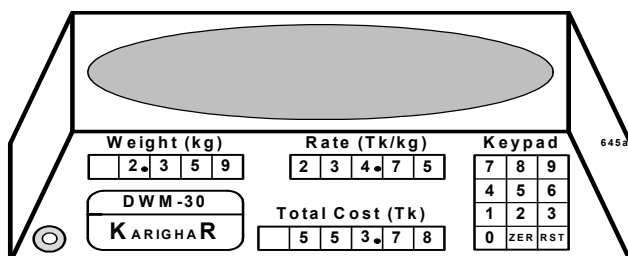


Fig. 1 Hypothetical pictorial view of the Digital Weighing Machine

II. BRIEF FUNCTIONAL DESCRIPTION

At power up and without any load on the pane, the Rate Field shows 0.00, Weight Field shows 0.000 and Cost Field shows 0.00. In case, the weight field shows any idle weight, the user may press the ZERO button to nullify it and bring the weight field showing 0.000.

The Control Program (CP) allows the user entering the rate digit-by-digit in the 'Right Entry Left Shift' type display unit of the 'Rate Field (Fig. 2)'. A digit always appears at the right-most position of the display. At each entry of the digits of the rate, the CP takes the 'reading of rate field' as a current rate, receives weight from the interrupt subroutine (Fig. 6), computes cost and then shows them on the display unit. The weight of the goods is sampled at every 2-sec interval after being interrupted by the 'Timer-1 Overflow Flag'.

The display is of multiplexed type (Fig. 2) and hence the MCU must refresh it once a while. This has been achieved by careful design of the CP, which periodically refreshes the display (Fig. 6) during its looping in the processing chain.

III. THINGS TO BE DESIGNED AND DEVELOPED

Besides from the mechanical fixtures (Fig. 8) of the DWM, the design involves the development and implementation of the following hardware and software tasks:

- Calibration of the input device against two known weights says 1Kg, 1.5Kg so that the machine measures the unknown weight correctly.
- Acquire rate of the goods interactively from a Keypad.
- Compute cost and display it on 7-segment display unit along with rate and weight.
- Generation of 2-sec Time Tick to acquire weight at every 2-sec interval and then computing the cost based on the entered rate.
- Discrete components based Keypad and associated electronics (Fig. 2) [5].
- 16-digit direct driven multiplexed display unit using CC-type 7-segment display devices.
- Interfacing of the Load cell and 24-bit serial ADC; these have been borrowed from a commercial DWM (Fig. 7, 8).

IV. REQUIREMENTS OF DEVELOPMENT TOOLS

The following tools/resources would be required to design and develop various hardware modules and software routines of the ATmega32 based DWM.

- IBMPC with WINXP Operating System: It offers a platform for developing software and downloading binary codes into the flash of the target MCU.
- AVR Studio 4: Assembler for creating source codes, binary codes for the ATmega architecture.
- RMCKIT [6]: Trainer/Write Kit for Atmega32. It comes with a SCC (Serial Communication Controller), target MCU, edge connectors and breadboards. The SCC receives binary code from the IBMPC and fuses them into the flash of the target MCU. The edge connectors and breadboards provide the facilities for circuit prototyping.
- TOP2005: A commercial ROM programmer to configure the internal Fuse Bits of the ATmega32 microcontroller.

V. HARDWARE/SCHEMATIC DIAGRAM OF THE DIGITAL WEIGHING MACHINE

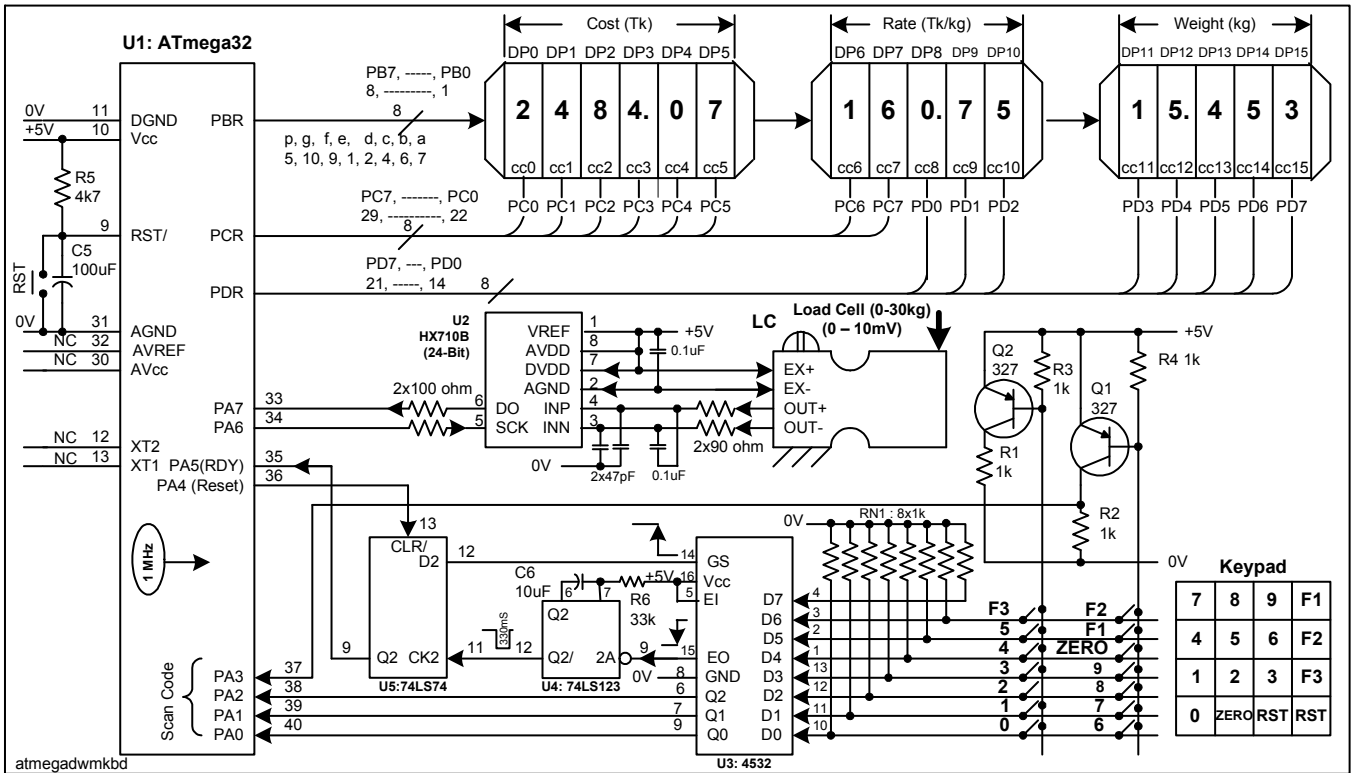


Fig. 2 Hardware/Schematic diagram of the ATmega32 based Digital Weighing Machine

A. Calibration of Input Device

The input device is composed of (i) the pane that holds the goods (Fig. 8), (ii) the holding frame that supports the pane (Fig. 8), (iii) the load cell, (iv) the wiring between the load cell and the ADC, and (v) the ADC. The calibration tasks finds the gain and offset against known weights W_1 (say, 1Kg) and W_2 (say, 1.5Kg) and then deriving equation (1) for the unknown weight (W) in terms of the ADC output. Calibration transforms a 'real system' into an 'ideal system' by imposing gain and offset on the real response (Fig. 3).

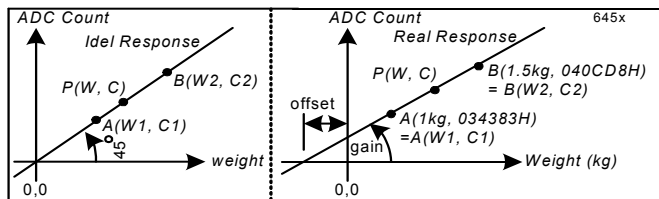


Fig. 3 Ideal and Real Responses of a System

The counts for points A(1Kg, 034383H) and B(1.5Kg, 040CD8H) are the averages of 10 readings each sampled in 2-sec interval. The equation for W based on Fig. 3 is:

$$\begin{aligned} \frac{W_2 - W_1}{C_2 - C_1} &= \frac{W_2 - W}{C_2 - C} \\ \Rightarrow \frac{1.5 - 1}{040CD8H - 034383H} &= \frac{1.5 - W}{040CD8H - C} \\ \Rightarrow 1W &= 1.5 - \frac{0.5 * (040CD8H - C)}{C955H} \\ \Rightarrow W &= 1.5 - \frac{0.5 * (265432 - C)}{51541} \\ \Rightarrow 1.5 - 2.5750 + 9.701015 * 10^{-6} * C \\ \Rightarrow W &= 9.701015 * 10^{-6} * C - 1.075 = 970 * 10^{-8} * C - 1.075 \\ \Rightarrow W * 10^8 &= 970 * C - 1.075 * 10^8 \\ \Rightarrow W * 10^8 &= 03CAH * C - 066851E0H \end{aligned} \quad (1)$$

Validity check of Eqn. (1):

$$\begin{aligned} 10^8 * W &= 03CAH * 040CD8H - 066851E0H \text{ (for 1.5Kg)} \\ \Rightarrow 10^8 * W &= F58AA70H - 066851E0H \\ \Rightarrow 10^8 * W &= 8F05890H = 149969040 = 1.49969040Kg \\ \Rightarrow W &= 1.500 Kg \text{ (after rounding)} \end{aligned}$$

B. Weight Acquisition

Now, we may chalk out the following algorithm to evaluate the Eqn. (1) and display the weight on DP11-DP15 positions of the 7-segment display unit.

```
L1-L1A: Initialize Stack Top by instructions
Initialize the rest as needed: rcallINIT
L1B: Refresh multiplexed display: rcallCCX7SD
L1C: if (2-sec interrupt has not occurred)
    Goto L1B
L1D: Acquire weight, save BINWT and show BCDWT at:
DP11-DP15 positions of display in the format with
rounding xx.xxx : rcall ACQWT
L1E: Acquire next sample : rjmp L1B
```

ACQWT: ; at every 2-sec Time Interval on Interrupt Request

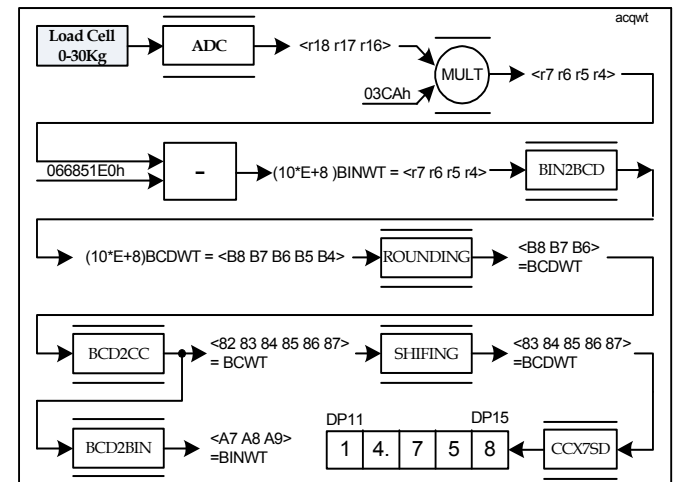


Fig. 4 Data structure for weight acquisition, calibration and display

C. Keyboard Hardware

The schematic of the keypad of the DWM has been borrowed from the literatures of MIDAS Engineering Co. of S. Korea [5]. The interfacing of the keypad and its acquisition software are the sole works of the author of this paper. The working principle of the keypad:

- When a key is pressed down, its 4-bit code called 'Scan Code' goes to the MCU via PA3-PA0 port pins.
- The lower 3-bit of the scan codes comes from the output (Q2-Q0) of the Priority Encoder, U3. The upper 1-bit of the scan codes comes from the emitter point of transistor, Q1. For example: Scan code for key 3 is 0011, for key 4 0100 and the like for other keys.
- The de-bouncing of a pressed down key comes from the $\approx 330\text{ms}$ delay of the oneshot, U4.
- The availability of the jitter free 4-scan codes is announced by LH-state of the Q2 of U5 latch, which is connected with PA5-pin of the MCU. After reading of the scan code, the user resets the keypad by clearing the U5 latch using the PA4-pin of the MCU.

D. Right-Entry Left-Shift Display Unit for Rate Field

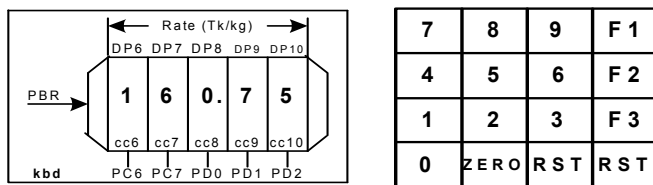


Fig. 5 Right Entry Left Shift Display Unit for Rate Field and the Keypad

In a right-entry left-shift display unit, the digit entered from the keypad always appears at the right-most position of the display. Before the digit is displayed, the digits at the left are shifted to the left by one position with the provision of suppressing the leading zeros and keeping the blanks.

Initially, 'Rate Field' shows 0.00. Now, let us enter a rate of Tk 160.75. We press the digit '1' on the keypad and it appears at the right-most (DP10) position of the display. The rate field display shows 0.01. Next, we press '6'; the display shows 0.16. Next we press '0'; the display shows 1.60. Next we press '7'; the display shows 16.07. Next we press '5'; the display shows 160.75.

After each entry of the digits of the rate, the Control Program takes the reading of the 'Rate Field' as the 'Current Rate' and multiplies it with the 'Weight' and then displays the 'Cost'. Thus, as the digits 1, 6, 0, 7 and 5 are entered one after another, we see the following figures on the 'Rate Field' and 'Cost Field'. Assume that the weight is 15.453 Kg.

Digit Entered	Current Rate	Weight	Cost Displayed
0.00 (initial)	0.00	15.453	0.00
1	0.01	15.453	0.15
6	0.16	15.453	2.47
0	1.60	15.453	24.72
7	16.07	15.453	248.33
5	160.75	15.453	2484.07

E. Control Structures to Acquire Rate

L21: Initialize everything as needed
 L22: Clear U5 (Fig. 2) and then Enable U5 : rcall CCX7SD
 L23: Refresh display
 L24: if (PA5 != LH)
 Goto L22
 L25: Read Scan code from PA3-PA0
 L26: Use Scan code as 'key' and consult LUT2 to collect CC-code for the digit entered from keypad
 L27: Shift digits of Rate Field to left by one position
 L28: Display the digit at DP10 position

L29: Compute BCDRATE from the CC-code data of rate field by consulting LUT3
 L30: Convert BCDRATE to BINRATE

LUT2 : Location Vs CC-code							
Location						Content (CC code)	Digit to Print
Adr	Base	Scan Code					
		PA3	PA2	PA1	PA0		
90	90	0	0	0	0	3F	0
91	90	0	0	0	1	06	1
92	90	0	0	1	0	5B	2
93	90	0	0	1	1	4F	3
94	90	0	1	0	0	66	4
95	90	0	1	0	1	6D	5
98	90	1	0	0	0	7D	6
99	90	1	0	0	1	07	7
9A	90	1	0	1	0	7F	8
9B	90	1	0	1	1	6F	9

LUT3 : Location Vs Unpacked Digit of Form 0X (X = 0 to 9)				
Location			Content (Unpacked BCD)	Digit
Adr	Base	CC-code of Digit of Rate Field		
013F	0100	3F	00	0
0106	0100	06	01	1
015B	0100	5B	02	2
014F	0100	4F	03	3
0166	0100	66	04	4
016D	0100	6D	05	5
017D	0100	7D	06	6
0107	0100	07	07	7
017F	0100	7F	08	8
016F	0100	6F	09	9

VI. CONTROL PROGRAM FLOW CHART

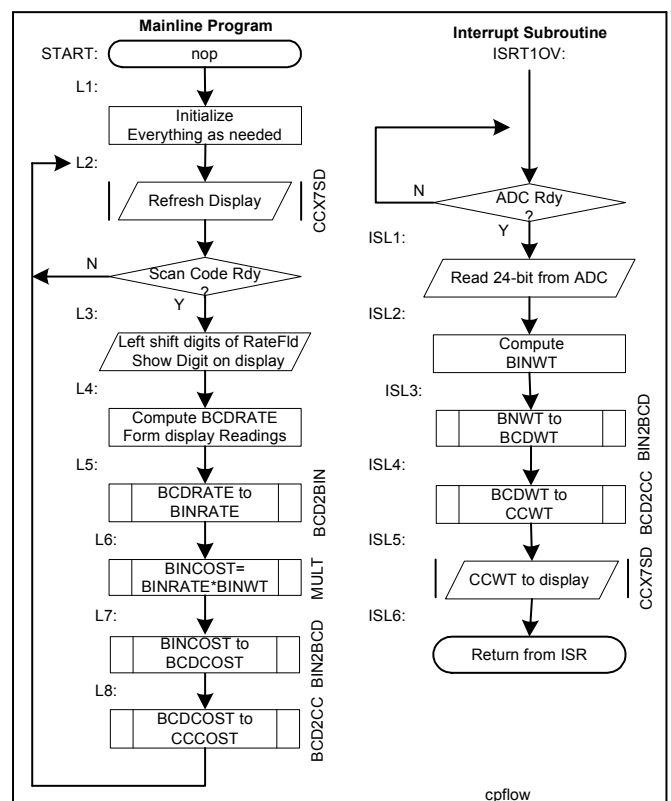


Fig. 6 Flow Chart for the Control Program of DWM (MLP and ISR)

In the Mainline Program (MLP), the MCU regularly, at 2-sec time interval, computes and displays the product cost by taking rate from the Keypad and the weight from the interrupt subroutine.

Timer-1 of ATmega32 has been configured to generate 2-sec Time Tick interrupt. On interrupt, the MCU goes to ISRT1OV (Interrupt Subroutine due Timer-1 Overflow), acquires weight, computes BINWT, and saves it into data structure (Fig. 4) as an input variable to be used by the MLP.

VII. DEVELOPMENT METHODOLOGY

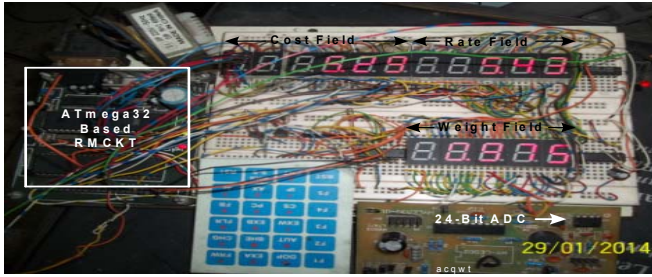


Fig. 7 RMCKIT – ATmega32 Based Trainer/Writer Kit



Fig. 8 'Pane Holding Frame', 'Load Cell' and 'Pane with Load'

To develop the hardware circuits and the software routines of the DWM, we shall follow 'Small Start Strategy (SSS)' approach. In SSS methodology, a basic circuit and its associated software are tested and then another incremental amount is added with the previous one. This process keeps going on until the whole system comes into picture.

The RMCKIT (Fig. 7) provides an excellent platform to develop the DWM. We can build the hardware on the breadboards as per Fig. 2, write the codes for the routines/subroutines using ASM/C in the IBMPC as per Fig. 4-6, create binary codes for the programs, and then transfer the program codes into the flash of the MCU for execution.

A. Data Structure-I

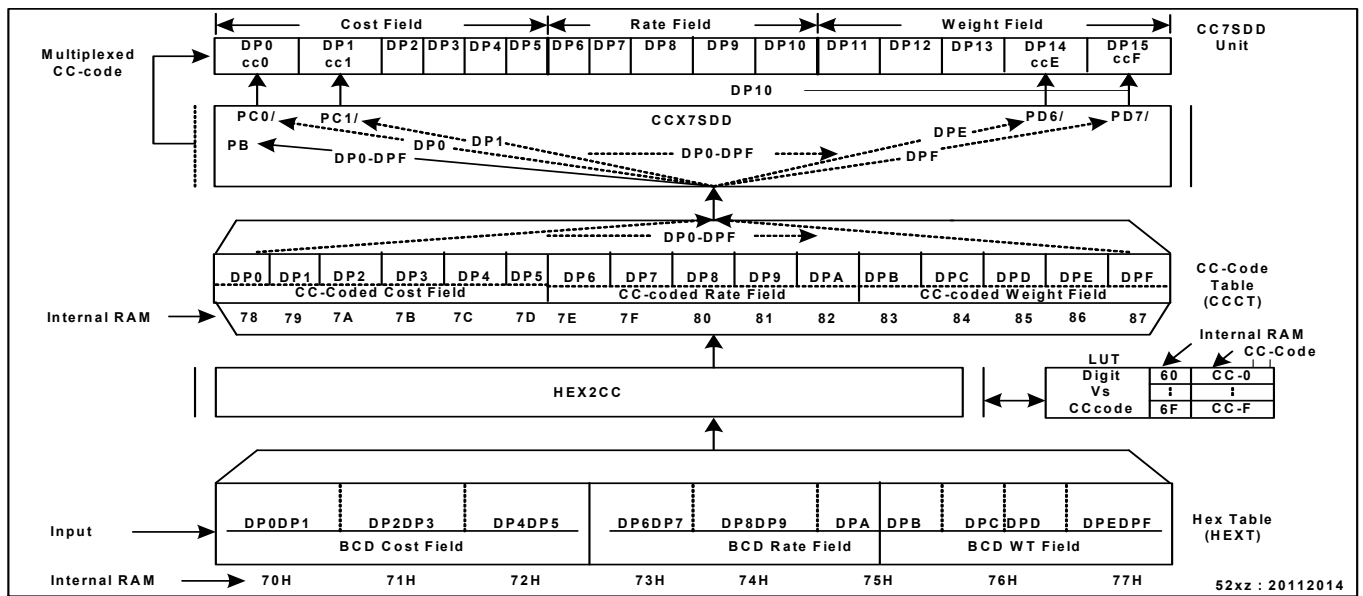


Fig. 9 Data Structure-I for the ATmega32 Based Digital Weighing Machine

B. Data Structure-II

Register File Structure

Register File (32 Symbolic Names)	Data RAM Address
R0	\$0000
R1	\$0001
R26	X
R27	Pointer
R28	Y
R29	Pointer
R30	Z
R31	Pointer

Data RAM Structure

IO Register (64 Symbolic Names (Port Address) Valid with in/out Ins)	IO Register (64 Data RAM Address Valid with lds/sts Ins)
TWBR (\$0000)	\$0020
SREG (\$003F)	\$005F
User Data RAM	
\$0060	LUT1 Digit Vs CCcode
\$006F	LUT2 Scode Vs CCcode
\$0090	System RAM
\$0099	
\$0800	Stack
\$085F	Filled Up

Fig. 10 Data Structure-II of Digital Weighing Machine

C. Binary to BCD Conversion Algorithm using Horner Rule

$$\begin{aligned}
 \text{BIN} &= b_{39}b_{38}b_{37}\dots\dots\dots b_2b_1b_0 \\
 \text{BCD} &= b_{39}2^{39} + b_{38}2^{38} + b_{37}2^{37} + \dots + b_22^2 + b_12^1 + b_02^0 \\
 \text{BCD} &= ((\dots((\text{IPBCD} \times 2 + b_{39}) + b_{38})2 + \dots + b_2)2 + b_1)2 + b_0
 \end{aligned}$$

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Where: IPBCD stands for Initial Partial BCD and is 00.

Control Structure for Computation:

for ($x=39$ to 0)

```

{
    IPBCD * 2 + bx → IPBCD (incorrect BCD)
    Perform BCD adjustment on the incorrect BCD
}

```

VIII. CONCLUSIONS

This project has produced realistic results, which will work as a reference to learn the methodology of system design and the programming of ATmega32 and HX710B. Use of a single ATmega32 MCU for simultaneous handling of the display unit, load cell, and Keypad is a remarkable achievement.

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Development of an 8-Bit Microcontroller Learning System Using 89S52 Architecture

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Abstract – The 89S52 microcontroller (MCU) is very popular among the students, teachers, amateurs, engineers, scientists, and hobbyists for building electronics projects and consumer products. Successful realization of MCU based products depends on the sound understanding of architecture, instruction and interfacing of the MCU, which can be effectively acquired with the help of a ‘Microcontroller Learning System (MLS)’. The MLS provides facilities to enter program codes and data into memory; program execution and debugging; checking and changing memory/register contents; breadboard/edge connector for prototyping interfacing circuits; and viewing the result on display devices. This paper presents the design and development procedures of an 89S52 MLS system along with examples to establish its effectiveness in learning microcontroller hardware and programming. We designed, developed, and tested a prototype MLS system named MicroTalk-8051 Trainer.

Keywords – 89S52, CISC architecture, microcontroller learning system, MLS, MicroTalk-8051, GUI interface.

I. INTRODUCTION

The 8051 is the generic name for the microcontrollers 89C51, 89S51, 89C52, 89S52, 89S8252, and the like. All these MCUs share a common basic architecture and an instruction set. The 8051 is a single chip IC housed in a 40-pin DIP package. It contains hardware resources, which are: 8-bit and 1-bit processors known as Byte Processor and Boolean Processor, Clock Oscillator, Special Function Register Set, Code Memory (Flash) with Security Bit, Data Memory (EEPROM) with Security Bit, RAM, Parallel IO, Asynchronous Serial IO, Counter/Timer, High Speed Serial Peripheral Interface (SPI), In System Programming Interface, and Parallel Programming Interface.

To learn the functionalities of all these resources and their interfacing with IO devices, one requires owning a ‘Microcontroller Learning System’, where he can experiment all kinds of trial-and-error approached hardware interfacing and the software programming. The user may own a learning system through purchase or by making it himself. In this paper, we have documented design methodology of an MLS, which the serious readers may adopt to build their own Trainers and spend times to learn everything (!) of the 8051 microcontroller.

There are many companies [1] around the world, who manufacture and distribute 8051 trainers among the educational and industrial institutions for learning purposes. These trainers accompany user manual, laboratory experiments, and schematics of the hardware but no information at all on the working principle of ‘Monitor Program (MP)’ and its design methodology. MP is a specialized program, which usually resides in the EPROM of the trainer and decodes the keypad commands for code/data entry, program execution, memory/register check and change, and finding the erroneous instruction (s) of a faulty program.

II. RESOURCE REQUIREMENT

- IBMPC with WINXP Operating System.
- MIDE-51 [2]: Assembler for 8051 microcontrollers.
- TOP2005 [3]: ROM Programmer.
- MicroTalk-8085 [4]: Optional item. It provides hardware of the MCU trainer as the peripherals of the 8085 microprocessor are fully compatible with 8051.

III. EXPECTED FEATURES OF THE MCU TRAINER

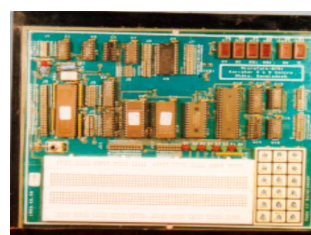
The trainer will have the following hardware and software features to perform experiments for learning the 8051 interfacing and its programming.

A. Hardware Features

- 89S52 MCU with internal Flash disabled.
- 2x8-Kbyte EPROM for holding the codes of the Monitor Program and the Subroutines.
- 8-Kbyte external RAM for loading users’ programs.
- 8279-based 6-digit CC-type display unit.
- 8279-based Keypad with 17 Command/Data keys and 1 Reset key arranged in 6x3 matrix.
- RS232/USB interface for downloading program codes and data from the IBMPC.
- IO connectors and breadboard for circuit interface.

B. Software Features

- EXA Command to open the address of an external memory location.
- CHG Command to change the content of an external memory location.
- FRW/BKW Commands to check next/previous memory and register location.
- DOP Command to execute a program.
- PCR & S/S Command to execute only one instruction.
- GPR Command to check General Purpose Register.
- CSR Command to check Control & Status Registers.
- TCR Command to check Timer & Counter Registers.
- DPR Command to check Data Pointer Register.
- PSR Command to check Processor Word Register.
- SPR Command to check Stack Pointer Register.
- ICR Command to check Interrupt Control Registers.
- Subroutines to convert data & access display/keypad.



D	E	F
DOP 01	EXA 09	FRW 11
A	B	C
AUT 02	BKW 0A	CHG 12
7	8	9
TIR 03	GPR 0B	CSR 13
4	5	6
SPR 04	PCR 0C	DPR 14
1	2	3
RAM 05	ICR 0D	PSR 15
RST	BKS	0
	S/S 0E	PRT 16

Fig. 1 Pictorial view of MicroTalk-8051 with Keypad Layout

IV. HARDWARE BLOCK DIAGRAM AND DESCRIPTION

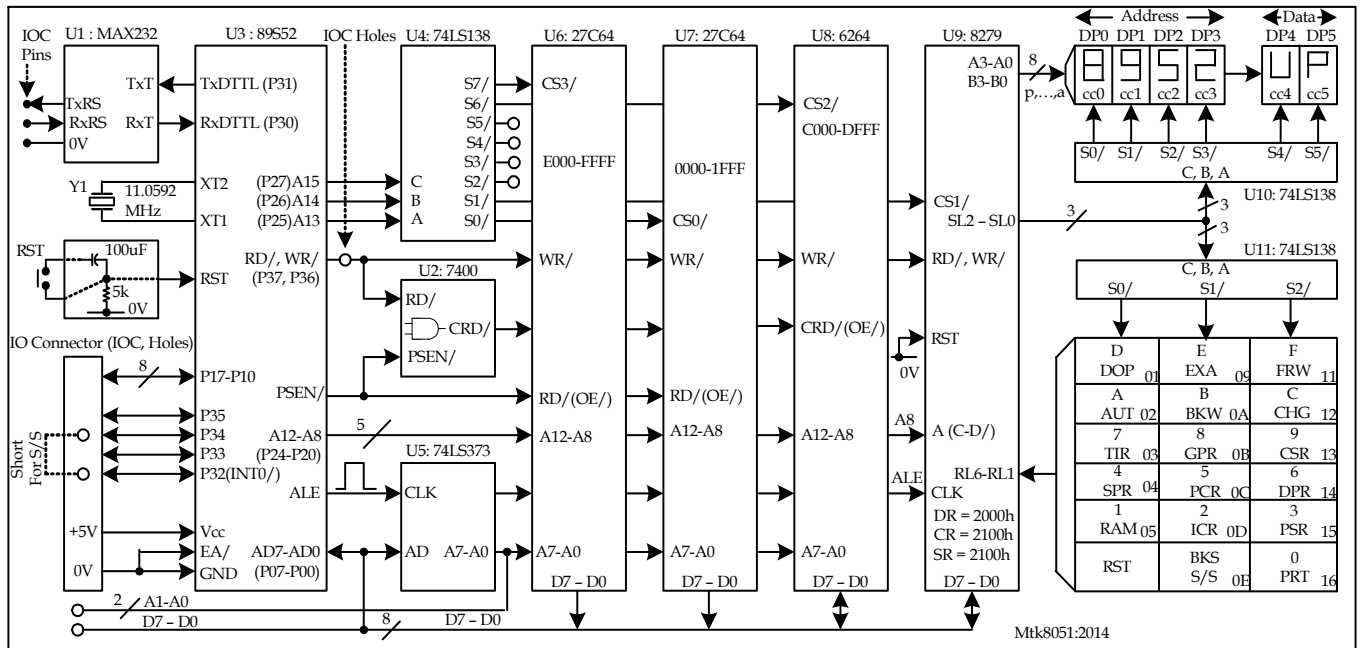


Fig. 2 Hardware Block Diagram for the 89S52 Microcontroller Learning System

A. Overall Functional Description

After power up, the MCU (U2) boots up at location 0000H and performs the necessary initialization tasks. It then keeps polling the Keypad for commands like EXA, or DOP or S/S. EXA command allows a user to enter program code/data into RAM space (C000-DFFF, U8) using Keypad. Memory address and its content appear at the 'Address Field' and 'Data Field' of the display unit. DOP command helps the user executing his application program. There are IO connectors and breadboard in the trainer, which facilitate developing and testing interfacing experiments. The serial communication controller (SCC, U1) establishes link with IBMPC to receive program code/data from the host computer to the RAM space of the trainer.

In the given system, the 8051 is working in the 'bus mode' meaning that it now asserts standard bus signals like address, data and control signals to access external memory (port) locations. The 8051 has 16 address lines (A15-A0), 8 data lines (D7-D0) multiplexed with 8 lower address lines (A7-A0), RD/, WR/ and PSEN/ signals to communicate with memory (port) devices. The 8051 automatically turns into bus mode while it executes instructions like *MOVX* and *MOVC*. In Fig. 2, we have indicated with () signs, which port line, transforms into which bus signal during bus mode operation.

B. Memory and Port Space Map

There are three memory chips and one IO chip in the trainer. To write codes for the monitor and example programs, one must assign addresses to these devices. Linear decoding scheme has assigned 8K space for the three internal registers of U9. We have indicated within the block of U9 of Fig. 2 the specific addresses for the registers of this IO.

FFFF	U6	Monitor Program (EPROM)
E000		
DFFF	U8	RAM
C000		
3FFF	U9	Keyboard Display Controller (Port)
2000		
1FFF	U7	Monitor Program (EPROM)
0030		
002F		
IE0 0003 DFEE		
TF0 000B DFF1		
IE1 0013 DFF4		
TF1 001B DFF7		
RI+TI 0023 DFFA		
TF2 002B DFFD		
0001		
0000	LJMP 0030H	

Fig. 3 Memory and Port Map

C. Linear Decoder U3

In bus mode, the total addressable locations including memory and port of the 8051 are 64Kbyte. In system of Fig. 2, a 3-to-8 linear decoder (U3) is just good enough to allocate spaces for the storage devices of the system. The truth table of the decoder is:

TABLE 1: TRUTH TABLE OF THE LINEAR DECODER, U3

Address Signals			Decoded Space	Connected at
A15	A14	A13		
0	0	0	0000 - 1FFF (8K)	CS0/-pin of U7
0	0	1	2000 - 3FFF (8K)	CS1/-pin of U9
0	1	0	4000 - 5FFF (8K)	CS2/-pin at IOC
0	1	1	6000 - 7FFF (8K)	CS3/-pin at IOC
1	0	0	8000 - 9FFF (8K)	CS4/-pin of IOC
1	0	1	A000 - BFFF (8K)	CS5/-pin of IOC
1	1	0	C000 - DFFF (8K)	CS6/-pin of U8
1	1	1	E000 - FFFF (8K)	CS7/-pin of U6

D. Microcontroller Subsystem

The MCU (U2) operates in multiplexed bus mode along with the availability of important IO operations like Counter, Interrupt, Serial IO, ISP, and Digital IO. U4 has conditioned the RD/ and PSEN/ signal to synthesize CRD/ signal, which allows the MCU to fetch program codes from the RAM chip. The ALE signal clocks the auxiliary latch (U5) to hold the address signals (A7-A0) from the AD7-AD0 bus before the bus turns into D7-D0 data lines. U2 is directly replaceable by the advanced 89S8252 chip, which has internal EEPROM and SPI interface. Bus signals, decoded lines, and IO lines are available at the IO connectors for interfacing purposes.

E. Keyboard/Display Subsystem

The 8279 controller [2] can handle 16 cc-type 7-segment display devices and 8x8 'walking 0' Keypad in multiplexed mode. The controller has built-in scan circuitry to generate timing signals to refresh the display, catch the scan code of a pressed down key. At the closure of a key, the MS-bit of SR assumes Logic-H state and its 8-bit scan codes (numerical values shown in Keypad) enter into keyboard FIFO, which could be read via DR.

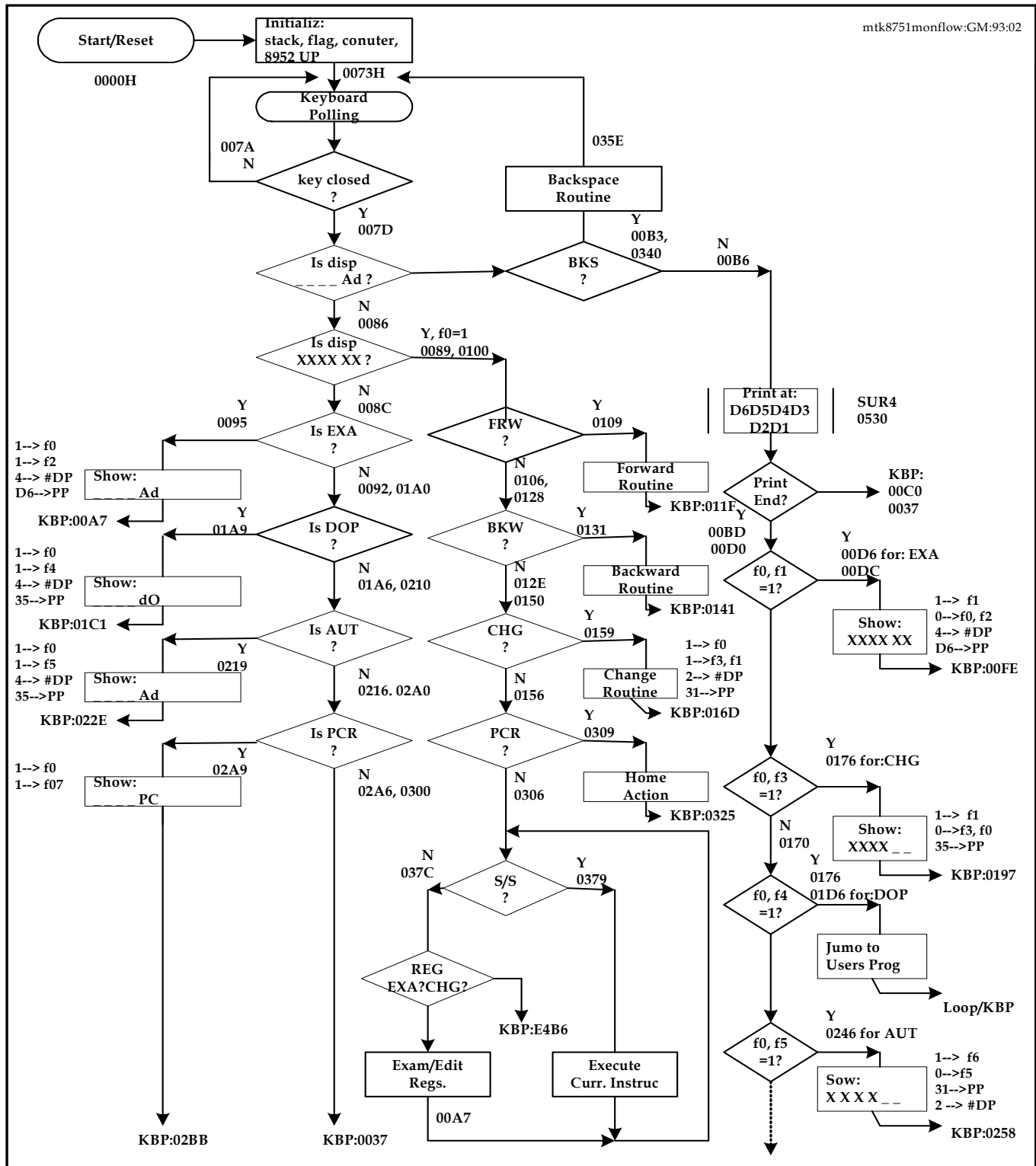


Fig. 4 Flow Chart for the Complete Monitor Program of MicroTalk-8051 Trainer

A Microcontroller Trainer interacts with a user by virtue of its EPROM-resident Monitor Program. The following are the essential components of a MP, which once developed could lead to the easy development of other components:

- Sensing EXA command and responses by showing the message `_____ Ad` in the display unit meaning that the user may now enter 16-bit address (as 4 hex digit) of a RAM location. Assume that the RAM location is C000.
- At the end of address entry, the display shows C000 XX.
- Senses the CHG command to show message C000 __ on

the display. Now, the user may enter 1-byte (2 digits, say D2) program code in the memory location. After code entry, the display becomes C000 D2.

- Sense FRW command and show the next memory location and its content on the display as C001 XX. Now, the user can employ FRW and CHG commands to finish the entry of all the program bytes in memory.
- Sense DOP command to accept the beginning address of the loaded program, execute the program, and show result on the display.

VI. DEVELOPING HARDWARE CIRCUITS

We shall follow ‘Small Start Strategy (SSS)’ approach to develop the hardware circuits and the software routines for the trainer. In SSS methodology, a basic hardware circuit and its associated software codes are developed and tested. After that, we add another incremental amount of hardware and software with the previous one and test them. We will try to develop the hardware on breadboards. However, at some stage of development, the trainer might stop working due to too many jumpers and noise. At this time, either we have to use the ready-made hardware of the MicroTalk-8085 trainer or placing the components on manually PTHed double layered PCB (Fig. 5).

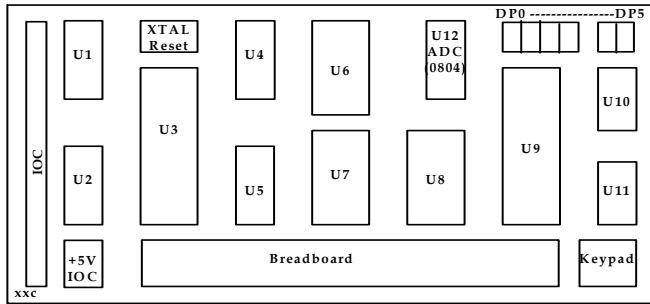


Fig. 5 Components Layout for the Manual PTHed Double Layered PCB

A. Building and Testing the Startup Circuit

- Place the display devices DP0-DP5, U9, U10 on the breadboard and connect them together using jumper wires as per circuit of Fig. 2. We need to consult data sheets of the chips to find the pin numbers.
- Place 89S52 MCU on the breadboard and complete bus connections with the 8279. Connect EA/-pin of the MCU to +5V and CS1/-pin of 8279 to 0V.
- Convert the following Pseudo Codes into Binary codes and fuse them into the flash of the 89S52 MCU.
Listing - A
L11: 0000 → 0030H ; LJMP 0030H
L12: Initialize 8279 for 8-digit, left entry, 2-key lock
L13: Set cursor position at DP0-position with auto index
L14: write characters 8 9 5 2 U P onto display RAM of 8279
L15: Loop HERE ; LJMP HERE
- Place the loaded MCU on breadboard and press Reset key. We will see 8952 UP message on the display.

B. Placing EPROM Memory (U7) with the Trainer

- Place U7, U5, and U3 on breadboard. Connect them with the system and the CS/-pins of the U7 and U9. Connect EA/-pin of MCU to 0V.
- Fuse the codes of Listing - A into the U7 using ROM Programmer and place it back on breadboard. Apply power. Press Reset key. The display will show 8952 UP.

C. Placing RAM (U8) with the Trainer

- Place U8 on breadboard and connect its pin with other IC.
- Place U4 on breadboard and connect its output signal (CRD/) with the RD/-pins of the RAM and EPROM.

D. Placing Keypad with Trainer

- Place U11 and switches on a breadboard.
- Connect the terminals of the switches with the row lines of the 8279 and the column lines of the U11 decoder.

VII. DEVELOPING SOFTWARE ROUTINES

To develop software routines for the MP, as we will experience, there are needs of data structures to store variables and lookup tables for code conversion purposes.

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A. Data Structure and Lookup Tables

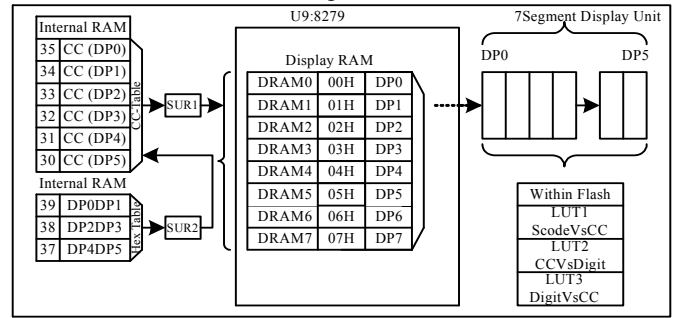


Fig. 6 Data Structure and Lookup Tables

B. Sense EXA Command and Show ___ _ A d on Display

```
Listing - B
L21: If (scan code != 09h)
    Goto L21
L22: LH → flag1
    Write cc-codes for _ s into CC-Table
    Call SUR1 to transfer CC-Table to display RAM of U9
    Enable printing logic for address field
```

C. Forming 16-Bit Address and Showing Location Content

```
Listing - C
L31: Read CC-codes from location 30H-33H of CC-Table
    Consult LUT2 and collect unpacked digits
    Process unpacked digits to form 4-digit hex address, Adr
L32: Pass the address into DPTR register
    Execute MOVX A, @DPTR to collect present data of Adr
    Put Adr and Data into Hex-Table
    Call SUR2 to convert Hex-Table to CC-Table
    Call SUR1 to show CC-Table to display unit via 8279
```

D. Sense CHG Command and Show XXXX __ on Display

```
Listing - D
L41: If (scan code != 12h)
    Goto L41
L42: Write cc-codes for _ s at DP4 & DP5 positions of CC-Table
    Call SUR1 to show CC-Table to display
    Enable printing logic for data field
```

E. Writing 8-Bit Data into Memory Location

```
Listing - E
L51: Read CC-codes from location 30H-35H of CC-Table
    Consult LUT2 and collect unpacked digits
    Process unpacked digits to form 4-digit hex address, Adr
    Process unpacked digits and form 2-digit data
L32: Pass the address into DPTR register
    Execute MOVX @DPTR, A to write present data into RAM
    Put Adr and Data into Hex-Table
    Enable FRW command
```

F. Sense FRW Command and Show Next Location and Data

```
Listing - F
L61: If (scan code != 11h)
    Goto L61
L62: Execute partial codes of Listing - C to form address
    Pass address into DPTR register
    (DPTR) + 01h → DPTR
    A ← ((DPTR))
    Pass DPTR and A into Hex-Table
    Call SUR2 and SUR1 to show address and data on display
```

G. Program Execution

```
Listing - G
L71: if (scan code != 01h)
    Goto L71
L72: Read Adr from Hex-Table and keep in DFDE and DFDD of RAM
    Make a jump to location DFDEH : LJMP DFDEH
```

VIII. CONCLUSIONS

The design methodology of an 89S52 Microcontroller Trainer, which once belonged to the business houses, is now available to the public. The interested readers may build their own Trainers and learn many more things like ‘Single Stepping’, ‘Register Check’, ‘MCU Based Low Cost Industrial Control Circuits’, ‘Consumer Products’, and ‘Creating Effective Lessons for Class Room Lectures.’

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Microcontroller Based Home Automation System Using Bluetooth, GSM, Wi-Fi and DTMF

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Abstract—Automation plays an important role in the global economy. Engineers combat to combine automated devices with mathematical and organizational tools to create complex systems for a rapidly expanding range of applications and human activities. This paper represents a low cost, flexible and standalone home automation system. We can control home appliances from outdoor also from indoor locations. We have used a PIR (Passive Infrared Sensor) in this system that detects an intruder. All Electrical loads are automated in this system. There is also an internet module that is used to notify the status of home appliances using Twitter. The advantage of the internet module is that the home appliances can be controlled from anywhere in the world. The systems also uses the DTMF (Dual-Tone Multi-Frequency) technology for the calling system, so the home appliances can be controlled by calling and pressing buttons of mobile devices. Using this system, physically challenged people can control home appliances from the bed or the wheelchair. This system can be implemented in houses, offices, hospitals, industries, or even in universities. The user interface is kept simple, so users from beginner to advanced level internet users can use the system without problem.

Keywords—microcontroller; gsm modem; PIR sensor; internet module

I. INTRODUCTION

We can imagine how helpful it will be to be able to switch on our air conditioning system ten minutes before we get home on a hot afternoon. When we leave our home for some work without realizing that some appliances such as fans, air conditioners, and lights are on; then by using our mobile phone or internet, we are able to turn off power to those devices. It will be even more useful if the system detects unauthorized movement in the house and alerts us or sends messages on our mobile phones or we can know the status of our house anytime by Twitter. These are some features of home automation and there is no end to its application. In fact, sophisticated home automation systems are now being developed that can maintain an inventory of household items.

The home automation market is primarily driven by growing need for effective solutions in various domestic applications such as lighting, safety and security and energy management. The development of smart cities and high-tech homes in the Middle East and Asia Pacific region is also influencing the growth of overall home automation market. The home automation market is primarily driven by growing need for effective solutions in various domestic applications such as lighting, safety and security, energy management,

entertainment (audio and video) and HVAC (heating, ventilation, and air conditioning) [4]. High degree of internet connectivity and penetration in developed countries such as the U.S., the U.K., France, and Germany is also contributing to the growth of home automation market globally. Safety and security was the highest revenue generating application segment in 2015.

Smartphone based home automation solution is also observing a significant growth in demand. Increasing usage of Smartphone based home automation apps are also contributing to the growth of overall home automation market globally [1]. North America led the home automation market in 2014 and the region is expected to continue its dominance in 2020. Moreover, consumer awareness and affinity towards new technology is further contributing to the growth of home automation market in North America, Europe and Asia Pacific [2].

This market research study analyzes the home automation market on a global level, and provides estimates in terms of revenue (USD million) from 2014 to 2020 [2]. It recognizes the drivers and restraints affecting the industry and analyzes their impact over the forecast period. Moreover, this project identifies the significant opportunities for growth in South Asian market in the coming years.

II. SYSTEM DESCRIPTION

Our designed home automation system is a central system that can control and create communication between nearly all aspects of the house. Our automated home can be called a smart home. If one forgets to switch off the lights or other appliances while going out, it allows you to turn off the appliance with your cell phone. By using twitter, you can get status of your home and also can control your home power devices. This is a simple automation system which contains remote mobile host controller and several home appliances.

We have five features in this project. We implement home automation using Bluetooth, GSM-SMS, GSM-DTMF, PIR (passive infrared sensor), and an internet module for Twitter. At first, we set up Bluetooth then GSM-SMS then GSM-DTMF (Dual-Tone Multi-Frequency) then PIR and the last part is the internet module using twitter.

The main features of this system are:

- Control through android mobiles.
- Control by SMS where device is capable of recognizing the user.

- Through GSM modem, appliances can be controlled by DTMF (Dual-Tone Multi-Frequency) tone. All devices can be controlled using phone calls.
- The user can control appliances through Twitter.
- Another feature is detecting humans or intruders by using motion detector or PIR sensor.

While the development of the first three features of this system has been completed, the last two features are still under development.

III. METHODOLOGY

We have combined the system with input processing and output that are shown in Fig. 1.

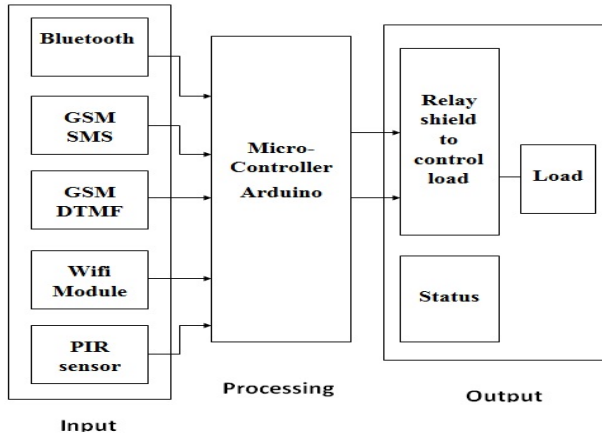


Fig. 1. Block diagram of home automation system

What follows is a 5-step methodology that has been proven to work in dozens of successfully completed home automation projects.

A. Bluetooth

Our first work is using Bluetooth. To create the Bluetooth connection, we needed one Bluetooth device, one Arduino Uno, two input one channel relay shield, one android app to control the system and two bulbs to be able to demonstrate how the system works. Now we describe how the whole system was combined by us. The Bluetooth device has 4 pins, which are Vcc, ground, Rx, and Tx. Rx is the receiver and Tx is the transmitter. We connected Bluetooth Rx, Tx with Arduino Uno Rx, Tx.

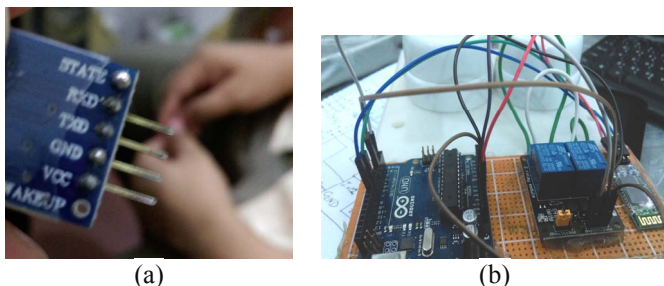


Fig. 2. Pictorial view of Bluetooth setup.
(a) Bluetooth pin (b) Full Bluetooth setup.

Bluetooth device has two more pins. Ground is connected with Arduino's ground and Vcc is connected with Vcc. Now we describe how the two relay shields were combined. There are also 4 pins: ground, In1, In2, Vcc. In 1 is input 1 which goes to signal and In2 is for second load. Input power 220V is incoming from the other side.

B. GSM-SMS

The system allows household appliance control using cell phone through global system for mobile communication (GSM) technology. The cellular communications is a potential solution for such remote controlling activities. SMS (short message service) technology can be used to control household appliances from distance. Remotely, the system allows the home owner to monitor and control the home appliances via mobile phone set by sending commands in the form of SMS messages and receiving the appliances status as well. The proposed system makes use of wireless control hence can be effectively used in systems where unwired connections are desired. The system uses the user's mobile handset for control and therefore the system is more adaptable and cost-effective and also providing ubiquitous access for appliance control.

Actually using the GSM module for the SMS system is the simplest part of the project. We set the GSM module over the Arduino Uno microcontroller. Inside the GSM module we put a necessary SIM (Subscriber Identification Module) card.

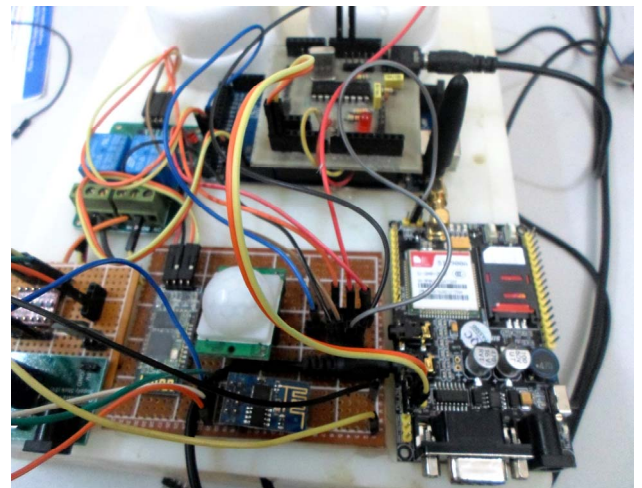


Fig.3. GSM based SMS and calling system

C. GSM-calling

For this feature, we use a DTMF (Dual-tone Multi Frequency) module. The DTMF module is a very old device. We have found this device after a long search.

DTMF is a tone composed of two sine waves of given frequencies. Individual frequencies are chosen so that it is quite easy to design frequency filters, and so that they can easily pass through telephone lines (where the maximum guaranteed bandwidth extends from about 300 Hz to 3.5 kHz). DTMF was not intended for data transfer; it is designed for control signals only. In GSM modules is a DTMF encoder

included. The DTMF tones are controlled by AT commands and we are working on this and we hardly believe that we will do this within a few days.

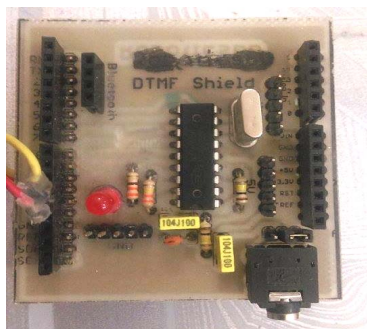


Fig. 4. DTMF device

D. PIR sensor

We use a passive infrared (PIR) sensor, which is an electronic sensor that measures infrared (IR) light radiating from objects in its field of view. They are most often used in PIR-based motion detectors. The main work of PIR is that when people come home, some appliances need to be turned on and when they leave, they need to be turned off.

PIR sensors are more complicated than many of the other sensors, because there are multiple variables that affect the sensor input and output.

The PIR sensor itself has two slots in it; each slot is made of a special material that is sensitive to IR. The lens used here does not do much and so the two slots can 'see' out past some distance (basically the sensitivity of the sensor). When the sensor is idle, both slots detect the same amount of IR, the ambient amount radiated from the room or walls or outdoors. When the warm body leaves the sensing area, the reverse happens, whereby the sensor generates a negative differential change. These change pulses are what is detected.

The IR sensor itself is housed in a hermetically sealed metal can to improve noise/temperature/humidity immunity. There is a window made of IR-transmissive material (typically coated silicon since that is very easy to come by) that protects the sensing element. Behind the window are the two balanced sensors.

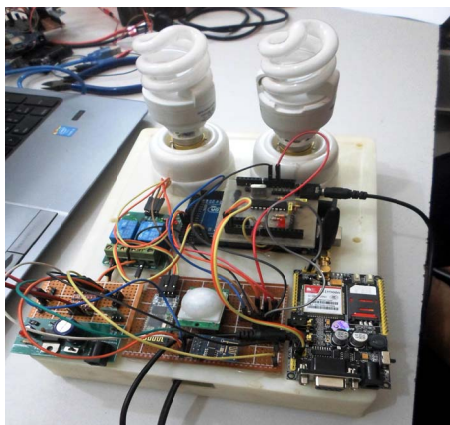


Fig. 5. Full circuit connection with PIR sensor.

E. Internet Module (For Twitter-Based Home Automation))

A Smart home is not a new term for science society, it has been used from decades. As electronic technologies are advancing, the field of home automation is expanding vastly. There were various smart systems have been proposed where the control is via Bluetooth, internet etc. Bluetooth capabilities are good and most of current laptop/desktops, tablets, notebooks and cell phones have built-in adaptor that will indirectly reduce the cost of the system. But it limits the control to within the Bluetooth range of the environment while most other systems are not so feasible to be implemented as low cost solution. In Wi-Fi based home automation system is presented. It uses a PC (with built in Wi-Fi card) based web server that manages the connected home devices. The system supports a wide range of home automation devices like fans, lights, other home appliances. A similar architecture is proposed in where the actions are coordinated by the home agent running on a PC or internet based devices. Other papers such as also presented internet controlled systems consisting of a web server, database and a web page of websites for interconnecting and handling the devices.

These systems utilize a PC which leads to a direct increase in cost and power consumption. On the other side, the development and hosting of the web page will also cost high. Lately few researchers have also presented use of Web services, Simple Object Access Protocol (SOAP) and Representational State Transfer (REST) as an interoperable application layer to remotely access home automation systems introduced a smart home management scheme over the Ethernet network based on XML SOAP standards [7].

The disadvantage of using SOAP based Web a service is that it is hard to handle and adds overhead to the client and server when parsing the message, resulting in slow operation and higher Bandwidth. REST has been represented as a Web-based interaction for controlling household devices using Web techniques such as HTTP (Hyper Text Transfer Protocol) caching and push messaging technique. Also a Web-based graphical user interface has been developed to manage the home devices [7]. Home automation using Cloud computing that has also been proposed by where users were able to control various lights and appliances within their home. All above mentioned systems has made tremendous contributions in home automation system. However, a PC is used as a server that increases the cost and power consumption while others require web page hosting that adds up the extra cost.

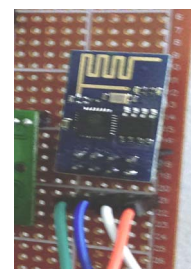


Fig. 6. The Internet Module

We have implemented a Twitter based home automation system with the help of an internet module (shown in Fig. 6). The internet module has a private twitter account, by which we can control our home appliances from anywhere and anyplace Fig. 7 shows the hardware connection for implementing the Twitter-base home automation system.

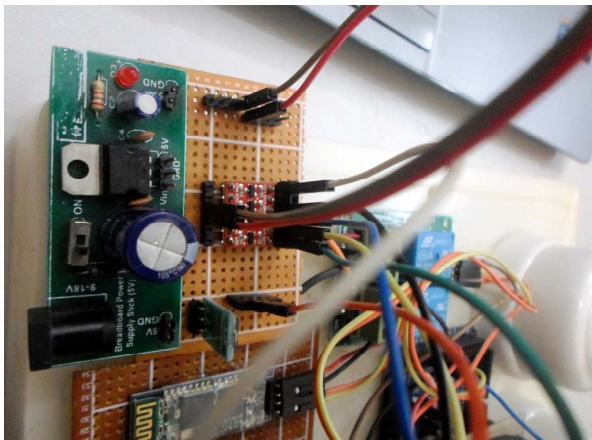


Fig. 7. Twitter connection board

We use the thingspeak website to complete the Twitter-based control. From Twitter, we tweet some words like, “#thingspeak #bulb1on or #thingspeak #bulb2off”. For this tweet, a URL message goes on thingspeak and by using that URL we can communicate with our home appliances remotely. This part of the design is also completely functional.

IV. RESULTS

All the implemented five features of this project are functionally working. This project includes Twitter based control and PIR sensor based detection of home intrusion that is actually the main part of security and privacy mechanisms. Others are Bluetooth system, GSM-SMS and GSM-calling system that are also helpful for security and privacy. We are successfully done with our hardware implementation of the project. All five features are working as expected.

V. CONCLUSION

This paper describes the design and implementation of a home automation system using various modern communication technologies. The system also provides a security system. The project uses Android and Arduino platforms, both of which are FOSS (Free Open Source Software). Therefore the overall implementation cost is very cheap and it is very affordable.

ACKNOWLEDGMENT

This project has been done as a part of an undergraduate Capstone Design course in the Department of Electrical and Computer Engineering in North South University, Dhaka, Bangladesh. We would like to thank our advisor Dr. Hasan Uz Zaman who has always given us good suggestions in the technical side and encouragement in all aspects.

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Performance Analysis of Ternary Content Addressable Memory (TCAM)

Behavior Analysis of Positive-Feedback Match-Line Sensing Schemes in High Speed TCM

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Abstract— With the advancement of scientific achievement, electronics sector has proved a radical improvement both in terms of efficiency and high speed applications. A TCAM is a memory that implements the lookup-table function in a single clock cycle using dedicated comparison circuitry. TCAMs are widely preferred in network routers for packet forwarding and packet classification, but they are also beneficial in a variety of other applications that require high-speed table lookup. The main challenge in TCAM design is to reduce power consumption without sacrificing speed or memory density. During match detection, switching of highly capacitive Match lines (ML) and Search lines (SL) cause huge power consumption. Energy efficient match detection or ML sensing schemes try to reduce this power consumption. In this paper, different ML sensing schemes are compared in order to find out the better scheme in terms of better power and speed performance. In conventional scheme large amount of energy is wasted as number of mismatched word is higher. Several further modified schemes are discussed and compared later on. The dedicated comparison circuitry of proposed ML sensing schemes reduces power consumption by minimizing search time and limiting voltage swing of MLs. A 16×16 bit TCAM powered by 1.8V supply voltage is designed with 0.18μm CMOS for this purpose.

Index Terms— TCAM, ML, CMOS, SRAM, etc.

I. INTRODUCTION

Ternary Content Addressable Memory (TCAM) is a type of associative memory that offers ternary storage and supports partial data-matching. Each ternary bit can be a “0”, a “1”, or a “don’t care” state. It is a key technology to enable the true power of the next-generation networking equipment and many lookup-intensive applications. TCAMs are hardware-based parallel lookup tables with bit-level masking capability, so gaining importance in high-speed intensive applications. For example, in the application where TCAM is doing the role of a co-processor in network processor[1][3]. In this application, Network processing unit provide great value to metro network equipment by providing flexible and adaptable processing solution. However, memory intensive packet searches that serve purposes as: routing, access control and Quality of Service (QoS) can quickly— as consume power and bus bandwidth. In classification of network routers application, TCAM is often used in network routers where each address

has two parts: Network address and Host address [2]. Network address varies in size on subnet configuration whereas Host address occupies remaining bits. Routing is done by consulting routing table maintained by router which contains-Known destination network address and Information needed to route packets to destination[3].

II. RELATED STUDY

A. TCAM Basics

TCAM consists of cells each containing 2 SRAM cells. SRAM elaborates Static Random Access Memory. SRAM is a type of semiconductor memory that uses bi-stable latching circuitry to store each bit. The term “static” differentiates it from dynamic RAM (DRAM) which must be periodically refreshed. SRAM exhibits data remanence. It holds data as long as power is applied. SRAM has 3 operations: (1) Hold (2) Write and (3) Read

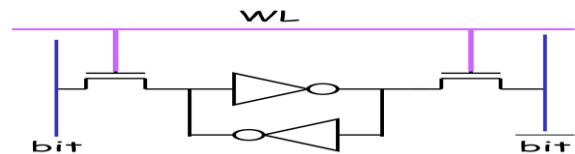


Figure 1: Basic 6T SRAM cell [3]

Hold: word line = 0, access transistors are OFF data held in latch; **Write:** word line = 1, access tx are ON new data (voltage) applied to bit and bit bar data in latch overwritten with new value; **Read:** word line = 1, access tx are ON bit and bit bar read by a sense amplifier. [7][8]

B. Mismatch-Dependent Power Allocation Scheme:

Mismatch-Dependent Power Allocation Scheme is proposed by [3]. Here feedback unit has been added to the basic CR MLSA to detect mismatched MLs during charging and reduce current to these MLs. Fig. 3.5 shows the proposed Mismatch-Dependent Power Allocation Scheme.

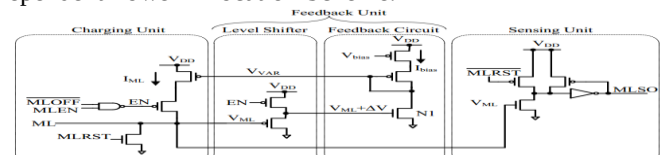


Figure 4: MLSA in Mismatch-Dependent Power Allocation Scheme. Dummy word isn't shown here, [3], [9]

feedback unit contains a level shifter and a feedback circuit to implement feedback. In speed optimized setting, the MLSO, VML and VVAR are pre-discharged to ground to first. MLEN starts the charging of all MLs. Initially both VML and VVAR are increased by current IML and I_{bias} respectively. As VVAR increases IML decreases. But with increasing VML, the level shifter output also increases. VML in a matched ML increases faster than that in mismatched ML. the level shifter output becomes significantly high in a matched ML to turn on N1 and VVAR starts to discharge. As a result IML for matched ML increases again. For a mismatched ML, VML rises slowly and to a lower value which depends on the number of mismatches presents in word. So for mismatched MLs N1 may get weakly turned on or it may remain off depending on the number of mismatches. This results small or no reduction in VVAR. As a result IML in a mismatched ML keeps decreasing [12]. So matched MLs get higher current than mismatched MLs. Same dummy word in CR MLSA is used to generate MLOFF signal to control the ML charging time. This scheme shows significant energy reduction over CR scheme. But the level shifter and feedback unit consume considerable amount of energy. In case of a mismatch N1 sometimes remains weakly turn on which means there is a conducting path between VDD and ground. So for a large number of mismatches, VML is small since ML has multiple path to ground. This causes a static power consumption. Also capacitive loading on the ML increases due to large transistor size in level shifter. Again transistor count in this MLSA is higher than CR MLSA. Finally feedback action is very sensitive to V_{bias} and transistor sizes. So the transistor sizes need to be tuned for better performance [21][22].

C. Active Feedback Scheme

To overcome the problems of MD scheme Active Feedback ML sensing scheme is proposed. The proposed scheme does not consume any static power and it outperforms the MD-MLSA in speed, energy, area and robustness [2]. Fig. 5 shows the proposed MLSA with active feedback.

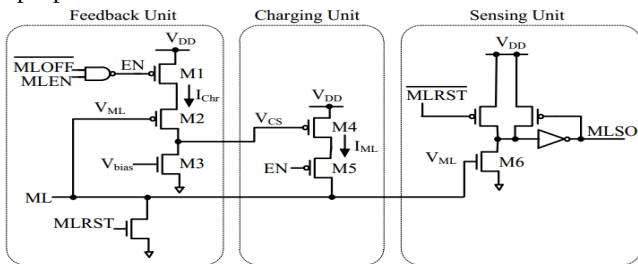


Figure 5: MLSA in Active Feedback Scheme. Dummy word isn't shown here, [2], [10]

In active feedback scheme VML and MLSO are reset to zero by MLRST signal. Search starts by asserting MLEN. M1 and M2 turns on making gate capacitance of M4 to charge. Initially IML and I_{Chr} are same. So, V_{cs} and VML increase with time. As a result IML and I_{Chr} decrease, this makes M2 to turn off. M3 is always on due to dc V_{bias}. In matched MLs V_{cs} discharges through M3. That makes M4 to conduct and IML to increase again. In mismatched MLs, VML rises slowly. This keeps M2 on and makes V_{cs} to increase which makes M4 to turn off. As a result IML decreases. So matched

MLs get higher current than mismatched MLs. Energy consumption in mismatched words become small. VBIAS and MLSA transistor sizes need to be tuned for better performance. As M3 is always on energy is wasted due to current flow through it in matched MLs. So further energy consumption reduction is needed.

D. Resistive Feedback Scheme

Fig. 6 shows the proposed MLSA with resistive shielding. It uses an nMOS transistor (M3) in the triode region to decouple the ML and its MLSA. The N3 channel resistance shields the sensing point [SP in Fig.4] from the highly capacitive ML. This way, the current source (I_{Chr}) can be sized down to save power without sacrificing the sensing speed. Note that due to the body effect and the decreasing gate-to-source voltage (V_{GSN3}) the N3 channel resistance increases when the ML voltage is rising up [2]. The N3 channel resistance depends on the number of mismatch bits.

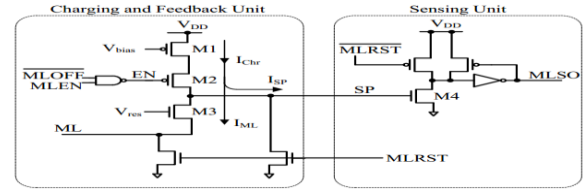


Figure 6: MLSA in Resistive Feedback Scheme. Dummy word isn't shown here, [2], [10]

Voltages of ML, sensing point (sp) and MLSO are reset to zero by MLRST signal. The MLEN signal (shown in fig. 8.5) enables the MLSA by activating EN thus turning on M2. VBIAS controls the total charging current I_{Chr}. Initially IML and I_{Chr} are same in all MLs. VML is increased by IML, so drain to source voltage of M3 decreases. Therefore IML decreases [11]. Reduction in IML causes increase in ISP. Increase in ISP causes increase in VSP. This, in turn, increases drain to source voltage of M3 again which increases IML again. As a result feedback continues between ML and SP. In case of a mismatch VML rises slowly. Therefore VSP also rises slowly [16-19]. VSP in a matched ML reaches threshold voltage much quicker than that in a mismatch ML and produce a high output at MLSO. VSP of mismatched words do not get the sufficient time to charge up to the sensing threshold voltage. Therefore MLSOs remain low. This technique of ML sensing is extremely simple in construction. Two additional transistors are required than CR MLSA. Faster sensing of the dummy word also reduces energy consumption because the ML current sources are shut down sooner. The charging current of mismatched ML is less affected by the M3 resistance because it has a larger and a weaker body effect than that of a matched ML. Energy and delay can be further reduced by decreasing V_{res} (shown in fig. 6) [2]. The sensitivity of the feedback action to process variation is also low since the feedback is weak.

III. SIMULATION & RESULTS

A. Current Race Scheme

Simulated search time in CR scheme is 5.0446ns which is measured by subtracting the starting time of a matched MLSO

(in figure DMLSO) and starting time of MLRST signal. Fig. 7 shows the search time of current race scheme.

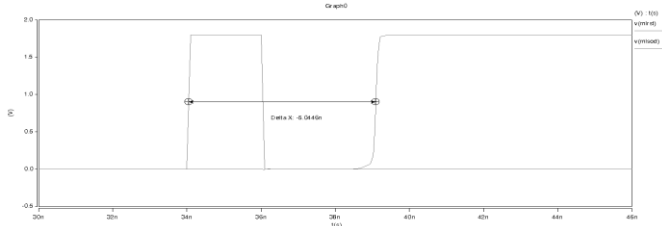


Figure 7: Search time measured between VMLRST and VMLSOD at 0.9V (50% of V_{in}) in CR scheme

Again simulated voltage margin in CR scheme is 1.112225V which is measured by subtracting the amplitude of crossing between matched ML (ml1) and MLSO (mlso1) and maximum magnitude of 1 bit mismatched ML (ml3). Fig.8 shows the voltage margin of current race scheme.

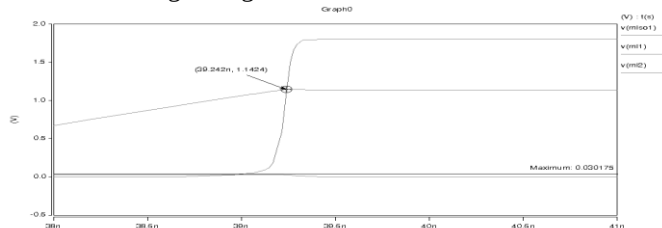


Figure 8: voltage margin measured between crossing of matched ML (ml1) and MLSO (mlso1) and maximum magnitude of 1 bit mismatched ML (ml3) in CR scheme

B. Mismatch-Deendent power Allocation

Simulated search time in MD scheme is 2.5927ns which is measured by subtracting the starting time of a matched MLSO (in figure DMLSO) and starting time of MLRST signal. Fig. 9 shows the search time of mismatch-dependent power allocation scheme.

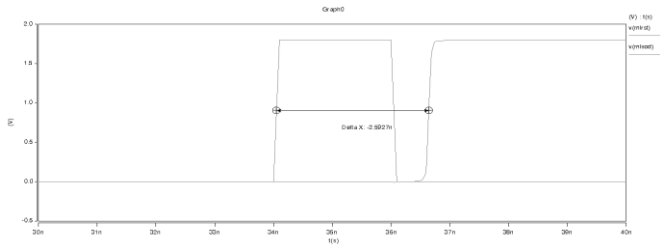


Figure 9: Search time measured between VMLRST and VMLSOD at 0.9V (50% of V_{in}) in MD scheme

Again simulated voltage margin in MD scheme is 0.81093V which is measured by subtracting the amplitude of crossing between matched ML (ml1) and MLSO (mlso1) and maximum magnitude of 1 bit mismatched ML (ml3). Fig. 10 shows the voltage margin of mismatch-dependent power allocation scheme.

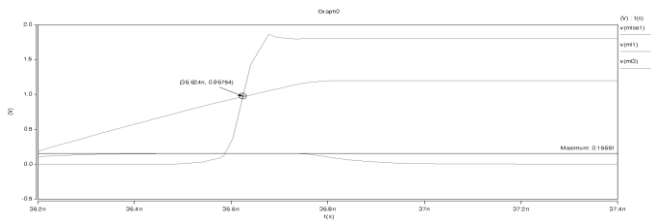


Figure 10: Voltage margin measured between crossing of matched ML (ml1) and MLSO (mlso1) and maximum magnitude of 1 bit mismatched ML (ml3) in MD scheme

C. Active Feedback Scheme

Simulated search time in AF scheme is 2.6361ns which is measured by subtracting the starting time of a matched MLSO (in figure DMLSO) and starting time of MLRST signal. Fig. 11 shows the search time of active feedback scheme.

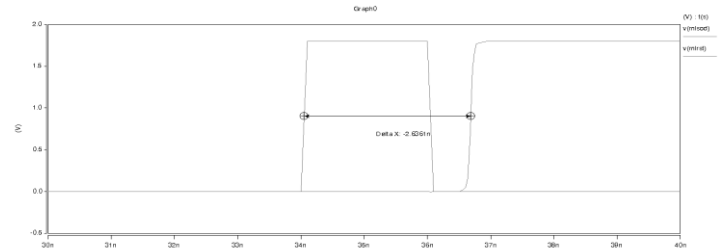


Figure 11: Search time measured between VMLRST and VMLSOD at 0.9V (50% of V_{in}) in AF scheme

Again simulated voltage margin in AF scheme is 0.859028V which is measured by subtracting the amplitude of crossing between matched ML (ml1) and MLSO (mlso1) and maximum magnitude of 1 bit mismatched ML (ml3). Fig. 11 shows the voltage margin of active feedback scheme.

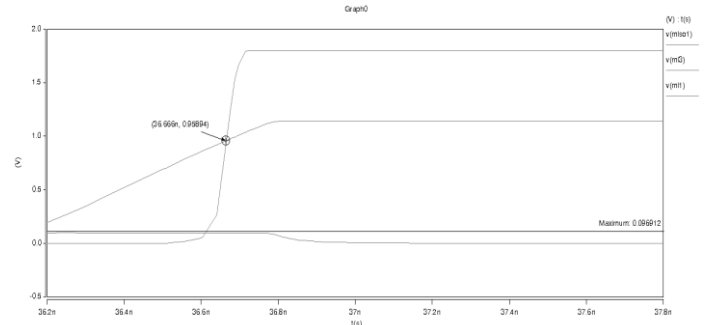


Figure 12: Voltage margin measured between crossing of matched ML (ml1) and MLSO (mlso1) and maximum magnitude of 1 bit mismatched ML (ml3) in AF scheme

D. Resistive Feedback Scheme

Simulated search time in RF scheme is 2.5266ns which is measured by subtracting the starting time of a matched MLSO (in figure DMLSO) and starting time of MLRST signal. Fig. 4.4(a) shows the search time of resistive feedback scheme.

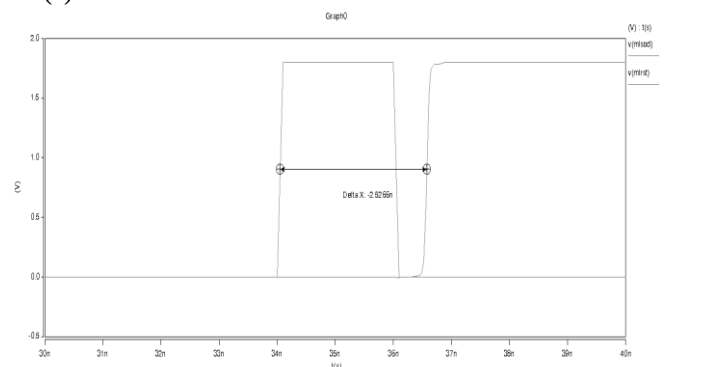


Figure 13: Search time measured between VMLRST and VMLSOD at 0.9V (50% of V_{in}) in RF scheme

Again simulated voltage margin in RF scheme is 0.67165V which is measured by subtracting the amplitude of crossing between matched ML (ml1) and MLSO (mlso1) and maximum magnitude of 1 bit mismatched ML (ml3). Fig. 13 shows the voltage margin of resistive feedback scheme.

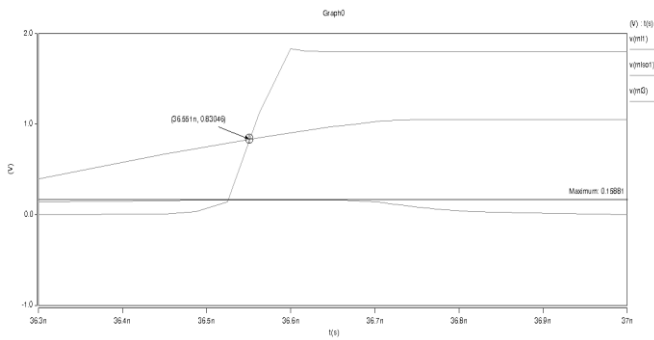


Figure 14: Voltage margin measured between crossing of matched ML (ml1) and MLSO (mlso1) and maximum magnitude of 1 bit mismatched ML (ml3) in RF scheme

Schemes	Search Time (ns)	Voltage Margin (V)
Current Race(CR)	5.0446	1.112225
Mismatch Dependent(MD)	2.5927	0.81093
Active Feed-back(AF)	2.6361	0.859028
Resistive Feedback(RF)	2.5266	0.67165

Table 1. Comparison among different schemes

IV. CONCLUSION

The proposed positive feedback schemes improve search speed at the cost of slight degradation of voltage margin. These MLSAs use the positive feedback to reduce the power consumption in TCAMs [3]. MD scheme allocates power to match decisions based on the number of mismatched bits in each TCAM word. With allocating less power to mismatched MLs and with most MLs being in this category, this scheme results in a considerable power reduction. Again AF and RF schemes require less transistors than the MD scheme. The active-feedback MLSA improves energy savings and voltage margin without consuming any static power. Energy measurement results of the two MLSAs show reductions of 56% and 48%, respectively, over the conventional CR-MLSA.

ACKNOWLEDGMENT

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Earthquake Monitoring and Warning System

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Abstract—The project aims at designing an earthquake monitoring and warning system that is capable of detecting earthquakes as well as warning people to take necessary precautions. This system will not only attempt to save human lives, but will also store the data for later use by professionals working at this sector. The objective of this project is to design a LabVIEW based control system by collaborating MEMS accelerometer sensor with Arduino microcontrollers in a ZigBee wireless network. The system also has telecom VAS capabilities to alert people and control machines. Existing earthquake systems are moreover designed for aftershock management and they hardly deal with real-time data. However, this system is mainly designed to react as soon as the sensor nodes come across a p-wave.

Keywords—Earthquake, P-wave, LabVIEW, XBee Radios, GSM Modem, Arduino, ADXL-345, Piezoelectric Sensor.

I. INTRODUCTION

Earthquake is one of the most disastrous and catastrophic natural adverse event [1]. Naturally earthquake is generated due to the movement of earth's crust or tectonic plates. Though the movements of these plates are quite slow but the effect of the collision is dangerous, a huge amount of energy released that creates seismic wave. Earthquakes show themselves as vibration, shaking and sometimes displacement of the ground. It is also caused by volcanic activity, landslides, mine blasts, and nuclear test. Earthquake rarely kills people or any living organism. But the event they trigger such as building collapse, fire, Tsunami etc. are the main reasons behind the loss. It is very tough to predict the time of earthquake. Although technology has grown so much in the last few decades but detecting earthquake before it happens is still a mystery to be unfold. Researchers and engineers, all over the world, are working extremely hard to design earthquake early warning system that could essentially help mankind to get sufficient time to move out to safe zones before an earthquake strikes.

Talking about early warning, with the help of this system people might get about half a minute to a minute or two depending on how far the epic centre lies. But scientifically it is not an early warning system as this project aims to detect p wave of earthquake. P wave is the primary shock wave that is released from the epic centre and it can be referred to sound waves. The movement of p wave is parallel to the direction of propagation of the seismic wave and hence it is not destructive for infrastructures. A series of destructive wave that follows p wave are known as s waves or secondary waves. S wave has a speed of about 60% that of p wave [2]. The movement of s wave is perpendicular to the direction of propagation of the seismic wave and hence it is extremely risky to infrastructures. Therefore, the difference in speed of p wave and s wave gives rise to a time delay which is about ten to fifteen seconds, will

actually work as an added advantage to this system. If the system is set in the fault zone, earthquakes can be detected earlier and people living far away from the fault zone can be alerted, so that they can evacuate buildings and move to a safer zone like open spaces. Simple vibration sensors and accelerometers are used to detect earthquake. The detection of p-wave is harder because of the low frequency which is about 1 to 3 Hz. There are two other waves as well, Love and Rayleigh wave [3].

Researches have been carried out which are relevant to this system. Although some of them are entitled to post-earthquake detection but one of them used abnormal behaviour of animals as mobile biological sensors for pre-earthquake sensing [4]. The idea of pre-earthquake sensing is preferable but such systems containing biological sensors will not come good when it comes to measuring the magnitude of the quake or datalogging for instance. On the other hand, another research work used MEMS sensor for sensing earthquake and also had an alarm feature [5]. But the system did not have a wireless sensor network. None of the earlier designed systems used ZigBee network topology or LabVIEW as a software for monitoring and datalogging purposes. The fact that also makes this project a novel one is the use of piezoelectric sensor especially tuned for p-wave detection and triggering the sensor nodes for data transmission.

II. SYSTEM DESIGN

An XBee Network is created with two sensor nodes and a coordinator [6]. At the moment of earthquake, piezoelectric sensor will trigger each node to send ground acceleration - from ADXL-345 digital accelerometer - and piezoelectric sensor data to the coordinator. The coordinator, after receiving the data, will pass it to the laptop via serial communication. Data acquired from both the nodes will be instantaneously processed and compared in National Instruments LabVIEW software. If there is a match, the laptop will automatically store the data in a CSV file and immediately send warning SMS via GSM communication to the receiving end. The receiving end is installed at tolls of a bridge. It contains an LCD screen, a magnitude indicator RGB LED, a servo motor for toll, a push button for toll control, a buzzer and a GSM modem. Depending on the Richter Scale Magnitude different messages are displayed on the LCD screen. The RGB LED glows with different colors assigned to it for different magnitudes. The buzzer will turn on to alarm people from going over the bridge. The push button, for toll control, will stop receiving user input and the toll (servo motor) will eventually come to a halt if the sonar sensor finds the toll spacing decreased i.e. when something is under the toll. The complete system design is shown in a flow chart in Fig. 1.

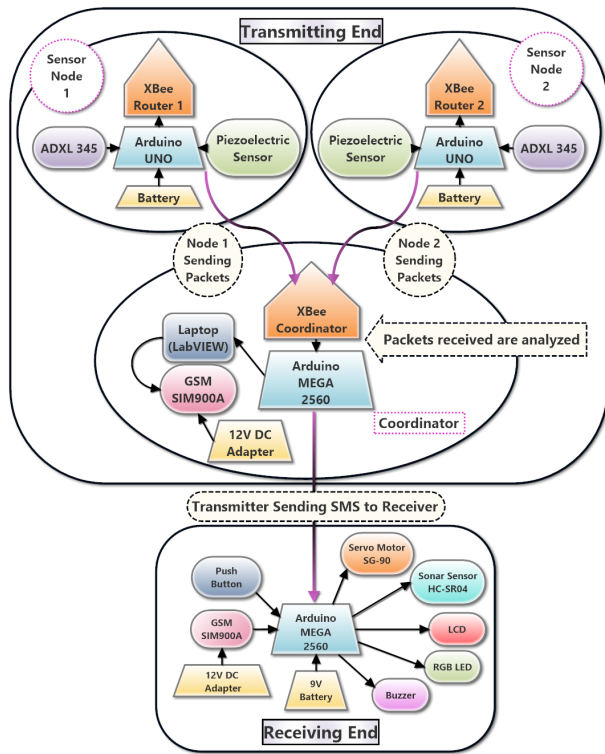


Fig. 1: Project Flowchart

III. CIRCUIT DESIGN

This section briefly explains how the sensor nodes, the coordinator and the receiving end of the project works.

A. Sensor Node

Each of the sensor nodes has got two sensors attached to it. The piezoelectric sensor is used to trigger each node to transmit data whenever there is an earthquake, which intern makes the battery that is powering the nodes long lasting. Each of the data packets sent to the coordinator contains the name of the node, ground acceleration from ADXL-345 digital accelerometer and piezoelectric sensor data. Fig. 2 shows the circuit diagram of the sensor node. An Arduino UNO board is used as microcontroller that handles all the sensor data [7]. An XBee series 2 radio is used to transmit data to the coordinator.

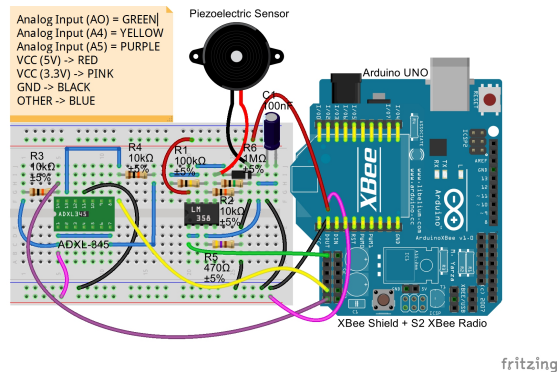


Fig. 2: Sensor Node Design

B. Coordinator

The coordinator is the key to the analysis part of the system. Packets received from the sensor nodes are opened and all the data are pushed to the laptop via serial communication for further analysis with the help of an application. It also contains a GSM modem to send immediate warnings via SMS which will take place automatically right after the software analysis. Fig. 3 shows the circuit diagram of the coordinator.

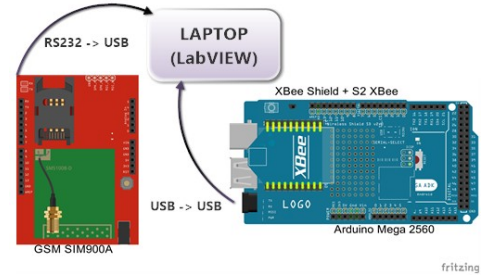


Fig. 3: Coordinator Design

C. Receiving End

Receiving end or in other words this is the application part of the project. A variety of loads can be controlled using the SMS, for example, shutting off power stations, gas stations and even lifts of high rise buildings. In this project, toll stations of bridges are controlled using the SMS. When the warning SMS is received the tolls automatically switches off to avoid any accidents during shaking. An alarm system is there to warn people to stay put. Fig. 4 shows the circuit diagram of the receiving end.

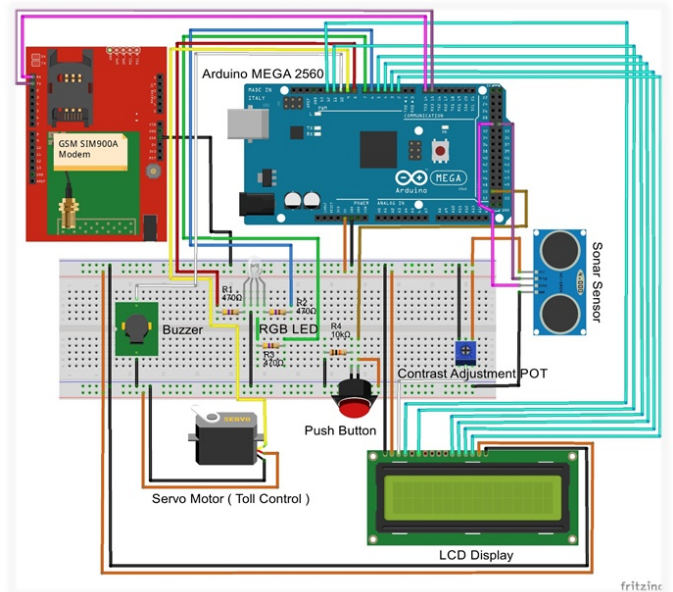


Fig. 4: Receiving End

IV. RESULTS AT THE TRANSMITTING END

The complete system is designed in National Instruments LabVIEW software [8]. To make the application user friendly

four tabs are created. The first tab represents ADXL-345 digital accelerometer, the second tab represents Resultant Acceleration, the third tab represents piezoelectric sensor and the last tab represents GSM SMS Mode. Each of the sections below briefly explains the matching solutions at the time of earthquakes.

A. ADXL-345 Digital Accelerometer Tab

Individually, ADXL345 data are selected from the serial read buffer. The absolute value of x, y and z axis are used to plot each graph in Fig. 5 for both the sensor nodes and the corresponding values represent the right most spikes of each graph.

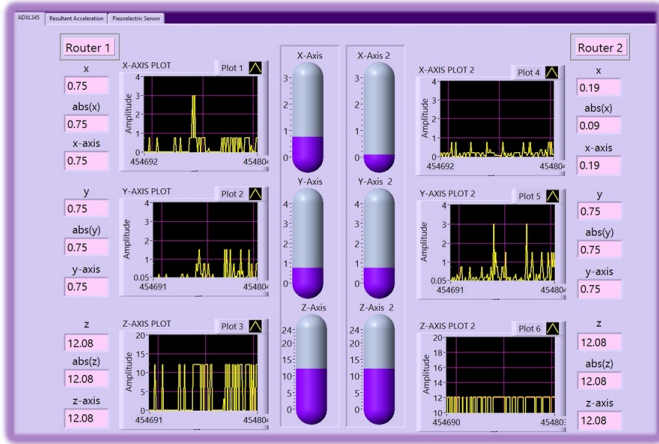


Fig. 5: ADXL 345 Tab

The figure shows the situation when both the routers are responsible for sending the data. In other words, a situation when the ground is shaking.

B. Resultant Acceleration Tab

The goal of this project is to detect p-wave of earthquake and it happens to be in the XY plane. In XY plane, Z is constant, so there is no point to carry the Z axis data forward into this tab. X and Y axis data are selected from the serial read buffer separately and applied in Eq. 1. The magnitudes of acceleration M_R , for both the routers (nodes), are plotted separately on each graph which is shown in Fig. 6. The magnitude meters indicate the exact magnitude as it rotates clockwise. The matching LED - in between the magnitude meters - glows only when there is a match in between the resultant acceleration of both the routers.

$$M_R = \sqrt{x^2 + y^2} \quad (1)$$

This is surely an effect of earthquake as both Router-1 and Router-2 are picking up data. Both the magnitudes are exactly matching and therefore, the matching LED is glowing. The magnitude meters are pointing at the same reading. In this case, it is not compulsory to check the piezoelectric sensor tab but to avoid any confusions, it can always be checked what sort of data the corresponding piezoelectric sensors are picking.

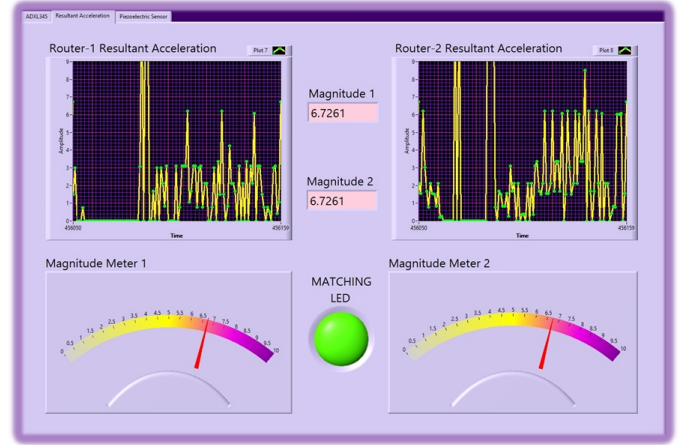


Fig. 6: Resultant Acceleration Tab

C. Piezoelectric Sensor Tab

Piezoelectric sensor data are selected individually from the serial read buffer which is shown in Fig. 7. This tab will have a great deal of importance in projects where simultaneously a bunch of routers, in a wider mesh network, will be communicating with the coordinator at the same time.

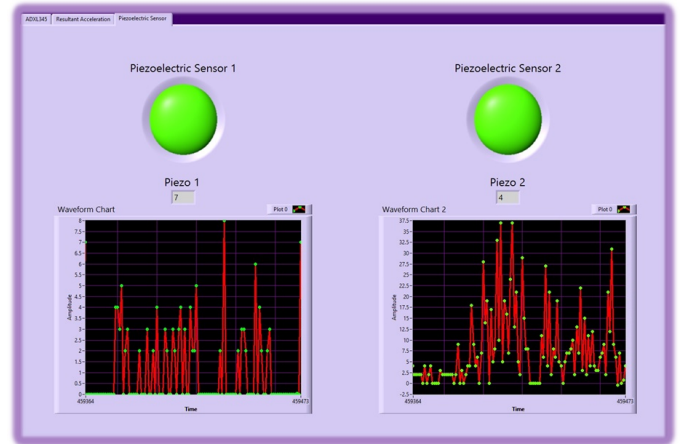


Fig. 7: Piezoelectric Sensor Tab

This is surely an effect of earthquake as both Router-1 and Router-2 are picking up data. Both the LEDs are glowing. Although the piezoelectric sensor readings are not matching but we are not bothered about the readings as both the sensors are attached with different amount of mass.

D. GSM SMS Mode Tab

The end of the monitoring part leads to the warning part which is completely handled by the GSM SMS Mode tab. Initially the users mobile number is set. If there is a match then instantly an SMS will be send to the receiving end. In Fig. 8 the status blank shows the character that is written in the serial write buffer.

Table 1 shows the Richter Scale Magnitude corresponding characters that will be sent.

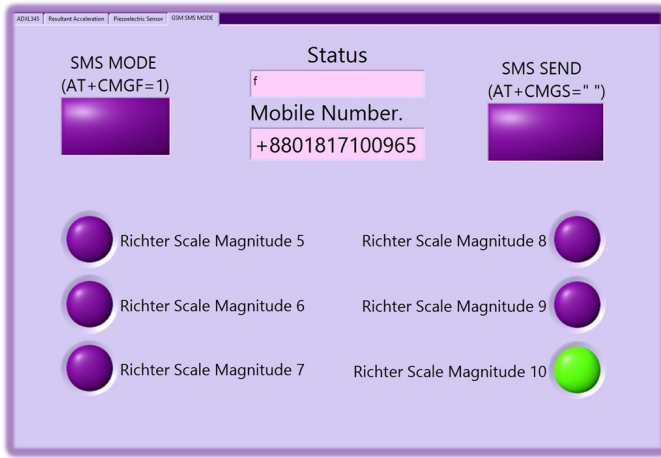


Fig. 8: GSM SMS Mode Tab

TABLE I: Richter Scale Magnitude Corresponding Characters

Richter Scale Magnitude	Acceleration Range (m/s^2)	Write Character
1	< 0.017	x
2-3	0.017 – 0.14	x
4	0.14 – 0.39	x
5	0.39 – 0.92	a
6	0.92 – 1.8	b
7	1.8 – 3.4	c
8	3.4 – 6.5	d
9	6.5 – 12.4	e
10	> 12.4	f

V. SAVING DATA

Every time the run button in LabVIEW project is pressed, a comma-separated values (CSV) file is automatically created to the desired location in the laptop. This intern helps to get track of the specific date and time of the earthquake. Even though the monitoring part is made as easy as possible, it is quite difficult to keep track of every data as a human being. Transitioning from one tab to another consumes time and the system holds the data for 507 milliseconds to be precise. Although the person will be able to see the peak from the graphs but the corresponding data will be gone by then. For ease of analysis, we have successfully collaborated LabVIEW with Microsoft Excel to save all the data at the specific date and time. Figure 9 shows a list of few data stored while the system was running.

	A	B	C	D	E	F	G	H	I	J	K	L
	Date	Time	x-axis(R1)	y-axis(R1)	z-axis(R1)	x-axis(R2)	y-axis(R2)	z-axis(R2)	Resultant Acceleration (R1)	Resultant Acceleration (R2)	Piezo (R1)	Piezo (R2)
1	5/5/2015	1:50:11 PM	0.19	0.75	12.08	0.19	0.75	12.08	0.773692	0.773692	2	2
2	5/5/2015	1:50:16 PM	2	0	0	2	0	0.19	2	2	0	-0.09
4	5/5/2015	1:50:25 PM	0.19	-4.5	12.08	0.19	1.5	12.08	1.511985	1.511985	13	13
5	5/5/2015	1:50:32 PM	0.19	12.08	3	0.19	12.08	7	12.081494	12.081494	0	0
6	5/5/2015	1:50:32 PM	0.75	0	0	0.75	0	0	0.75	0.75	0	0
7	5/5/2015	1:50:35 PM	12.08	3	0	12.08	3	0	12.446943	12.446943	0.05	0.75

Fig. 9: Data stored in a CSV file

VI. RESULTS AT THE RECEIVING END

Each of the individual pictures in Fig. 10 shows the warning message along with the type and magnitude of earthquake. The RGB LED glows with different colors assigned to it at different magnitudes.



Fig. 10: Results at the Receiving End

VII. CONCLUSION

This project is intended to raise awareness among people living in Bangladesh. With a massive population and high rise buildings everywhere someone's got to step up and design a system that can give people time to take precautions before the earthquake strikes. And we are extremely lucky that we can stand beside them at their desperate times. As the name suggests, earthquake monitoring and warning system, this project will not only let earthquake engineers to monitor earthquakes in real-time but will also let the system send warnings all by itself. The reason behind designing it completely automatic is to cut off any delay that might creep in due to human involvements. This project can be further improved by adding more sensitive sensors. Geophones can be a probable alternative vibration sensor. Adding GPS monitoring network will improve the systems effectiveness.

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Influence of Ellipsoid Particle Size Distribution on Microwave and Millimeter Wave Frequencies

At Sand and Dust Storm Physical Measurement

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Abstract – Determining the particle size distribution (PSD) of a solid and liquid particles is of great practical interest. An experimental research was conducted to explore the influence of PSD on propagation properties of electromagnetic wave specifically in microwave and millimeter wave bands. Majority of earlier studies have primarily concentrated on the homogeneous spherical dust particle in mathematical models and equations. Beside of the lack of information on particle shape is one of the main causes for inaccurate calculation and extrapolation. I found that dust grain is of random shape but it is analogical to the ellipsoid in the real environment of dust media in literature. In this paper, emphasis of shape of particle at Sand and Dust Storm (SDS) phenomena in the microwave and millimeter wave with the actual physical measurement was presented and discussed. Moreover, collected of samples used in the demonstration was dust grains taken from the North East of Sudan were inspected. The findings of the experiment are essential of many belongings namely size, shape, number, and surface characteristics. Finally, we found that exact particle distribution with all the parameters could be used in volume fraction and consternation, many applications in image radar, and remote sensing as well as telecommunications links.

Keywords - Particle Size Distribution, Sand & Dust Storm, Microwave, Millimeter Wave, Ellipsoid particle.

I. INTRODUCTION

Purpose of particle size distribution is critical in many straightforward processes in signal propagation and radio wave prototypical and model scheme. Researchers have been studied diverse mathematical and experimental PSD measurements and techniques that was developed to facilitate several functions and applications. On the other hand, SDS and dust particles is an important constituent of atmospheric aerosols and plays an essential role in cloud condensation in many areas around world. Moreover, the impact of (SDS) at microwave (MW) and millimeter wave bands (MMW) in telecommunication has been investigated by many researchers

in the past decades [1-3]. Dust particle is a solid grain matter (1-150 μ m) suspended in the atmosphere. However, as the wavelength of millimeter wave becomes comparable with atmospheric in-homogeneity, a relatively large atmospheric effect on propagation might be expected from particle mass, particle number concentration, and particle size with high time resolution and high accuracy. The major obstacle faced in using the frequencies of 30-300 GHz millimeter wave band for terrestrial radio relay applications is the influence of the atmospheric conditions on the propagation of the electromagnetic waves. In particular, attenuation far in excess of that at long wavelengths is experienced at these frequencies. Furthermore, this paper is aimed at understanding the influence of ellipsoid PSD on radio waves in arid and desert land. The study involves the continuous experimental lab exam from four samples brought from North East Sudan. Hence, the hypothesis of study effects of various meteorological parameters on wave propagation can be deduced.

Propagation of the electromagnetic wave (EMW) with the earth's atmosphere depends on the wavelength of signal radiation, and the nature of the medium being traversed into the atmosphere is affected by the scattering and absorption of sand and dust particles at frequencies of above 3 GHz [5]. This medium consists of different molecular species and aerosol particles. Dust media are single atmospheric components that are ubiquitous in nature, such as the non-spherical particle, and are frequently the determining factor in the amount of signals transmitted and received in all stages of the operational process. Additionally, awareness of the propagation of millimeter wave through the lower atmosphere arises from the possible application of these frequencies in many communications, despite their severe limitations. However, many reports are mainly concerned with particle absorption and scintillation, either along a path at or near vertical incidence (between ground and outer space), or on a

link with terminals located in the dust storm environment [7-8]. It is clear that there is a strong need for quantitative microwave and millimeter wave regime observations that roughly correspond to EMW in free space including: 1)attenuation due to precipitation as a function of frequency and rate of precipitation; 2) the amplitude of scintillation on line-of-sight links; 3)absorption due to dustiness and humidity; and 4)any other clear-air effects.

The particles treated in this paper may have arbitrary forms and they are small compared to the wavelength based on particle size. As identified, the “Accurate” measurement of size is important because quality and performance of most particulate products are closely related to size distribution of dust particles. The past decade has seen rapid evolution and growth of applications for measuring size and shape. The aim of this paper is to examine the radio signals that are lost as it propagates through dust media and the influence of PSD. This paper is organized as follows: Section II includes the literature review of related works, Section III explains the methods and materials involved, and Section IV discusses the results and conclusion.

II. LITERATURE REVIEW

Recently, researchers have shown that dust conditions can be divided into three main types: dust storm, floating, and blowing. [8] is also an important parameter that allows the understanding of the heterogeneity of particle sizes in a mixture:

- Particle size distribution;
- Particle concentration and orientations;
- Chemical composition of the dust particles; and
- Surface bound or adsorbed.

The investigation of all parameters is very interesting from the point of view of the applications of PSD and the synthesis of the particles, as it serves to understand what happens during the optimization process, when a few synthesis parameters are adjusted.

A. Definition & Formulation

Dust is the suspension of solid particles produced by mechanical disintegration of materials such as crushing, grinding, and blasting. Solid aerosols are generated by the handling, grinding, abrasion, or cutting of a bulk material or solid and/or liquid, in gas floating particles within a particle size ranged between 0,001 μm to 100 μm . Dust particle size is related to the amount of verve involved in the construction; the higher the verve, the smaller the particle produced; the lower the verve, the larger the particle made [5]. First, dust storm happenings are random; a strong wind vigorously blows ground-based dust particles and makes the atmosphere very hazy producing horizontal visibilities of less than 1 km or even less than 2 m. Next, the floating dust is a meteorological process in which dust particles are uniformly suspended in the atmosphere and the horizontal visibility is less than 10 km. Lastly, blowing sand conditions occur when the wind blows dust particles from the ground into the air, decreasing the horizontal visibility to between 1 and 10 km. Moreover, the size of the dust particles is an imperative parameter as there is an optimal size for each application. Depending on the size and number of particles, the concentration of where they accumulate is different. Thus, particles between 10 to 150 μm

are retained. The dimensions of an ellipsoid along three orthogonal axes may be different. Therefore, the ellipsoid geometry has one more degree of freedom than the commonly used spheroid geometry, and the former is a better approximation to the shapes of realistic dust particles [8].

An accurate calculation of wave attenuation in sand and dust storms needs detailed information such as sand and dust particle dimension, shape, complex permittivity, size distribution functions, visibility, frequency, etc. Based on previous research experience [2], Sharif [3] and Ahamed, [10] fitted their dust particle size measurements to exponential distribution although their measurement depended on only one sample. They fitted a limited number of dust particle samples to the normal distribution. The problem becomes complicated as some authors have claimed that the particles are dependent on the height and spread of the storm [12–14] while others have reported that the distribution is independent of wind speed [9]. Similarly, more than one distribution may exist during a storm [14]. Considering the spherical dust particle shape, Adel [6] demonstrated that the attenuation and phase shift coefficients from dust or sand storms at certain wavelengths may be expressed for a mono-dispersive system (equal particle radii). The initial collection process may also bias the dust particle size distribution and produce artificial dust. Real dust particles that are smaller than about 50 μm in diameter do not last very long by themselves after being collected by vacuum cleaners [9].

It is obviously clear from the preceding section that the formula used for predicting the attenuation of microwave signals due to a dust storm has a parameter called particle size or particle diameter, which is of great importance in the process of estimation [13]. For each specific condition, measurement and proper fitting of the probable distribution of dust particles are needed to realize the correct impairment calculations to signify the distinction in non-spherical particles during the dust storm to enhance the accuracy of both the dust storm attenuation prediction models.

III. METHODS AND MATERIALS

The limitation of the use of millimeter wave bands (MMW) for many telecommunication purposes is mainly due to high attenuation due to precipitation, particularly, rain, fog, snow, and dust storm. Thus, to design and plan such links, it is desirable to have comprehensive information about the relationship between attenuation and climate environment characteristics. It is particularly useful to establish relationships between attenuation experienced over a link of several kilometers in length and the prediction measured at one point.

A. The Spherical Particles

The formulae for spherical particles, which are used in 95% of all former models and equations, are very simple, even if arbitrary polarization of the incident light is admitted. Spherical particles have $S_3 = S_4 = 0$. Thus, two complex amplitude functions occur for any direction; these functions are $S_1(\theta)$ and $S_2(\theta)$, and they depend only on the scattering angle θ . Here, we consider the electromagnetic wave with a certain frequency and traveling in one direction. We choose a plane of reference through the direction of propagation. Using $\mathbf{r}$, we shall denote the unit vector along the normal of this plane (sense arbitrary); and $\mathbf{1}$ will be used as the unit vector in this plane that is perpendicular to the direction of the

propagation. The sense is chosen so that $r \times 1$ is in the direction of propagation as clarified in the following Figure 1.1. Notice that the direction of movement in the three particles (sphere, two ellipses) is in the same direction as the reference wave. The single most striking observation is the analysis of the shape in the particle size distribution for data evaluation in section IV.

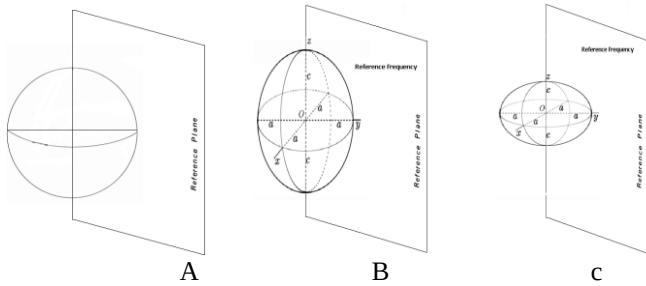


Figure 1.1 Shapes of particle sphere and tow ellipsoid

B. Ellipsoid Dust Grain

Many previous research findings have argued that sand particles are assumed to be spheres, [5] although the shapes of realistic sand particles are quite different.

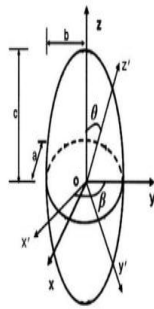


Figure 2 Shapes of particle ellipsoid all axis

The estimated aspect ratios of the three axes (hereafter, indicated by a, b, and c) of sandstorm particles are approximately $c:b \frac{1}{4} 1:0.71$ and $b:a \frac{1}{4} 0.71:0.53$. For simplicity, in the following discussion, the two aspect ratios, $c:b \frac{1}{4} 1:0.71$ and $b:a \frac{1}{4} 0.71:0.53$, are indicated in a concise form given by $c:b:a \frac{1}{4} 1:0.71:0.53$. Size distribution ellipsoids are used with an axis-ratio of $a_1:a_2:a_3 \frac{1}{4} 1:0.71:0.53$, where the parameters a_1 , a_2 , and a_3 are defined. In the present scattering calculations, this axis-ratio is used for non-spherical sand particles that are assumed to be ellipsoids. Ellipses rotate around one axis; this axis is the longer axis and it is called prolate, or otherwise, oblate. The geometry of these particles is defined by two parameters: one parameter for the shape and one for the size. If particles are assumed to be ellipsoids, then one needs to know the axes' ratios. Microscopic investigation revealed that the average axes' ratios for the best fit ellipsoid are $1:0.75:(0.75)_2$ [8].

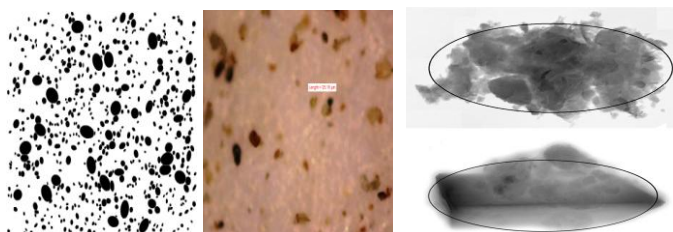


Figure 2. Electron micrograph snapshots of ellipse particles; the approximation of ellipse image axes are $6.3\mu m$ (upper figure) and $11.6\mu m$.

IV. RESULTS AND DISCUSSION

A. Components Measurement

Classical conditioning was assumed prior to using direct measurement in most of the studies and various techniques for calculating PSD was exits. However, the method of numerous direct measurements of shape or particle distribution in dust storm is a complicated task. A 3-dimensional surface is drawn per our samples up to collected samples as shown in Figure 4.B the graph shows similar functional distribution relative to the value of shape and size. Additionally, for the purpose of analysis, ellipse dust particle falling has two situations known as oblate and prolate, as discussed before, assuming that the eccentricity is small. Different methods for particle size analysis are applicable to different size ranges of the particles.

B. PSD diffraction Analytical

Experimental studies of the effects of PSD has been in existence since the past twenty years [8]. Partical diffraction by laser is the most widely used technique for particle size analysis. The methods have developed considerably over the years as had the capability to analyze over broad size range in a variety of dispersion media. Moreover, various methods of analyzing these techniques exist that look at the measured property via a cloud or assemblage of particles and several other forms of particle sizing techniques that do not count or assign a size to individual particles. Determination of particle size distribution and understanding how it affects your products and processes can be critical to the success of drug development, as it has large impact on reactivity, dissolution rate, flow ability. Conversely, to solve this issue, laser diffraction within the sedimentation has been used for the past ten years. The particles pass through an expanded and collimated laser beam in front of a lens where the focal plane is positioned with a photosensitive detector consisting of a series of concentric rings [3]. Distribution of scattered intensity is analyzed by a computer to yield the particle size distribution. Comparison is later made of a "snap-shot" of each speckle pattern with another taken at a very short time (microseconds). The time dependent change in the position of the speckles relates to the change of position of the particles and hence, particle size. The dynamic light signal is sampled and correlated with itself at different time intervals using a digital correlation and associated computer software [5,11]. The relationship of the auto-correlation function obtained at time intervals is processed to provide estimates of the particle size distribution.

Particles pass through a laser beam using a low power and the light scattered by them is collected over the laser beam in a range of angles in the forward direction. It is typically fast at less than 2 minutes to take a measurement and analyze, and it is simple to use. The angles of diffraction are, in the simplest case, inversely related to the particle size [8].

The particles pass through an expanded and collimated laser beam in front of a lens where the focal plane is positioned with a photosensitive detector consisting of a series of concentric rings. Distribution of the scattered intensity is analyzed by a computer to yield the particle size distribution. The PSD of sample 1 is measured using the laser diffraction particle size analysis (SALD, 2300) while the volume of the

PSD is given in four different measurements for each sample (number, size, shape, and cumulative results).

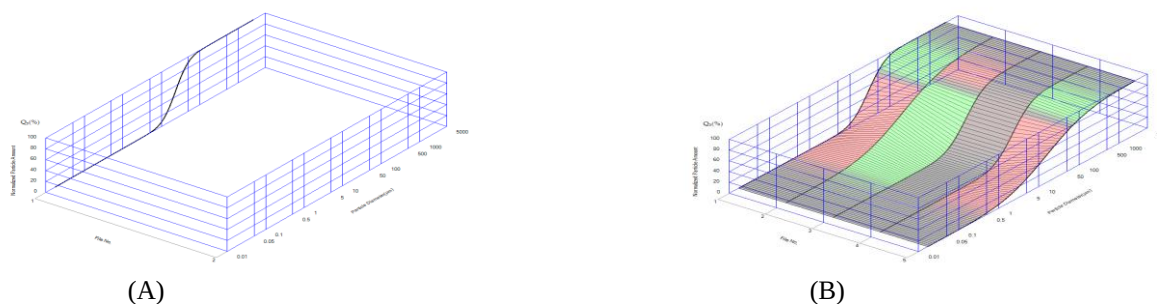


Figure 4 show fragments and dust size distributions at limited time grinding. B) 3D normalized particle amount. B

Our calculation using real dust particles produced many types and ranges of meteorological parameters and percentages of moisture content; Table 1 shows an area of the PSD for dust grains. There are various approaches of performing a particle size analysis and in obtaining the results. Many researchers prefer the direct number answer about the average size. A new and better approach to getting the results is to use both the central points of the distribution along with one or other values to describe the measurement of the distribution [10]. The particle analysis was found to provide a good dust fit particle for a variety of soils.

V. CONCLUSIONS

The knowledge of the behaviors of an ellipsoid particle in dust storm has been of interest in several new fields of Information and Communication Technology such as remote sensing system, interconnection, radar, and advanced industrial devices in monitoring the environment. The particle size distribution analysis tools provide the ability to measure and report the PSD of the sample. Microwave, millimeter wave, and the sub-millimeter band technology frontier are rapidly advancing in meeting the demand of a high level of quiet performance and reduced cost. This includes the increasing number of satellite and terrestrial links that need SDS such as many military and commercial applications and numerous scientific tenders at high frequencies. The extremes of the grain-size distribution were also well adapted to the experimental results, which may help to understand better the research of dissimilar expedients, such as the advanced 5G and surveillance technology, which has been based on the MMW band. Our future work interest is to investigate the impact of PSD and dusty media on radio wave propagation in the optical band.

ACKNOWLEDGMENT

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Design and Analysis of A Low-Profile, Elliptical Patch Super Wide Band (SWB) MIMO Antenna

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Abstract—In this paper, design of a low-profile, coplanar waveguide fed elliptical patch Super Wide Band (SWB) multiple input multiple output (MIMO) antenna for various applications are presented. Two planar elliptical patch antennas with Microstrip-feed are used as the elements of the MIMO antenna. PTFE (Polytetrafluoroethylene) is used as dielectric substrate whose relative permittivity is 3.5. The proposed antenna occupies a wide usable fractional bandwidth of more than 172.12% (2.2473-30 GHz) which covers frequency range for all types of UWB applications along with some other common wireless communication systems such as ISM, Wi-Fi, WiMax, part of LTE and so on. Performance analysis of designed antenna is carried out in free space in terms of reflection coefficient, radiation pattern and voltage standing wave ratio (VSWR) and the results show that, this antenna could be a very suitable one for above mentioned applications.

Keywords—Super Wide Band (SWB); coplanar waveguide; elliptical patch; MIMO; microstrip; PTFE; fractional bandwidth

I. INTRODUCTION

In 2002, the release of the 7.5 GHz of unlicensed spectrum by the U.S. Federal Communications Commission [1] for commercial usages and applications sparked a renewed interest of UWBRT (Ultra Wideband radio technology) in industries, universities and governments. Today “ultra-wideband” usually refers to UWBRT (Ultra Wideband radio technology) where electronic systems should be able to coexist with other electronic users and other legacy technologies [2]. Ultra Wide Band and Super Wide Band systems provide a means for short range high data rate wireless transmission between electronic devices. SWB comprises all of UWB's advance features but with more channel capacity, higher time-precision and super resolution in communication, ranging and screening, respectively [3]. UWB and SWB applications in particular create a considerable challenge, with exceptionally demanding requirements on size, bandwidth and efficiency [2-5].

In recent years, there has been an increasing amount of literature on SWB antenna [2-8]. For example, a novel SWB antenna that achieves a frequency band between 5 to 150 GHz is proposed in [2]. In spite of the huge bandwidth it is not suitable for lower frequency bands such as WiMax and S-band communication.

All wireless technologies face the challenges of signal fading, multi-path, increasing interference and limited spectrum [9-12]. Digital communication using MIMO has emerged as a breakthrough for wireless systems of revolutionary importance. MIMO technology exploits multi-path to provide higher data throughput, and simultaneous increase in range and reliability all without consuming extra radio frequency [13-14].

There are several definitions of bandwidth circulated among the antennas and propagation society; the two definitions which most frequently used, are the percent bandwidth and the ratio bandwidth [5, 7]. The ratio bandwidth is defined as follows: $B_R = BW / f_L$; Where BW is the nominal bandwidth defined by $BW = f_H - f_L$. Here, f_H , f_L are the maximum and minimum frequency at -10 dB, respectively, B_R is the ratio bandwidth, commonly noted as $B_R = R: 1$, where R is the ratio defined as $R = f_H / f_L$. The term SWB is used to indicate a greater bandwidth than a decade bandwidth that is it is often used for describing a ratio bandwidth of 10:1 or larger [5-6]. The operating frequency of designed antenna is 2.2473-30 GHz which exhibits a ratio bandwidth of 13.35:1. Therefore this antenna can be considered to be a super wide band (SWB) MIMO antenna.

FCC has allocated (3.1-10.6 GHz) frequency band for Imaging, Communication and Measurement Systems and (22-29 GHz) band for Vehicular Radar Systems which are all UWB applications [1]. ISM bands are (2.45, 5.8 and 24.125) GHz, Wi-Fi uses 2.4 GHz and 5 GHz radio bands, WiMax technology operates on 2.3, 2.5 and 3.5 GHz frequency bands and some part of LTE technology operates on greater than 2.3 GHz frequency band [3]. The designed antenna covers all these frequency bands so, it can be said that this antenna is a very suitable one for these applications. In this paper, a low-profile, elliptical patch SWB MIMO antenna is designed with relative permittivity 3.5. All the design and simulation have been done with CST Microwave Studio Suite Software.

II. ANTENNA DESIGN AND IMPLEMENTATION

The designed antenna operates from the frequency of 2.2473 GHz to 30 GHz. The antenna is designed on Teflon as substrate which is RO3035TM series high frequency circuit material of Rogers Corporation with $\epsilon_r = 3.5$, named PTFE (Polytetrafluoroethylene). Properties of substrate material

include high tensile strength, high resistance to stretching, rigid, non flammable, durability. It can be used in applications up to 30-40 GHz [15]. Each radiating disc of the antenna is fed by a 50 ohm Microstrip feed line. A conductive ground plane with a length of 21 mm is placed on the same side of the disc. The patch and ground planes are made of copper threads. Fig. 1 and 2 show the top and side view of the proposed antenna.

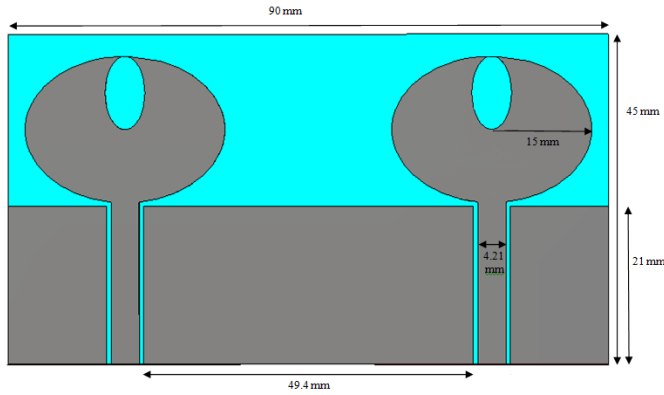


Fig. 1. Top view of designed Elliptical Patch SWB MIMO Antenna.

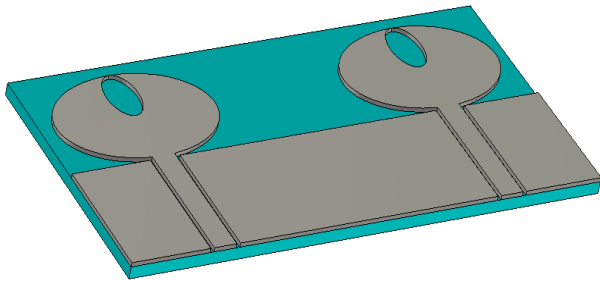


Fig. 2. Side view of designed Elliptical Patch SWB MIMO Antenna.

In addition, there is an elliptical shape slot on the upper hemisphere of each radiating patch. In Table I the antenna measurements are shown.

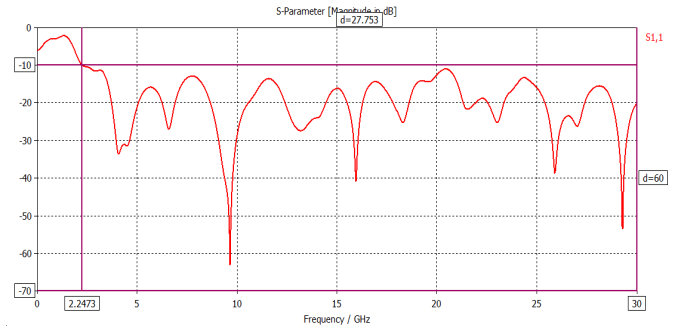
TABLE I. MEASUREMENTS OF ELLIPTICAL PATCH SWB MIMO ANTENNA

Parameters	Dimensions (mm)	Material
Substrate:		RO3035™ series circuit material ($\epsilon_r = 3.5$ and $\tan \delta = 0.0017$)
Length	45	
Width	90	
Thickness	2	
Elliptical Patch:		Copper
Major Radius	15	
Thickness	0.7	
Ground Plane (Left & Right):		Copper
Length	21	
Thickness	0.7	
Ground Plane (Middle):		Copper
Length	21	
Width	49.4	
Thickness	0.7	
Feed Line:		Copper
Width	4.21	
Thickness	0.7	

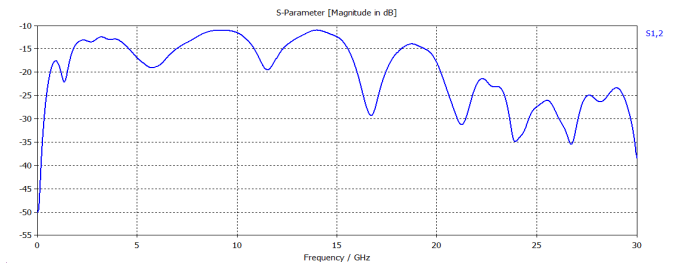
III. SIMULATION RESULTS AND ANALYSIS

A. Reflection Coefficient

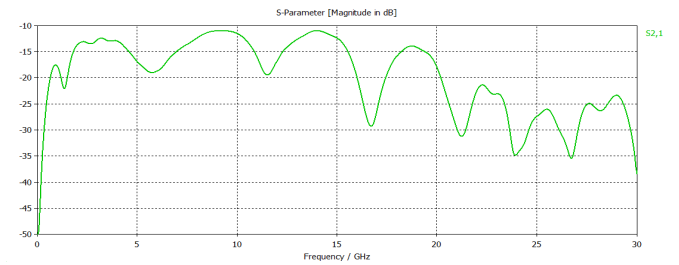
The S-parameter plot of designed antenna in Fig. 3 shows that a wide bandwidth of almost 27.7527 GHz starting from 2.2473 GHz to 30 GHz has been achieved for all $S_{11} < -10$ dB. The isolation between patches is quite good as the transmission co-efficient (S_{12} & S_{21}) are below -10 dB [11]. It has got a good impedance matching all over its occupied bandwidth.



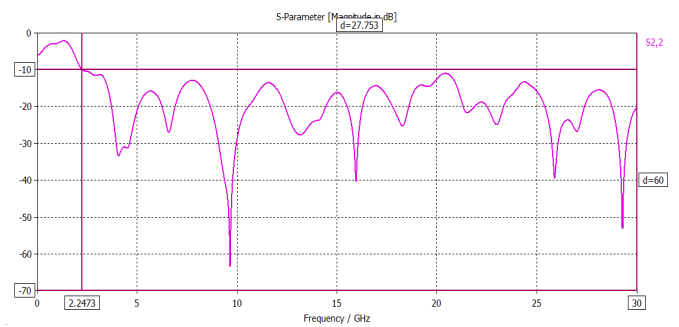
(a) S_{11}



(b) S_{12}



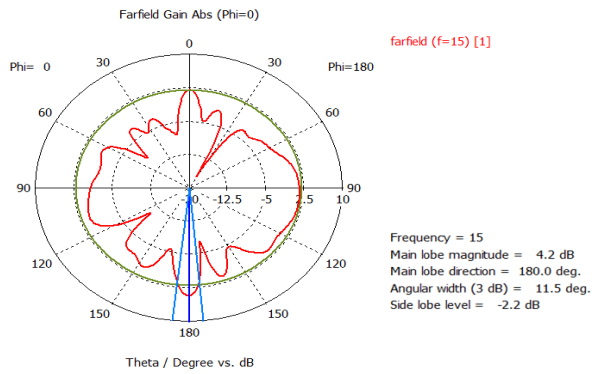
(c) S_{21}



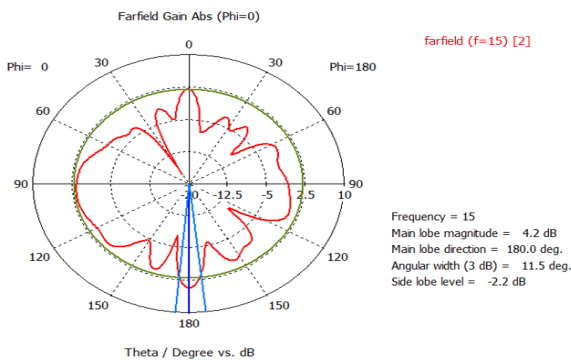
(d) S_{22}

Fig. 3. Simulated S parameters ((a) S_{11} , (b) S_{12} , (c) S_{21} , (d) S_{22}) of designed Elliptical Patch SWB MIMO Antenna.

B. Radiation Pattern



(a) For left side antenna



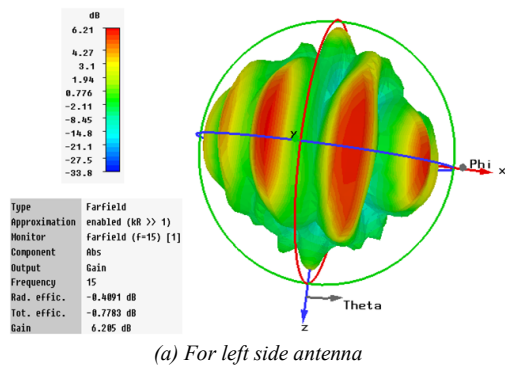
(b) For right side antenna

Fig. 4. Radiation Pattern of E-Plane of (a) left side antenna and (b) right side antenna of designed Elliptical Patch SWB MIMO Antenna.

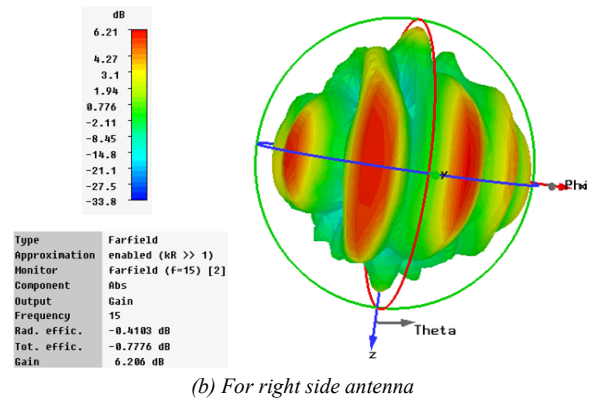
Fig. 4 (a) and (b) show the power for both antennas, is radiated mostly in the lower and upper hemisphere; the power is directed towards 180 degree with main lobe magnitude of 4.2 dB. 3 dB angular beamwidth (i.e., HPAW) is 11.5 deg. Omni directional radiation pattern is observed.

C. 3D Radiation Pattern and Gain

The simulated result of Omni directional radiation pattern over free space is shown in 3D in Fig. 5. The result shows the gain of left side antenna is 6.205 dB and right side antenna is 6.206 dB. Directivity and total efficiency of the antennas are quite efficient.



(a) For left side antenna

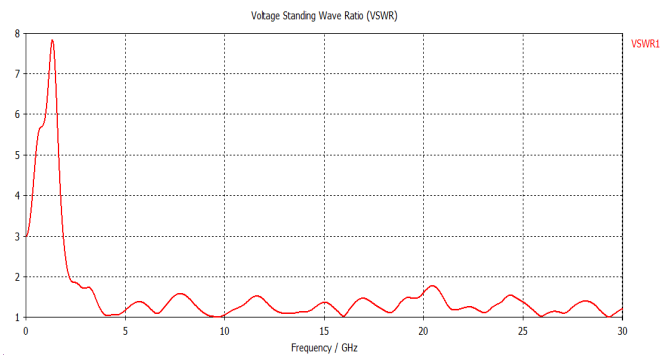


(b) For right side antenna

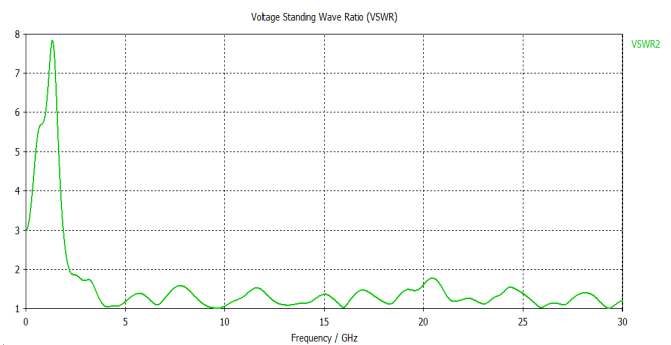
Fig. 5. 3D Radiation Pattern of (a) left side antenna and (b) right side antenna of designed Elliptical Patch SWB MIMO Antenna.

D. Voltage Standing Wave Ratio (VSWR)

Fig. 6. Shows VSWR of proposed antenna are below 2 which are considered well for an antenna [5, 16]. For designed antenna, VSWR below 2 means a good impedance matching between antenna and transmission line. Thus a very less number of incident waves reflect back to the source.



(a) For left side antenna



(b) For right side antenna

Fig. 6. Simulated VSWR of (a) left side antenna and (b) right side antenna of designed Elliptical Patch SWB MIMO Antenna.

IV. VALIDATION OF DESIGNED ANTENNA

In this paper, proposed antenna (90mm×45mm) has a ratio bandwidth of 13.35:1 (2.2473-30 GHz) with low-profile, planar configuration. In [17], a SWB MIMO antenna is proposed whose ratio bandwidth is 6:1 (2-12 GHz) with a very complex slotted patch shape; in [2], an elliptical patch shaped SWB antenna with a high ratio bandwidth that is 21:1 (0.41-8.86 GHz) is proposed but it has a very large patch dimension; in [18], a rectangular shaped MIMO antenna (40mm×42mm) is proposed whose ratio bandwidth is 2.3:1 (3.1-7.2 GHz), and it does not even cover the UWB frequency range. Table II illustrates these comparative analyses of validation.

TABLE II. COMPARATIVE ANALYSIS OF VALIDATION

Antennas	Patch Shape	Ratio Bandwidth
Proposed SWB MIMO Antenna	Slotted Elliptical	13.35:1 (2.2473-30 GHz)
Ref [17]	Complex slotted ellipse-shaped	6:1 (2-12 GHz)
Ref [2]	Elliptical shaped	21:1 (0.41-8.86 GHz)
Ref [18]	Rectangular	2.3:1 (3.1-7.2 GHz)

V. CONCLUSION

In this endeavor, designing a low-profile, elliptical patch SWB MIMO antenna is proposed which has a wide usable fractional bandwidth of more than 172.12% (2.2473-30 GHz). The aim of this paper is to observe the performance and characteristics of the proposed antenna in free space for UWB applications as well as some other wireless communication systems such as ISM, Wi-Fi, WiMax, part of LTE and so on. The designed antenna provides good performance over free space throughout the occupied bandwidth. The key finding of this paper lies in the fact that, this single SWB MIMO antenna can be used for all types of UWB devices which is described by Federal Communications Commission (FCC) in 2002 because the designed antenna covers both (3.1-10.6 GHz) and (22-29 GHz) frequency ranges and also frequency bands for ISM, Wi-Fi, WiMax and part of LTE system. Reducing the size with improved radiation performance could be the focus towards the further step of this design.

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Molecular Simulation of Hydrogen Plasma Irradiation into Bubble-formed Tungsten Material

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Abstract—Binary-collision-approximation-based simulation is performed for the investigation of the penetration depth and the reflection rate on tungsten material with helium bubbles on the surface. Simulation results show that helium bubbles prevent penetration and increase the absorption rate.

Keywords—plasma-material interaction; tungsten; helium bubble; molecular simulation; binary-collision-approximation

I. INTRODUCTION

Tungsten material is attracting many researchers because of the use of a plasma facing material in nuclear fusion device because of its excellent features such as high-melting point, high threshold energy of physical spattering, and high reflection rate of hydrogen isotope and impurities. For example, tungsten material is planning to employ for divertor plates in ITER (International Thermonuclear Experimental Reactor). Because the divertor plates directly contacts with energetic plasmas, the erosion of the surface of the divertor plates is inevitable. Some changes of the characteristics as plasma facing material are caused by the erosion. Therefore, it is necessary to reveal the change of the characteristics to design the divertor system.

Many experimental studies reported bubble formation on the surface of tungsten materials under helium plasma irradiation in wide range of incident energy [1-4]. The lowest incident energy for such defect was found to be about 6 eV [2]. Helium ash is generated when nuclear fusion between deuterium and tritium is caused. Therefore, the bubble formation might be observed on the divertor plates in ITER. Although the plasma temperature around the divertor plates is relatively low in the steady state operation, the hydrogen plasma is injected into divertor plates with higher energy when transient thermal events such as disruption occur. In this study, therefore, binary-collision-approximation-based (BCA) simulation is performed for the investigation of the effects of the existence of helium bubbles on the penetration depth and the absorption rate of incident hydrogen atoms under the condition that the incident energy is less than 1 keV.

II. SIMULATION METHOD [5-8]

Figure 1 shows the schematic diagram of BCA simulation. A projectile is injected into a target material. The projectile collides with the nearest neighbor atom. The other interaction between the projectile and its surrounding atoms is neglected in BCA simulation. The momentum of the projectile changes at the collision, then the projectile collides with the next target atom. In the collision process of BCA, the motion of target atoms which collide with the projectile depends on their received kinetic energy E_t . When E_t is less than the threshold of displacement energy of target atom E_d , the target atom does not move while the target atom recoils when E_t is larger than E_d . The motions of recoil atoms are solved in the same procedure as the projectile. The scattering angle of the projectile and the target atom at each collision is obtained analytically in a two-body interatomic potential $V(r)$, where r is the distance between the projectile and the target atom. In our simulation code ACVT, the following Moliere approximation to the Thomas-Fermi potential [10] is employed for $V(r)$:

$$V(r) \equiv \frac{Z_1 Z_2 e^2}{r} \Phi\left(\frac{r}{a}\right), \quad (1)$$

$$\Phi(x) \equiv 0.35e^{-0.3x} + 0.55e^{-1.2x} + 0.10e^{-6.0x}, \quad (2)$$

where Z_1 and Z_2 are the atomic numbers of the projectile and

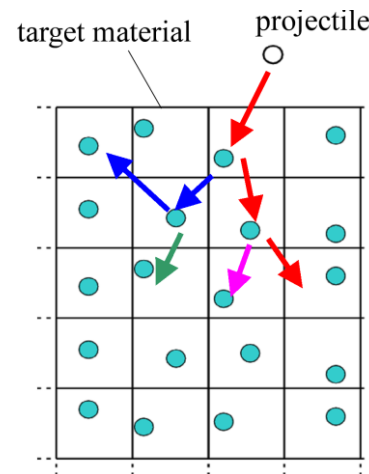


Fig. 1. Schematic diagram of binary collision approximation. [6]

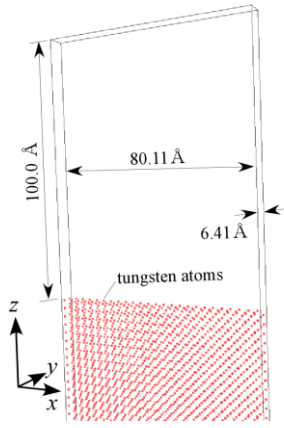


Fig. 2. Simulation model with the xyz coordinates of our simulation. The red dots denote the positions of tungsten atoms. [13]

the target atom, e is the electronic charge, and a is the screening length, which depends on Z_1 and Z_2 [11].

Our BCA simulation code ACVT [5-9] can perform two kinds of simulation. One of them is “cumulative simulation” which takes into account structural changes of the target material and accumulation of incident atoms by injection. The other one is “independent simulation” which refreshes the target material to be the initial target material before each injection.

III. SIMULATION MODEL

For the preparation of the investigation, we performed BCA simulation to generate the helium bubble structure on the surface of tungsten materials at first. Then we employed another BCA simulation to estimate the penetration depth and absorption rate of incident hydrogen atoms for the bubble formed tungsten material. These two processes are explained in this section.

A. Generation of Helium Bubbles

Figure 2 shows the system and xyz coordinates of our simulation. Tungsten atoms are initially arranged as perfect bcc

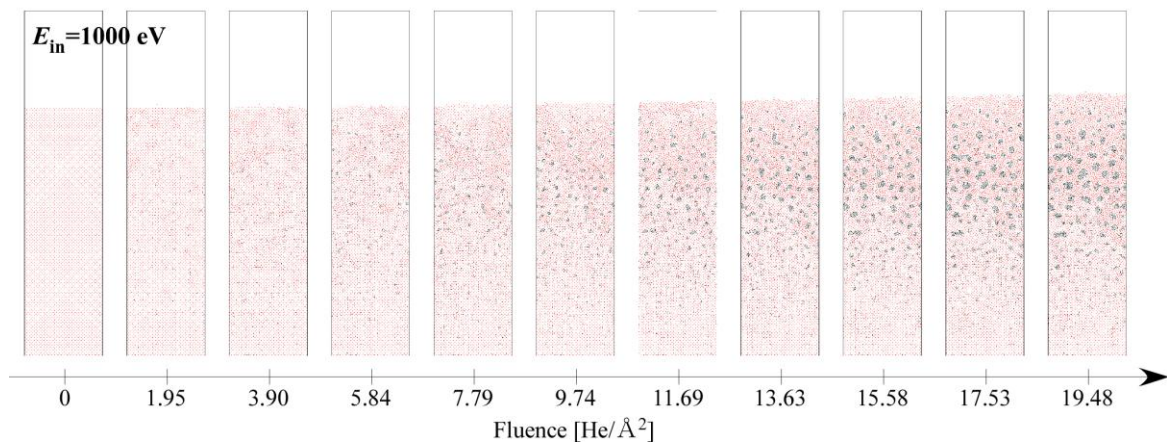


Fig. 3. Time evolution of surface of tungsten material irradiated by helium plasma. The red and black dots denote the positions of tungsten and helium atoms, respectively.

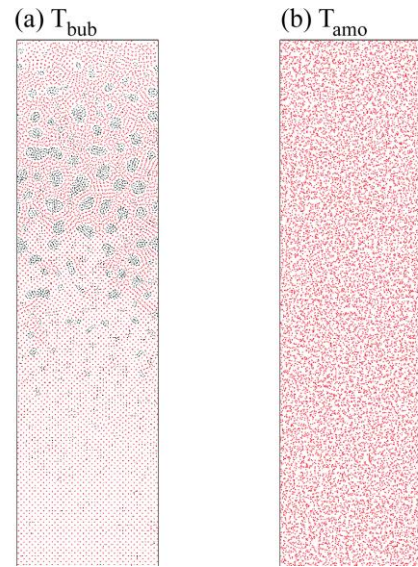


Fig. 4. Structure of two types of target materials.

crystalline structure. The lattice constant is set to 3.2046 Å. The size of target material is set to 80.11 Å × 6.41 Å × 8011.5 Å. The z-axis of the simulation box is set parallel to the edge of the target material whose length is 8011.5 Å. Periodic boundary conditions are used in x- and y-directions. The temperature of the target material is set to 0 K. The threshold of displacement energy of tungsten atom E_d is set to 38 eV. The incident energy is set to 1000 eV. A total of 10,000 helium atoms are injected one by one into the target material. The incident angle is set to parallel to z-axis, i.e., perpendicular to (100) surface of bcc crystal. The x- and y-coordinates of the starting position of the helium atom are set randomly. The z-coordinate of the starting position is set 100 Å above the surface of the target material. “Cumulative simulation” is employed to take into account the structural changes of the target material and accumulation of helium atoms by injection. The calculation of structural relaxation of the target material is performed every 1000 injections. For the calculation of the relaxation, the following tungsten-helium potential [12] is employed.

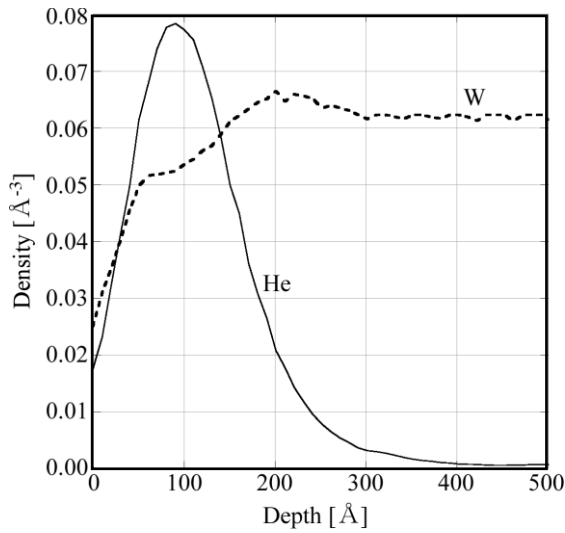


Fig. 5. Depth profile of helium and tungsten densities of T_{bub} .

$$U(\{\mathbf{r}\}) \equiv \frac{1}{2} \sum_{i,j \neq i} V_{Z_i Z_j}(r_{ij}) - \sum_i^W F(\rho_i(\{\mathbf{r}\})) \quad (3)$$

where $r_{ij} = |\mathbf{r}_i - \mathbf{r}_j|$ is the distance between the i -th and j -th atoms. The two-body potential function $V_{Z_i Z_j}(r_{ij})$ is a polynomial functions based on AFS potential. The atomic numbers (Z_i) are 74 for tungsten atoms and 2 for helium atoms. The function $F(\rho_i(\{\mathbf{r}\}))$ is an embedded energy. The symbol $\sum^W$ indicates summation over only the tungsten atoms. For more detail of the relaxation process, please see [13].

B. Calculation of Penetration Depth and Absorption Rate

The target material at fluence of 19.48 Å^{-2} in previous process is used for the target material of another BCA simulation for the estimation of the penetration depth and absorption rate of incident hydrogen atoms. To clarify the effects of helium bubbles, the tungsten in amorphous structure (i.e., structure which tungsten atoms are placed at random positions) is also used for the target material. Let us use T_{bub} and T_{amo} to indicate these two types of target materials. Figure 4 shows the structure of T_{bub} and T_{amo} . Figure 5 shows the density of helium and tungsten atoms in T_{bub} . The helium bubbles are spread over the depth of approximately 200 Å from the surface. The tungsten density in the bulk of T_{bub} is 0.0623 Å^{-3} . The tungsten density of T_{amo} is set to the same value as T_{bub} .

“Independent simulation” is performed to obtain statistical values for multiple injections into the fixed target structures T_{bub} or T_{amo} . A total of 10,000 hydrogen atoms are injected into each target material for fixed incident energy E_{in} from 10 eV to 1000 eV. The incident angle is set to parallel to z-axis. The mean penetration depth and the absorption rate are calculated for 10,000 injections.

IV. RESULTS

Figure 6 (a) shows the incident energy dependence of the mean penetration depth. The penetration depth of T_{bub} is shorter than that of T_{amo} because helium bubbles prevent the motions of hydrogen atoms. The difference is approximately 2-

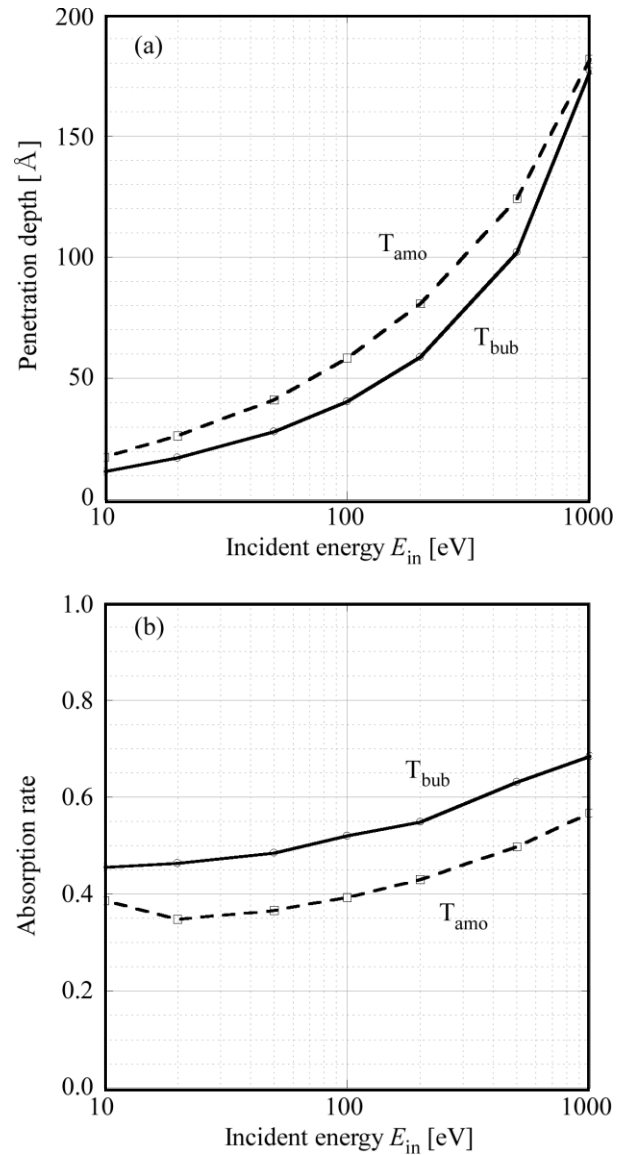


Fig. 6. Incident energy dependence of (a) mean penetration depth and (b) absorption rate for T_{bub} and T_{amo} .

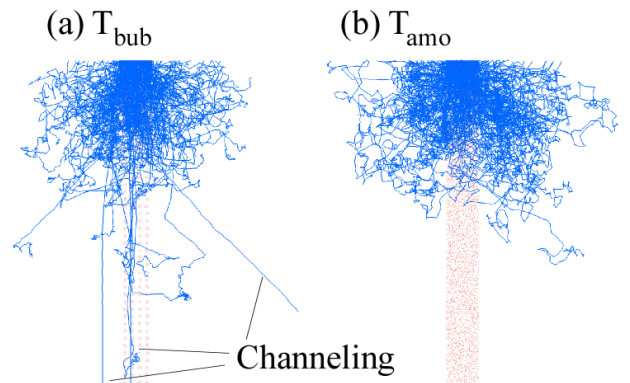


Fig. 7. Trajectory of incident hydrogen atoms into T_{bub} and T_{amo} .

40 %. The transferred energy from the projectile into target atom for each binary collision becomes larger when the mass ratio of the projectile and target atom approaches to 1. Therefore, the hydrogen atom loses more kinetic energy when it collides with helium atoms than the case of collision with tungsten atoms. Therefore, the penetration depth becomes shorter when helium bubbles exist in tungsten materials. The difference of penetration depth for T_{bub} and T_{amo} becomes small as incident energy increases because the rate of the hydrogen atoms, which pass through the bubble-formed area (i.e., approximately 200 Å from the surface), increase as incident energy increases. When incident energy is 1000 eV, the penetration depth of T_{bub} is almost the same as that of T_{amo} .

Figure 6 (b) shows the incident energy dependence of the absorption rate. The absorption rate of T_{bub} is higher than that of T_{amo} because incident hydrogen atoms easily stop in helium bubbles as explained above. The difference is approximately 15-25 %. Different from the case of the penetration depth, the difference between the absorption rate of T_{bub} and T_{amo} does not become small even though the incident energy is 1000 eV, because hydrogen atoms which pass through the bubble formed area hardly return to the surface.

Figure 7 shows 200 trajectories of incident hydrogen atoms in the case of $E_{\text{in}}=100$ eV. Most of hydrogen atoms stop near the surface where helium bubbles exist in the case of T_{bub} . However, the other some hydrogen atoms cause channeling and reach deep place because of the bcc crystalline structure of bulk of T_{bub} . The ratio of the channeling atoms increases as incident energy increases.

Figure 8 (a) shows the plots of stopped positions of incident hydrogen atoms in T_{bub} . From this figure, it is clear that helium bubbles prevent the motions of hydrogen atoms when incident energy is low. As a result, most of hydrogen atoms stop in the bubbles when $E_{\text{in}}=100$ eV, while most of hydrogen atoms do not stop in the bubbles when $E_{\text{in}}=1000$ eV.

V. CONCLUSION

BCA simulation is performed to estimate the penetration depth and absorption rate of incident hydrogen atoms for the case of hydrogen plasma irradiation into bubble formed tungsten materials. The results show that the penetration depth and absorption rate of T_{bub} is lower and higher than that of T_{amo} , respectively. Therefore, it is expected that retention of hydrogen isotopes increases when the helium bubbles are formed on the divertor plates.

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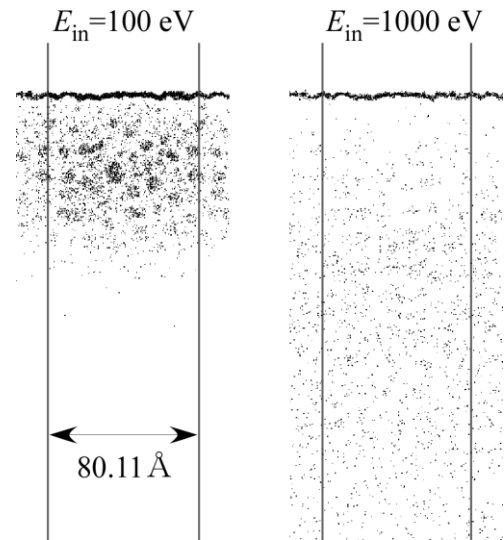


Fig. 8. Plots of stopped positions (black dots) of incident hydrogen atoms.

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Demonstration on Performance Characteristics of Microstrip Low-Pass Filter for Various Substrates Employing Tschebycheff Distribution

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Abstract— In modern microwave design sector, Electromagnetic Band Gap (EBG) structures which is a unique innovation by virtue of their high potency and low complexity in designing with integrated circuits. This article examines the relationship between operating performance of different substrates and Tschebycheff polynomial distribution. A number of methods, method of moments (MOM) for numerical analysis and Tschebycheff polynomial distribution for designing the traditional circular pattern on ground plane are used. In like manner, the parametric comparison has been demonstrated in view of the selectivity, 20-dB insertion loss bandwidth (20-dB IL BW) and 10-dB return loss bandwidth (10-dB RL BW). The consequence of that is, two substrates FR4 and RT/Duroid 5870 prevail with good matching upto 70 dB, negligible ripple less than 0.7 dB in the passband and wider, deeper stopband performance upto 6-7.5 GHz. Improving compactness of the circuit with high efficiency, this complication is solved by Tschebycheff polynomial distribution.

Keywords—EBG Structures; Tschebycheff Polynomial Distribution; Selectivity; 20-dB IL BW, 10-dB RL BW.

I. INTRODUCTION

Recently, subset of metamaterials Electromagnetic Band Gap structures (EBGs) have grown in popularity because of some quite extraordinary, regular patterned, sophisticated properties. However, such high level properties are quite uncommon in a single natural material [1]. EBGs exhibit wider bandpass, steeper and deeper band rejection properties. In this present day, EBG assisted microstrip filters are frequently playing a role in modern microwave and RF engineering. Nevertheless, the spurious transmission other than fundamental frequencies, play a negative role on the highly efficient broadband system. To minimize the spurious transmission and register the compactness, numerous microwave and millimeter wave devices have been fabricated. To design EBG structure on a particular substrate, some representative parameters need to be preserved effectively. Such as the element spacing, operating frequency, width, height of the substrate, thickness, irregularity of the surface, volumetric ratio of the EBG element, pattern of the elements which have influence on bandpass and band rejection properties. In addition, other complication in EBG circuit due to the radiation from aperiodic etching with no proper formulation as a result of the inappropriate selection of dielectric substrate [2].

In this exegesis, applying Tschebycheff polynomial distribution with traditional circular patterned EBG structures

inscribe on the ground plane of a microwave transmission line on the discrete RT/Duroid substrates is proposed. On the whole, microwave assisted low pass filter return loss, insertion loss and selectivity enhanced performance where Tschebycheff polynomial distribution are employed to fabricate the filter in various substrates, is the novelty of the exegesis. Developing wider, deeper stopband and suppressing ripples in the passband are great features of the Tschebycheff polynomial. A unique measure of the Tschebycheff polynomial the side lobe level (SLL) is exploited to manage the width of stopband insertion loss, the depth of passband return loss, ripples in both passband and stopband, selectivity. Consequently, with the Tschebycheff distribution ripple, width and depth of the stop bandwidth can be mitigated with the improvement of rate of roll-off of the transfer function at cut-off [3]. The effective value of the permittivity will change with the changing of the substrate dielectric constant. Furthermore, changing permittivity will influence the course of the guided wave length and circular pattern spacing will alter according to the formulation of Bragg's theorem. Likewise, substrate height will play a vital role in case of width for the standard 50 Ω transmission line.

The low pass filter utilizing the Tschebycheff polynomial distribution not only has some specific ground of originality but also some interesting features, which includes the following:

- a. Total 5 substrates are used:
 - 1) Same dielectric constant with different thickness
 - 2) Distinct dielectric constant with same thickness
 - 3) A mid-level dielectric constant with thickness matching with another substrate
- b. Tschebycheff-1 and Tschebycheff-2 distribution are used and traditional circular pattern is simulated for all substrates.
- c. Stop bandwidth is steeper, wider and deeper than a conventional uniform distributed low pass filter [3].
- d. The insertion loss, ripple is negligible and significant improvement in matching in the passband.
- e. Compactness of the circuit can be varied as per dielectric constant (ϵ_r) and periodic circular pattern is engraved with conventional polynomial formulation.

Consequence of the Tschebycheff distributions with circular pattern in five distinct RT/Duroid substrate are being demonstrated, compared and observed by interpretation.

II. DESIGN THEORY OF STRUCTURES AND POLYNOMIAL DISTRIBUTION

Tschebycheff polynomial has a convenient characteristic of equal ripple characteristic which is observed for a selective side lobe level (SLL). The Tschebycheff excitation coefficient are determined by the SLL. The recurrence formula [4] for Tschebycheff polynomial:

$$T_m(z) = 2zT_{m-1}(z) - T_{m-2}(z) \quad (1)$$

Every polynomial can be evaluated using

$$T_m(z) = \cos[m \cos^{-1}(z)]; -1 \leq z \leq +1 \quad (2)$$

$$T_m(z) = \cosh[m \cosh^{-1}(z)]; z < -1, z > +1 \quad (3)$$

As major to minor lobe ratio for 26 dB [4], the amplitude of coefficient will be:

$$0.36 \ 0.49 \ 0.71 \ 0.89 \ 1 \ 0.89 \ 0.71 \ 0.49 \ 0.36$$

The volumetric ratio of each circular element will be:

$$FF = \frac{r}{a} \quad (4)$$

Where, Filling Factor (FF), radius of each element (r), spacing between two elements (a).

By varying dielectric constant (ϵ_r) of the substrates and radius of the element will adjust according to the polynomial coefficient.

III. ANALYSIS OVER DISTINCT SUBSTRATE WITH POLYNOMIAL DISTRIBUTION

The conventional circular frame of design will be delineated among the five substrates which are (a) RT/Duroid 5880, $\epsilon_r = 2.20$, $h = 20$ mil (b) RT/Duroid 5870, $\epsilon_r = 2.33$, $h = 20$ mil (c) FR4, $\epsilon_r = 4.4$, $h = 20$ mil (d) RT/Duroid 6010 LM, $\epsilon_r = 10.2$, $h = 25$ mil (e) RT/Duroid 6010 LM, $\epsilon_r = 10.2$, $h = 50$ mil. Besides, both Tschebycheff type-1 and type-2 are applied, center frequency of 10GHz remain constant throughout the inspection. Selectivity, 20-dB IL BW and 10-dB RL BW, this three significant parameters will be taken account to find out the substrate with the best, reliable and superior efficacy.

TABLE I : COMPARATIVE STUDY AMONG THE PERFORMANCE OF SUBSTRATES WITH TSCHEBYCHEFF TYPE-1 DISTRIBUTION; $FF = 0.375$ REMAIN CONSTANT.

Substrate	Selectivity	Stopband isolation	20-dB IL BW	10-dB RL BW	3-dB Cut-Off	Frequency	1 st Null & Peak
ϵ_r ; h [mil]	[dB/GHz]	S_{21max}	f_0	[GHz]	[GHz]	Frequency	Deviation
		[dB]	[GHz]			[GHz]	%
2.20; 20 mil	50	62.5	9.94	6.26	6.98	6.74	0.60
2.33; 20 mil	44.74	75.21	10	6.34	7.15	6.93	0.00
4.40; 20 mil	94.44	81	10.40	6.40	7.70	7.62	4.00
10.2; 25 mil	20.48	39.98	10.44	3.70	8.29	8.02	4.40
10.2; 50 mil	17.17	20.16	9.82	0.57	8.78	8.44	-1.80

A. Effect of Substrate with Tschebycheff Type-1 Distribution :

In Tschebycheff type-1 polynomial distribution, amplitude of the coefficient are proportional to the semi diameter of the EBG circle. Besides, various substrate is simulated with type-1 distribution by *Zeland IE3D*. From Table-1, FR4 substrate shows profound result and more superior operation.

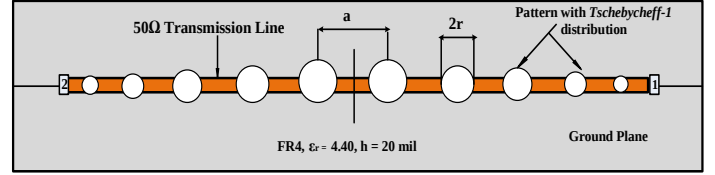


Fig. 1: Schematics of designed FR4 substrate structure low pass filter with the proposed Tschebycheff-1 distribution, when spacing a and semi diameter r are 320 mil and 120 mil.

Demonstration of FR4 substrate with Tschebycheff-1 reveals superior efficacy in selectivity, deeper and wider stopband isolation.

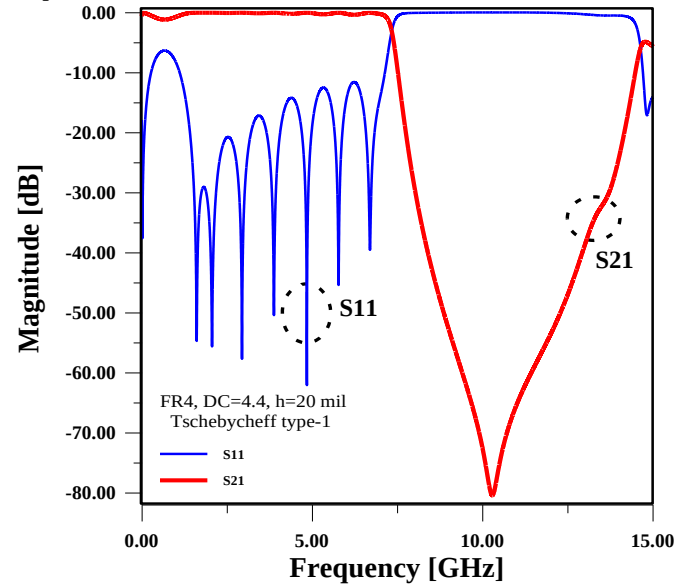


Fig. 2: Simulation result for Tschebycheff-1 distribution for FR4 substrate.

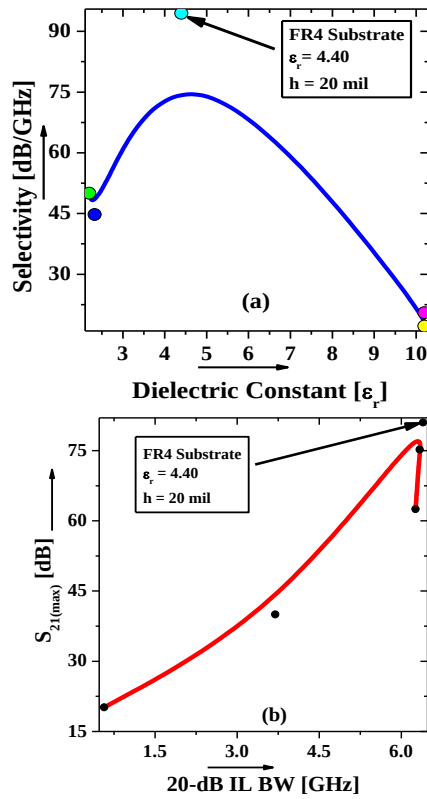


Fig. 3: Fluctuation of (a) selectivity vs ϵ_r ; (b) maximum insertion loss vs 20-dB IL bandwidth for Tschebycheff type-1 distribution for distinct substrates.

In Fig. 2, passband has very small ripple less than 0.3 dB, high impedance matching upto 63 dB, stopband has wide bandwidth with very good reflection. In Fig. 3(a), FR4 has the comprehensively maximum selectivity among the substrates and Fig. 3(b), insertion loss is elevated proportionately with the 20-dB IL bandwidth. Indeed, FR4 prevails amid all the substrates.

B. Effect of Substrate with Tschebycheff Type-2 Distribution :

In Tschebycheff type-2 distribution, coefficient's amplitudes will commensurate according to the circumference (πr^2) of EBG pattern. From Table-2, between type-2 distribution in various substrates, RT/Duroid 5870 prevails with less intricacy passband and superior efficacy.

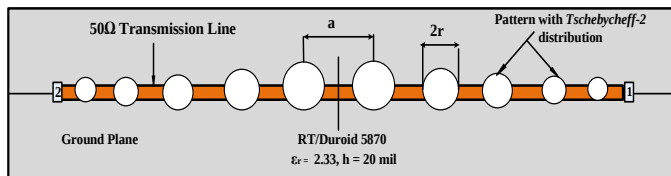


Fig. 4: Schematics of designed RT/Duroid 5870 substrate structure low pass filter with the proposed Tschebycheff type-2 distribution, when spacing a and semi diameter r are 418 mil and 157 mil.

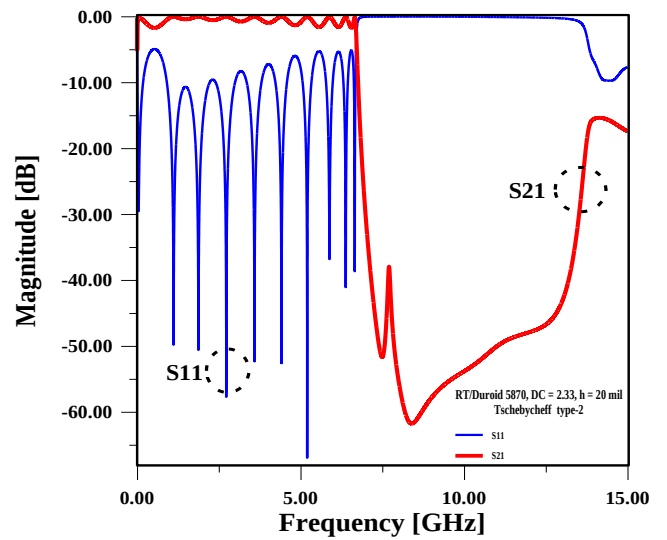


Fig. 5: Simulation result for Tschebycheff-2 distribution for RT/Duroid 5870.

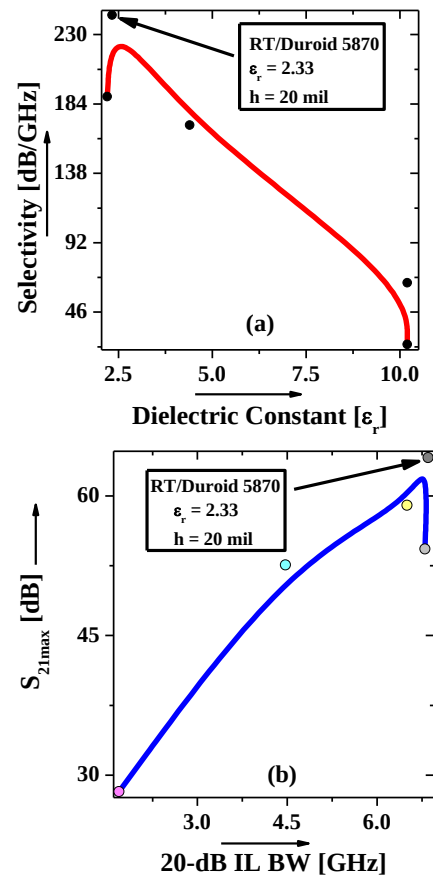


Fig. 6: Fluctuation of (a) selectivity vs ϵ_r ; (b) maximum insertion loss vs 20-dB IL bandwidth for Tschebycheff type-2 distribution for distinct substrates.

In Fig. 5, impedance matching upto 70 dB and ripple is less than 0.8 dB. RT/Duroid 5870 shows very sharp selectivity in Fig. 6(a) and Fig. 6(b) shows high efficacy in stop bandwidth. It shows deeper and wider stopband which is main concern.

TABLE II: COMPARATIVE STUDY AMONG THE PERFORMANCE OF SUBSTRATES WITH TSCHEBYCHEFF TYPE-2 DISTRIBUTION; $FF=0.375$ REMAIN CONSTANT.

Substrate ϵ_r, h [mil]	Selectivity [dB/GHz]	Stopband isolation S_{21max} [dB]	f_0 [GHz]	20-dB IL BW [GHz]	10-dB RL BW [GHz]	3-dB Cut-Off Frequency [GHz]	Frequency Deviation %	1 st Null & Peak [dB]
2.20; 20 mil	188.89	54.31	10	6.80	6.91	6.91	0.00	Null 29 Peak 4
2.33; 20 mil	242.86	64.11	10.05	6.85	6.93	6.93	0.50	30 6
4.40; 20 mil	170.00	69	10.60	6.50	7.70	7.60	6.00	29 5
10.2; 25 mil	65.39	52.58	10.55	4.47	8.20	8.05	5.50	31 10
10.2; 50 mil	24.64	28.23	9.91	1.69	8.52	8.41	-0.90	38 16

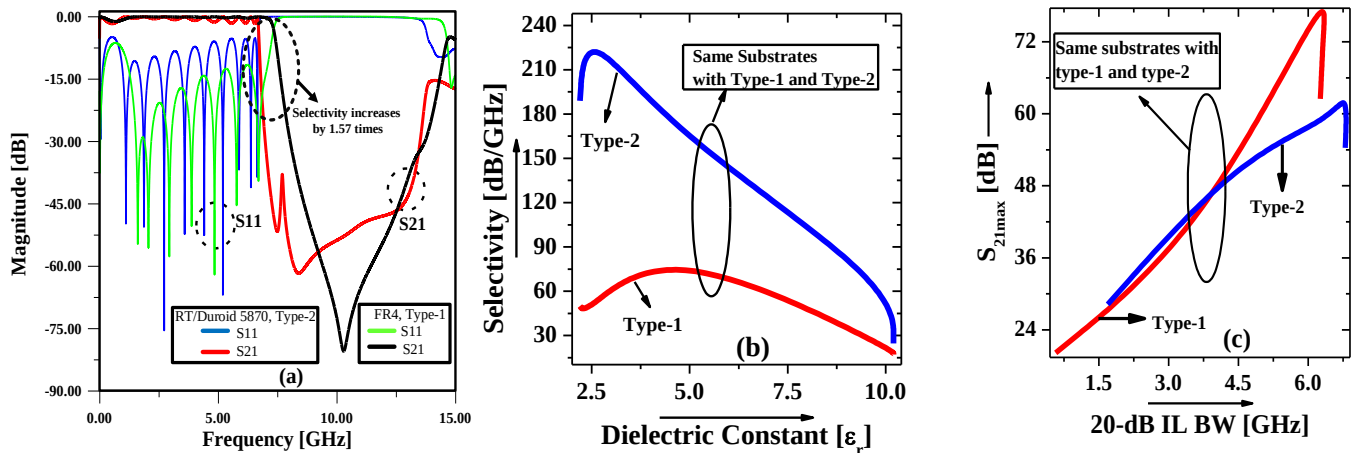


Fig.7: Simulated S-parameters of a 50Ω transmission line on a Tschebycheff circular pattern EBG on different substrates both type-1 and type-2; (a) type-1 vs. type-2; (b) selectivity vs dielectric constant; (c) maximum IL vs 20-dB IL bandwidth. For all cases, $FF= 0.375$ remain constant.

IV. ACCOMPLISHMENT

Designing the filter by Tschebycheff distribution in separate substrates is observed. In type-1 and type-2 distribution, substrates FR4 and RT/Duroid 5870 exhibit selectivity, 20-dB IL and 10-dB RL bandwidth of 94.44 dB/GHz, 6.4 GHz, 7.7 GHz and 242.86 dB/GHz, 6.85 GHz, 6.93 GHz respectively. Naturally, at high frequency FR4 is a high lossy substrate [5]. Still, for using type-1, it displays negligible insertion loss in the passband with ripple less than 0.3 dB, sharp selectivity, good matching and very wide stop bandwidth upto 7 GHz. Likewise, having low dissipation factor RT/Duroid 5870 also reveals very wide return loss bandwidth upto 7 GHz with ripple not greater than 0.7 dB, excellent matching, sharper cut-off and wider stop bandwidth in type-2. However, as the designing in type-2 distribution is complicated which relates coefficients to circumference, so substrates will give more lossy performance as dielectric constant extends. On the other hand, type-2 displays sharper selectivity than type-1 which is 1.57 times higher from Fig. 7(a) and 7(b). In Fig. 7(c), type-1 exhibits wider and deeper stop band properties for all substrates. In brief, FR4 and RT/Duroid 5870 exhibits more superior performance at type-1 and type -2 respectively.

V. CONCLUSIONS

In this paper, applying the Tschebycheff distribution to enhance the operating performance of substrates is demonstrated. Due to the high efficacy in passband and stopband of this two substrate but their compactness needs to be improved. The complication of trade-off can be solved by polynomial distribution. It is well applicable to active applications for microwave and millimeter-wave integrated circuits.

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Real-Time Hand Detection and Tracking with Depth Values

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Abstract— Aim of this paper is to propose a methodology of real-time hand detection based on skin color model and background subtraction under any complex background with extracting the depth information. With the use of stereo camera calibration and disparity mapping, the depth information of the hand is extracted. Reasonable selection of threshold of skin color model and combining with background difference segmentation, then a bitwise-and operation results hand's binary outline feature clearly. Combining this result with the distance information experiments show accurate detection and tracking of hand with depth measuring from the camera.

Keywords- *Hand detection, Skin color model, Background difference, Contour, Convex Hull, Stereo vision, Camera calibration, Disparity mapping.*

I. INTRODUCTION

Recognition or Pattern recognition is a branch of machine learning that work on recognition of patterns in data those are taken as input. This algorithm provides a methodology that finds out results for all possible inputs and then find out the most matching input based on their individual statistical data. There are many branches under pattern recognition. One is gesture recognition that interprets human gestures using mathematical algorithms.

In past few years hand gesture recognition [1] has been an intriguing technique in different application fields of computer vision. It has been a great approach in human computer interaction, robotics, gaming, system for sign languages and so on. Detecting and tracking human body parts like hand and face and extracting 3D space information [2], [3] from the 2D observation of the tracked models provide position and location of the part to be modelled and recognized. Both in 2D and 3D spaces this type of applications have their categorized potential. Most of the previous methodology preferred RGB cameras [4] and HSV color model to detect and track the hands. In these works, skin color and motion information were used. In [5], first the hand was segmented by merging motion and skin color information. They used motion information because they thought the region is static. But however this methodology may fall down in case of moving skin-colored objects or in the moving complex background. Also, it is difficult to fetch out a general skin color as human skin color varies from

different race to race with different level of illumination properties. Some works relied on skin color found in [2], [6], [7], [11]. Also, in [4], the ToF camera was used to obtain depth images and tracked hand using depth and color information. The problems regarding skin color segmentation are like it has same color pixels in face, arms, clothes or any other skin color object in the background and since human skin color ranges vary, so detecting exact color or maintaining a skin color dataset is not efficient to detect only the hand. Background difference algorithm may extract the tracked hand from any complex background, but the problem with the color cannot be solved so easily. In [7], they proposed methodology that extract static skin color based on YCbCr skin color model and was able to detect the hand from any complex background. But still, this method cannot work properly only for the hand, since sometimes it takes the arm and face color. In this paper, we proposed a methodology that calculates an efficient threshold selection of Cb and Cr that can only detect the hand without counting any other color pixel even near to the color of the hand or the palm. But this also does not solve all the problems. There still remains some problem with distance.

Users interact with the machines using their hands in front of their bodies as easily as any background. Thus the lack of information on the position or location of the hand may disturb the segmentation and other color may also leave a mark of inefficiency. In this case depth measuring methodologies can provide a good solution. One way to implement the method is stereo vision [8]. Elmenzain et al. Combined skin color and 3D depth information to the segment hand region [9]. Argyros et al. developed a 3D hand tracker [2]. This was done through the coupling of the results of two 2D skin color trackers that run easily on two video streams. In [10], a comprehensive calibration algorithm of all parameters of the camera pair was proposed. They worked with correspondence of color and depth images, but distortion of depth values was not corrected. Smisek et al. [12] proposed a depth distortion correction component that is the average of all the remaining values in metric coordinates. We propose a complete calibration for all parameters of the camera pair using image rectification [13] and a disparity distortion correction based on the observed disparity of the nearest object from the camera pair.

II. PREVIOUS WORK

Most of the previous work showed either only color segmentation and then tracking the hand by extracting skin color information using combining color models one by one or with the help of depth measuring. But effectively discriminating other objects from the hand, those have a color like human skin in a complex background combining with depth measuring has not done so far so effectively. Our proposed methodology does not take reasonable threshold of Cr and Cb directly. There is a subtraction between Cr and Cb. It forms a new color channel that efficiently detects the skin color without noises. But as human skin color information varies from different race to race, so depending on the skin color only is not an efficient way to detect the hand in a complex background. That's why depth information plays an important role here. In [4], they used depth images and color information to detect the hand, but problem regarding skin color segmentation is that other parts of the body may have same color pixels and also due to a variety of human skin color ranges. So only maintaining a skin color dataset is not enough. Thus, along with depth information, we proposed a new color channel to distinguish hand color, from other parts of the body as efficiently as possible. In [10], they proposed a calibration algorithm for all parameters of the camera pair. They also worked with color model and depth images, but the distortion of the obtained depth values was not given a proper correction. Also in [12], though they proposed a depth distortion correction component, still some distortion remains, and results camera noises. Our methodology proposes a better depth mapping model using image rectification and disparity distortion correction.

III. PROPOSED METHODOLOGY

The process comes out with the result of foreground binary information and segmented binary information for the reasonable selection of the threshold of the YCbCr skin color model. We did not work with any color sample or classify it rather we took a color range, which is unique and can track human color only. We omitted other colors and objects with the help of background subtraction. Meanwhile, on the other side camera calibration forms rectification matrix and a disparity distortion correction based on the disparity of the nearest object from the camera which minimizes distortion in depth values and reduces camera noises as much as possible. Therefore, distance information and segmented result are merged to form the final output image of the hand in the complex background.

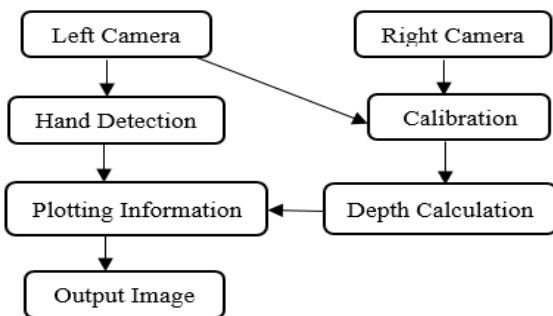


Figure 1: Flow chart of the proposed methodology

IV. DESCRIPTION OF WORK

A. Background Subtraction:

Background difference algorithm [14] is a technique to detect the foreground. An image of the background is taken and set fixed as the background is static here. An image of the object was taken that comes in front of the camera. The background image was subtracted from that image and thus the foreground was detected. Applying threshold a new binary image is extracted out. If the background is denoted by I_b and the input image is denoted by I_n , then foreground may come out from the equation below:

$$I_f = I_n - I_b$$

B. Skin color segmentation:

Skin color segmentation is used to find out what that is in front of the camera is in range of human color or not. We did it based on a skin color model that is used in video or video photography system named YCbCr. It is acronym for Y=luminance, the light intensity which is non-linearly encoded based on gamma corrected RGB, Cb=blue difference and Cr=red difference chrominance. Chrominance is a signal that is used in video to convey color information from the picture, separating those from luminance (Y) signal. Chrominance has two color difference components, U=B-Y (blue – luminance) and V=R-Y (red – luminance). However YCbCr is not an absolute color space rather a way of encoding RGB color information. It can be used as an efficient human skin color model that we noticed here in our work. We totally rejected the luminance part that is Y and worked with CbCr. A new color channel image was extracted by subtracting Cb from Cr and taking the original Cr color image. Applying a threshold on these two color images and merging those, resulted a binary of a mostly maximized mask. Region filling algorithm was used to reduce low density noises. But in small ratio to fetch out a completely segmented binary mask with almost no noises.

The pseudo code for background subtraction and color segmentation is as follows:

```

for each  $I_{no}(x,y) \leftarrow$  capture image
do
    a.  $I_f = I_{no} - I_b$ 
    b.  $I_{n1} = C_r - C_b$ 
    c. Threshold  $I_{n2} = 0 \leq I_{n1} \leq 8$ 
    d. Threshold  $I_{n3} = 0 \leq C_r \leq 134$ 
    e.  $I_{nn} = \text{bit-wise not}(I_{n2}) + \text{bit-wise not}(I_{n3})$ 
    f.  $I_{mask} = I_f \ \&\& \ I_{nn}$ 
    g.  $\text{hand\_detection}(I_{mask})$ 
end for

```

Here,

I_f = foreground image

I_{no} = input image

I_b = background image

I_{mask} = final binary mask

C. Hand detection using Convex Hull and Contour:

Hand detection using contour approach is effective. Applying the methodology the biggest contour and its information can be extracted (figure-2). Approximating it the hand was detected smoothly and efficiently. Using the contour information as input to find the convex hull. Convex hull [15] [16] is used to detect the surface of the object and other components. Using the information from the convex defects and calculating those with top points the finger information of the hand was filtered.

The positional information can be calculated by using the moment of the hand. The center point or moment of the hand can be detected using the area or total intensity and the information about its orientation. The height and the width of the hand from the moment were calculated. We drew lines from its moment to all the top of the fingers to find out the thumb to determine right or left hand. We also measured the angle between these lines. We used the Euclidean distance formula to determine the angle and distance between two points in two dimensional space.

$$Distance = \sqrt{(X1 - X2)^2 + (Y1 - Y2)^2}$$

Here,

$(X1 - X2)$ = Position of the first point.

$(Y1 - Y2)$ = Position of the second point.

D. Final Binary:

A bitwise 'and' operation was performed into those two binary images which were generated from color segmentation and background difference algorithm. The final binary achieved a full maximization and thus reduced all external noises. We found out contour, convex hull and convexity defects from that binary image. Pseudo code for hand detection is as follows:

```
hand_detection(Imask)
{
  a. mpoint = moment (Imask)
  b. C = contour (Imask)
  c. Capprox = contour_approx (C)
  d. Chull = convex_hull (Capprox)
  e. Cdefects = convex_defects(Capprox, Chull)
  f. If the angles between two fingers ≤ 90°
  g. Then Hpoints = defects
  h. Hthumb = thumb_location (Hpoints, mpoint)
  i. If Hthumb == left
  j. Then Right hand
  k. If Hthumb == right
  l. Then Left hand
}
```

Here,

m_{point} = moment point

C_{approx} = contour approximate

C_{hull} = convex hull

C_{defects} = convexity defects

H_{points} = hand points

H_{thumb} = thumb

E. Stepwise analysis with screenshots:

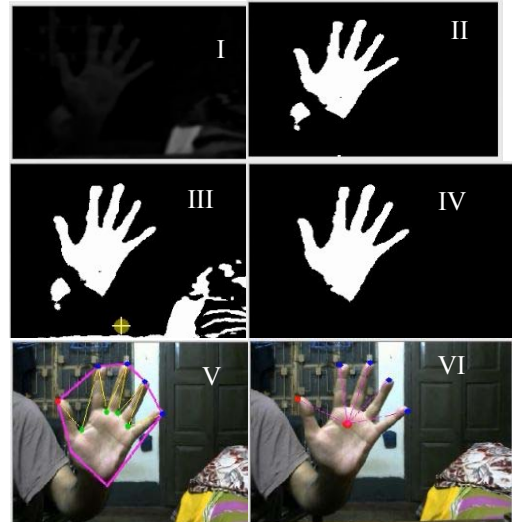


Figure 2: (I) $Cr - Cb$. (II) $(Cr - Cb)$ & background subtractor. (III) $(Cr - Cb) + Cr$. (IV) Final binary. (V) Convex hull. (VI) Moment.

F. Stereo Calibration:

Stereo calibration recovers depth information from camera images comparing more than one view of the same scene [17]. Applied this methodology the output resulted a 3D scene where each 3D point corresponded to an individual pixel of in one of the images. The parameters obtained from calibration were used to rectify stereo pair of images and to reconstruct the corresponding 3D scene. One way to calibrate the stereo system is to capture multiple pair of images of a calibration pattern from different angles. The typical calibration patterns are an asymmetric checkerboard (figure-3) where one side must be the even number of squares and other side must be odd numbers. The method was applied to calibrate and this allowed our experiment to construct the orientation of the checkerboard efficiently and unambiguously.

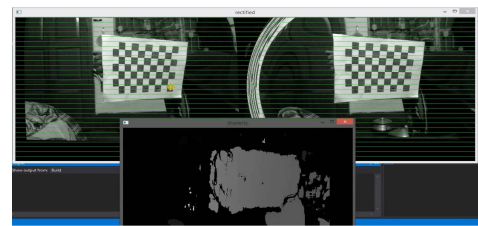


Figure 3: Calibration and Rectified images

Rectification is a very important step to compute disparity as it reduces the 2D stereo correspondence to 1D stereo. After that the disparity computation for the 3D reconstruction comes forth. Disparity is the distance in the pixels of the picture with corresponding points in the rectified image. Using the disparity mapping methodology the depth information (figure 4) of the hand from the camera was computed. Plotting those points accordingly the distance information was mapped. Thus the distance of the nearest object (the hand) in front of the camera was found out.

G. Depth estimation:

For calculating the 3D position of the hand we used this formula,

$$\begin{aligned}d &= Xr - Xl \\X &= Xl * Q[0,3] \\Y &= Yl * Q[1,3] \\W &= d * Q[3,2] + Q[3,3]\end{aligned}$$

Where,

$$\begin{aligned}X &= X/W \\Y &= Y/W\end{aligned}$$

Here,

Xr = Right camera X point

Xl = Left camera X point

Yl = Left camera Y point

Q = disparity-to-depth mapping matrix of size[4,4].

Obtained from camera calibration.

For calculating distance from the camera, we found out Z from the equation below:

$$Z = \frac{bf}{d}$$

Where,

b = distance between two camera lenses.

f = camera focal length

d = disparity

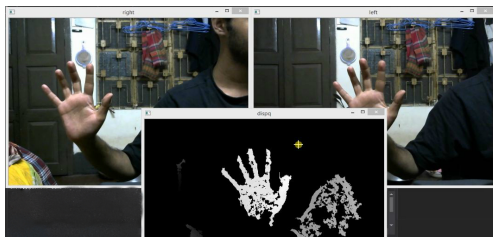


Figure 4: Disparity mapping

Performing threshold and then merging the binary image with the final binary image that was constructed using detection and color segmentation. After that, our experiment resulted a full maximized reformed binary image of the hand that was meant to be detected and tracked using our methodology.

V. CONCLUSION

To advance the methodology in the future, we would like to add movement in the 3D virtual hand and fingers and make it more real time and user friendly. It will benefit developing many apps and games with the help of a little amount of resources with much less cost, but yet much more effective, efficient, real-time and user-friendly. We also believe this system would benefit the augmented reality running apps and games or other accessing system. In addition, the dataset of disparity mapping would benefit in robotics and other virtual 3D hand or any other gesture movement development and application.

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Improving Facial Recognition Accuracy by Applying Liveness Monitoring Technique

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Abstract—In a typical transmission control protocol and Internet protocol (TCP/IP) suite, there are several ways, such as using photograph, three-dimensional (3D) model, and video clip of a valid user to mock the facial recognition. Studies show that there has been significant improvement in detecting photograph and 3D model spoofing. However, there is no such improvement in detecting video spoofing. Recent studies suggest that liveness monitoring using the facial features has potential to improve security, especially in biometric systems. In this paper, a liveness monitoring technique is introduced to assist facial recognition. In the proposed system, an eye blinking pattern is used through a real-time generic web-camera. An eye blinking password system using ultrasonic range sensing module is developed for liveness detection of the users. The system is tested by conducting experiments using 15 valid users and 100 different user appearances. According to the experimental results, the proposed system achieves 99% face recognition accuracy and 99% liveness detection performance.

Keywords—Eye blinking password, facial recognition, liveness monitoring, security in biometric systems, video spoofing

I. INTRODUCTION

It is frustrating to use passwords for protecting personal information and/or personal identification numbers for automated teller machines (ATMs). Facial recognition method is proven to be very useful to providing user authentication which helps one to protect information. Research in this field has a growing encouragement from many government agencies and universities. The facial recognition uses digital image or video processing techniques for authentication. Facial recognition techniques have proven to be very useful for contemporary security applications such as home security and video surveillance [1]. However, complex algorithms are required for processing images and videos in facial recognition techniques. Moreover, facial recognition encounters several challenges in providing accurate results. The most common challenges occur due to improper lighting, viewing angles, and spoofing alerts. When the user is subjected to improper lighting conditions, the same user can appear noticeably different [2]. There is also a difference between the indoor and outdoor lighting conditions [3]. This makes the system hard to identify the users. A constant lighting is required so that the facial recognition system can identify the user in any situation. The system also requires proper viewing angles for user identification, as the user may appear different from various

viewing angles [4]. Additionally, the user must stand in front of the system at an appropriate angle to get accurate results.

Spoofing attacks are considered as a major threat because they are implemented with the sole purpose of fooling the system while obtaining the required authentication information from the users [5]. A photograph of a valid user can be bent, rotated, and/or shifted accordingly before the camera to spoof the authentication system. That makes photographic spoofing a simple and cheap method of fooling the facial recognition system [6]. Video spoofing [7], on the other hand, poses a greater threat than any other spoofing techniques as it provides many physiological clues to the system such as eye blinking, head movement, etc. Another technique of spoofing is through a 3D model of a valid user [8]. This technique of spoofing does not pose much of a threat because a 3D model lacks the physiological information of the face and the realism of a valid user (appearance of a live person).

On a face, there are distinct points (called nodal points) that make them distinguishable. Common nodal points include the distance between the eyes, width of the nose, depth of the eye sockets, shape of the cheekbone, and the length of the jaw line. These nodal points are measured using numerical code called faceprint [9]. They can be measured and processed in three different approaches: Image input normalization, geometric approach, and elastic face matching. The filtering algorithms used do not check if the image or video is reliable as it cannot distinguish between a live feed of the camera and a video clip placed in front of the camera. A number of algorithms have been provided to identify spoofing attacks [10]. Some of these algorithms have been successful in identifying photograph spoofing with great accuracy. However, algorithms used to identify video spoofing attacks do not much improvement. When a video clip of a valid user is placed in front of a camera, it can easily deliver the necessary data to measure the nodal points of a face which makes it a challenge to identify such spoofing attacks. Therefore, a hardware-based liveness monitoring technique is proposed to improve the facial recognition accuracy.

Some related published articles are summarized in Section II. A number of selected anti-spoofing techniques are discussed in Section III. Section IV describes the proposed video spoofing detection method. Experimental results are presented In Section V. Finally, this work is concluded in Section VI.

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II. BACKGROUND AND MOTIVATION

A two-dimensional (2D) facial recognition system processes a single 2D image of a face [11, 12]. In a 3D system, various technologies such as patterned illumination light or paraxial viewing can be used to develop a 3D representation of a face [13]. One advantage of 3D facial recognition is that it is not affected by changes in lighting like other techniques [14]. Facial recognition systems utilize several popular recognition algorithms including principal component analysis using eigenfaces, local binary patterns, and elastic bunch graph matching [15]. A computational model of face recognition is developed in [16]. They suggest that an information theory approach of coding and decoding of face images may provide an insight in the facial features. Another facial recognition algorithm using facial bunch graph matching is proposed in [17]. In the elastic bunch graph matching algorithm, the visual features based on Gabor wavelets and dynamic link matching are necessary for facial recognition. The user is marked as a known valid user and is authorized if there are enough reasonable convincing similarities between the two wavelets.

III. POPULAR ANTI-SPOOFING TECHNIQUES

A. Photographic Spoofing

Any image of a valid user can be used to fool the facial recognition system in photo spoofing. In [18], a blinking-based liveness detection method is proposed where the blinking of an eye is used to prove that it is not an imposter image. For a normal eye-blink, three positions are necessary to identify that the blinking criteria is satisfied. The three positions are: Open, ambiguous, and closed as shown in Figure 1.



(a) Open (b) Ambiguous (c) Closed
Fig. 1. Three necessary eye positions

Eye closity algorithm is used for quick and accurate recognition of eye blinking. An average person blinks every 2 to 4 seconds. Any generic camera can capture in no less than 15 frames per seconds. This means that the camera can capture about two or more frames per eye blink. This makes eye blinking a more prominent feature to identify a live face and it is a more non-intrusive approach in identifying spoof attacks. This study shows promising results in identifying the eye blinking. Three different approaches used to measure the detection rates are: One eyed detection rate, two eyed detection rate, and clip detection rate. The system could be affected by strong glass reflection, making it difficult to determine the eye blinking. In [19], an improved method to detect face spoofing using micro texture analysis is proposed by Maatta et al. In this method, they are able to distinguish the real image from the fake photograph by using the differences in the reflection of light. They analyze the surface properties which show the difference in pigments. They provide results from an

experimental setup, where images of 15 subjects are used as separate sets for training and test purposes. They setup two sessions for training and one for the test purpose. During the test session, only 3 clients are qualified. In order to add complexity to the testing, they introduce 6 more clients. Their results are promising in determining the spoofing; however, their system can be flawed if a high resolution image is used to fool the system.

B. Video Spoofing

A typical video based face detection system automatically detects the face regions, extracts facial features, and recognizes facial identity. In [20], a method to identify video spoofing through visual rhythm analysis is proposed by Allan da Silva Pinto et al. In any image or video processing technique, the data has to be sent through an analog to digital converter to process the signals. During this conversion, it is more likely that noise would be generated in the system. These noise signals are distinct to the camera from which the video is obtained and the surrounding it captures. If a live video is captured, then from the noise generated we can identify its source by using a generic camera. In case of video spoofing, a video clip from a different camera is given as live feed to the system. After analog to digital conversion, the noise generated is likely to vary as the noise is subjected to camera feed. This variation in noise rhythm is used to identify the spoof threat. Their experimental results are encouraging because of their high accuracy. However, there is a problem – if spoofed video is created using the same camera as the one used for the facial recognition system, then high accuracy may not be achieved.

IV. A FACIAL RECOGNITION SYSTEM WITH THE PROPOSED LIVENESS MONITORING TECHNIQUE

The purpose of eye blinking module is to set a password with the blinking effect. For instance, four blinks can be set as a password for a particular user and the system only then recognizes that person as a valid user with four blinks in front of the camera. The eyeball depth is an important nodal point in distinguishing the user, because it is unique to every human being. Prior to running the facial matching algorithm, we identify the eyeball depth of the user in front of the camera. Ultrasonic range sensing modules are said to have better accuracy and precision to determine the distance between the two points. The purpose of measuring these values is to make sure that a valid user is standing in front of the generic camera by checking the depth irregularities with the depth of an eyeball. These depth irregularities cannot be spoofed by any other video displaying devices like a cell phone or mini digital display devices. Thus, if the video clip of any person is placed in front of the camera, this design will detect a video spoofing alert. In such a way, user authentication is improved and spoofing threats are reduced. Although this approach seems to be valid, adding a contingency to the existing technique would provide better accuracy and would completely eradicate the spoofing threats. The proposed model utilizes hardware-based security to eliminate spoofing threats. Figure 2 illustrates the inter-actions of various components of the proposed system using a block diagram. Table I shows major hardware and software modules used in the experiments.

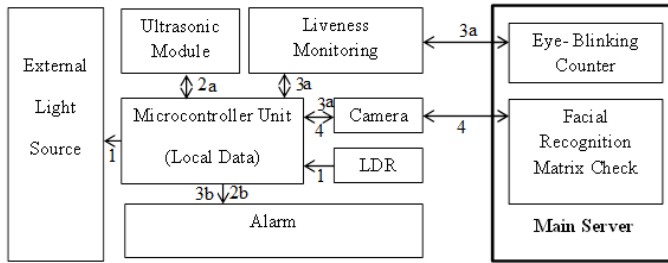


Fig. 2. Block diagram used in this study

TABLE I. HARDWARE AND SOFTWARE USED IN THE SYSTEM

Components	Type/Description
Microcontroller	Arduino Leonardo
Ultrasonic Ranging Module	HC-SR04
Servo Motor	Futaba S3003
Light Depending Resistor	LDR07
Software	Arduino IDE 1.0.5

A. Work-Flow Diagram

The work-flow of the proposed system is illustrated in Figure 3. The microcontroller is preloaded with the measurement values of the valid users and an accuracy factor. The ultrasonic range sensing module measures the eyeball depth of the valid users. This measurement data is sent to the microcontroller unit for further processing. If the values (ultrasonic distances) match, the liveness detection monitoring comes into picture and goes to the eye blinking phase. (If the measurements do not match, then the microcontroller unit signals the buzzer indicating that it is a threat.)

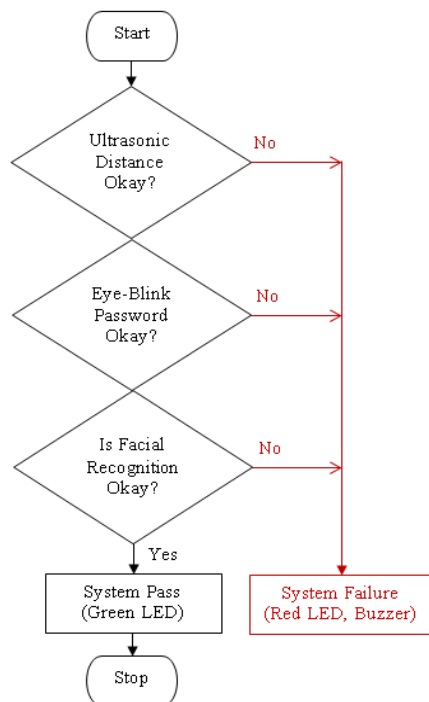


Fig. 3. Flow chart of the proposed model

Users are given a different eye-password according to their eye blinks. The user, after the acceptance of eyeball value measurements, is validated for liveness monitoring. The liveness monitoring detects the eye blinking of the valid user. The liveness monitoring algorithm requires better processing power and thus, it is sent to a nearest server that contains the database of the valid user information related to the facial recognition system. Clip based detection is used to identify the eye blinking as it promises better accuracy and efficiency. If the eye blinking password does not match then it again buzzes, indicating a threat. Only after the eye blinking criterion is met, the system moves to the facial recognition method for the true authentication of the user. In a way, it is acting as a security measure to better improve the reliability. The ultrasonic module requires only the processing power of the microcontroller and thus it can be kept isolated from the database. In this regard, it is providing proper isolation to restrict threats such as someone is hacking the main server to access information. For the liveness monitoring, the proposed algorithm includes identifying the outline of an eye (the left and right most positions of the eye are identified), matching the position of the eye (with the standard eye position), and calculating the difference between the position of the top of an eye outline and that of a standard opened eye.

The system depends on ultrasonic range sensing, making proper arrangement of these sensors are necessary for efficient operation. Moreover, it uses eye blinking as a password, hence proper lighting conditions are required to provide optimum lighting to detect the eyelid movement. These challenges can be addressed by proper design and operations.

V. EXPERIMENTAL RESULTS

In this section, we evaluate a facial recognition system with the proposed eye blinking based liveness monitoring technique. For the measurement of eyeball depth, 15 users are taken into consideration. Users are considered for both training and testing stages. The eyeball depth measurement values (top, center, and the bottom depth) of the valid users are: User 1 (12, 20, 10), User 2 (13, 10, 11), User 3 (12, 12, 11), User 4 (14, 10, 13), User 5 (14, 11, 12), User 6 (15, 13, 12), User 7 (11, 10, 10), User 8 (13, 11, 12), User 9 (12, 10, 11), User 10 (12, 9, 11), User 11 (11, 11, 9), User 12 (15, 11, 10), User 13 (13, 12, 13), User 14 (11, 9, 10), and User 15 (14, 13, 10). The values are unique for each user. The values of eyeball depth for User 1 are (12, 20, and 10). After these values are taken, they are kept aside in a database along with the facial matrix of each users. These eyeball values are preloaded in the microcontroller system for testing.

Once the training stage is completed, the system is ready to test users. First, it is verified whether the microcontroller unit can identify the valid users with the preloaded values of the eyeball depth measurement. During the test stage, all the valid users show full accuracy and the microcontroller system is capable of identifying the users every time. One hundred different attempts are made to authenticate the system by measuring their eyeball depth values. These values and experimental results are summarized in Table II. The system recognizes the valid users.

TABLE II. COMPARISON BETWEEN VALID AND INVALID USERS

Appearance	Ultrasonic Distance	Pass/Fail Check	Valid/Invalid User or Error
1	(14,10,13)	Pass	Valid (User 4)
2	(13,10,11)	Pass	Valid (User 2)
3	(12,11,12)	Fail	Invalid
4	(15,13,12)	Pass	Valid (User 6)
5	(11,12,11)	Fail	Invalid
6	(12,12,11)	Pass	Valid (User 3)
7	(14,11,12)	Pass	Valid (User 5)
8	(12,20,10)	Pass	Valid (User 1)
9	(11,11,12)	Fail	Invalid
10	(10,12,12)	Fail	Invalid
11	(10, 10, 11)	Fail	Invalid
.....			
98	(12, 12, 12)	Pass	Error
99	(14,13,10)	Pass	Valid (User 15)
100	(13,13,12)	Fail	Invalid

Once a spoofing attempt is made in the system by showing the video of a valid user, the ultrasonic ranging module measures for the eyeball depth but it receives distance values between the video display screen and the module whose values are different from those of the eyeball depth of the user. The system identifies it as a spoofing attempt and declines authentication. The eye-password authentication logic for the liveness monitoring method is illustrated in Figure 4.

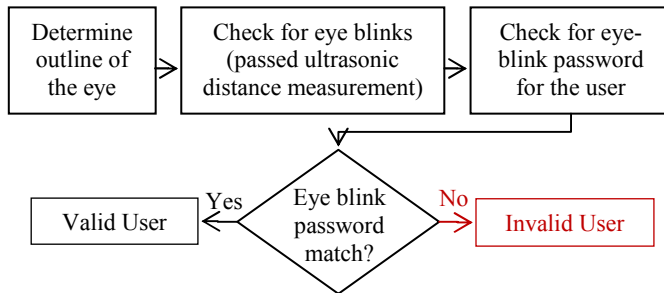


Fig. 4. Liveness monitoring eye-password authentication

VI. CONCLUSIONS

In this work, an eye blinking based liveness monitoring technique is used to enhance the performance and accuracy of a facial recognition system. An eye blinking password system is developed using ultrasonic range sensing module for liveness detection of users. A real-time generic web-camera is used in the system to implement the local facial features such as eye blinking. It is expected that the proposed system should offer high accuracy as it can identify video spoofing and perform an improved facial recognizing algorithm. In this experiment, 15 different valid users are used for training and testing purposes; 100 different user appearances are tested in the experiments. Based on the experimental results, 99% face recognition accuracy and 99% liveness detection performance are achieved. The proposed facial recognition system is robust in natural environments in the presence of noise and illumination changes. Future security systems are expected to have the capability of recognizing people in real-time with less constrained conditions.

We plan to enhance the facial recognition system for faster execution of the liveness monitoring in our next endeavor.

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Fabric Defect Classification with Geometric Features Using Bayesian Classifier

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Abstract—Fabric defect inspection is the pivotal part in the production of textile products. Since manual inspection is tedious and erroneous, automated fabric inspection has been topic of research for past years. Automation of fabric inspection involves two major aspects: defect detection and defect classification. We focused on classifying defects based on geometric features of defects. The features are obtained by applying statistical technique on an image dataset. Classification of defects is accomplished using simple Bayesian classifier, which delivers a pleasing accuracy.

Keywords — fabric defect, defect detection, defect classification, Bayesian classifier, accuracy.

I. INTRODUCTION

Textile goods are the topmost items for exporting in Bangladesh. Over the years, it has been the most beneficial sector in economy of Bangladesh. To sustain in the ever-growing competitive markets of the world, production of quality items is of maximum priority. Maintaining quality includes isolating defective fabrics from the production line for removal. Isolation of defective fabrics is done through manual inspection in the present time. However, it is a long, tedious process with not so good result. Manual inspection is prone to error since defects can be too tiny or hideous to be ignored by human eye.

Researchers have agreed upon a better way to deal with this problem. The solution is automation, i.e. machine vision based fabric inspection. Automatic fabric inspection systems mainly possess two aspects of concern: defect detection and defect classification.

In this paper, we have taken a dataset with 128 fabric samples each reprising of 8 (eight) geometric features. A statistical feature extraction process has been applied in order to find the features. The features were extracted from images of knitted fabrics converted into suitable format for extracting through applying essential effects. After obtaining the features from the samples, we applied Bayesian classifier to classify the defects based on those features. The results have been found good enough.

The rest of the paper is arranged as follows. Section II discusses the related works in this field. Section III is used for describing the methodology of our classification model. Section IV contains the implementations. Experimental results and comparative analysis are given in the section V and finally we discuss the future prospect of our work along with conclusion in section VI.

II. LITERATURE REVIEW

Feature selection plays a vital role in the process of classification of defects. However, defects themselves are not well stated in any literature. The researchers have stated the definition of defects and their classes differently. We primarily choose the definitions stated in [1]. Fig. 1 depicts the defects.

Researchers tried different fabric inspection techniques over past years [3]-[21]. Though most of them concentrated on defect detection, some have approached forward for defect classification. Classifying techniques include Neural Network (NN), Support Vector Machine (SVM), clustering and statistical inferences. The works [4]-[7], [10]-[12], [19] dealt with feature selection and scene analysis featuring multiclass scenarios.

Statistical inference was used only for binary classification, i.e. categorization of only defective and defect-free in [14], [15]. Binary classification does not address the problem of defect classification. Murino et al. [6] implemented SVM for classification. However, the feature extraction process was too complex to be implemented in real-life case. Model-based clustering used by Campbell et al. [13] is not suitable for real-time systems like automated fabric inspection systems. NN has been used for classification in number of works. Backpropagation learning algorithm was used in [4], [7], [10] and [11]. Saedi et al. [4] trained their network for multiclass problems, but was functional only in frequency domain. Karayiannis et al. [6] used NN for defect classification but analysis of defects and justification of features were not done up to the mark. Kuo et al. [9] used an NN to deal with multiclass problem but eventually the defect analysis and feature justification fell short. Moreover, their sample space was too

small to be adequate. Classification errors are the outcome of these approaches. Mitropulos et al. [11] used an NN classifier and their results were satisfactory since their sample size was small. Islam et al. [5] and [19] used resilient back propagation to train the NN addressing the multiclass problem. However, the number of features was small and the defect analysis was insufficient. Because the sample size was small, their approach worked to some extent. Shady et al. [12] trained NN using Learning Vector Quantization (LVQ) algorithm to address multiclass problem. Nevertheless, the results were not satisfactory. Other notable uses of NNs in defect detection are in [8] and [9] where two NNs were used. Even then, the results could not reach the satisfaction mark. In [1] they were more concerned about extracting and selecting features rather than classifying defects. Moreover, the data size i.e. the number of images was comparatively smaller than our work. NN was used for classifying but not for detailed investigation, it was used for demonstrating the utility of the features.

Since the application of NNs did not cut the deal, we expected that Bayesian classifier could be a better option. Bayesian classifiers have been used in several classification scenarios [22], [23]. Keerthika et al. [22] used Naïve Bayesian classifier for feature selection. In addition, Bayesian classifier is used for binary classification, i.e. categorization of defective and non-defective by Jeong et al. [23].

III. METHODOLOGY

Our approach consists of statistical defect detection and classification based on geometric features. We use a statistical technique, i.e. thresholding technique, for defect detection and deploy Bayesian classifier for defect classification. It starts with inspection of image of single-color knitted fabric, which is converted into a gray-scale image. Then the image is filtered in order to smooth it and remove noises. The gray-scale histogram of the image is formed and two threshold values are calculated from the histogram using histogram peak technique [24]. Using these threshold values, the image is converted into a binary image. This binary image contains object (defect) if any exists, background (defect-free part), and some noises. These noises are

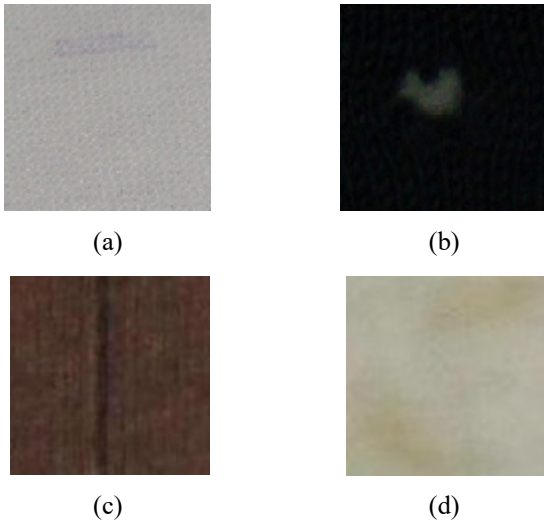


Fig. 1. Different types of defects occurred in knitted fabrics.

(a) Color yarn, (b) Hole, (c) Missing yarn, (d) Spot.

removed using thresholding based on the minimum size of defects. Then a feature vector is formed calculating a number of features of the defect. An already trained Bayesian classifier classifies the defect based on the feature vector.

A. Defect Types

In our paper, we propose a classification of four types of defects. Choosing these defects was done considering frequently occurring in knitted fabrics in Bangladesh. They are color yarn, hole, missing yarn, and spot. Fig. 1 shows the four types of defects. They are described in detail in [1]. The minimum defect size we considered is 3mm×1mm.

B. An Appropriate Set of Geometric Features

An appropriate set of geometric features is selected for classifying the defects. The set consists of the following geometric features:

- i) Height of defect window, H_{DW} .
- ii) Width of defect window, W_{DW} .
- iii) Height to width ratio of defect window, $R_{H/W} = H_{DW} / W_{DW}$.
- iv) Number of defective regions, N_{DR} .

All of these four features can be found in detail in [1].

C. Bayesian Classifier

Bayesian classifier uses traditional Bayes theorem to classify unknown samples based on maximum likelihood. Bayesian classifier is a probabilistic framework for solving classification problems. The Bayes theorem:

$$P(event B|event A) = \frac{P(event A|event B)P(event B)}{P(event A)} \quad (1)$$

- $P(A)$ and $P(B)$ are the probabilities of event A and B without regard to each other.
- $P(A | B)$, a conditional probability, is the probability of event A given that event B is true.
- $P(B | A)$, a conditional probability, is the probability of event B given that event A is true.

In plain English, it can be represented as:

$$posterior = \frac{prior \times likelihood}{evidence} \quad (2)$$

A naïve Bayesian classifier assumes the attributes to be independent. Therefore, it is fast and easy to compute. Moreover, it is not sensitive to irrelevant data. From Bayes Theorem, it can deduced that,

$$P(B | A) \approx P(A | B)P(B) \quad (3)$$

Naïve Bayesian classifier is reputable for classification and has been used for many years. As per the assumption by naïve Bayesian classifier that the attributes are independent, they will not affect the probability of each other. However, there are some challenges like, bias, variance and training data noise. Training data noise is subject to feature extraction and can be reduced by selecting distinguishing features. The formal definition of Naïve Bayesian classifier can be represented as:

$$C = arg \max_{k \in \{1,2,...,K\}} P(C_k) \prod_{i=1}^n P(x_i | C_k) \quad (4)$$

where,

- $P(C_k)$ = probability of class k
- $P(x_i|C_k)$ = probability of feature x_i given class k

Now, since the attributes are continuous in this case, probability density estimations were used. Assuming the distribution as normal, the density estimation function was defined as:

$$\varphi_{\mu,\sigma}(x) = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-\frac{(x-\mu)^2}{2\sigma^2}} \quad (5)$$

Density function expresses the relative probability of a point where μ is the mean, σ is the standard deviation. $\varphi_{\mu,\sigma}(x)$ is used for calculating $P(x|C_k)$.

The model was trained by calculating estimation of prior $P(C_k)$ and for every attribute A_i , for every attribute value v of A_i estimating $P(A_i = v|C_k)$.

Applying the probabilistic model for a given sample with $(v_1, v_2, v_3, \dots, v_n)$, the class picked that maximized

$$P(C_k) \prod_{i=1}^n P(A_i = v_i|C_k) \quad (6)$$

IV. IMPLEMENTATION

The image processing part was accomplished using Java program we developed. The program was based on Java Standard Edition platform (Java SE 6). The input images were 512×512-pixel images of single-color knitted fabric, which were then converted into grayscale images with 256 grayscale levels. The images were captured using a 9.5-megapixel Fuji camera of model FinePix S9500. In our approach, we intend to detect defects of minimum size of 3mm×1mm. So, any object smaller than the minimum defect size in terms of number of pixels is eliminated from the binary image.

The sample dataset was input into the classifier and trained with different ratios of training and testing samples. The sample data set consists of 128 fabric samples with feature values in the form of numerical values. The numerical values were obtained through image processing and collected in a excel sheet. The sample dataset was divided between training dataset and test dataset. The means and standard deviations were calculated using standard formulae and stored in the “pTable”. Once the classifier is initialized and the mean, standard deviations of features are stored, the classifier is ready to classify.

After building the classifier, test samples were input into the classifier one at a time for classifying. We used around fifty percent of the sample data for training the classifier and the rest fifty percent was subject to be tested. For classification using Naïve Bayesian classifier, the posterior probability of each class for every test sample was calculated. Then, the class with maximum posterior probability was selected for each test sample.

We trained the classifier with five classes: color yarn, vertical missing yarn, horizontal missing yarn, hole and spot. The sample dataset contained samples of each classes and some defect-free samples as describe in Table I.

TABLE I. SAMPLES AND THEIR DISTRIBUTIONS

Sample	Sample size	Training samples	Test samples
Color Yarn	16	10	6
Vertical Missing yarn	19	10	9
Horizontal Missing yarn	19	11	8
Hole	23	10	13
Spot	18	9	9
Defect free	33	20	13
Total	128	70	58

V. RESULTS AND DISCUSSION

We have applied our method over the dataset obtained from the work on feature selection using NN [1]. For calculating accuracy, we followed the simple technique of correctness ratio.

$$accuracy = \frac{\text{no. of samples correctly classified}}{\text{no. of samples tested}} \times 100\%$$

We tested around fifty percent of the sample data with classifier built on only the geometric features and then again with both the geometric and statistical features. Table I. describes the training to testing comparison. We achieved a promising result with accuracy of 99.19%.

In the result, there was betterment after adding the statistical features. Table II clearly justifies the promise. Since there have been a variety of classification of defects those include variety of researcher perceived classes, a major drawback for assessing the performance among those remains. Kumar [25] reported in his review of fabric defect detection that accuracy of more than 95% is industry standard. In this context, our achieved accuracy is above standard.

We have tested the samples with different test-train ratios and found positive result in increasing the training samples. Fig. 2 demonstrates the accuracy curve rising along with the increase of training samples. When we had 80% of the sample data separated for testing while only 20% left for training, the accuracy was not up to the mark.

TABLE II. DEFECT CLASSIFICATION PERFORMANCE

Class	Accuracy
Color yarn	98.02%
Vertical missing yarn	87.26%
Horizontal missing yarn	98.48%
Hole	95.47%
Spot	97.39%
Defect free	100.0%
Total	99.19%

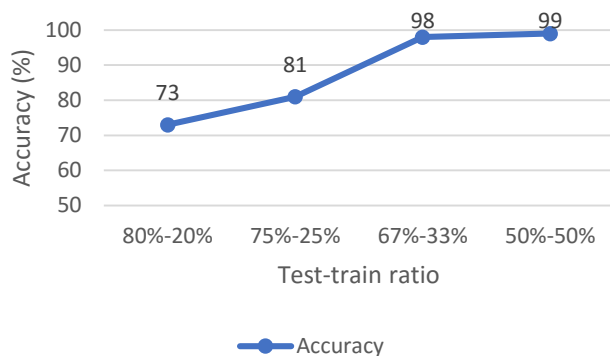


Fig. 2. Performance curve

Increasing the training sample percentage to 25%, the accuracy escalated a little. Nevertheless, increasing the training sample to 33% and to 50%, the accuracy had a drastic change. It quickly ascends above 95%.

VI. CONCLUSION AND FUTURE WORKS

In this paper, we have demonstrated a comprehensive approach to classify fabric defects. We have considered eight features of geometric characteristics. We completely relied on the geometric features of the data set. Although we have found very good results, the classification cannot be stated as final. The result that we achieved can simply be regarded as good enough compared to other works.

There remain other works. Expanding the sample dataset and improving image quality are two of the major aspects to be focused. Besides, research can be done on reducing the number of features required for classification. Work is in progress to combine other features with these and reduce the number of features for classification.

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Risk Analysis of Buildings using an Expert System: A Case Study in Bangladesh

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Abstract—This paper represents an expert system which determines to what extent a building is risky. In this purpose, all the available systems to analyze the risk of a building are first examined. The existing systems are unable to handle uncertainty, which is actually a big challenge for analyzing the risk of a building. However, the rule based expert system, proposed in this paper is capable of capturing all the uncertainties during the risk analysis of a building. The proposed system is developed by using Rule based Inference Methodology Using Evidential Reasoning (RIMER) Approach. Signs of a risky building are considered by RIMER Approach to infer the risk level of a building. Experimented results show that the proposed system analyzes level of the risk more efficiently than the existing systems.

Keywords—Risk Analysis; Uncertainty; Signs of a risky building; Rule based Inference Methodology Using Evidential Reasoning (RIMER); Expert system

I. INTRODUCTION

Risk analysis of building is very important because building is considered as one of the essential ingredients of modern civilization. This topic is more significant for Bangladesh because building collapse happens in Bangladesh very often. In 2013 Rana Plaza collapsed, which killed 1129 with 2515 injuries. It was identified as a risky building but the experts could not identify how risky it was. According to RAJUK there are more than 3000 risky buildings in Dhaka city. This number is not small in other cities of Bangladesh. Experts try to identify the risky buildings. For this purpose, a number of expert systems have been developed but none of these systems can exactly determine to what extent a building is risky. Moreover, these systems cannot handle uncertainties. So, the results provided by these systems are inexact. However, this paper represents an expert system that exactly determines how risky a building is.

Section II to V arrange for background study, signs of a risky building, proposed methodology and risk identification of a building. Section VI and VII offers implementation of the RIMER Approach and results and discussion. Section VIII concludes the paper.

II. BACKGROUND STUDY

Building fault detection or safety assurance has been a great concern for many years. Many researchers have been conducted to improve the methodology for analyzing the risk of the buildings. Many researchers suggest quantitative model based method to be used to determine the risk of a building [1] whereas some other researchers emphasize qualitative model based method. But quantitative method was found complex and computationally intensive [1]. So, qualitative method has been preferred in many cases.

Some rule based methods have also been developed [2]. On the other hand, some researches proposed process history based method. Artificial neural network [3], fuzzy logic [4] and data mining techniques [5] are also used to identify the risk of a building. A number of embedded systems [6] and expert systems [7] have been developed. But none of these systems can exactly say how risky a building is.

III. SIGNS OF A RISKY BUILDING

Building collapse is defined as the act of falling down, cracking in or crumbling of a building. Buildings may collapse partially or completely. Partially collapsed buildings are often said to be damaged ones whereas major damaged buildings are considered to be collapsed.

Buildings may collapse due to natural disasters, say, and earthquakes or faulty construction. Deficient

structural design is another reason for a building to be considered as perilous one.

Researcher categorizes the damage of a building into six ranks: (i) light, (ii) minor, (iii) medium, (iv) major, (v) partially collapsed and (vi) totally collapsed[8]. Some researchers classify the damage of a building into three levels [8]: (i) Operational damage: covers light to major damage, (ii) Heavy damage: where medium to major damages are enclosed and (iii) Collapsed: which includes partial and total collapse of a building.

However, experts agreed on this point that, before being totally collapsed every building shows some signs faster or late. So, if the signs that a building shows before being damaged or collapsed can be observed, then it is possible to determine that the building is chancy.

Any risky building shows some cracks in its beam, column or slab [9]. Risk indication can also be found from the structural design of the building. So, to ensure the safety of a building or to determine the insecurity of a building, an expert needs to check if there is any crack in column, beam or slab of the building [9]. Structural design of the building should be analyzed for this purpose. Another most important question which should be considered for analyzing the risk of a building is instability of the building. Building instability can be uniform or differentiated [10].

IV. PROPOSED METHODOLOGY

An expert system consists of two basic components: knowledge base and inference engine. A knowledge base is constructed base on human knowledge. Some If-Then rules are used to construct the knowledge base. The rules are provided by the experts. An inference engine makes the decision based on the knowledge base.

A rule base handles uncertainties while making a decision. The uncertainties come from vague, incomplete and imprecise information [11]. Rule based inference methodology using evidential reasoning approach [11], which handles uncertainties and provides the most exact results.

The knowledge base of RIMER Approach is established based on some belief rules, which are provided by the domain expert. The inference engine is constructed by Evidential Reasoning algorithm.

[11], [12]. The basic knowledge base is represented as:

IF {(cracks in beam is High) ^ (cracks in column is High) ^ (cracks in slab is Low)} Then building cracks is {(High, 0.9), (Medium, 0.1), (Low, 0.0)}

In RIMER approach inputs that are provided by the experts are transformed into referential values using the following formula

$$S(A_i^*) = \{(a_{ij}, \gamma_{ij}); j = 1, \dots, J_i\} \quad (1)$$

$$\text{Where, } \gamma_{ij} = \frac{a_{i(j+1)} - A_i^*}{a_{i(j+1)} - a_{ij}}, \quad \gamma_{i(j+1)} = 1 - \gamma_{ij} \text{ if } a_{ij} \leq A_i^* \leq a_{i(j+1)} \text{ and } \gamma_{ik} = 0 \text{ for } k = 1, 2, \dots, J_i, k \neq j+1$$

Then, the activation weight w_k for the packet antecedent A_k in the k^{th} rule is generated by using the following formula:

$$w_k = \frac{\theta_k \alpha_k}{\sum_{i=1}^L \theta_i \alpha_i} \quad (2). \quad \text{Where, } \theta_k = \prod_{i=1}^{T_k} (\alpha_i^k)^{\bar{\delta}_{ki}} \text{ And } \bar{\delta}_{ki} = \frac{\delta_{ki}}{\max\{\delta_{ki}\}} \text{ } i = 1, \dots, T_k \text{ and } 0 \leq \bar{\delta}_{ki} \leq 1$$

Here, w_k is the activation weight of k^{th} rule. θ_k is the relative activation weight for the k^{th} rule. α_i is the degree to which the input A_i^* belongs to the referential value and δ is the relative weight for antecedent attribute.

After generating the activation weight the belief degrees are updated if the belief degrees are not complete. The belief degrees are updated using the following formula:

$$\beta_{ik} = \bar{\beta}_{ik} \frac{\sum_{t=1}^{T_k} (\tau(t, k) \sum_{j=1}^{J_t} \alpha_{tj})}{\sum_{t=1}^{T_k} \tau(t, k)} \quad (3); \quad \text{where, } \tau(t, k) = \begin{cases} 1 & \text{if } U_i \text{ is used in defining } R_k \text{ } (t = 1, \dots, T) \\ 0 & \text{otherwise} \end{cases}$$

After the accomplishment of updating the belief degree the belief rules are expressed using a matrix. The Evidential Reasoning (ER) approach is used to aggregate the rules and generate final conclusion. ER approach is used to transform the degree of belief β_{ij} into basic probability mass [13], [14] which is done by $m_{j,k} = w_k * \beta_{ij}, j = 1, \dots, N$ (4) and $m_{D,k} = 1 - \sum_{j=1}^N m_{j,k}$ (5). Here, $m_{D,k}$ is the probability mass unassigned to any individual consequent. After that, rules aggregation is done. All the combined rules provide the final results. The final results are the representation of some confidence

level of how much risky a building is. For example, if the final result for a certain building is $\{(H, 0.6), (M, 0.3), (L, 0.1)\}$ then, the system is 60% sure that the risk level of the building is High, 30% sure that the risk level is Medium and 10% sure that the risk level is Low.

V. RISK IDENTIFICATION OF A BUILDING

For implementing an expert system that will identify the risk level, Rule based Inference Methodology using Evidential Reasoning (RIMER) is used. The following figure shows the architectural design of the proposed system. The BRB architecture consists of presentation layer, application layer and data management layer.

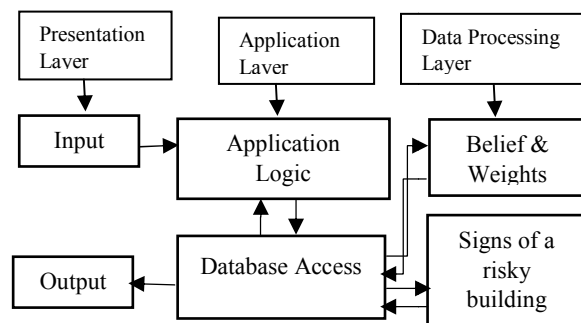


Fig. 1. System architecture to determine the risk of a building

The presentation layer provides the interface of the system where application layer contains the basic application logic and data management layer deals with database.

VI. IMPLEMENTATION OF THE RIMER APPROACH

A building risk is determined using belief rule based expert system. For that, the user needs to provide the inputs which are actually some signs of a building. Then the inputs are transformed into three referential values: high, medium and low. After input transformation the activation weights are calculated which requires some additional parameters, say, rule weight and attribute weight. Belief degree is updated if the degree of belief is not complete. After that, evidential reasoning is used to infer the risk of the building.

TABLE 1: INPUT TRANSFORMATION INTO REFERENTIAL VALUES

Attributes	Input	Referential Values		
		High	Medium	Low
Cracks in beam	0.8	0.6	0.4	0.0
Cracks in column	0.98	0.96	0.04	0.0
Cracks in slab	0.4	0.0	0.8	0.2

A BRB framework has been developed to construct a BRB knowledgebase. The knowledge base constructed by the help of some experts. Experts' belief is collected from the civil department of Chittagong University of Engineering & Technology, Bangladesh.

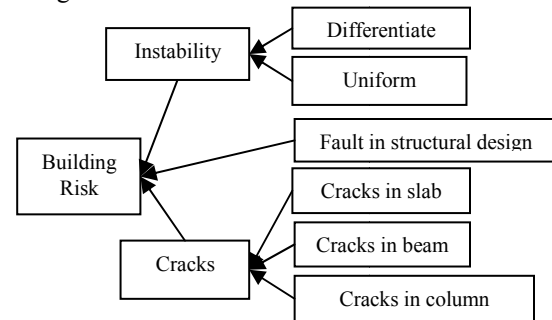


Fig. 2. BRB framework to analyze the risk of a building

VII. RESULTS AND DISCUSSION

The proposed system decides the risk level and shows the risk in percentage, which easily makes the people understand if the building is usable or not. But the manual system which is currently being practiced cannot analyze the building risk efficiently and is not reliable to the people in most of the cases. The system is tested over 43 risky building in Chittagong City. Sample results of a building are provided in the next figure.

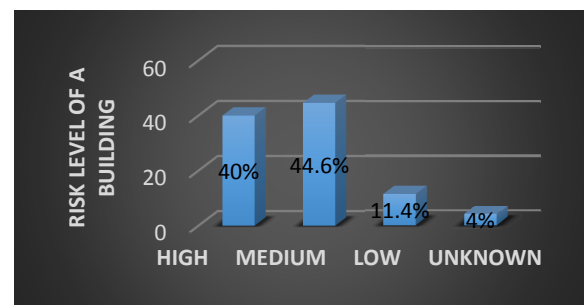


Fig. 3. Risk of a building

The results from the manual system for these buildings are collected with the help of two experts. Comparing the manual system and BRB expert system it is found that the proposed system is more efficient in analyzing the fault of buildings than the manual systems.

The performance of the expert system is presented using a ROC curve.

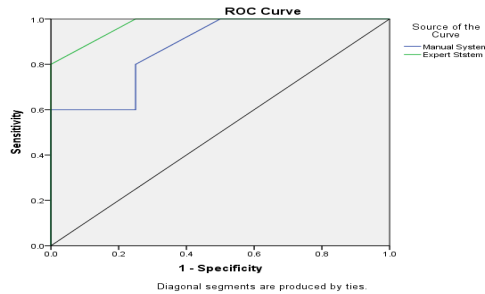


Fig. 4. ROC Curves Illustrate the Performance of BRB Expert System and Manual System

The above figure shows two ROC (Receiver Operating Characteristics Curve) curves. The green curve represents the expert system's risk analysis performance where the blue curve shows the performance of the manual system. The AUC (area under curve) of manual system is 0.875 where AUC of expert system is 0.979. So, it becomes clear that expert system provides the better performance than the manual system in the risk analysis of a building.

VIII. CONCLUSION AND FUTURE WORK

The objective of the research is to make the human life safe. For this purpose, an expert system is developed for determining building faults, which is considered the most devastating subject for human life. The developed system handles uncertainty while determining the risk level of a building and provides the most exact results.

In future, an image processing technique will be applied to acquire inputs for the proposed system. At that time the system will not require humans to observe the condition of the building to provide the inputs.

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Performance Studies of Different Closed Loop Glucose Controllers for Treating Type 1 Diabetes Mellitus

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Abstract—This thesis deals with the performance analysis of different closed loop controllers for blood glucose level control of type 1 diabetes patient as a nonlinear model, which was simulated in MATLAB Simulink environment. PI controller, Fuzzy logic controller and optimal regulator were used to control glucose level by infusing remote insulin. A basic proportional-integral controller was modified using trial & error method. To enhance the performance of insulin infusion inside the human body Fuzzy logic controller and optimal regulator were also utilized with proper customization. For fuzzy logic, membership functions and fuzzy rules were defined. Also for optimal regulator study, matrixes were built for input & output equations. Finally, the performances of three different controllers were compared.

Keywords—Type 1 Diabetics; PI controller; Fuzzy logic controller; Fuzzy rules; Optimal regulator; membership function; Close loop Glucose Controllers;

I. INTRODUCTION

Diabetes mellitus is one kind of metabolic disease in which human body can't produce enough insulin to control glucose level moderate or can't use insulin properly. The hormones insulin and glucagon release from β -cell and α -cell respectively are responsible for keeping glucose level moderate. There are three types of diabetics, those are type 1, type 2 and Gestational diabetics. This thesis deals with only type 1 diabetics. Type 1 diabetics is such kind of diabetics where body does not produce enough insulin. Insulin is a hormone which is needed to convert glucose into energy that is needed for our daily work. So, for this situation insulin need to be pushed from external source to keep glucose level moderate. Otherwise it can cause Hyperglycemia.

PI is the most frequently used controller. PI control stands for Proportional-Integral control. Each of these, P and I controller has basic algorithm and both are implemented for individual purpose. The main duty of PI controller is to calculate the error between actual output and measured output. There are different ways to tune the PI controller but trial & error method was used in this thesis. In trial & error method the gains of proportion and integral were adjusted randomly.

To enhance the performance of insulin delivery system scientists have improved the steady-state and transient performance of PI controller by introducing fuzzy theory named PI type fuzzy controller [1-4]. In this controller design triangular membership function was used for input depending

on the output response of controller. Following that fuzzy rules were defined for our system.

A quadratic behavior standard is normally considered in glucose level control algorithms on the basis of glucose deviation and amount of remote insulin. In this thesis Work, a Linear Quadratic glucose Control (LQC) problem is stated for reducing the cost function.

II. NONLINEAR PATIENT MODEL

In this section, the nonlinear patient model [5] is discussed which will be introduced and simulated in MATLAB Simulink environment. In the regarded model, Equation (1), (2) and (3) are the governing Equations of blood glucose system. There are three principal state parameters in this set of equations. G , X and I which are glucose concentration, insulin concentration and remote insulin concentration respectively. Table 1, shows the parameters with their corresponding values in these equations.

$$\frac{dG}{dt} = -p_1 G - X(G - G_b) + \frac{G_{meal}}{v_1} \quad (1)$$

$$\frac{dX}{dt} = -p_1 G + p_3 I \quad (2)$$

$$\frac{dI}{dt} = -n(I + I_b) + \frac{u}{v_1} \quad (3)$$

TABLE I. PARAMETER VALUES IN EQUATIONS 1 TO 3.

Parameter	Value	Description	Units
$G_{nominal}$	81.5	Initial glucose concentration	$\frac{mMol}{dL}$
$I_{nominal}$	10.5	Initial Insulin concentration	$\frac{mU}{L}$
$X_{nominal}$	0.00546	Initial remote insulin concentration	$\frac{1}{min}$
G_b	4.5	Basal injection level of glucose	$\frac{mMol}{L}$
p_1	0	Insulin-independent rate constant of glucose uptake in muscle, liver and adipose tissue	$\frac{1}{min}$

p_2	0.025	Rate for decrease in tissue glucose uptake ability	$\frac{1}{min}$
p_3	0.00001 3	Insulin-independent increase in glucose uptake ability in tissue per unit of insulin concentration above I_b	$\frac{mU}{L.min}$
n	0.093	First order decay rate for insulin in plasma	$\frac{1}{min}$
I_b	4.5	Basal injection of insulin	$\frac{mU}{L}$
v_1	12	Amount of taken meal	L
G_{meal}	5.54	Amount of glucose intake with meal	$\frac{mMol}{L.min}$

Diabetes is a disease that either the patient has a high blood glucose concentration due to the lack of insulin production, or the body's cells do not respond correctly to the insulin that exists. The lack of insulin production is regarded as type 1 diabetes, where the patient needs to take regular insulin infusion and blood tests to ensure if the blood glucose level is normal. Fig. 1 shows the implementation of a virtual patient through nonlinear model in MATLAB Simulink environment.

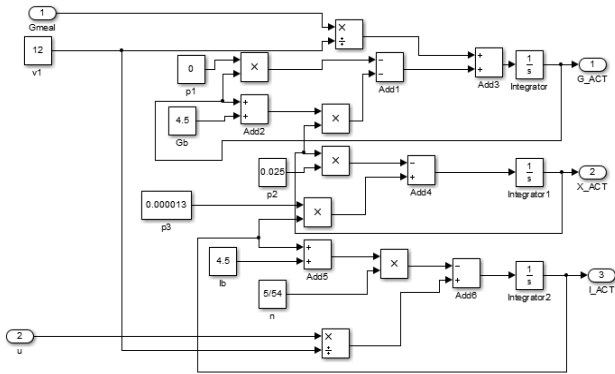


Fig. 1. Virtual Patient Model

III. PI CONTROLLER

The combination of proportional and integral terms is important to increase the speed of the response and also to eliminate the steady state error. Proportional + Integral (PI) controllers were used because of the desirable property that systems with close loop transfer functions have. Our goal was to have zero steady state error with respect to a step input.

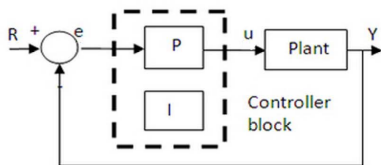


Fig. 2. Proportional Integral (PI) Controller block diagram

The proportional and integral terms [6] is given by:

$$U(t) = k_p * e(t) + k_i \int_0^t e(\tau) d\tau \quad (4)$$

Where K_p = Proportional gain, K_i = Integral gain, Error = Set point – Measured value, t = time, τ = variable of integration

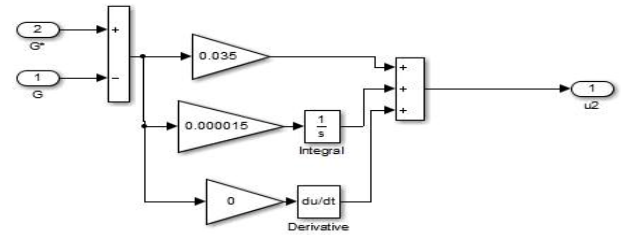


Fig. 3. PI Controller Design

IV. FUZZY LOGIC CONTROLLER

Fuzzy logic has rapidly become one of the most successful of today's technologies for developing sophisticated control systems. Fuzzy set theory and fuzzy logic establish the rules of a non-linear mapping [7]. It became a standard tool when process control emerged in 1940s. The use of fuzzy sets provides a basis for a systematic way for the application of uncertain and indefinite models. Fuzzy set is based on a logical system called fuzzy logic controller. It is much closer in spirit to human thinking and natural language than classical logical systems.

Fuzzy sets were first proposed by Lofti A. Zadeh [8] in his 1965 paper entitled none other than: *Fuzzy Sets*. Fuzzy logic can serve as a useful tool dealing with complex systems which are faced with challenges and issues that are associated with reasoning and decision making. Selecting a system totally depends on its internal complexity.

In this paper, a fuzzy set, based on MAMDANI, [9] with two inputs as error and one output. In Figure 5, this organization in MATLAB FUZZY TOOLBOX is shown.

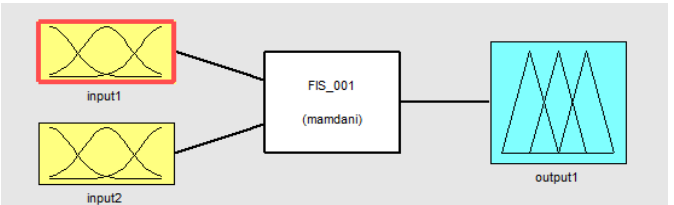


Fig. 4. FUZZY set

Fuzzy control, which directly uses fuzzy rules is the most important application in fuzzy theory [10]. Using a procedure originated by Ebrahim Mamdani in the late 70s, three steps were taken to create a fuzzy controlled machine. Those steps

are fuzzification, Rule evaluation, Defuzzification. Fuzzy rules, which are defined in fuzzy rule box is given below

	Δe_G							
	NB	NM	NS	ZO	PB	PM	PS	
NB	PB	PB	PB	PM	PM	PS	ZO	e_G
NM	PB	PM	PM	PM	PS	ZO	NS	
NS	PB	PM	PS	PS	ZO	NS	NM	
ZO	PM	PM	PS	ZO	NS	NM	NM	
PS	PM	PS	ZO	NS	NS	NM	NB	
PM	PS	ZO	NS	NM	NM	NM	NB	
PB	ZO	NS	NM	NM	NB	NB	NB	

Input, output graph of fuzzy logic:

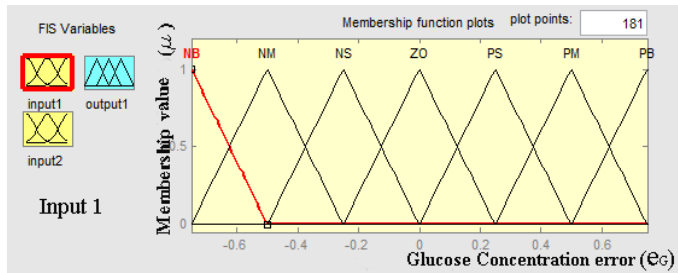


Fig. 5. Membership function of glucose Concentration error

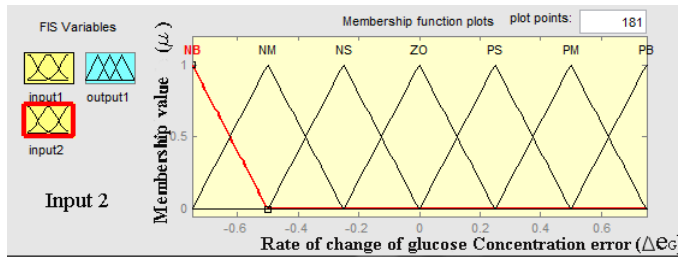


Fig. 6. Membership function of the rate of change of glucose concentration error

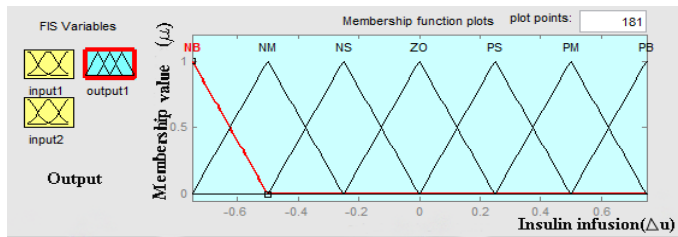


Fig. 7. Membership function of rate of change of insulin infusion

Block diagram of fuzzy logic controller is:

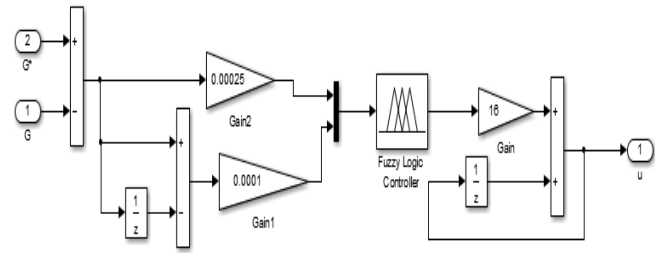


Fig. 8. Fuzzy Logic Controller

V. OPTIMAL REGULATOR

Linear Quadratic Regulator (LQR) is an optimal multivariable feedback control approach that minimizes the excursion in state trajectories of a system while requiring minimum controller effort. At the beginning, virtual patient model was solved under steady state condition and matrixes were for input & output equations.

The input, output equations are:

$$\dot{X}(t) = AX(t) + Bu(t) + Ed(t) \tag{5}$$

$$y(t) = CX(t) + Du(t) \tag{6}$$

Where,

$$A = \begin{bmatrix} -p_1 - x_0 & -G_0 - G_b & 0 \\ 0 & -p_2 & p_3 \\ 0 & 0 & -n(1 - n_b) \end{bmatrix}$$

$$B = \begin{bmatrix} 0 \\ 0 \\ \frac{1}{v_1} \end{bmatrix} \quad C = [1 \ 0 \ 0] \quad D = [0]$$

$$E = \begin{bmatrix} 0 \\ 0 \\ -n_b \end{bmatrix}$$

Block diagram of optimal regulator is:

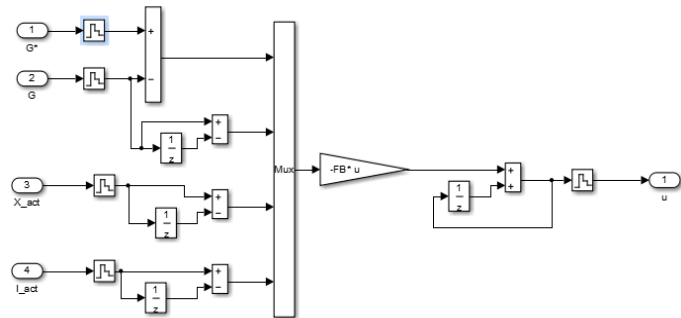


Fig. 9. Optimal Regulator

VI. SIMULATION AND RESULT

Output response of three different types of controller are:

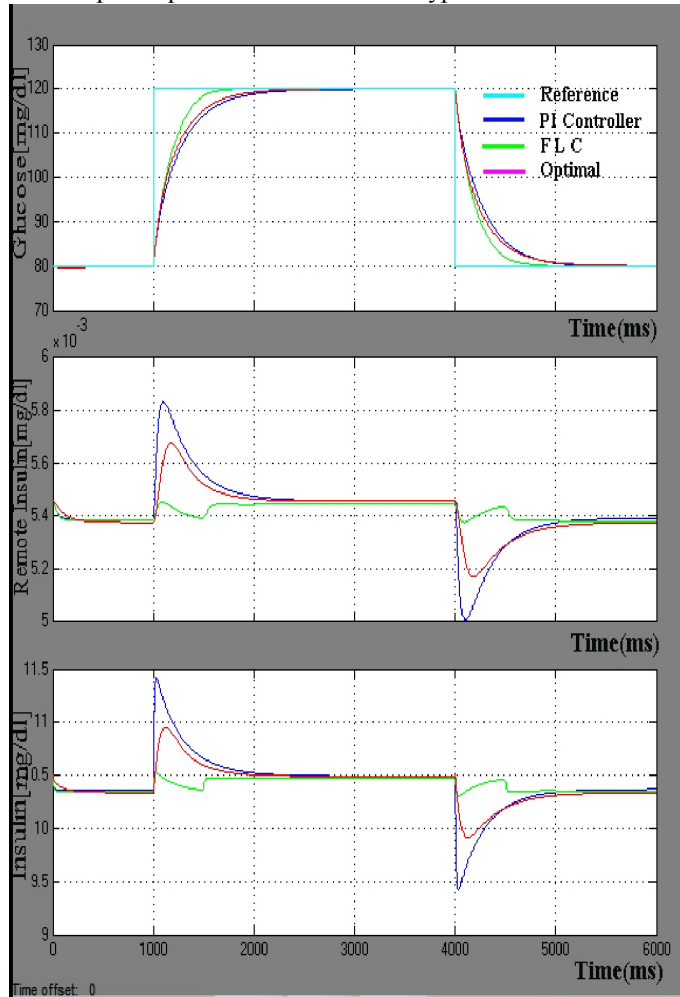


Fig. 10. Combine Responses of three Controllers

In the above graph, we can see that there are three slots. First one is for glucose second one is for remote insulin and third one is for insulin infusion rate. When blood glucose level of a patient is increased above a nominal level, insulin needs to be pushed otherwise a patient can die. In above graph each slot has four wave shapes. Blue represent reference wave, purple represent PI controller wave, green represent fuzzy logic controller wave and pink represent optimal regulator wave.

We calculated a nominal value for glucose level in human body and put this value as reference wave. Now, when one takes his/her meal then his/her glucose level is increased. As a result, insulin level is also increased but not in proportion with glucose. There is some shortage in between glucose and insulin in human body. To meet this shortage, we need to push extra insulin. In above graph remote insulin is insulin which is actually given from external source. Third slot of our graph shows that which controller need how much time to compensate the level of body insulin. We can see in graph that PI controller require maximum amount of insulin and maximum time to push. Optimal regulator need less amount of

insulin and less time than PI controller. Fuzzy logic controller requires minimum amount of insulin and minimum time than optimal regulator and PI controller.

From the analysis of the above graph we can say that fuzzy logic controller is more efficient and give better performance than other two controllers here.

VII. CONCLUSION

In this thesis performances of three different closed loop controllers have been studied. To control blood glucose level of a type 1 diabetes patient we used three controllers and try to analyze that which one gives us the better performance. Those controllers were PI controller, Fuzzy Logic controller and optimal regulator. The main act of the controllers were how efficiently they can push insulin. To control glucose level a virtual patient model was first designed. Then three controllers were implemented using different transfer functions and logics. The performance graph in fig. 10 shows that, when a person takes his/her meal the glucose level is increased. If glucose level crosses the limit of reference value, controller will start to push insulin to keep the glucose level moderate. Finally, from the analysis, it can be said that Fuzzy logic controller is the best. Thus Fuzzy logic controller is more efficient than PI and optimal regulator. After that optimal regulator provide better response and then PI controller.

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Implementation of An EOG Based Security System by Analyzing Eye Movement Patterns

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Abstract— In this paper, a new security system is proposed based on human computer interface by using Electrooculography (EOG) signal keeping concentration on physically disabled people. Nowadays security is the main concern in every sector of our life. Most of the system requires a secret password to get access of the system under protection. In our work, a password based security system is proposed where password can be given by eye via EOG signal. The EOG interface detects the eye movement through electrodes placed on the face around the eyes. A unique pattern of EOG signal obtained by eye movement in certain directions can give the most secure password. An algorithm is proposed to detect the eye movement and extract the features of EOG signal. These features are further used to recognize the pattern for password using string matching technique. This system can be used as an added feature of EOG based assistive system. This is an offline approach which is highly accurate and reliable.

Keywords—*electrooculography (EOG); human computer interface (HCI); eye movement features; saccadic patterns; string matching.*

I. INTRODUCTION

Today's world is based on technology. This technology is intended to ease our daily life as well as that of the people suffering from several disabilities. In recent times, a great advancement has been observed in communication systems between humans and computers. The communication between machines and humans, by means of graphical user interfaces or multimedia applications have been developed mainly. But these advances have been limited only in using keyboards, mouse, joysticks or tactile screens [1]. Recently new assistive technologies in human computer interaction (HCI) based on voice recognition, visual information, brain-computer interfaces BCI and infrared head-operated joysticks [2] etc. has been emerged.

People with severe nervous system injuries or disorders like amyotrophic lateral sclerosis (ALS), lose their ability to speak and move their extremities [3]. But their eyes work actively. Thus eye movement can be used effectively as an interfacing element for assistive technology. Transferring information from eye motion of users to HCI system can help users to communicate with outside world. So, analysis of eye motion is necessary to improve these systems. In present,

many HCI systems already exist that utilize eye movements, including systems of infrared oculography (IROG), dual Purkinje image (DPI) and search coils (SCs) [3]. These systems are very helpful to users. But they are very bulky which restrict the movement of patients.

Electrooculography (EOG)-based HCI system on the other hand is one of the most effective method to detect the eye movement by measuring the difference of potential between the cornea and the retina. EOG signal is used for HCI system because it has relatively large potential differences (amplitudes range between +0.40 to +1.0 millivolts), and their relationship to eye movements is linear (i.e., an angle of -30° to $+30^\circ$). For the above reasons, EOG-based HCI systems are preferred. This system finds its application in controlling television [4], wheelchair navigation [5], robot arm movement [6], mouse, keyboard control [7] etc. These systems serve as rehabilitation for disabled people.

This paper involves a proposal of EOG based security system where password can be provided with a unique eye movement pattern. By detecting eye movement and extracting features the pattern can be recognized. Then this pattern is used to unlock the system. In section II, EOG signal, its features and the method for recording the electrical activity of eye-movement is discussed. In the next section, the algorithm of our proposed method is illustrated. The various tests and investigational results are given in section IV. Finally, the conclusions are drawn in section V.

II. BACKGROUND

A. Electrooculography signal

Electrooculography (EOG) is the corneal-retinal potential produced due to depolarization and hyper polarization existing between the retina and cornea. The higher metabolic rate at the retina compared to the cornea causes the eye to maintain the voltage of about +0.40 to +1.0 millivolts with respect to the retina and with a frequency range of about DC to 100 Hz. At resting state, when user is looking straight, a zero potential signal is obtained. When the direction of gaze is changed, the corneal-retinal potential will also be changed at the same time. The behavior of EOG signal is linear for gaze angles of $\pm 30^\circ$ and changes approximately 20 μV for each degree of eye movement [1].

B. Signal Acquisition

As the bipotentials are rarely deterministic, measuring EOG signal depends on many factors. To minimize the artifacts due to these factors such as EEG (electroencephalogram), EMG (electromyogram), the positioning of the electrodes, skin-electrode contacts, head and facial movements, lighting conditions, blinking, etc., ideal condition is maintained[8]. The subject is hold in a position of minimum movement. The environment is also kept less noisy. BIOPAC MP36 data acquisition system is used to attain the EOG data for this experiment. The acquisition system and the placement of five electrodes are shown in Fig. 1.

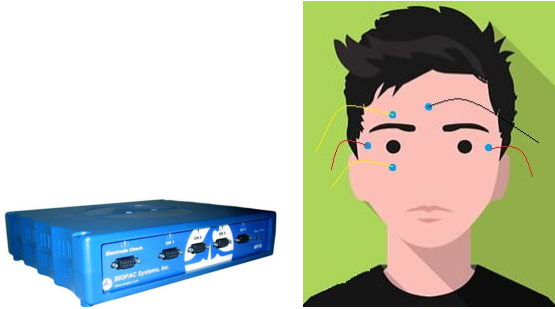


Fig. 1. (a) BIOPAC MP36 data acquisition unit and (b) placement of electrodes (Yellow=vertical electrode, Red=horizontal electrode, Black=reference electrode)

III. SYSTEM ARCHITECTURE

The overall system architecture is shown in Fig. 2.

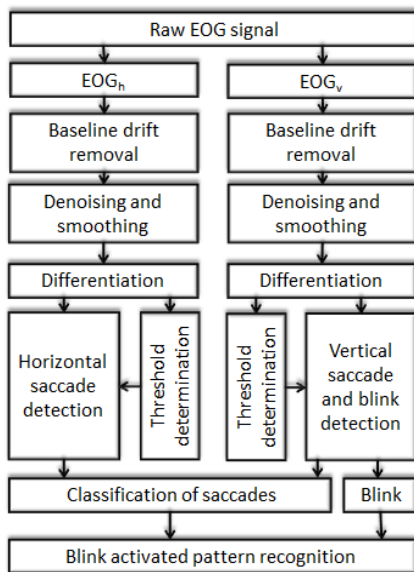


Fig. 2. System block diagram.

A. Preprocessing

1) *Baseline drift removal*: User concentration is required while using eye movements to control graphical interfaces. Signal extraction for a long time may make the users tired and so it becomes more difficult for users to perform eye movements, this phenomenon is known as „jitter“ [1]. For this reason, users’ capacity to control their ocular motor system

decreases inversely proportional to tiredness. This event shows drift of signal from baseline. It can be eliminated by modeling the extracted EOG signal with a high degree polynomial. It is then subtracted from the original signal to cancel out the baseline drift. A higher order Polynomial increases the probability to estimate the baseline [9]. A polynomial of degree 20 has been chosen to model the baseline [10].

Feature Detection

1) *Automatic threshold determination*: Subject independent analysis can be achieved by an automatic threshold determination algorithm which eliminates the need of calibration. In this algorithm at first the local maxima of the derivative of EOG horizontal and vertical components has been located and categorized in ascending order. Large artifact peaks are removed by eliminating peaks above the points where the difference between sequential peaks is less than the typical maximum amplitude difference between blink and saccadic peak i.e. 0.35 mV. From these peaks the appropriate peaks are selected (depending on the feature being desired) using a universal threshold process. The universal threshold process separates the saccadic peaks from the blink peaks. The maximum value of the peaks corresponding to blinks and saccades are the desired threshold values of blink and saccade respectively. Thus blink and saccade thresholds are determined for both EOG_h and EOG_v. The selection of valid peaks is shown in Fig. 4.

2) *Blink Detection*: Blink features are detected from the vertical EOG signal (EOG_v). Blinks are characterized in the EOG_v signal by high amplitude peaks of short duration. Derivative of EOG_v characterizes the blinks as a distinct signal feature having two peaks with a zero crossing between them as shown in Fig. 5. So this distinct characteristic is detected. Involuntary eye blinks are eliminated by thresholding the detected blink amplitudes with the blink threshold determined earlier from the automatic threshold determination algorithm. Involuntary eye blinks has smaller amplitudes and smaller duration than voluntary eye blinks.

3) *Saccade Detection*: From the EOG signal characteristics it can be concluded that when the signal amplitude increases then the corresponding eye movement is towards the rightward direction for EOG_h or towards the upward direction for EOG_v. Similarly decreasing amplitude indicates eye movement in opposite direction. So saccades can be detected by simply observing the change in amplitude i.e. by observing the derivative of the signals. The saccadic peaks for horizontal saccades are shown in Fig. 6.

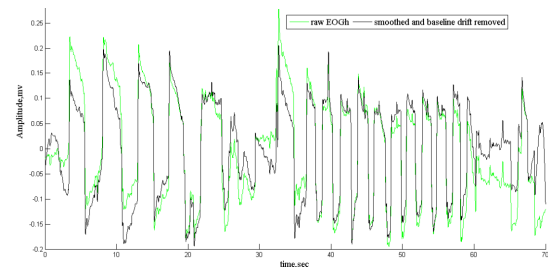


Fig. 3. Baseline drift removed and smoothed EOG (Horizontal) signal along with raw signal.

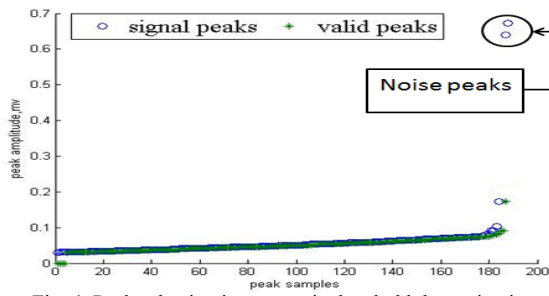


Fig. 4. Peak selection in automatic threshold determination

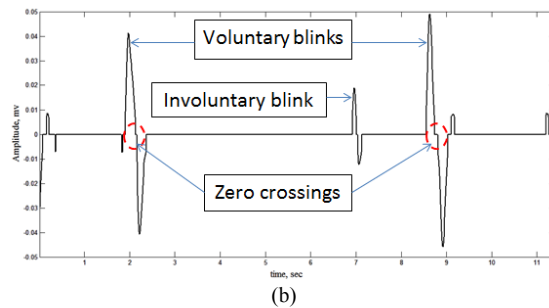
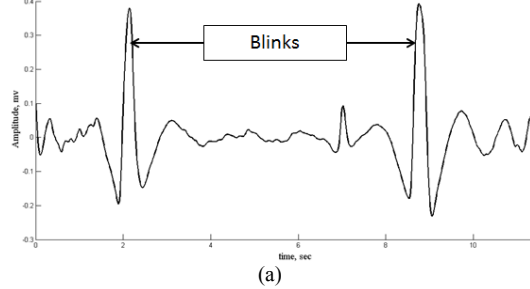


Fig. 5. (a) Blinks in EOGv and (b) Voluntary and involuntary blinks.

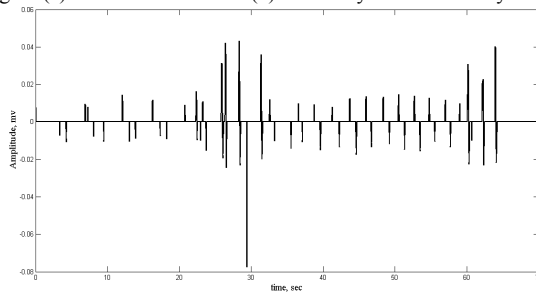


Fig. 6. Saccadic peaks for horizontal saccades.

C. Classification

After the features of interest are extracted from both the horizontal and vertical signals, they are classified. Horizontal and vertical saccades are extracted from the derivative of EOGh and EOGv respectively. An oblique saccade is found when a horizontal and a vertical saccade occur simultaneously. So a total of eight directional saccades are found.

D. String Encoded Pattern Recognition

The eye movement features corresponding to eight directions are encoded by character strings as shown in Fig. 7. The characters 'r', 'l', 'u', 'd', 'e', 'f', 'g', 'h' are assigned to the directions right, left, up, down, right-up, right-down, left-up, left-down respectively. A string is formed from the detected saccadic features. A particular eye movement pattern can then be found from the encoded string by simply comparing it with the desired eye movement string pattern.

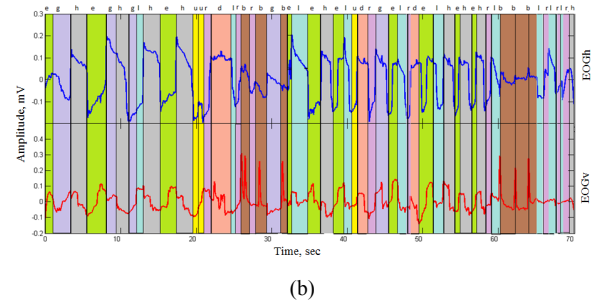
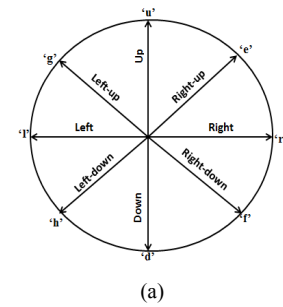


Fig. 7. (a) Directions of eye movement and (b) Movements encoded by character strings in the EOG signal.

IV. RESULTS

The EOG data were acquired using BIOPAC MP36 data acquisition system in Biomedical Engineering Lab of KUET. The data were then analyzed according to the method proposed using MATLAB. The eye movement features were detected and classified for different data sets corresponding to different subjects. The features were string encoded and later used for pattern recognition.

A. Experimental Setup

Three subjects were used (two male and one female; two of them wore glasses, age: 22-25) for the experiment where they were asked to move their eye in certain predefined patterns with a blink at the start and end of each movement pattern. Video-oculography i.e. video recording of eye movements through digital camera was also collected during the experiment. The experiment was conducted in a controlled environment where they were instructed to keep calm and to keep their body and head movements as minimum as possible to minimize EMG noise. Fig. 8 shows the measured EOG signal.

B. Extracted Features

Total five eye movement features were extracted namely: Left, Right, Up, Down and Blink. The oblique saccades were then found. The extracted features from a recording of 100 seconds for the three subjects are shown in Table: 1. The detected features are shown in Fig. 9.



Fig. 8. EOG signal extraction and extracted raw EOG signal.

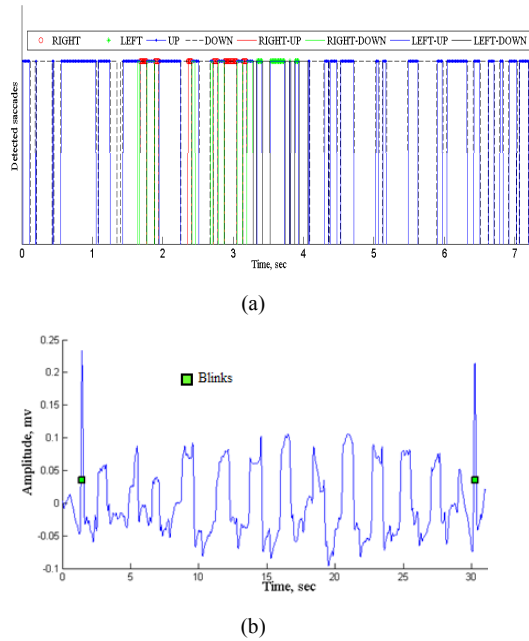


Fig. 9. Detected features for a small data sequence (a) saccades and (b) blinks.

TABLE I. NUMBER OF DETECTED FEATURES FOR THREE SUBJECTS

Subjects	Right	left	Up	Down	Right-up	Right-down	Left-up	Left-down	blinks
S1	92	85	448	447	89	93	85	83	22
S2	37	61	418	409	39	37	61	57	13
S3	138	205	264	256	141	133	204	195	13

B. Application to Security System

Various patterns of eye movement can be used as passwords for the security system. Some sample eye movement patterns is shown in Fig. 10. Each pattern of eye movement starts and ends with a blink in order to differentiate normal eye movement from the desired eye movement pattern. The proposed system encodes the eye movement features detected earlier into strings as described in previous section and try to match the desired string with this detected string. If a match is found, the device under security is granted operation otherwise ask for the password again. Sample detection for a particular eye movement pattern is shown in Fig.11.



Fig. 10. Sample eye movement patterns.

f r l d u d u d u d u d u d u l r f r u l r f r u d u d u l r

Fig. 11. Detection of a particular pattern (udud).

V. CONCLUSION

In this work an EOG based security system is proposed and implemented offline. People with severe brain injury can use this system with EOG based assistive system. The system is triggered by eye blinks and uses detection of saccades for eye movement pattern recognition. The detection accuracy of blinks and unidirectional saccades was 100% but for oblique saccades the accuracy dropped to 51.7% measured through video-oculography data. The accuracy of the system can be improved with better acquisition system. The proposed method can be implemented practically in real-time using improved wearable EOG acquisition system[11]. Normal people can also use this system with wearable EOG acquisition system. It can be used to protect devices such as personal computers, smart phones, etc. This EOG based approach is preferable than other approach such as involving image processing due to its accuracy, cost effectiveness and simplicity.

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Development of Rhythm Practice Supporting System with Real-time Onset Detection

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Abstract—We are developing a rhythm practice support system with onset detection. The system produces guiding metronome sounds and processes the performed sound to detect onsets in real time. Spectral flux and dynamic thresholding are used to find onsets in the audio signal. We made three types of prototype display for the system; a survey of user display preference was conducted. An evaluation of the accuracy of the onset detection is also provided.

Keywords—Music playing, Rhythm practice, Onset detection, ASIO, Chuck, Processing

I. INTRODUCTION

When playing music, a sense of rhythm is critically important. Although the performed pitch can be measured and visualized using a musical tuner, training for rhythm is usually carried out using a metronome and human judgement. For the beginner player, this judgement is difficult and not as precise as that of skilled players. Even for skilled players, the practice of rhythm is non-negligible because slight differences in note production timing, such as groove, have a great effect on the expression of musical feelings.

Our system aims to support rhythm practice by visualizing the timing of onsets, which is the starting time of the musical note in the audio signal. The system produces tick sounds similar to those of a metronome, and derives the performed sounds. The onsets in the audio signal is graphically displayed along with the timing of the metronome sounds in real-time.

II. RELATED WORKS

There are several studies of systems that assist practicing musical instruments [1]-[11]. The IMUTUS (Interactive Music Tuition System) [1] is an open platform for training beginner students, which enables playback, recording, and performance evaluation for simple music. The system uses real-time signal analysis for score following and off-line signal analysis, including onset detection, to provide detailed feedback. IMUTUS evolved into the VEMUS (Virtual European Music School) project [4]. In addition to self-learning opportunities, VEMUS offers group learning and distance learning. Tambouratzis states that the system relies on the participation of music teachers, such as regular face-to-face lessons on a weekly basis, to provide both the foundation of the concept

and teachers' experience [4]. This kind of system is used in a supporting role for music training.

In more specific musical instrument training situations, systems for piano [2], [7], drum [3], [5], violin [8], and cello [10] have been studied. Several other pieces of proprietary software are listed in [9] and [11]. Most of these applications make use of MIDI (Musical Instrument Digital Interface) for input user-interfaces and/or to use with a particular piece of music. On the other hand, our system inputs audio signals and processes arbitrary patterns of monophonic musical instrument. Furthermore, real-time processes regarding rhythm, which are merely treated in the previous studies, are implemented.

III. ONSET DETECTION

The onset is the starting time of a note in a musical sound. To detect the onset, various methods have been proposed [12]-[18]. These methods commonly use STFT (Short Time Fourier Transform). The time-frequency representation $X(n, k)$ of the audio signal $x(t)$ is expressed as

$$X(n, k) = \sum_{m=0}^{N-1} x(nh + m)w(m)e^{-\frac{2j\pi mk}{N}}, \quad (1)$$

where n is the time in the spectrogram, k is the frequency, N is the FFT (Fast Fourier Transform) frame size, h is the hop size of the STFT, and $w(m)$ is the FFT window. We should distinguish the discrete time n in the spectrogram from the time t in the audio signal. The time indicator n in the time-frequency representation differs from the audio signal time t as follows:

$$n = \left\lfloor \frac{t - N}{h} \right\rfloor, \quad t \geq N \quad (2)$$

Because we consider the real-time environment, we must wait until the data of the first FFT frame arrives.

Generally, onset detection consists of three steps: pre-processing the input signal, making detection function, and peak picking. In the first step, normalization is applied to the original signal. In the next step, a detection function, also called reduction function, is calculated. There are several methods for deriving the detection function; we use the spectral flux

(SF) method. SF exploits the difference of adjacent frequency bins in the power spectrum; it is represented as follows:

$$SF(n) = \sum_{k=1}^{\frac{N}{2}-1} \max(0, |X(n, k)| - |X(n-1, k)|) \quad (3)$$

The operator “max” considers only the case of an increase in the power of the adjacent spectrogram bin. In the final step of onset detection, we used peak picking with dynamic thresholding. The local maximum above the threshold value is considered to be the onset. The dynamic thresholding level $TH(n)$ from the following equation is employed.

$$TH(n) = \delta + \lambda \text{ median}(SF(n - u_1 : n + u_2)) + \alpha \text{ mean}(SF(n - u_1 : n + u_2)) \quad (4)$$

where δ is a constant factor, λ and α are weights for median and mean respectively. The parameter δ works for avoiding noises at low input audio signals, while λ and α considers great dynamics, which is common in a music signal, as shown in Rosão [15]. The parameters u_1 and u_2 are the preceding and succeeding target widths of the dynamic thresholding, respectively. The detection function $DF(n)$ and the onset time $OT(n)$ are expressed in the following equations.

$$DF(n) = SF(n) - TH(n) \quad (5)$$

$$OD(n) = \begin{cases} 1 & DF(n) > 0 \text{ and } \underset{v_1 \leq m \leq v_2}{\operatorname{argmax}} DF(m) = n \\ 0 & \text{otherwise} \end{cases} \quad (6)$$

where v_1 and v_2 are the preceding and succeeding target widths for peak picking used to obtain the local maxima.

IV. SYSTEM CONFIGURATION

Fig. 1 shows an overview of our system. We constructed this system on Windows OS. Chuck and Processing are used for programming languages. Chuck, a real-time music programming environment developed at Princeton University, is used as the main part of the system; it generates metronome sounds and processes sound signals from microphone input. On the other hand, Processing is used to provide visual feedback to the user. Windows' standard audio device has a relatively long latency (100 to 200 ms) for A/D conversion; therefore, it is not suitable for our real-time purposes. Thus, to acquire sound data from the microphone input with low latency, we used ASIO4ALL. ASIO (Audio Stream Input/Output) is a

commonly used device driver interface for PC audio devices. Although the original ASIO API (Application Programming Interface) requires a compliant sound device hardware on a PC, the ASIO emulator ASIO4ALL can be used with ordinary sound cards or devices.

The system generates metronome sounds at a specified tempo and outputs it from the speaker. The downbeat and upbeat are distinguished by pitch, 880 Hz and 440 Hz, respectively. The user of this system plays a musical instrument while hearing the metronome sounds. The timing of the metronome ticks and detected onsets is transferred via Open Sound Control (OSC) from Chuck to Processing. OSC is a communication protocol for sending and receiving musical performance data between PC applications and electronic musical instruments. With OSC, performance data are conveyed in text form like HTML.

V. EXPERIMENT

To exclude the influence of the room's sound reflection, all experiments are conducted in an anechoic room. We set the input signal as monaural, the sampling frequency to 44.1 kHz, and 1024 samples as FFT frame width. A Hanning window is applied to the frame. The STFT hop width is 10 ms, which is equal to the time resolution of the system. The onsets were detected with the SF method on spectrogram, which is a graphical representation of the time-frequency information.

To confirm the results of onset detection, we input the generated metronome sounds only. As the result, there is a delay of roughly 20 ms; however, all of the onsets of the metronome beats are correctly detected.

We tested the system with three kinds of musical instruments: flute, trumpet, and French horn. The test pattern is shown in Fig. 2; it is a one-octave scale with tonguing and legato (total 48 notes) at tempo 60 (approximately 15 seconds) in B-flat major. The parameters in section 2 were set to $u_1, v_1, v_2=50\text{ms}, u_2=0$. The ground truth data were the average of three listener judgements from the audio signal envelope and the spectrogram. When the detection falls between 25 ms before and 55 ms after a ground truth datum, we classified the onset a true positive (TP), which means that it was correctly detected. Usually the range of tolerance is set to 25 ms before and after the ground truth data for the fast-attack instruments such as piano; however, we extended the latter tolerance to 55 ms in order to accommodate the slow rise of the attack sounds of these wind instruments. Falsely detected onsets are classified as false positive (FP). Onsets that were not detected are classified as false negative (FN). To evaluate how the onsets are accurately detected, the F-measure is commonly used, which is the harmonic mean of precision P and recall R .

$$F - \text{measure} = 2PR(P + R) \quad (7)$$

$$P = TP / (TP + FP) \quad (8)$$

$$R = TP / (TP + FN) \quad (9)$$

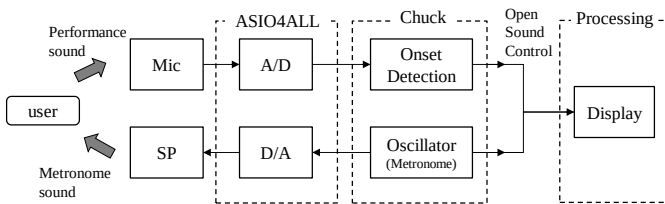


Fig. 1. System Diagram



Fig. 2. Test pattern (in Bb)

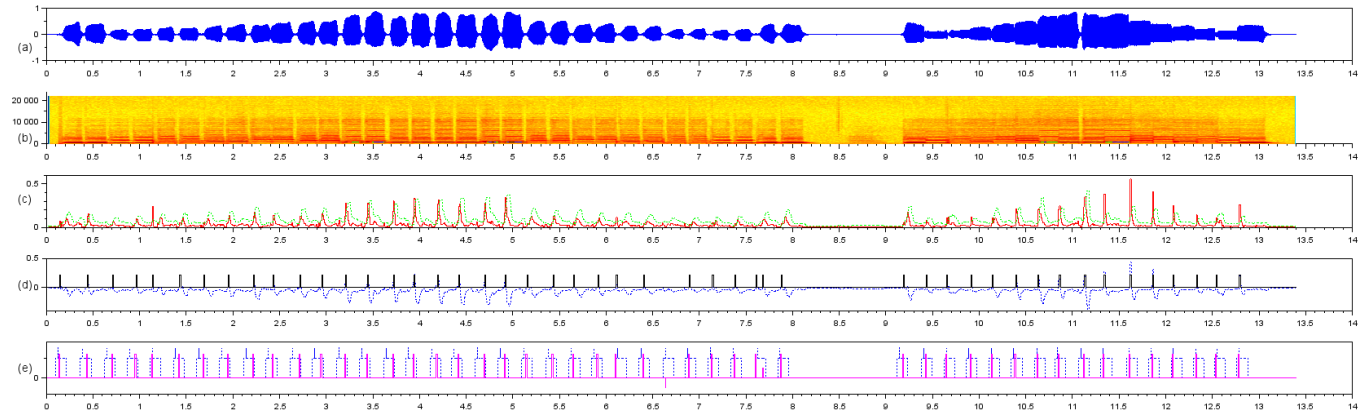


Fig. 3. The process of onset detection. (a) audio signal, (b) spectrogram, (c) spectral flux (red) and threshold (green), (d) detection function (light blue), detected onset (black), (e) ground truth data (blue), judgement (magenta)

TABLE I
ONSET DETECTION RATES

Instrument	Precision	Recall	F-measure
Flute	0.980	0.980	0.980
Trumpet	0.980	0.980	0.980
French horn	0.941	1.000	0.970

Table I shows the result of the onset detection. The precision of French horn is relatively low. That's because its input level was lower than that of other two instruments and it cause lower S/N ratio, since the bell of French horn faces backward, which direction is opposite to the microphone. Fig. 3 shows the process of onset detection in detail; it is aligned roughly to fig. 2. This figure was created on Scilab using recorded data for the test pattern shown in Fig. 2 with a flute. Fig. 3 (a) shows the normalized audio signal, and (b) is its spectrogram, which corresponds to eq. 1. In fig. 3 (c), the red line shows the SF in eq. 3, and the green line represents the dynamic thresholding level TH in eq. 4. In fig. 3 (d), the light blue line indicates the DF in eq. 5, and the OD in eq. 6 is indicated by the black line. In fig. 3 (e), the correctness of the onset detection is indicated by the magenta line, which is compared to the ground truth data indicated by the dotted blue line. When the magenta line rises upward long, it represents a TP. The line going short upward indicates a FP. The short downward change is a FN. Each span of blue line indicates the tolerance range, containing the ground truth point, represented by the upward pulse. In fig. 3, there are 47 TPs out of 48 notes, one FP around 7.7 ms, and one FN around 6.6 ms.

Fig. 4 shows prototype output user interfaces for the system.

Fig. 4 (a) mimics a pendulum-swing type metronome. The swinging arm represents the beat, and a line is recorded at the positions at which an onset is detected. Fig. 4 (b) is a cycloid form output. A radial bar rotates clockwise. The upward direction corresponds to the start of the first beat, and the rightward to the second beat. The position of the numbers express the timing of the detected onsets and their pronounced order. In fig 4 (c), the metronome bar proceeds to right in the first row, which corresponds to the first beat. Then, the bar moves to the second row during the second beat. The red dots show the detected onsets.

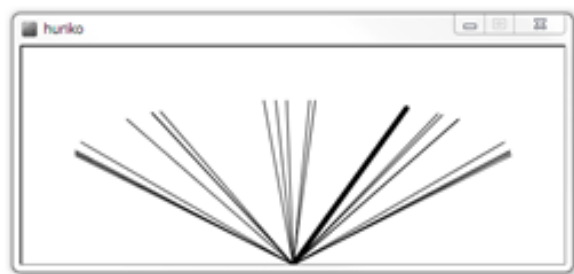
We carried out a survey to determine which display form was preferred by certain musicians. The number of the participants is 15: flute, oboe, bassoon, B-flat clarinet (2), bass clarinet, tenor saxophone (2), trumpet (2), French horn (2), trombone, tuba, and snare drum. Each player performed the test pattern in fig. 2 at first; then, they used this system as they liked. Table II shows the result of the questionnaire. The cycloid style was preferred.

VI. CONCLUSION

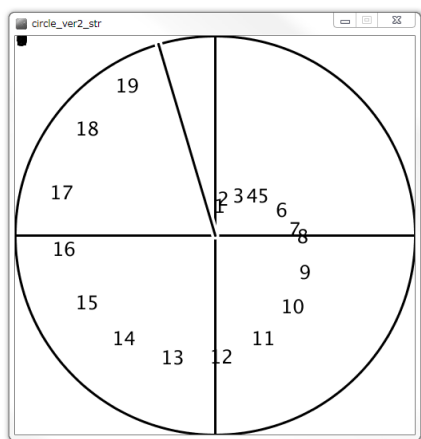
In this paper we have presented a system for rhythm practice. The system outputs metronome tick sounds; players

TABLE II
USER'S PREFERENCE

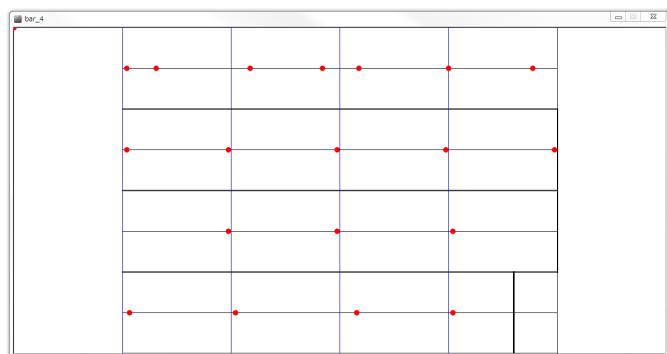
Display style	1st	2nd	3rd
(a) Pendulum-swing	1	2	1
(b) Cycloid	9	5	12
(c) Time-bar	5	8	2



(a) Pendulum-swing display



(b) Cycloid display



(c) Time-bar display

Fig. 4. Three types of display of detected onsets

can perform while listening to the ticks. The onsets from a monophonic musical instrument sound can be derived in real-time using SF and dynamic thresholding. The detected onsets are displayed in three ways. In a user survey, the cycloid style display was preferred. The parameters for onset detection are set manually in this work; automatic determination of them is left as the subject of future studies.

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Database Optimization of the Source Code to Recommend Software Developers using Canonical Order Tree Algorithm

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Abstract—Recently frequent and sequential pattern mining algorithms have been widely used in the field of software engineering to mine various source code or specification patterns. In practice software evolves from one version to another is needed for providing extra facilities to user. This kind of task is challenging in this domain since the database is usually updated in all kinds of manners such as insertion, various modifications as well as removal of sequences. If database is optimized then this optimized information will help developer in their development process and save their valuable time as well as development expenses. Some existing algorithms which are used to optimize database but it does not work faster when database is incrementally updated. To overcome this challenges an efficient algorithm is recently introduce, called the Canonical Order Tree that captures the content of the transactions of the database and orders. In this paper we have proposed a technique based on the Canonical Order Tree that can find out frequent patterns from the incremental database with speedy and efficient way. Thus the database will be optimized as well as it gives useful information to recommend software developer.

Keywords—Frequent pattern mining, Canonical Order tree, Incremental mining, AFPIM (Adjust Frequent Pattern Incremental mining).

I. INTRODUCTION

The problem of mining association rules and the more general problem of searching frequent patterns from huge databases have been the subject of numerous studies. These studies can be classified into two categories. First category is the functionality where the major question is what kind of rules or patterns are needed to compute. While some studies [1], [2], [3] in this category are considered as data mining exercise in isolation, whereas the data mining can work best with the database management system [4], [5] and the thing that human use. Second category is performance where the major question is considered that how to determine the association rules or frequent patterns as perfectly as possible. Studies in this category again classified into several subgroups. This subject of mining frequent pattern also used in software development or update process. Software development process for a system is originally done by writing a program based on the specific design specification. At present it can be done by various techniques. The main purpose of the software development or update is to speed up the development process. For this developers in the software industries usually create new methods, attributes or modifies existing or reuse previously define

methods, attributes. One of the processes to develop software is reusing the existing libraries and frameworks. Developers can be highly benefited if they get appropriate guidelines and accurate recommendations. It can decrease development cost as well as save valuable time. Scanning and Recommending Application Programming Interface (API) is a process to search relevant items from an existing repository.

This process is used widely across all over the world. Some approaches of these API searching processes are based on searching online. Developers generally search for API usage pattern using online repositories provided by various tools.

Code search engines often deals with the extra large files. The extra large files often contain a large number of API methods, attributes. Developers face the difficulty is to choose the appropriate API methods, attributes to use. As the database contains a large amount of data and takes a long time to find appropriate methods, attributes and also it does not provide specific frequent pattern of attributes. However, to solve the problem of finding frequent pattern an efficient algorithm called Canonical Order tree [6], [7] can be applied and it perform well than some existing algorithms. This task of finding frequent pattern can be applied in other field like optimization of database to gain some valuable information and based on this information provide appropriate guidelines to developers for evolving software from one version to another. It is needed for software developer to take decision which portion of existing version is needed or which portion can be ignored and modified to make the software more user friendly. Such information can be obtained from software database and normally that portion which is frequent is needed to add with the update version of the software. Based on this information database of the software is optimized. But if there are any system or technique which helps to gain this optimized information of the software database that will be very useful for software developer. In this paper we show this technique and the experimental result shows that our proposed technique are performed better than the existing algorithm [8], [9] based on the criteria of time complexity.

II. BACKGROUND AND RELATED WORKS

There are some existing algorithms which mine frequent pattern and optimize database as perfectly as possible. Studies

in this category again can be classified into several subgroups. The first subgroup is made of fast algorithms based on the level wise Apriori framework [10] and second subgroup focuses on the performance enhancement mechanism like hashing and segmentation [11], [12] for speeding up Apriori based algorithms. In summary, those Apriori based incremental mining algorithms can not be easily adoptable to FP-tree based [13], [14] incremental mining. To overcome the situation the Frequent/Large patterns mining (FELINE) with the Compressed and Arranged Transactional Sequences (CATS) tree [8] was invented. But it was mainly designed for interactive mining, where the build once, mine many principle holds. However, such mechanism does not necessarily hold for incremental mining. The another algorithm called AFPIM algorithm [9] was proposed to reduce this problem but it is unable to eliminate the task of rescanning of the database when it is change dynamically. In our proposed technique which is based on canonical order tree can mine frequent pattern from the incremental database and rescanning is not needed when database is changed dynamically.

III. PROPOSED TECHNIQUE

We use the sample database which is taken from open source repository and apply our proposed technique to find the frequent pattern and thus optimize the database of the software. Here we have considered a sample database called google Maps API in version 2 and version 3. The following table and figure shows our implementation of the techniques. Here it can be seen that the frequent pattern (e.g. methods, properties, events use in various classes) of the database of the software in version 2 is also used in updated version 3. Here the minimum support we use 50% that means minimum support value=2. In V2 frequent pattern are mouseout, mouseover, clickable and getBounds(). In V3 the frequent pattern are clickable, getBounds(), mouseover, getmap(), rightclick and visible.

A. Find frequent items

Here the term frequent item means the item which satisfying the minimum support.

B. Applying canonical order tree to find frequent item

We have applied canonical order tree as sequential or frequent pattern mining algorithm to optimize the item of database Table I and III. The result is represented with Figure 1 and 2 respectively. It will scan the items as per the minimum support and sort out the expected items.

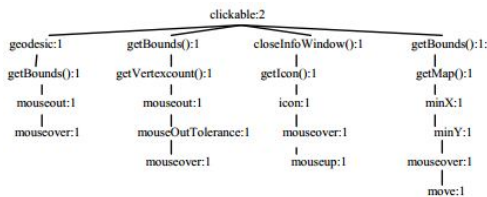


Fig. 1. Implementation of Canonical Order Tree for Google Maps API V2

The algorithm is applied on the Table I and after optimization Table II is generated.

The algorithm is applied on the Table III and after optimization Table IV is generated.

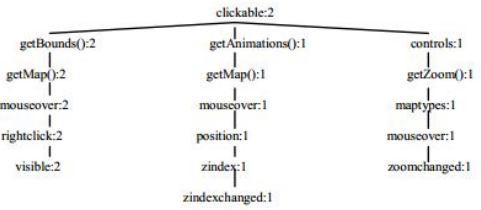


Fig. 2. Implementation of Canonical Order Tree for Google Maps API V3

TABLE I. SAMPLE DATABASE OF GOOGLE API V2

Serial	Transaction	Item set
Item1	google.maps.Map	getMap(),getBounds(),mouseover,min X,min Y,move
Item2	google.maps.Marker	closeInfoWindow(),getIcon(),icon,title,mouseover,mouseup
Item3	google.maps.Polyline	getBounds(),mouseout,mouseover,clickable
Item4	google.maps.Polygon	getBounds(),mouseout,mouseover,clickable

TABLE II. OPTIMIZED SAMPLE DATABASE OF GOOGLE MAP API V2

Serial	Transaction	Item set
Item1	google.maps.Map	getBounds(),mouseover
Item2	google.maps.Marker	Mouseover
Item3	google.maps.Polyline	getBounds(),mouseout,mouseover,clickable
Item4	google.maps.Polygon	getBounds(),mouseout,mouseover,clickable

TABLE III. SAMPLE DATABASE OF GOOGLE MAP API V3

Serial	Transaction	Item set
Item1	google.maps.Map	getZoom(),setOptions(),mapTypes,zoomchanged,mouseover,controls
Item2	google.maps.Marker	getAnimation(),getMap(),position,zindex,mouseover,zindexchanged
Item3	google.maps.Polyline	getMap(),getBounds(),visible,rightclick,mouseover,clickable
Item4	google.maps.Polygon	getMap(),getBounds(),visible,rightclick,mouseover,clickable

TABLE IV. OPTIMIZED SAMPLE DATABASE OF GOOGLE MAP API V3

Serial	Transaction	Item set
Item1	google.maps.Map	mouseover
Item2	google.maps.Marker	getMap(),mouseover
Item3	google.maps.Polyline	getMap(),getBounds(),visible,rightclick,mouseover,clickable
Item4	google.maps.Polygon	getMap(),getBounds(),visible,rightclick,mouseover,clickable

IV. ALGORITHM ANALYSIS

In our proposed approach Canonical Order Tree [6], [7] is applied over the other Frequent Mining Algorithms. From the investigation it is found that Canonical Order Tree has the maximum efficiency comparing with other algorithms like GSPAN, PrefixSpan, SPADE etc. The implementation of Canonical order tree is given here.

Step-1: Input transaction number, data item sets in 2D array, minimum support.

Here the transaction number represent items of the database table. Minimum support is an integer value and depends on this value the canonical order tree algorithm find the sequential or frequent item from the database. Data sets input in two

dimensional array. Here two dimension is represent database table row and column. Step 1 is illustrated in the Figure 3.

```

transaction_number = input.nextInt();
t=t+transaction_number;
for(k=n+1;k<=t;k++)
{m= input.nextInt();
for (i=1;i<=m;i++)
a[i]=inData.readLine();
for (i=1;i<=m;i++)
for (j=i+1;j<=m;j++)
{ q= a[i].compareTo(a[j]);
if(q>0) {
temp = new String(a[i].toString());
a[i] = new String(a[j].toString());
a[j] = new String(temp.toString());
}}
}

```

Fig. 3. Code Snippet of Primary Input

Step-2: Sorted data items into canonical order (for example lexicographic order). Step 2 is illustrated in the Figure 4.

```

for(i=1;i<=p1;i++)
{ for(j=1;j<=p1;j++)
{if(i==j) continue;else
{k=1;while(k<=1)
{q = b[i].equalsIgnoreCase(c[k]);
k++;}
}
}

```

Fig. 4. Code Snippet of Canonical Orde

Step -3: Support count of every items of Sorted data item. Step 3 is illustrated in the Figure 5.

```

if(flag==1)
{q= b[i].equalsIgnoreCase(b[j]);
if(q==true){count++;}
}

```

Fig. 5. Code Snippet of the support count of sorted data items

Step- 4: Checked if item set satisfied minimum support.

If yes then item set is stored in the array and considered it as frequent item. If no then ignore that item set. Step 4 is illustrated in the Figure 6.

```

if(flag==1)
{q= b[i].equalsIgnoreCase(b[j]);
if(q==true){count++;
c[l] = new String(b[i].toString());
l++;}
}

```

Fig. 6. Code Snippet of checking item satisfy minimum support

Step- 5: Checked if all sorted item sets is checked?

If yes then frequent item set is stored in the array and ready to print.

If no then continue from step 3 to 5. Step 4 is illustrated in the Figure 7.

V. EVALUATION

To evaluate the applicability and the efficiency of our proposed approach, we have compared it with the other existing algorithms like AFPIM, FELINE. Before, analyzing the result a dedicated experimental environment is set up.

```

if(count>=min_support)
{System.out.println("\n");
System.out.print(" "+b[i]);}

```

Fig. 7. Code Snippet of checking all items of database

A. Environmental Set up

The result found are simulated from WEKA [15] MATLAB, Netbeans IDE and Turbo C compiler in system of core-i5 processor with 4GB RAM along with the internet connectivity of local operator.

B. Time Complexity Evaluation

The algorithm is implemented in Turbo C compiler as well as Netbeans7.3 compiler(first 3 is Netbean and last 3 is C compiler and time calculate in millisecond). This is shown in the Table V.

TABLE V. TIME COMPLEXITY EVALUATION

Task	Canonical order tree	AFPIM	FELINE
1	255060	373932	405429
2	299515	373998	773455
3	373752	573997	999465
4	255165	405419	573999
5	299717	433799	905997
6	373852	583999	999715

Here the Y-axis represent time in ms and X-axis represent iteration of applying three techniques.

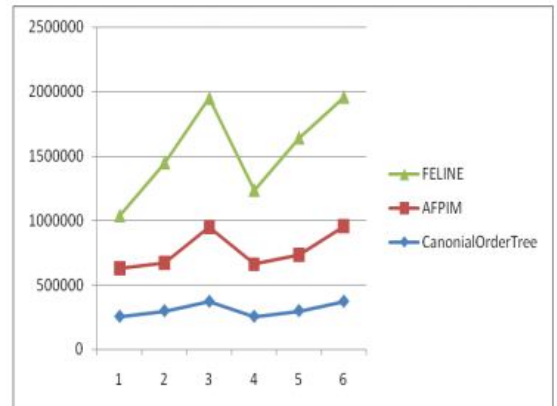


Fig. 8. Runtime evaluation in graphical representation

From the Figure 8 it can be seen that our propose technique based on canonical order tree is comparatively more than the others.

C. Empirical Evaluation

To see that if the proposed approach is helpful to software developer, an experimental investigation is held through empirical evaluation. Before starting up the empirical evaluation, four different teams consisting of three members are formed, and they are said to work with the different systems environment including the Can Tree Approach along with the other existing algorithm. The errors comparison found from the Empirical Evaluation is shown in Table VI.

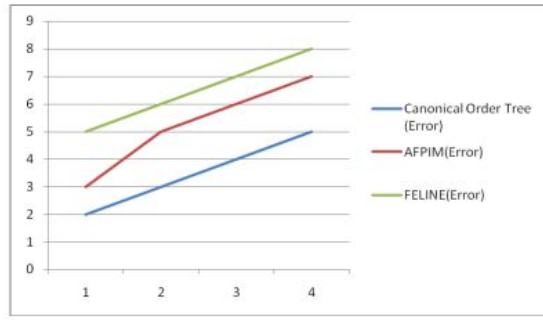


Fig. 9. Graphical representation of empirical evaluation

The result is given below.

Here Y-axis represent error and X-axis represent team number

From the Figure 9 it can be seen that the proposed approach brings less error comparatively than others.

TABLE VI. EMPIRICAL EVALUATION

Team	Canonical order tree (Error)	AFPIM(Error)	FELINE(Error)
1	2	3	5
2	3	5	6
3	4	6	7
4	5	7	8

VI. THREAT TO THE EVALUATION AND LIMITATION

Here in the experimental evaluation google API is considered that is not standard repository. If different repository is used the result may be varied. If team member of the empirical evaluation is changed the result may be changed. Learning curve of the particular team member influences the evaluation result, as seen from the experimental evaluation.

VII. CONCLUSION & FUTURE WORK

Software development is getting changed so rapidly. Existing repository can help software developers. If the existing repository is optimized then developers can be guided accurately. The frequent pattern, found from the incremental database using the proposed approach is more relevant and convenient than the others. Based on this frequent pattern database is optimized and this optimized information recommend software developer. We still have been working on the proposed approach to extend its applicability for the professional uses, through real time implementation, the proposed approach also can be applied along with the application of graph mining algorithm. The repository we use is not standard so we want to work with standard repository. Also we want to work with better simulation environment than current simulation environment.

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A Survey of Spectrum Sensing Techniques in Cognitive Radio Network

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Abstract—There is a huge demand of spectrum now a days, but in conventional system radio spectrum is very scared resource. To overcome this problem and improve spectrum utilization, Cognitive Radio (CR) concept has been proposed. CR is the greatest mechanism for use spectrum efficiently. The main objective of CR is to use scarce and limited natural resources efficiently without any interference to the primary users (PUs). This paper explains the several spectrum sensing methods that used in CR so far and also deals benefit-drawback of each sensing method. We also explained the cooperative spectrum sensing and it's all forms and also introduced the multi-dimensional spectrum sensing concept. Cooperative sensing shares its sensing data with others and utilizes the sensing outcomes of others as a result its decreases the probabilities of misdetection and probabilities of false alarm. Finally, provided an overview of CR for opportunistic spectrum access and related research topics.

Keywords—Cognitive radio, spectrum sensing, multi-dimensional spectrum sensing, cooperative sensing.

I. INTRODUCTION

Wireless communication system design requires higher data rate and larger channel capacity as well as better quality of service and spectrum utilization efficiency to meet the needs of wireless users. Radio spectrum requires assiduous planning and management for getting its maximizing value for all users. Radio spectrum is worldwide allocated to the radio services on the primary or secondary basis. This is reflected in the Radio Regulations published by the International Telecommunication Union (ITU) [1]. To solve the spectrum congestion problem CR is introduce the opportunistic usage of the frequency bands that are not heavily occupied by licensed and PUs. This paper presents the overview of CR system and corresponding research field explaining its basic functions using cognitive cycle concept. Firstly, we introduced the preliminary concept of CR in Section II. Various spectrums sensing methods for CR are explained in Section III, The cooperative spectrum sensing and its various forms are discussed in Section IV. Section V explained the multi-dimensional spectrum sensing concept. Challenges and Compressions of various spectrum sensing methods are explained in Section VI and conclusions are presented in Section VII.

II. COGNITIVE RADIO

CR technology has developed to the access on unused frequency bands. Firstly Joseph Mitola and Gerald Maguire was introduced the concept of CR [2]. CR is an intelligent wireless communication system that is aware of its environment and uses the methodology of understanding-by-building to learn from the environment and adapt to statistical variations in the input stimuli, with two primary objectives in mind: (1) highly reliable communication whenever and wherever needed; (2) efficient utilization of the radio spectrum. The main objectives of the CR is the efficiently use of natural resources including space, frequency, time without any interference to PUs. To achieve these objectives, it is required to sense and understand its local radio spectrum environment to identify spectrum holes in radio spectrum space, for taking decisions to accesses spectrum and the transmissions accordingly. A spectrum hole [3] (or it's also called white space) is a band of frequencies assigned to a PU, but at a particular time and specific geographic location, the band is not being utilized by that user. Fig.1 showed that the PU geographical region that are dynamically access spectrum holes or white space by secondary unlicensed users to improve spectrum utilization.

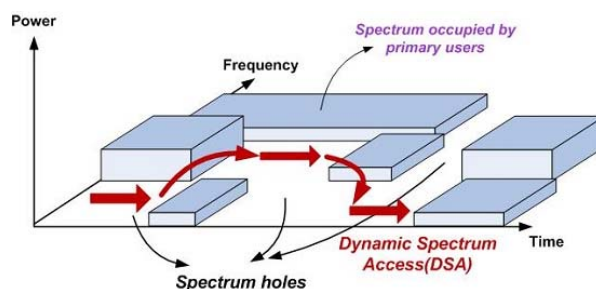


Fig.1 Illustration of spectrum holes [4]

In CR, PUs has higher priority or legacy rights on the usage of the spectrum. Contrariwise secondary users (SUs) have lower priority exploit this spectrum without any cause interference to PUs. Fig.2 showed the cognitive cycle of CR operation. The main functions of CR are spectrum sensing, spectrum decision, and spectrum sharing and spectrum mobility [5] are discussed below.

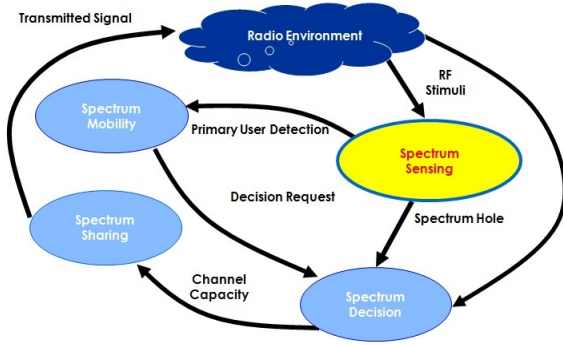


Fig.2 Cognitive cycle of CR [5]

Spectrum Sensing: Spectrum sensing is to obtain awareness about the spectrum usage and existence of PUs in a geographical area, detect usage statistics of the PUs and SUs and determine possible spectrum space holes. This awareness can be obtained by using geo location and database, by using beacons, or by local spectrum sensing at CR.

Spectrum Decision: After getting information from spectrum sensing, the aim objectives of CR are to send and receive information as far as possible with best quality of service. To improve its operation CR may use data from regulatory and policy database.

Spectrum Sharing: In CR there is ample numbers of SUs are participating to usage the available spectrum holes. To achieve balance between cognitive and non-cognitive users for transferring information CR share the available resources.

Spectrum Mobility: To avoid interference to PUs, CR has stop its current operation or empty its currently used radio spectrum and change the radio frequency when PU starts to operate. CR has ability to constantly investigate possible alternative spectrum holes.

III. SPECTRUM SENSING METHODS FOR COGNITIVE RADIO

At present the spectrum sensing methods are still in intermediate stage of development [6]. Some methods are also implemented and tested in laboratory environment, but they are still in preliminary stage. The motivation of this study is to point out the challenges and complexities still arise during the use of recently proposed methods. In this section the most common spectrum sensing methods used in CR system explained below.

1. Transmitter Detection Scheme

In transmitter detection based scheme, CRs catch out signal energy that is generated from the PUs. The sensing results are gathered using two performance matrices in terms of Probability of detection, P_d -which quantifies the probability of SUs detecting that incumbent is present. Probability of false alarm, P_f - which quantifies the probability of SUs declaring that an incumbent is present in the spectrum when the spectrum is in fact free [7].

a. Energy Detector

An energy detector method is known as radiometry or periodogram [8]. The main working principle of energy detector based sensing is to find the received signal energy and compares with the threshold. These signals are detected by comparing the output of the energy detector with a threshold which depends on the noise floor [9]. Energy detector based sensing also facing some challenges due to low signal-to-noise ratio (SNR) values. For detecting spread spectrum signals energy detectors do not work powerfully [10]. Fig.3 showed the block diagram of energy based spectrum sensing process.

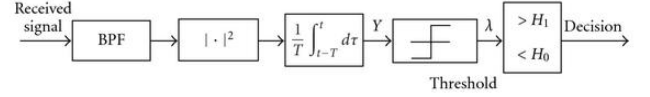


Fig.3 Energy based spectrum sensing [6]

The received signal of energy detector is:

$$y_i(n) = s_i(n) + w_i(n) \quad (1)$$

Where, $y_i(n)$ is the n -th sample of received signal at i -th SU, $s_i(n)$ with mean zero and variance σ_s^2 represents the signal transmitted by the PU, $w_i(n)$ is Additive White Gaussian Noise (AWGN) with mean zero and variance σ_w^2 at i -th SU and $n = 1, 2, 3, 4 \dots \dots N$ is the number of samples of received signal.

The decision metric for the energy detector can be written as

$$M_i = \frac{1}{N} \sum_{n=1}^N |y_i(n)|^2 \quad (2)$$

Where, N = Size of the observation vector. To choice of occupancy band can be obtain by the matching the decision metric M opposite to fixed threshold λ_E . Presence or absence of PUs can be distinguished from the two hypotheses as follows:

$$y_i(n) = w_i(n) \quad : H_0 \quad (3)$$

$$y_i(n) = s_i(n) + w_i(n) \quad : H_1 \quad (4)$$

Where, H_0 means frequency is ideal, there is no PU signal and H_1 means frequency is occupied, there is PU signal. There are two performance metrics is probability of detection (P_D) and probability of false alarm (P_F). P_D is probability of detecting a signal when it truly is present in spectrum. It can be expressed as

$$P_D = Pr(M > \lambda_E | H_1) \quad (5)$$

Similarly P_F is the probability of incorrectly test decides that signal is not occupied actually but considered frequency is occupied. It also can be written as

$$P_F = Pr(M > \lambda_E | H_0) \quad (6)$$

To get more benefits from CR we should be kept P_F is lower and P_D is higher. The decision threshold λ_E can be selected for finding an optimum balance between P_D and P_F . For energy detector, the P_D and P_F can be calculated [11] as given equations:

$$P_D = Pr(M > \lambda_E | H_1) = 1 - \Gamma\left(L_f L_t, \frac{\lambda_E}{\sigma_w^2}\right) \quad (7)$$

$$P_F = Pr(M > \lambda_E | H_0) = 1 - \Gamma\left(L_f L_t, \frac{\lambda_E}{\sigma_w^2 + \sigma_s^2}\right) \quad (8)$$

Where, $\Gamma(a, x)$ is incomplete gamma function and λ_E is denoted decision threshold. The region of convergence (ROC) for different SNR values showed below [10]. From Fig.4 it's clear that the performance of the threshold detector increases at highest SNR values.

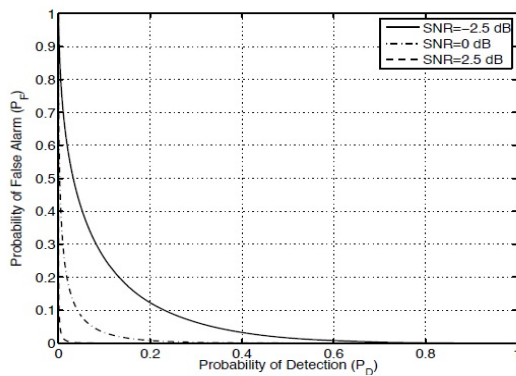


Fig.4 ROC curves for energy detector based spectrum sensing under different SNR values [10].

As a part of practical implementation so far, Wasmote sensor based CR detection method is analyzed in the paper [12]. This is basically a scheme for detecting 2.4 GHz WLAN channels and Bluetooth with the received signal strength in indoor lab environments. Another software defined radio (SDR) chip based energy detection method is described in [13]. Four different frequencies have considered to sense both in lab and outdoor environment using SDR. But the mobility is still an issue to resolve for these kinds of implementation.

b. Cyclostationary Detection

Cyclostationarity detection method is used for detecting PU transmissions by exploiting the built-in periodicity of the received signals. For detecting signal presence in the spectrum this method used cyclic correlation function instead of power spectral density (PSD). This algorithm differentiated PU signals from noise. Cyclostationarity based method more used for distinguishing among different types of transmissions and PUs.

c. Matched-Filter Detection

Matched-filter detection is the foremost method for detect the existence of PU if the transmitted signal is known [10]. Better detection performance and less time to achieve processing

gain is the main advantage of matched filtering method. Its need more consummate knowledge about PUs signaling features such as bandwidth, operating frequency, modulation type and order, pulse shaping, and frame format.

IV. COOPERATIVE SPECTRUM SENSING

In Cooperative spectrum sensing (CSS) CR shares its sensing data with others and utilizes the sensing outcomes of others to give a decision. CSS has robust to sensing errors due to hidden node or fading channels also its decreases the probabilities of misdetection and probabilities of false alarm. Additionally, cooperation can solve hidden PU problem and it can decrease sensing time. The main challenges of CSS include developing efficient information sharing algorithms and increased complexity. Cooperation can be implemented in three ways:

A. Centralized Sensing

In centralized sensing a central manager (BS or AP) collects all CR sensing data and makes a decision on channel state, i.e. idle or busy. The goal is to decrease the fading effects of the channel and increase detection performance [10].

B. Distributed Sensing

In distributed sensing, cognitive nodes share information among each other but each CR makes decisions itself. It's more advantageous than centralized sensing because there is no need any backbone infrastructure that's why it is reduced cost.

C. Cluster Based Sensing

In clustering method all CRs are grouped into cluster, where one CR act as cluster head (CH) and remaining CRs will act as cluster members. In CR all CRs sends their own information such as current location, current energy level and SNR of reporting channel to the Fusion center (FC). Firstly FC sorts all CRs in ascending order according to the SNR value of the reporting channel and then computes average energy level and makes CH for current round, which energy level is higher than average level.

V. MULTI-DIMENSIONAL SPECTRUM

The definition of spectrum opportunity is defined as "a band of frequencies that are not being used by the PU of that band at a particular time in a particular geographic area" [14]. CR sensing only related to these three dimensions of the spectrum: frequency, time, and geographical space. Fig.5 showed the several dimensions.

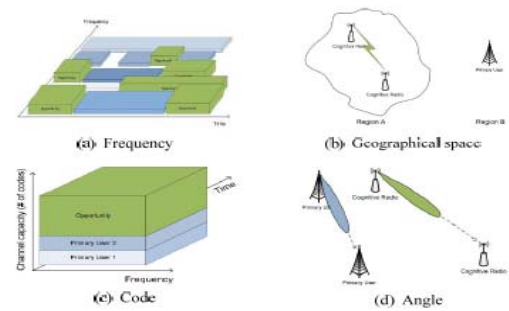


Fig.5 Multiple dimensional spectrum opportunity [10]

VI. CHALLENGES AND COMPARISON OF VARIOUS SENSING METHODS

There are various challenges are associated with the spectrum sensing for CR are given below:

Hardware Requirements: Spectrum sensing for CR applications requires high sampling rate (HSR) with large dynamic range, and high speed signal processors. Noise variance estimation techniques have been popularly used for optimal receiver designs like channel estimation as well as for improved handoff, power control, and channel allocation techniques.

Hidden PUs Problem: Several factors may cause this problem including multipath fading or shadowing observed by SUs while sensing for PUs transmissions.

Detecting Spread Spectrum PUs: PUs that use spread spectrum signaling, it's very difficult to detect as the power of the PUs is distributed over a wide frequency range even though the actual information bandwidth is much narrower.

Security Issues: Security is always challenges part for all technologies. CR has some selfish or malicious users they can modify its air interface to mimic PUs. For this reasons the spectrum sensing result is not well enough. Between, the centralized and distributed architectures and cooperation and non cooperation approaches, the centralized architecture and cooperation approach are more vulnerable to attacks.

Basic comparisons or benefit-drawback of various sensing methods are:

In energy detector based sensing there is no need any information of PU's signal that's why its design is complex free. It also used in wide band spectrum sensing. But the main drawback is its selection of threshold. In case of low SNR rate, this method is inefficient to discriminating interference from primary signal. Cyclostationary detection gives better performance than energy detection but its more computational complexity and takes longer observation time. The main benefit of matched filter detection is its better detection performance and less time to achieve processing gain. The disadvantages of matched filter detection method is it requires proper knowledge of the PUs signaling features such as bandwidth, operating frequency, modulation type and order, pulse shaping, and frame format. It is large power consumption method.

VII. CONCLUSIONS

In wireless communication system CR is most challenging and promising concept for efficient usage of radio spectrum. This paper discussed spectrum opportunity and spectrum sensing concepts considering various dimensions including simulated

and implemented techniques. We also discussed advantages and disadvantages of several sensing methods. High speed communication with low cost and dynamic spectrum access is the future goal of CR. In my future work I'll analysis energy detection based sensing method and will propose an adaptive threshold which is a function of the detection threshold and SNR of the primary user received at the sensing node.

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A comparative study of WLAN security protocols: WPA, WPA2

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Abstract—in this paper, we have made a comparison between different WLAN security protocols, WPA and WPA2, and we correlated the two with respect to performance. We have emphasized on IEEE 802.11n amendment. We performed an experiment to measure the maximum throughput achieved by different encryption protocols. We concluded that WPA2 has less reduction on network throughput than WPA due to its encryption algorithm, CCMP, which is highly improved compared to TKIP, which is adopted by WPA.

Keywords—WLAN Security Protocols, WPA, WPA2, WiFi, IEEE 802.11, IEEE 802.11i, TKIP, CCMP.

I. INTRODUCTION

The 802.11i, an amendment ratified on June 24, 2004, specifies security mechanisms for WLANs. IEEE 802.11 originally specifies a wired LAN equivalent data confidentiality algorithm Wired equivalent privacy (WEP). However, numerous researchers have shown that none of the data confidentiality, integrity, and authentication can be achieved by the use of WEP. WEP had many design flaws and is considered completely broken. As a result, the use of WEP for confidentiality, authentication, or access control is deprecated on later revision of the standard in 2012 [1].

Although WEP fails to satisfy the security requirements of the standard, a new standard will require a new hardware. It is not practical to easily discard the users with legacy devices supporting only WEP. Hence, in 2003 the Wi-Fi Alliance announced that WEP had been superseded by Wi-Fi Protected Access (WPA) which uses the legacy hardware. WPA was just an intermediate solution to cover the weaknesses of WEP.

WPA adopts Temporal Key Integrity Protocol (TKIP) for confidentiality and Integrity, which still uses Rivest Cipher 4 (RC4) for data encryption, but includes a key mixing function and an extended IV space to construct unrelated and fresh per-packet keys. WPA also introduces Michael algorithm [2], a weak keyed Message Integrity Code (MIC), for improved data integrity under the limitation of the computation power available in the devices. Furthermore, in order to detect replayed packets, WPA implements a packet sequencing

mechanism by binding a monotonically increasing sequence number to each packet.

Although TKIP addresses all known vulnerabilities, yet it had some limitations due to the use of legacy hardware. TKIP relies on Message Integrity Check (MIC) algorithm called Michael, which provided inadequate security [3, 4]. As a result, it has been replaced by the more secure encryption algorithm CCMP (Counter-Mode / Cipher Block Chaining Message Authentications Code Protocol). However, TKIP is still supported by many people [5] and should be discouraged.

Under the assumption of upgrading the hardware, IEEE 802.11i is proposed as a long-term solution to provide enhanced security in the MAC layer. It uses CCMP (Counter-Mode / Cipher Block Chaining Message Authentications Code Protocol) to provide confidentiality, integrity, and replay protection. On the other hand, it combines 802.1X authentication and key management procedures to provide mutual authentication and generate a fresh session key for data transmissions. Thus, IEEE 802.11i should provide a sufficient data confidentiality, integrity and authentication.

Several works discussed the effect of Temporal Key Integrity Protocol (TKIP), and Counter Mode/CBC-MAC Protocol (CCMP) security algorithms on network throughput for IEEE 802.11 networks. Yet, most of the works were based on older versions of the standard.

In this paper, we provide a theoretical comparison on different security protocols, WPA and WPA2, and evaluate the effect of these protocols on network performance using IEEE 802.11n amendment. To the best of our knowledge, there are very few works that have been carried under 802.11n.

II. SECURITY METHODS

Basic security mechanics, in the wireless world, entails the general capabilities of confidentiality, integrity, availability, authentication, authorization, and access control. Security mechanisms provide the means through technologies, protocols, and implementations to achieve the basic security mechanics. Some important key mechanisms to deploy in a wireless network include encryption protocols, digital

signatures, and key management. Security, for all practical purposes, is the combination of processes, procedures, and systems used to achieve the basic security mechanics.

The 802.11i specification defines two classes of security algorithms: Robust Security Network Association (RSNA), and Pre-RSNA. Pre-RSNA security consists of Wired Equivalent Privacy (WEP) and 802.11 entity authentication. RSNA provides two data confidentiality protocols, called the Temporal Key Integrity Protocol (TKIP) and the Counter-mode/CBC-MAC Protocol (CCMP), and the RSNA establishment procedure, including 802.1X authentication and key management protocols. Figure 1 summarizes the authentication, key generation, confidentiality and integrity in 802.11i.

Pre-RSNA algorithms originated from WEP which first appeared in the 802.11b amendment. WEP was introduced for the purpose of providing confidentiality, integrity and authentication. The RC4 stream cipher from RSA Security Inc. is used for confidentiality (encryption). A 32-bit cyclic redundancy check (CRC) is used for data integrity.

There are two pre-RSNA authentication methods; namely, open system and shared key authentication. With the exception of the open system authentication mechanism, all pre-RSNA algorithms have been deprecated.

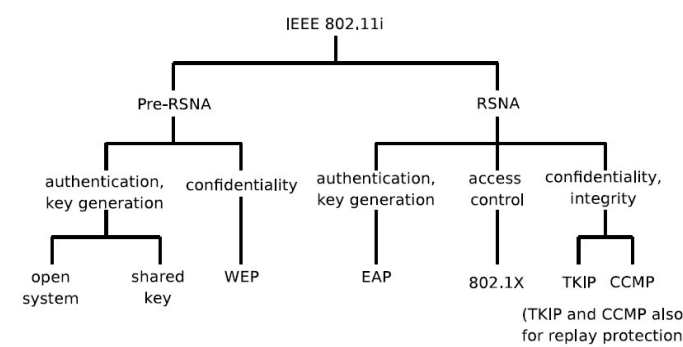


Fig. 1. Summary of 802.11i security [6]

III. WPA AND WPA2

In April 2003, the Wi-Fi Alliance released a recommendation in response to corporate concerns on wireless security they provided an intermediate solution which enhances the previous security standard WPA. However, WPA has shown significant vulnerabilities and the Wi-Fi Alliance introduced Wi-Fi Protected Access (WPA2) based on IEEE 802.11i standard ratified in June 2004.

Similarities of WPA and WPA2 include, they both use the same authentication system. Enterprise networks use 802.1X/EAP frameworks as part of the infrastructure that ensures centralized mutual authentication system. For home and small office environments Pre-Shared Key (PSK) is used.

On the other hand, one difference between WPA and WPA2 is that WPA2 uses a mixed mode that supports both WPA and WPA2 enabled devices on the same wireless network. Another difference is that; WPA uses TKIP

encryption as a Security Protocol which in turn uses RC4 cipher, while as WPA2 uses CCMP-AES as a Security Protocol. WPA uses Michael algorithm for data integrity but WPA2 uses more robust efficient and stronger algorithm, CBC-MAC. A comparison of WPA and WPA2 is shown in Table 1.

TABLE I. SUMMARY OF 802.11 SECURITY METHODS [7]

	WEP	802.11i Methods	
		WPA	WPA2
Security Protocols	RC4	TKIP	CCMP
Cipher	RC4	RC4	AES
Key Length	40 or 104 bits	128 bits encryption 64 bits authentication	128 bits
Key Life	24 bit IV	48 bit IV	
Key Generation	Concatenation	Two phase mixing function	Not needed
Data Integrity	CRC-32	Michael	CBC-MAC
Header Integrity	None	Michael	CBC-MAC
Replay Protection	None	Packet Number	
Key Management	None	EAS-based	
Authentication	Open or Shared key	802.11x or Pre-Shared Key (PSK)	

A. Confidentiality and Integrity

Since both WPA and WPA2 use the same authentication system, we will consider only Confidentiality and Integrity in our comparison of the two protocols.

“Confidentiality is the non-occurrence of the unauthorized disclosure of information, whereas Integrity is the non-occurrence of the unauthorized manipulation of information” [8]. To ensure security, confidentiality and integrity must be provided by the security standard. Three standards provide data confidentiality and integrity; WEP, TKIP and CCMP. WEP and TKIP are deprecated and no longer in use, leaving CCMP as the ruling standard in 802.11i confidentiality since 2004. CCMP provides stronger encryption than TKIP and it incorporates replay protection. In this section we will focus on TKIP and CCMP as WEP is obsolete and no longer in use. The standard also defines one integrity protocol for management frames: BIP (Broadcast Integrity Protocol).

Implementation of TKIP is optional for an RSNA and used only for the protection of data frames. A design aim for TKIP was that the algorithm should be implementable within the capabilities of most devices supporting only WEP, so that many such devices would be field-upgradable by the supplier to support TKIP. BIP is a mechanism that is used only when management frame protection is negotiated. BIP provides integrity protection for group addressed robust management frames. BIP is only used to protect management frames within the Basic Service Set (BSS). [1]

IV. EFFECT ON THROUGHPUT

A. Main goals

This part analyzes the impact of different security protocols on WLAN performance. The analyses of these protocols were performed in practical experiments on IEEE 802.11n wireless

networks. In the experiment we enabled and disabled WPA and WPA2 on different cases and scenarios and then analyzed the throughput variation of different TCP traffic over the network. We analyzed the initial results when no security protocol was used and then compared the results obtained for different security protocols.

Transmission Control Protocol (TCP) is a core protocol of the Internet Protocol Suite, because it originated in the initial network implementation in which it complemented the Internet Protocol (IP). Consequently, the entire suite is commonly referred to as TCP/IP which is the protocol that major Internet applications such as file transfer and email rely on. Applications that do not require reliable data stream service may use the User Datagram Protocol (UDP), which provides a connectionless datagram service that emphasizes reduced latency over reliability [9]. We have used TCP to measure the throughput of the network.

The main reason of measuring throughput is measuring the maximum data throughput in bits per second of a communications link. A typical method of performing a measurement is to transfer a 'large' file from one system to another system and measure the time required to complete the transfer or copy of the file. The throughput is then calculated by dividing the file size by the time to get the throughput. The throughput of communications links is measured in bits per second (bit/s), kilobits per second (kbit/s), megabits per second (Mbit/s) and gigabits per second (Gbit/s).

The Maximum bandwidth is calculated as follows:

$$Throughput \leq RWIN/RTT$$

Where RWIN is the TCP Receive Window and RTT is the round-trip time for the path. The Max TCP Window size in Windows Operating System is 65,535bytes. Example: Max Bandwidth = 65535 bytes / 0.220 s = 327675 B/s * 8 = 2.621 Mbit/s. Over a single TCP connection between those endpoints, the tested bandwidth will be restricted to 2.621 Mbit/s.

The main goal of this paper is to uncover the compared effect of different security configurations (WPA and WPA2) on the performance of the WLAN. This will be achieved by measuring the throughput of the network while traffic is pushed through the network. The traffic generated will be a 'synthetic' data which is generated by the LanTraffic/IPTraffic. Synthetic data will be used to evaluate the existing security configurations in order to make comparisons ranging from No-Security to WPA and the complete WPA2 specifications with AES-CCMP.

B. Network design

1) Network setup

A Prolink-PRN2001 access point is used in the experiment and the clients successfully authenticated with the AP. The access point was operating in the 802.11n 2.4GHz 54Mbps mode. There is little difference between 2.4GHz and 5GHz in terms of propagation in an indoor office environment [11]. For this reason, we have used 2.4GHz in our setup. Client isolation was disabled and re-authentication level in the router was defined in terms of seconds passed.

Two clients were used to send generated packets from Client1 to Client2.

Client1 was an Intel® Core™ i5-2410M CPU @2.30GHz with 4GB RAM, running Windows Technical Preview 64-bit (6.4, Build 9841) with wireless adapter Qualcomm Atheros AR9285 802.11 b|g|n Wi-Fi Adapter.

Client2 was an Intel® Core™ i3-3120M CPU @2.50GHz with 4GB RAM, running Windows Technical Preview 64-bit (6.4, Build 9841) with wireless adapter Realtek RTL8188CE Wireless LAN 802.11n PCI-E NIC.

None of the computers were running any programs which could significantly affect the processor or network utilization.

TABLE II. SUMMARY OF THE SYSTEMS USED IN THE EXPERIMENT

System	Processor	RAM	Interface
Client1	Intel® Core™ i5-2410M CPU @2.30GHz	4GB	Qualcomm Atheros AR9285
Client2	Intel® Core™ i3-3120M CPU @2.50GHz	4GB	Realtek RTL8188CE Wireless LAN 802.11n PCI-E NIC

All of the hardware used in the experiment were placed in the same room to ensure that the walls and other obstacles between them would not be a factor affecting the measurement. Hardware locations were the same and roaming was not allowed. Antennas were placed no closer than 1.5m (5feet) apart since wireless devices restrict signals that are too strong [10]. Figure 2 shows the network setup.

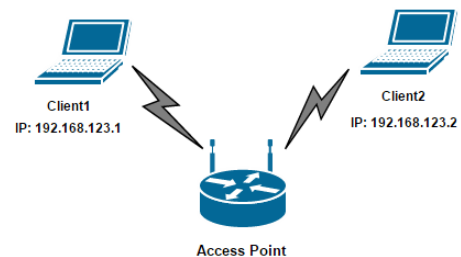


Fig. 2. Network Setup

2) Security Levels

The following three security layers were used to test the performance of the wireless network:

a) *No Security*: This is often default configuration of most wireless devices. The main reason for using this level is to insure the minimum effect on throughput. It is used as a reference for other security standards.

b) *WPA with PSK authentication and RC4 encryption:* This uses WPA configuration, Pre-Shared Key is used for authentication.

c) *WPA2 with PSK authentication and AES encryption:* This uses WPA2 configuration, Pre-Shared Key method of authentication is used.

3) *Network Measurement Methodology*

As mentioned earlier, the main goal of this paper is to compare to which extent the security mechanisms reduce the throughput in WLAN. In order to perform this analysis we need to generate traffic and send it through the Wireless channel and measure it from both the Sender (Tx) side and the Receiver (Rx) side and compare these results. A traffic generator is used to generate traffic on Client2 and Client3 to send to Client1. Several traffic generators were evaluated by [11] including Iperf, NetPerf and IPTraffic, using Ethereal for the comparison of these generators and to analyze whether each tool generated traffic according to the parameters given. They have found that IPTraffic was the only tool to generate reliably consistent traffic based on the parameters given. LanTraffic was the other software we used to measure the total send/received throughput of the network. IPTraffic and LanTraffic were both used to increase the accuracy of the measurement taken.

4) *Traffic Generator Configuration*

The traffic generators used in this study were IPTraffic and LanTraffic. It is a network testing tool that can create Transmission Control Protocol (TCP) and User Datagram Protocol (UDP) data streams and measure the throughput of a network that is carrying them. It is a network performance measurement and packet control tool.

IPTraffic and LanTraffic were configured as follows:

Total Number of Packets: The number of packets used has no effect in the throughput of the network. The experiment was conducted with Client1 receiving a minimum of 20,000 packets. We have observed the throughput for an average of 5min and 30seconds. And during this time the range was minimal, showing that the number of packets has no effect on the throughput.

Outgoing Bandwidth: We did not use any bandwidth limiting software. There was no any other process running which used the network. We configured the tool to push more traffic through the interface.

TCP Windows: It is Windows configuration dependent. TCP uses a receive window that is four times the size of the maximum TCP segment size negotiated during connection setup, it uses up to a maximum size of 64 KB (65535 bytes).

Traffic Protocol: TCP is the main protocol we used in the IP Traffic Generator. Since UDP is used for time sensitive applications, such as VOIP. UDP is slightly used in the Traffic generator, only in certain scenarios.

Traffic Type: The traffic type was not important for the test. So we depended on automatically generated traffic by the LanTraffic and/or IPTraffic.

V. RESULTS AND DISCUSSION

A. *Effect on Throughput:*

In this experiment, Client2 was configured to send the generated traffic data to Client1. We measured the respective Throughputs of TCP packets based on different security configurations (No Security, WPA and WPA2 respectively). We used LanTraffic in the measurement as a Traffic generator and receiver on Client2 and Client1 respectively.

All the statistics in this section, tables and graphs, were taken from the Receiver's side (Client1).

We used LanTraffic to measure the received throughput and put the values in graph. First off, the average throughput was measured disabling the security protocols as a reference; this helps us identify any reduction in throughput when security protocols are used in WLANs. A total of 16 connections were allowed to send TCP packets from Client2 to Client1. The Total Received Throughput was approximately 35Mb/s, Figure 3. But, when WPA is allowed there was a significant reduction in the throughput of the network and the Total Received Throughput after summing the 16 connections from Client2 to Client1 became 13Mb/s as in Figure 4. Finally, WPA2 has shown an improved throughput over WPA. Figure 4 shows the total received throughput for WPA2 configuration became 30Mb/s with only 5Mb/s reductions from the No Security configuration. Where the security levels are statistically different, the WPA throughputs are lower than the respective WPA2 throughputs. This suggests that the chipset in the WLAN adapters has better support for AES/CCMP than RC4/TKIP. This is most noticeable in the one to one 1460 byte TCP transfer. The transfers with two senders also support this; in the UDP and 6 byte TCP transfer, the WPA2 throughputs are slightly higher than the WPA throughputs.

From the comparison of these values, it is clear that the throughput of WLAN decreases with the use of complex security mechanisms. However, Figure 3 shows that WPA2 have improved performance compared to WPA which uses TKIP. Although several factors affect the performance, for example the chipset of the access point may have enhanced performance on AES over TKIP. For now, we can conclude that WPA2 has enhanced throughput and thus better performance than WPA [11].

Andrew [11] showed that wireless network throughput is largely unaffected when the various security levels are applied. But from our analysis it is clear that throughput is affected by the use of security protocols though the effect might be negligible, considering one-to-one transmission. This is similar to findings in [12] and [13], where throughput fell as more complex security protocol was used.

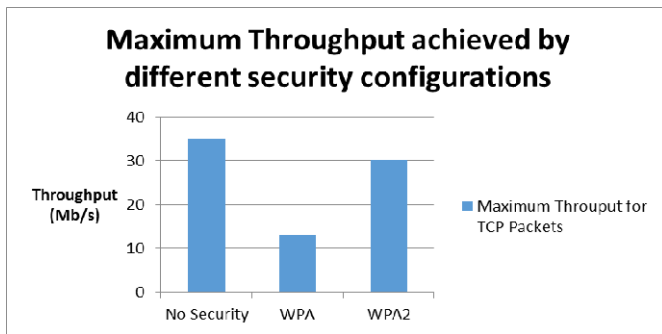


Fig. 3. Comparison of Maximum Throughput achieved by the three security levels.

Even though CCMP is largely improved compared to TKIP, yet there is a significant reduction in throughput. A possible explanation of this might be the cost of combining the encryption and integrity protocols in CCMP. "First, the encryption of the various message blocks can no longer be carried out in parallel since CBC-MAC requires the output of the previous block to calculate the MAC for the current block. This slows down the protocol. Second, CBC-MAC requires the message to be broken into an exact number of blocks. This means that if the message cannot be broken into an exact number of blocks, we need to add padding bytes to it to do so. The padding technique has raised some security concerns among some cryptographers but no concrete deficiencies attacks have been found against this protocol." [8]

Previous works have been carried on IEEE 802.11a/b/g wireless amendments. But, in this paper we used IEEE 802.11n amendment. We were unable to find the necessary hardware for testing the effect on throughput for the new IEEE 802.11ac since all the devices we use today are capable only on IEEE 802.11n and require hardware upgrade for the new amendment.

We did not consider the case for multiple clients since almost all other researches, [11] and [12], showed the same effect for multiple clients.

VI. CONCLUSION

There has been some works done in the area of analyzing the security protocols of wireless networks, but much of it remains focused on the old protocol, WEP, rather than on new encryption techniques available today (WPA and WPA2). The studies that incorporate these advanced encryption protocols did not correlate the two with respect to performance. In addition to drawing conclusions on observed data, this research aims to tie together previous work and examine similarities and differences among it.

Most of the work done in this area focused on the most basic security mechanisms employed in 802.11a/b/g. In our comparison we used the 802.11n amendment which is hardly mentioned in any research. 802.11n provides best security mechanism in comparison together with 802.11g [14] and it utilizes the CCMP encryption algorithm much better than 802.11g. In our study we have emphasized on IEEE 802.11n amendment and we proved that it provides similar outcome and effect on the wireless network.

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Interference Free Cooperative Spectrum Sharing in Cognitive Radio Networks using Spatial Modulation

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Abstract—Cognitive radio (CR) and spatial modulation (SM) are two exciting and emerging techniques to increase the spectrum efficiency for 5G. In this paper, we present a novel cooperative spectrum sharing protocol using SM. A two phases uncoded decode-and-forward (UDF) relaying scheme is considered. A secondary transmitter (ST) consisting of M_t transmit antennas act as UDF relay for the primary system (PS). In the first phase, a block of information bits of the primary transmitter (PT) is mapped into two information carrying units: a symbol which is chosen from a constellation diagram and a particular antenna number from a set of transmit antennas. The ST then uses iterative- maximum ratio combining (*i*-MRC) to demap the transmitted block of information bits. In the second phase, the ST forwards the estimated primary data by activating only one antenna based on the own secondary data. The primary receiver is then uses *i*-MRC to estimate the forwarded primary symbol and the secondary receiver uses *i*-MRC to estimate own secondary data by detecting only transmit antenna indices of the ST. As a result, mutual interference between the PS and secondary system (SS) is avoided. We investigate the bit error rate (BER) of the PS and SS. Simulation results show the efficacy of the proposed system.

Index Terms—cognitive radio, cooperative relaying, spatial modulation, interference free, spectrum sharing.

I. INTRODUCTION

Cognitive radio (CR) was first coined by J. Mitola III [1]; it has been widely employed by the wireless community to solve the spectrum inefficiency problem. Cooperative diversity has recently been incorporated into cognitive radio networks (CRNs) to combat fading and to improve performance. Moreover, deployment of multiple antennas at wireless nodes may increase capacity without causing bandwidth expansion and significantly enhance the transmission reliability [2].

In [3]–[8], the authors proposed the CR spectrum sharing methods that treat secondary users (SUs) as cooperative relays. Cooperative spectrum sharing protocols where secondary nodes equipped with single antenna can either cooperatively relay the primary traffic or act as secondary access transmitters are presented in [3], [4]. Multi-antenna based protocols are presented in [5], [6]. In [5], the authors proposed a cooperative overlay scheme in which a secondary transmitter (ST) base station equipped with multiple antennas cooperatively relays the primary traffic. In [6], the authors treat both the ST and the secondary receiver (SR) as relays to cooperate the primary transmission. In [5] and [7], the authors claim that their proposed schemes are interference free i.e., there is no mutual interference between the primary users (PUs) and SUs.

However, PU leases half of its time slots to the SU in exchange of cooperation in [5] whereas PR and SR use interference cancellation (IC) techniques to mitigate mutual interference in [7]. In [8], the authors propose a two-phase cognitive transmission (TCT) protocol for secondary spectrum access in CRNs. In TCT protocol, SUs employ IC technique to mitigate the interference from PUs as well as limit their transmit power to satisfy a given QoS requirements. However, due to the mutual interference between PUs and SUs, the overlay approach may still experience a secondary transmission performance degradation. In [9], the authors proposed a spectrum leasing strategy for cooperative CRNs. The PU leases a fraction of the licensed spectrum to the SU in exchange of cooperation. However, in all the existing works [3]–[9], the PU leases its fraction of time slots or spectrum or both the PR and SR use IC techniques to mitigate the mutual interference.

Spatial modulation (SM) [10], [11], a wireless transmission technique which is a very flexible mechanism to achieve higher spectral efficiency (SE) using multi-antennas compared to the conventional modulation schemes with same modulation order. In SM, only one transmit antenna is active at each time instant and other antennas are set to zero power. Therefore, the transmit antenna synchronization and inter-channel interference (ICI) are completely avoided. However, in SM each information bits carry both signal constellation and antenna information. A cooperative SM scheme is proposed in [12] but the authors did not consider the CRNs environment.

In this work, we have proposed an interference free cooperative spectrum sharing protocol in CR network using SM where PU does not need to lease fraction of its time slots or spectrum to the SU. Moreover, the ST does not need to transmit its own signal to the SR. It only forwards the primary information. The PR can recover own primary signal by estimating the forwarded symbol by the ST whereas the SR will recover its secondary signal by detecting only transmit antenna indices of the ST. As a result, there is no mutual interference between the PU and SU. In this work, our contributions can be summarized as follows:

- We propose a novel interference free cooperative spectrum sharing protocol, where PU can use whole portion of its time slots and spectrum without leasing to the SU.
- The ST does not need to transmit its own signal to its receiver. As a result mutual interference between the PU and the SU is avoided; no IC and management techniques are also needed at the PR and SR.

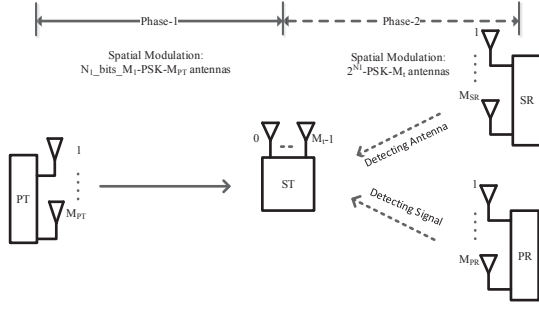


Fig. 1: System model under consideration.

- PU can achieve higher SE using SM compared to the conventional modulation scheme for the same modulation order. On the contrary, the secondary SE can be increased by using higher number of transmit antennas at the ST.

II. SYSTEM AND CHANNEL MODELS

We consider a cooperative CR spectrum sharing environment consisting of a primary system (PS) with one primary transmitter-receiver (PT-PR) pair and a secondary system (SS) with a secondary transmitter-receiver (ST-SR) pair. The proposed system model is shown in Fig. 1. Assume that direct link between the PT and PR is not strong enough due to fading/shadowing and is thus not considered for data communication. The ST is considered as a base station and equipped with $M_t = 2^{n_2}$ antennas; where $n_2 (= \log_2(M_t))$ is the number of bits in a secondary data symbol. The PR and SR are equipped with M_{PR} and M_{SR} antennas respectively whereas PT is equipped with $M_{PT} = 2^{n_1}$ antennas; where $n_1 (= \log_2(M_{PT}))$ is the number of antenna information bits which is added to the mapped block of information of the PT. No dedicated relays are required in the proposed protocol. Assume that the channel over all the links are subject to Rayleigh flat fading with additive white Gaussian noise (AWGN) and all the signals follow independent fading path to avoid correlation. The channel coefficients are independent and identically distributed Gaussian random variables with zero mean and unit variance. Additionally, assume that the channel coefficients remain static during both transmission phases and the receivers have full knowledge of the channel state information.

During cooperative transmission, primary transmission is performed over two phases via the help of the ST. At phase-1, the total number of bits that can be transmitted by the PT using SM is given as follows

$$N_1 = n_1 + m_1 \quad (1)$$

where $n_1 = \log_2(M_{PT})$, $m_1 = \log_2(M_1)$ and M_1 denotes the modulation order during phase-1; $M_1 = 4, M_1 = 8, M_1 = 16$ and $M_1 = 32$ for Q-PSK, 8-PSK, 16-PSK and 32-PSK respectively. For example, the PT can send 3 bits per symbol from two transmit antennas using Q-PSK modulation whereas in conventional modulation scheme only two bits could be sent using Q-PSK. Similarly, in order to transmit 4 bits, Q-PSK

Algorithm 1 Cooperative relaying and TA selection

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1: procedure INTERENCEFREE( $M_1, M_{PT}, P_{msg}$ )
2:    $M_t = 2^{n_2} \leftarrow$  Number of transmit antennas at ST
3:    $T \in \{t_1, t_2\} \leftarrow$  Two phases; half duplex transmission
4:    $P_{msg} \leftarrow$  estimated  $N_1$  bits of the PT's data by ST at  $t_1$ 
5:    $P_{buf} \leftarrow N_1$  bits buffer at ST to put PT data at  $t_1$ 
6:    $S_{msg} \leftarrow$  ST's own data divided into block of length  $n_2$ 
7:   if  $T = t_1$  then
8:      $P_{buf} \leftarrow P_{msg}$ 
9:   end if
10:  if  $T = t_2$  then
11:    for  $i=0$  to  $M_t-1$  do
12:       $T_A = \text{de2bi}(i, [], 2, \text{'left-msb'})$ ; where  $\text{de2bi}()$  is a
        MATLAB function to convert decimal to binary.
13:      if  $S_{msg} = T_A$  then
14:        Transmit  $P_{msg}$  by  $i_{th}$  antenna
15:      end if
16:    end for
17:  end if
18: end procedure

```

modulation and 4 antennas can be used by the PT and so on. The ST is then uses i -MRC to de-map the original information bits of the PT.

At phase-2, a block of information bits is mapped into two information carrying units: a symbol which is the estimated primary symbol of phase-1 and a particular antenna number of the ST from a set of antennas. It is noted that the ST has to use 2^{N_1} -PSK modulation scheme in this phase to forward the primary symbol. The ST forwards the primary symbol by activating particular antenna of the ST based on its own secondary data. The cooperative relaying scheme and the transmit antenna (TA) selection technique by the ST are given in Algorithm 1. Moreover, the Algorithm 1 is clearly explained by an example given in Table I. The number of bits that can transmitted during phase-2 by the ST is given as follows:

$$N_2 = n_2 + N_1 \quad (2)$$

where $n_2 = \log_2(M_t)$. The PR is then uses i -MRC to detect the symbol and SR uses i -MRC to detect the antenna. The SR can recover own secondary data from the antenna information. However, the SE of the PU and the SU is given respectively as follows

$$PU_{SE} = \frac{1}{2} \log_2(N_1) \quad (3)$$

$$SU_{SE} = \frac{1}{2} \log_2(M_t) \quad (4)$$

where the scaling factor $\frac{1}{2}$ is due to the fact that the overall transmission is divided into two phases for the PU and SR has to wait one time slot to recover own desired signal.

Let, $\mathbf{p}(k)$ be the N_1 bits binary vector of the PT to be transmitted. This vector is then mapped to another vector $\mathbf{x}_1(k)$ of size M_{PT} where only one element in the resulting vector

TABLE I: Example 1: cooperative Transmission: In phase 1, PT transmits 3 bits (left most bit represent antenna information) using QPSK and 2 transmit antennas. In phase-2, ST forwards the primary symbol using 8-PSK and two transmit antennas. For the better explanation of the example, we assume the correct estimation of the information bits by the receiver.

PT data to be transmitted (Mapped)	Expected data at ST at t_1	ST's own data	TA No. of ST	Transmitted data by ST	Expected data at PR at t_2	Expected data at SR at t_2
000	000	0	0	0000	000	0
001	001	0	0	0001	001	0
010	010	0	0	0010	010	0
011	011	0	0	0011	011	0
100	100	0	0	0100	100	0
101	101	0	0	0101	101	0
110	110	0	0	0110	110	0
111	111	0	0	0111	111	0
		1	1	1000	000	1
		1	1	1001	001	1
		1	1	1010	010	1
		1	1	1011	011	1
		1	1	1100	100	1
		1	1	1101	101	1
		1	1	1110	110	1
		1	1	1111	111	1

is different from zero [10]. The primary symbol transmitted from the antenna number l ($l \in [1 : M_{PT}]$) is x_l , T denotes transpose, H denotes Hermitian transpose and h_1 is the channel coefficient between PT and ST. Notation: bold letters denote vectors. The received vector $\mathbf{y}_1(\mathbf{k})$ at the ST is then given by

$$\mathbf{y}_1(k) = \mathbf{h}_1(k)x_l + \eta_1(k), \quad \text{For } k = 1 : M_l \quad (5)$$

where $\eta_1(k)$ is the AWGN vector. Therefore, the transmit antenna number and the transmitted symbol at the ST during phase-1 can be estimated by using i -MRC [10] as follows

$$g_j = \mathbf{h}_1^H \mathbf{y}_1, \quad \text{For } j = 1 : M_{PT} \quad (6)$$

$$\mathbf{G} = [g_1 \ g_2 \ \dots g_{M_{PT}}]^T \quad (7)$$

$$\tilde{l} = \underset{j}{\operatorname{argmax}} |G| \quad (8)$$

$$\tilde{x}_l = Q(G_{(j=\tilde{l})}) \quad (9)$$

where $Q(\cdot)$ is the constellation quantization function. Therefore, after correct estimates of $\tilde{l}$ and $\tilde{x}_l$ the ST can recover the desired information bits of the PT. The estimated information bits of the PT is then forwarded by the ST. The similar techniques described above are used to forward the primary data by the ST during phase-2. The PR and the SR is then recover their own desired information bits by using i -MRC during phase-2, similar to phase-1. If the PR and SR are unable to estimate the symbol and antenna number respectively, then it is considered as an error for the primary and secondary signal respectively. Therefore, the average overall BER for the PS and SS can be respectively expressed as follows

$$BER_{PU} = \frac{TNT_{bits} - TCE_{bits}}{TNT_{bits}} \quad (10)$$

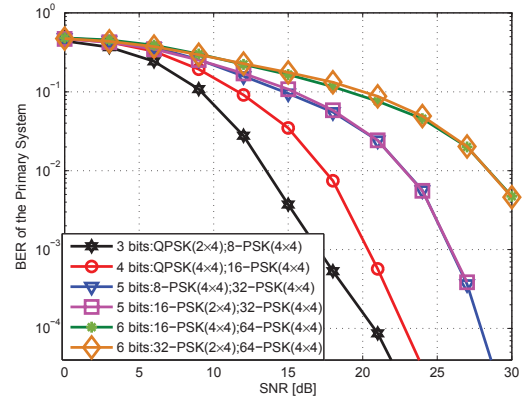


Fig. 2: BER of the primary system.

$$BER_{SU} = \frac{1}{n_2} \sum_{i=1}^{n_2} P_e \quad (11)$$

where TNT_{bits} and TCE_{bits} are the total number of transmitted bits by the PT and the total number of correctly estimated bits by the PR respectively. In addition, P_e is the probability that the antenna number estimated by the SR is incorrect.

III. RESULTS AND DISCUSSION

This section presents the Monte Carlo simulation results with the help of MATLAB to study the performance of our proposed cooperative spectrum sharing protocol. We investigate the BER of the PS and SS to show the efficacy of the proposed protocol. In Fig. 2, we depict the BER of the PS with cooperative relaying over Rayleigh fading channels as a function of signal to noise ratio (SNR) where SNR varies

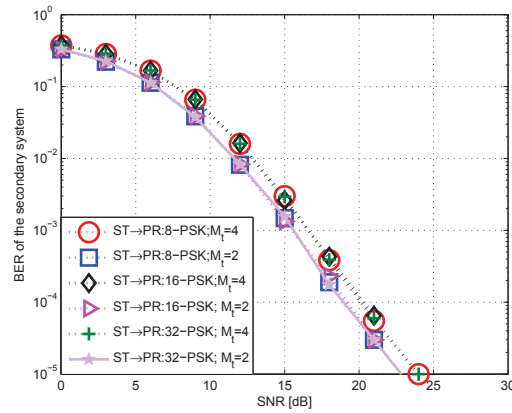


Fig. 3: BER of the secondary system.

from 0 to 30 dB in each case. The PT uses M_1 -PSK where as the ST uses 2^{N_1} -PSK to forward the primary information. Two cases of M_{PT} where $M_{PT}=2$ and $M_{PT}=4$ are considered. We let, $M_t=M_{PR}=4$ for all cases. We plot the BER of the PS for 3, 4, 5 and 6 bits transmission. In Fig. 2, n_1 bits: M_1 -PSK ($M_{PT} \times M_t$); 2^{n_1} -PSK ($M_t \times M_{PR}$) represents that the PT transmits n_1 bits by using M_1 -PSK modulation and M_{PT} transmit antennas at phase-1 and the ST forwards estimated primary symbol using 2^{n_1} -PSK modulation and M_t transmit antennas during phase-2. It can be observed that BER decreases with increasing SNR as expected. It is pointed out that the pairs 8-PSK (4×4); 32-PSK (4×4), 16-PSK (2×4); 32-PSK (4×4) and 16-PSK (4×4); 64-PSK (4×4), 32-PSK (2×4); 64-PSK (4×4) show the similar BER performance. This can be explained by noting that those pairs transmit the same number of bits. Lastly, it is observed that the BER of the PS increases with the increasing number of transmit bits.

Figure 3 shows the simulation results of BER of the SS as a function of SNR in our proposed protocol with different values of M_t . Here, SNR varies from 0 to 30 dB. We let, $M_{SR}=4$ for all cases. We consider two cases of M_t where $M_t=2$ and $M_t=4$ respectively. It is clear that BER decreases with increasing SNR. As we explained in section 2, only one antenna of the ST is active at each time instant. Furthermore, it is pointed out that BER of the SS increases as the number of antennas at the ST increases considering same number of antennas at the SR. This can be explained by noting that the SE of the SS increases with the increasing number of antennas at the ST. As a result BER increases. It is also found that whatever modulation scheme the ST uses by using the same number of transmit antennas to forward the primary symbol, the BER of the SU becomes almost similar. This is due to the fact that secondary BER depends on the number of the antennas at the ST, not the modulation order.

IV. CONCLUSION

We have proposed a novel interference free cooperative spectrum sharing protocol using SM in CRNs. It is pointed out that the PU does not need to lease its fraction of time slots or

spectrum to the SU. Therefore, the PU can use whole portion of its time slots and spectrum for its own data transmission. On the other hand, the ST does not need to transmit its own signal to its receiver. The SR receiver can recover the own secondary signal by only detecting the transmit antenna indices of the ST. As a result mutual interference between the PU and SU is avoided and IC techniques are no longer needed at the PR and SR. We have investigated the BER of both the PS and SS. It is found that SE of the PU depends on both modulation order and transmit antenna numbers whereas the SE of the SU only depends on the number of transmit antennas at the ST. However, closed-form solution of the BER will be investigated in the future work.

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Improved Quality of Service in ZigBee Network with Statistical Modeling

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Abstract—Wireless Sensor Network (WSN) comprises the less power consuming, light weight and effective Sensor Nodes (SNs) for higher network performance. ZigBee is the real time application for the WSN used for low-cost data and energy consumption characteristics. ZigBee is a robust network certain application of which has come out with a critically low designation for less data rate, long delivery delay and packet loss as well in order to lower the cost. However, ZigBee devices have many potential features to make a splendid wireless network. Hence, putting an improvement analogy inside the performance parameter is a way to overcome the limitations of ZigBee usage in certain scales. ZigBee follows a model for its node distribution, traffic routing and Quality of Service (QoS). The QoS indicates the efficiency and performance of any network. The increase in efficiency of QoS on the ZigBee network lies in the realization of maximizing throughput of the network. This paper is destined to represent the improved QoS in the ZigBee network applying Stochastic and Widest Path models. Riverbed Simulator v17.5 is used to investigate foremost QoS parameters in the ZigBee network. The optimized result of this study will be handy in the application field of the ZigBee network. And it also best fits for the deployment of the large-scale device to device communication.

Keywords—ZigBee Network, QoS, Stochastic Model, Widest Path model, Riverbed.

I. INTRODUCTION

ZigBee is a wireless, open, mesh networking standard with the application of home, office automation, pervasive computing. In a distributed networking environment, ZigBee devices work together for a higher networking compatibility as the communication medium. ZigBee is a secure standard in wireless sensor networks for its low-cost, reliability, wirelessly monitoring capability and less energy consumption applied mostly in large-scale implementation. Under the ZigBee alliance in 2004, ZigBee is standardized as IEEE 802.15.4 protocol [1], [2].

Moreover, for industrial usage a complex networking environment is required and performance is a key criterion in such fields. In a ZigBee networked industry, office, smart-home or other organization, a large number of devices are interconnected through ZigBee sensors and establishes a large-scale network [3], [4]. This huge number of devices are autonomously distributed in the network field for successfully network automation.

Since, the prospects are big, the number of devices is high, device interconnection and communication significantly influences on overall network performance as well as the

QoS for the applied ZigBee network. QoS includes higher throughput, simplified node distribution with less delay in Medium Access Control (MAC) Layer and in the Physical layer of the network. Modification of these properties improves the network performance. A conventional ZigBee network comprises three types of devices - Coordinators, Routers and End-devices [5]. A proper distribution of these devices through the network channel enhances the data transmission rate, network lifetime, latency and other QoS parameters.

Previously, numerous research studies are conducted on the different ZigBee network, design, routing path impedance for the improvement of wireless network and QoS analysis perspectives [6], [7], [8]. According to Sun et al. QoS measured adding traffic differentiation in the research [7]. Another study [8] illustrates QoS management for ZigBee network based on the subdivision of radio ranges in the research. Results from [9] holds Fuzzy based routing path and positioning techniques integrated with ZigBee to enhance the QoS. In the paper [10], query-based wireless CSMA network through stochastic geometry analyzed for characterizing the wireless network performance. According to the study [11], statistical models like Poison Point Process (PPP) adopted to improve network performance with its node distribution technique.

Yet, there has been a lack of QoS enhancement in the ZigBee network with the Statistical probability models. Hence, a more efficient network is needed that rises the QoS. The paper focuses on improving QoS performance for application traffic distribution based ZigBee network implementation. The different network performance metrics (e.g. End-to-End delay, MAC delay, Data dropped threshold, Queuing delay and Queue size, Throughput, Number of Hops) are analyzed to improve the QoS.

Further extension of this paper is organized as follows: Section II describes the methodology. Section III demonstrates an environment for ZigBee network simulation setup. Analysis and results of the improved ZigBee network is provided in Section IV. In the end, the paper is concluded in Section V with future perspectives.

II. RESEARCH METHOD

A literature review has been brought about ZigBee network and improvement its QoS performance. The study ends in exploring a ZigBee network with the enhanced operation. The literature review predicts that statistical models are more effective for network construction to enhance its performance.

To increase ZigBee network performance and QoS, two widely used statistical models - Stochastic and Widest Path model are applied to the more improvement of the proposed network.

An effective simulator was required for such network implementation. The further study references various simulators such as Network Simulator-2 (NS-2), Network Simulator-3 (NS-3), OMNET++ and Riverbed [12]. Among the various simulators, Riverbed has user friendly development features of ZigBee network. In this research work, Riverbed simulator is chosen to design, implement and complete the simulation of the ZigBee network for its sophisticated deployment features and the wide range of acceptance.

At the end of the simulation, data in different ZigBee layers are collected carefully. Appropriate graphs are plotted according to the simulated data. These graphs represent the enhanced QoS on ZigBee network based on Stochastic and Widest Path model.

III. SIMULATION

A. ZigBee Network Implementation

To maintain a large ZigBee network and acquire the QoS, a novel statistical approach has led. Several statistical models are useful for planning and designing networks those differ in size and aspects. In this study, among other statistical models, Stochastic and Widest Path modeling has carried out for ZigBee network formation. These models are chosen for ZigBee device's routing path selection process and for traffic distribution. Initially, the formation of ZigBee network was its own way, but distributed end device's request selection, execution, and routing is implemented according to Stochastic and Widest Path model later. A brief procedure of the techniques with these two models is described following:

1) *Stochastic Model*: Stochastic modeling is one of the most reliable processes of probability and statistics. A Stochastic process is combined of random variables X_n , indexed by parameter n , where n belongs to some index set Θ . In other words, a Stochastic process is simply a model running in time according to probabilistic rules. Θ is a set of integers which represents specific time points in discrete event simulation. In ZigBee network, the routing requests or application traffic are some random variables. The Coordinator processes these requests in discrete time intervals in the form of Θ , Stochastic model's discrete event simulation. The simulation of these devices Θ contains the routing requests or application traffic as random variables, X_n of specific time points n according to Stochastic model.

A Stochastic process importantly treats as a probabilistic model that evolve in time. Following the axioms of probability, a rich set of results may come to forward. These results are applicable to a ZigBee network, which develops real-world ZigBee network scenarios. The model mainly focuses on the ZigBee network for a well-developed taxonomy of routing selection and queuing model.

2) *Widest Path Model*: In the networking context, the Widest Path problem is recognized as "Bottleneck Problem". This model finds out the best-fitted routing path for weighted nodes. Weights represent the size of edges. If a weighted graph has vertex v and width to be the minimum of any

edge of the path, definitively $\text{widthto}(v)$ to be the widest path from s to v (i.e. $\text{widthto}(s) = \infty$). To present it in shortest path routing: $\text{widthto}(x) = \max [\min (\text{widthto}(v), \text{width}(e))]$ Where, $e = (v, x): v \in \text{tree}$. When this Widest Path model is applied in the ZigBee network, the application traffic in the network travels through this set of shortest or the minimum path and efficient routing selection occurs in the network. This lowers the transmission cost, increase the data transmission rate. Hence, QoS on the prospective ZigBee network improves.

To build an environment based on statistical models for ZigBee network necessary steps as follows: Executing "Extend attribute" command a new attribute "Model" is created for all ZigBee devices. The common attribute "Model" is configured with Riverbed modeler's available model files. The properties of "Model" attribute has been set with Stochastic and Widest Path model. The "Model" attribute has the privilege to set it in device configuration mode either Stochastic or Widest Path. This indicates when the attribute is set in any of these two models for all devices, the network is run with that model. According to this procedure, two scenarios are created for this ZigBee network. All devices are configured in Scenario_1 is on the Stochastic model. On the other hand, Scenario_2 is on the Widest Path model.

The both scenarios have the same parameters and values, including ACK status, data rate, packet interval time, traffic destination, packet size, etc. The parameters of the proposed ZigBee network are listed in Table I. Through this way, a novel statistical model based large-scale ZigBee network is simulated. After 600 sec simulation run, the proposed network routing layout is depicted as in Figure 1.

TABLE I. SIMULATION PARAMETERS IN CLOSE LOOK

Parameters	Description
ACK status	Enabled
Number of re-transmission	5
Data Rate	Auto Calculate
Packet Size	Constant (1024)
Packet interval time	Constant (1,0)
Implementation Model	Stochastic and Widest Path Model
Traffic Destination	Coordinator
No. of Nodes (overall)	32
No. of Coordinator	1
No. of Routers	5
No. of End Devices	26
Network Dimension	300 m * 300 m
Simulation time	600 sec

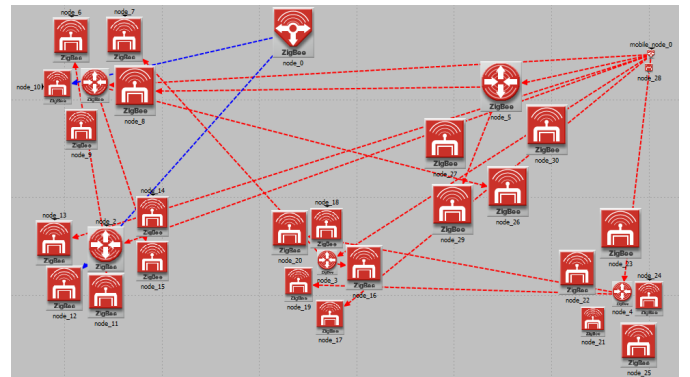


Fig. 1. ZigBee network routing simulation.

IV. ANALYSIS AND RESULTS

This section presents the detailed descriptions about the analysis of the obtained results from discrete event simulation. The graphs are plotted on the basis of various network performance measurement parameters in different layers.

A. End to End Delay

In the Application layer, end-to-end delay refers to the time taken for a packet transmission across the ZigBee network from source to destination or one end device to others. In this sense, the delay components:-

$$D_{end-end} = N[d_{trans} + d_{prop} + d_{proc}]$$

The total end-to-end delay depends on the number of links (N) and router's delay. Each router has its own transmission (d_{trans}), propagation (d_{prop}) and processing (d_{proc}) delay. Hence, this formula gives a rough estimate of end-to-end delay. Figure 2 shows, the end-to-end delay has increased about 28% for the Widest Path model than the Stochastic model in the ZigBee application layer.

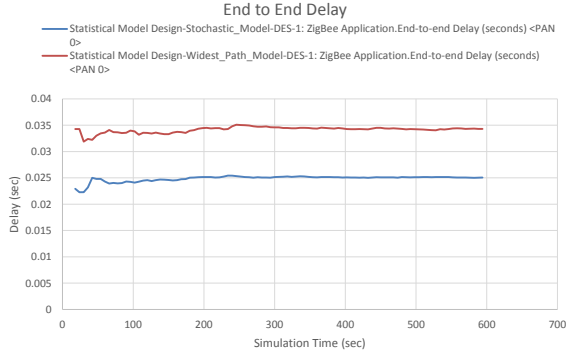


Fig. 2. End-to-End delay (sec) for global network.

B. Global MAC Delay

MAC delay represents the total end-to-end delays (sec) of all the packets received in the 802.15.4 MAC's of all wireless personal area networks (WPAN) nodes in the network and forwarded to the higher layer. Figure 3 illustrates, MAC delay for the global network is almost 41% higher in the Widest Path model than the Stochastic model. Besides, the MAC delay for the Coordinator (node_0) is almost same in both of the models.

C. Global MAC Data Dropped Threshold

ZigBee's 802.15.4 MAC layer drops higher layer's data traffic due to consistent transmission or retransmission failure. The data dropped statistic reports the number of the higher layer packets that are dropped because the MAC could not receive any ACKs for the transmission or retransmissions of those packets or their fragments, and the packets retry counts reached MAC's retry limit or failure to transmit a packet due to channel access failure (reaching maximum backoff limit). This data dropped statistic also reports the outstanding packets in the buffers that are dropped during roaming. In the Widest Path model, the data dropped retry threshold is approximately 2-times higher for the global network than the Stochastic model but, for the Coordinator (node_0) it remains same as illustrated in Figure 4.

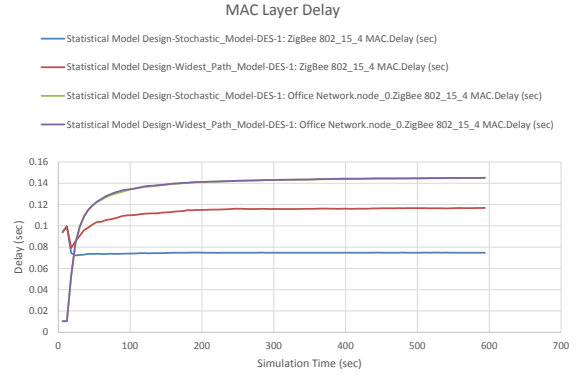


Fig. 3. MAC delay (sec) for global network and Coordinator (node_0).

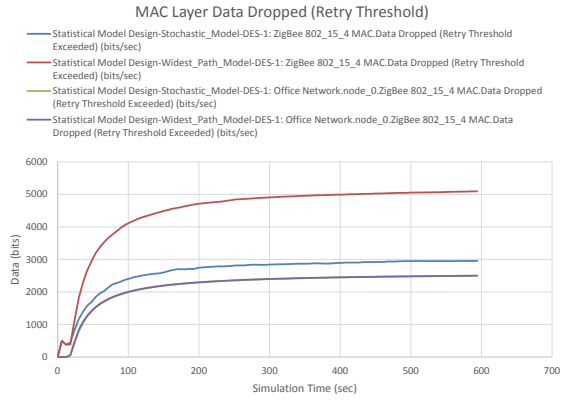


Fig. 4. MAC data dropped retry threshold (bits/sec) for global network and Coordinator (node_0).

D. Queuing Delay and Queue Size

All the data requests are sent for processing in the Coordinator (node_0). The Coordinator (node_0) queue holds these requests for further execution. Queue size represents the total number of packets outstanding in the MAC queue to be executed. The queuing delay represents the delay that packets from the network layer incur at the MAC queue until it can be executed. It is one of the vital components to increase overall network delay. In the Stochastic model, queue size for the Coordinator is nearly 3% smaller and also close to 10% lower queuing delay than the Widest Path model according to Figure 5.

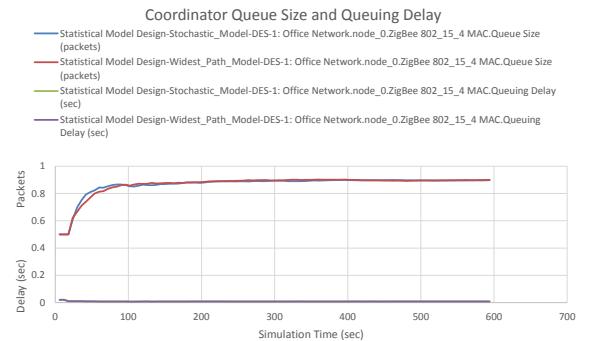


Fig. 5. Average queue size (packets) and queuing delay (sec) for Coordinator (node_0).

E. Throughputs

Throughput for the Coordinator (node_0) and global network is always higher in the widest path model than the Stochastic model. Figure 6 describes, about 80% throughput increases in the Widest Path model for the global network where for the Coordinator it is almost same in both Stochastic and Widest Path models.

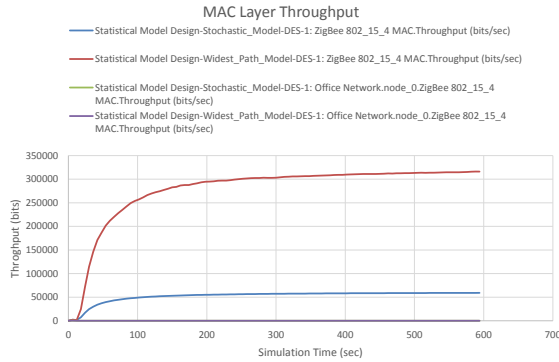


Fig. 6. Average throughput (bits/sec) for global network and Coordinator (node_0).

F. Number of Hops

In ZigBee PAN, hop number represents the average number of hops, application traffic travels in the PAN. It is the number of hops a packet travels from the source throughout the intermediate nodes to reach the destination. Figure 7 shows, the average traveled hops of application traffic in global network slightly higher in the Stochastic model than the Widest Path model. The Stochastic and Widest Path models are used in

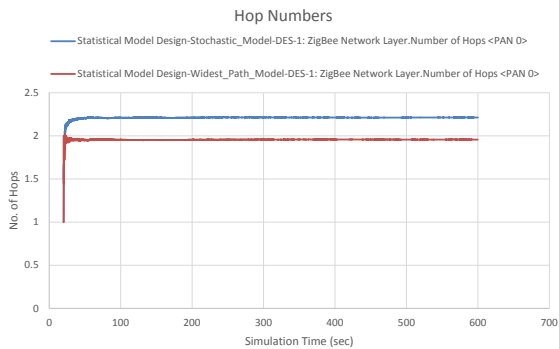


Fig. 7. Average hop numbers for global network and Coordinator (node_0).

a wide range of networking field. Using these two models, ZigBee network shows significant QoS characteristics. The Widest Path model always gives more end-to-end delay, MAC delay, data drop throughput than the stochastic model. In the contrary, the Stochastic model gives less delay, data dropped and throughput as well as smaller queue size and lower queuing delay for the Coordinator (node_0) than the Widest Path model. Moreover, for the Coordinator (node_0) MAC delay has roughly equal values in both Stochastic and Widest Path models.

V. CONCLUSIONS

This paper represents an improved QoS in ZigBee network using Stochastic and Widest Path model of probability. The simulation study shows, there is about 28% less end-to-end delay in the application layer, 41% less delay, two-times lower data drop threshold in the global MAC layer, 3% smaller queue size in terms of the Stochastic modeling rather than the Widest Path modeling. However, in terms of the Widest Path model, around 80% increase in overall throughput; wherein for the network layer, both scenarios have almost the same number of hops for traveling application traffics. These resulted in improved QoS parameter for ZigBee network.

This improved network has better performance with less end-to-end delay (latency), increased throughput, less data drop tendency and smaller queue size than the conventional one's. Improved QoS performance of this study can be applied to Machine-to-Machine communication and the Internet of Things deployment.

In the future, it would be interesting to investigate statistical model based device to device communication. It will also be examined and applied for the big data analytic issue. This can be an effective reference to implementing a ZigBee network with further maximized throughput.

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Dynamic Spectrum Allocation for Cognitive Radio Ad hoc Network

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Abstract—This paper proposes a novel cross-layer dynamic spectrum allocation for cluster-based cognitive radio ad hoc network. A primary focus is on the cluster formation and maintenance. The proposed clustering mechanism divides the network into clusters based on three values: spectrum availability, power level of node and current speed of the node. Therefore, clusters form with the highest stability and flexibility to avoid frequent re-clustering in the network. Moreover, the proposed method integrates the spectrum sensing at physical (PHY) layer with the packet scheduling at MAC layer to be more robust to Primary Users (PUs) activity as well as node mobility in a network. The simulation results show that the proposed protocol outperforms the conventional protocols in terms of throughput, power consumption and packet transmission delay.

Keywords— *cognitive radio network; cognitive radio ad hoc network; dynamic channel assignment; mobility management*

I. INTRODUCTION

Due to the expansion of wireless communications, it is almost impossible to cope up with allocation of spectrum as it is limited and finite resource. Therefore, it is vital to ensure the best usage of this limited resource to meet the recent demand of radio frequency. The traditional spectrum allocation methods that are authorized by different regulators are not intelligently handled. It is observed that the licensed spectrum remains unoccupied most of the time [1] where the regulator body is unable to allocate the frequency for upcoming new wireless devices. To utilize these unoccupied band, resolve the spectrum scarcity for new wireless application and make spectrum allocation more dynamic, cognitive radio (CR) technology is introduced [2, 3]. This intelligent technology has introduced opportunistically access of secondary or unlicensed user, known as SU, to use licensed band without interfering the primary or licensed user known as PU.

Cognitive radio network (CRN) can be classified as the infrastructure-based CR network and the Cognitive radio ad hoc networks (CRAHNs) [4]. The infrastructure-based CRN has a central network entity to manage the network such as base-station (BS) in cellular network. Instead, CRAHN does not have any infrastructure backbone and each CR user can communicate with other CR users through ad hoc connection. In contrast with infrastructure-based CRN, a distributed cognitive radio network can be a more appropriate selection because of its lower system complexity, faster align positioning, and lower cost of implementation.

II. RELATED WORKS

In recent years designing an efficient MAC protocol for CRN has taken a lot of attention and many researches have been proposed to attain this goal [5,6,7]. CogMesh proposed by Chen et al. [8], is a most well-known cluster-based MAC protocol CRN which opportunistically utilizes different spectrum holes for smooth peer-to-peer communication. CogMesh forms clusters with nodes sharing local common channels, where multi-channel multi-access network is considered. A node forms a cluster using the common channel of the cluster, which is called the master channel, and invites the neighboring nodes which have the same channel to join the cluster. For intra-cluster communication, CogMesh MAC protocol uses guaranteed access period and random access period for control message exchange. Although CogMesh shows promising performance in statistic CRN, due to lack of mobility handling mechanism it face performance degradation in dynamic network. Due to its popularity, we will consider it as the conventional candidate of our proposed scheme for performance comparisons.

III. SYSTEM MODEL

In this paper cognitive radio sensor network is considered. We assume each node is aware of its location and each node has single half-duplex cognitive radio transceiver which is capable of detecting and utilizing spectrum holes in a distributed and efficient way. A given number of non-overlapping orthogonal channels are available and each channel has a unique channel ID.

Channel availability for each node is related to the node physical location. Similar to IEEE 802.22 standard, secondary users (SUs) use available channel only when it is not occupied by primary users (PUs). By detecting PU presence, SU vacate the channel. For avoiding interference between PUs and SUs a simple interference avoidance model is used in proposed network.

This paper consider the clustering architecture for CRAHN. We do not consider the global control channel for the network. However, after clustering the network, each cluster has its own control channel. Nodes forming the cluster become Cluster Head (CH), which is responsible for inter-cluster communication as well as intra-cluster channel access control. Inter-cluster communication is relayed by Gateway nodes (GW). Gateways are the nodes which are in the boarder of two neighboring cluster and can hear both cluster beacons. Two SUs node consider 1-hop neighbor if they are within transmission range of each other.

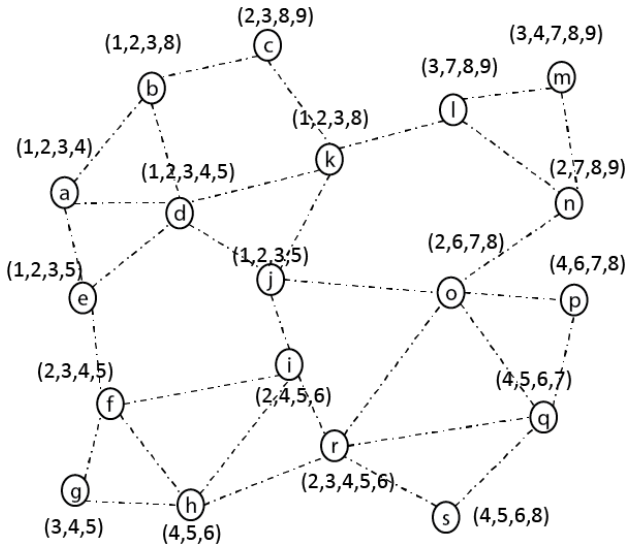


Fig. 1: Connectivity graph of a Cognitive Radio Network with available channel list in brackets.

IV. CLUSTER FORMATION AND MAINTENANCE

Once the neighbor discovery phase is completed, cluster formation phase starts. The proposed clustering mechanism divides the network into clusters based on three values: spectrum availability, power level of node and current speed of the node. In the proposed clustering scheme, clusters are formed with the neighboring nodes in an ad-hoc topology. The procedure is explained in the rest of this section.

Nodes in the proposed architecture exchange their available channels list (ACL) based on the sensed spectrum availability and generates its neighbour list N_i , where $i = 1, 2, 3, \dots, n$. Afterwards the cluster formation phase starts. Since proposed method tries to make less amount of cluster, cluster formation is defined as a maximum vertices biclique problem to include more nodes inside each cluster while make sure of existing enough number of idle channels for intra-cluster communication. At first, based on neighbour list, N_i and accessible channels set, every CR_i creates an undirected bipartite graph $G_i (A_i, B_i, E_i)$. Graph $G (V, E)$ is called bipartite if vertices set V can be splitted into two disjoint sets A and B where $A \cup B = V$, such that all edges in E connect vertices from A to B . Here, $A_i = CR_i \cup N_i$, and $B_i = C_i$ where, C_i is the ACL of CR_i . An edge (x, y) exists between vertices $x \in A_i$ and $y \in B_i$ if $y \in C_i$, i.e., channel y is in the available channels of CR_i . Fig. 1 shows the connectivity graph of a CRN with available channel list in brackets. The set of vertices A_d corresponds to the 1-hop neighbors is $A_d = a, b, e, k, j$ while the set of vertices B_d corresponds to the accessible channels set of CR_d , is $B_d = 1, 2, 3, 4, 5, 8$. Here, vertices 1, 2, 3 are common to all vertices in A_d . With neighboring nodes (a, b, e, k, j) and channels $(1, 2, 3)$, node d forms its maximum vertices biclique. Thus, every node in the network constructs its own maximum vertices biclique graph.

To carry out this objective, we have introduced a parameter named Cluster Head Election Value (CHEV), where CHEV of a node is the product of neighboring nodes and idle common channel found from the biclique. Every CR calculates CHEV based on (1) as follow:

$$CHEV = \frac{E_{current}}{V_{current}} \times CH_{sh} \log(N_{ngh}) \quad (1)$$

where, $E_{current}$ is the current energy level of node, $V_{current}$ is the current speed of the node, N_{ngh} is the number of 1-hop neighboring nodes of CR_i and CH_{sh} is the number of shared channel with all the 1-hop neighbors found from the biclique for CR_i .

Node with higher CHEV value comparing with its neighbors forms the cluster and becomes the cluster head (CH). For example both node a and d has 3 shared channel while node d has higher number of neighbors compare to node a . Therefore, considering both nodes are static and having same amount of power node d would have higher number of CHEV and becomes CH. If the CHEV value of a node CR_i is lesser than its neighbor, CR_i joins the neighbor, which has the highest CHEV value, as cluster member (CM). Once the clusters are formed, CHs prioritize other cluster members based on CHEV for the reserved cluster head (RCH) selection. CM with the highest CHEV becomes the RCH for the cluster. The RCH takes charge of the cluster if current CH moves out, which shrinks the possibility of re-clustering

After the cluster formation, CH determines and maintains list of operating frequencies for the cluster. Member nodes then check neighbor list to find the existence of any other cluster heads in the neighborhood. If located, member node informs the CH about having a neighbor from different cluster. CH will receive this information from all the nodes and then assign the node with higher CHEV as a gateway node (GW) with each particular cluster to maintain the inter-cluster communication. As clusters in the architecture are formed in a distributed way and independently, it is possible of having loops in the architecture. Loops increases the robustness of the architecture in case of link failure.

A. Neighbor discovery

Following the cluster formation procedure, the network is formed and divided into several clusters. Neighboring clusters are connecting through gateway nodes.

We define a bit for node status known as in-cluster bit, to differentiate the nodes those already are in the clusters and those are joining the clusters. If in-cluster bit is TRUE then node tunes to cluster control channel to receive the cluster beacon. If a node receives the beacon from a CH, then the node engages in cluster action. If a node does not receive any beacon in the cluster CCC, it starts channel hopping among the cluster's common channel by a sequence which is already defined by CH (CnHS). If a node receives beacon in any of those channel, it engages in the cluster action otherwise it changes the in-cluster bit to false and goes to neighbor discovery phase.

In the neighbor discovery phase, a node starts channel hopping to hear beacon in any of its free channel. If the node receives any beacon from any CH, it compares its own set of free channel with the information it receives from the CH through the beacon packet. If the node has at least γ common channels with the cluster then it requests to join the cluster. If there is a free mini-slot for new joining node then CH accepts the request and define a mini-slot for new node, otherwise CH reject the request.

If the node does not receive any beacon in that particular channel or does not have γ common channel with the cluster or its request get rejected by CH, it tune to the next free channel to detect appropriate cluster in its neighborhood. The value of γ can be changed based on the application or QoS requirements. Finally if node cannot join any cluster it forms its own cluster and broadcast its beacon.

Thus, proposed architecture reorganizes the ad-hoc architecture and divides the network into clusters where clusters consist of Cluster Heads (CHs), Secondary Cluster Heads (SCHs) and Cluster Members (CMs). Within a cluster, single-hop communication exists among the CMs and CH. Two clusters are connected by the Cluster Gateways (GWs). In other words, GWs are the communication hub for clusters. Communication within a cluster is done through the cluster head using local common control channels.

V. SIMULATION RESULTS

To evaluate the proposed protocol, comparison study has been done using simulation in MATLAB with a cluster-based approach known as CogMesh. Three major metrics for MAC protocol (energy efficiency, throughput and delay) have been considered in the simulation.

To simulate the protocols 100 CRAHNs are randomly deployed in a 100m×100m area. The communication range of each node is set to 10 meters and the available primary radio channels is equal to 10. The simulation duration is 1000 seconds (100 superframes) and the result is averaged of 50 runs.

Fig. 2 to 4 show the impact of nodes speed as well as network dynamicity on the performance of protocols. Network dynamicity has been defined as the changes in number of mobile nodes in the network. As it can be seen in Fig. 2 by increasing the speed of mobile nodes from 2 to 10 m/s throughput of CogMesh and proposed method drop 13% and 7%, respectively. Proposed method yet shows better performance since it uses adoptive data period to handle upcoming nodes in the cluster. Fig. 2 shows by increasing the number of mobile nodes in the

network throughput is fallen sharper for CogMesh while proposed method perform better. The reason is proposed method uses adoptive scheduling to handle the mobile nodes in the network.

Fig. 3 illustrates the effect of node speed on the energy consumption of nodes. As it has shown CogMesh shows higher energy consumption (17% in static network and around 30% in high speed) compare to proposed method. By increasing the node speed it is more likely a cluster head move out of cluster and reclustering have to be performed. Moreover, due to this movement the previous common channel might not be available for all nodes so nodes need to do channel sensing and neighbor discovery more often which can cause higher energy consumption. Since proposed method is using RCH and higher number of common channel as well as informing neighboring cluster about approaching mobile nodes it is consuming less energy. It can be seen also proposed method can outperform the CogMesh in terms of energy consumption (up to 20% less power consumption) especially in higher network dynamicity.

In terms of delay, Fig. 4, frequent reclustering in higher speed as well as high number of nodes in each cluster, cause up to 25% (3 millisecond) higher delay in CogMesh protocol compare to that in proposed method protocol. It also shows proposed method can outperform the CogMesh in terms of transmission delay (up to 10% less delay) especially in higher network dynamicity.

As it can be seen from the results by either increasing the speed of mobile nodes or having higher number of mobile nodes in the network performance of CogMesh dropped more compare to slight performance degradation in proposed method. As it can be seen from the result, proposed method could improve CogMesh throughput, power consumption and delay, up to 5%, 30% and 25%, respectively. Results show the both protocols can handle the low speed and low mobility in the network, while proposed method outperform CogMesh in the more dynamic network.

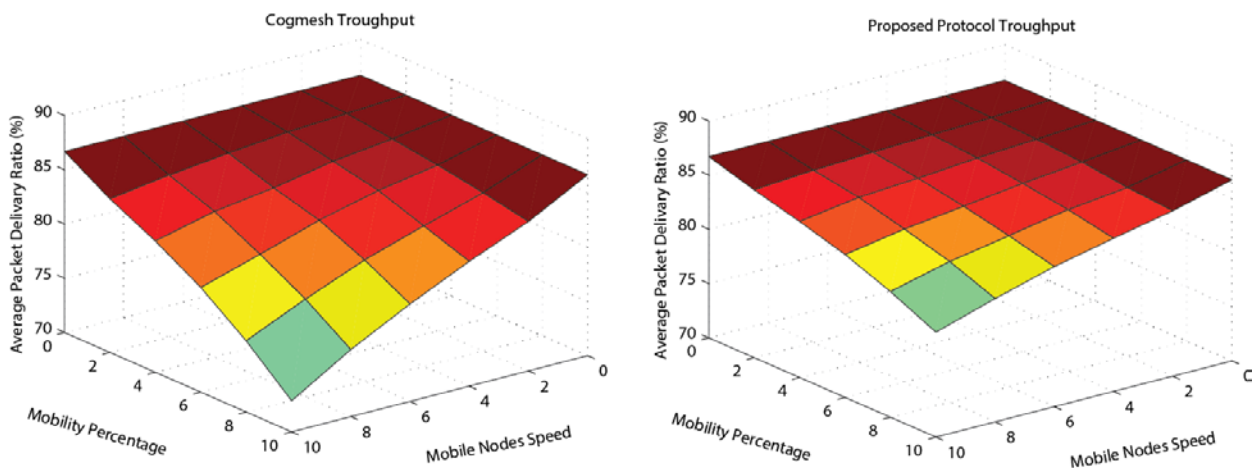


Fig. 2: Impact of nodes speed and network dynamics on the throughput of protocols.

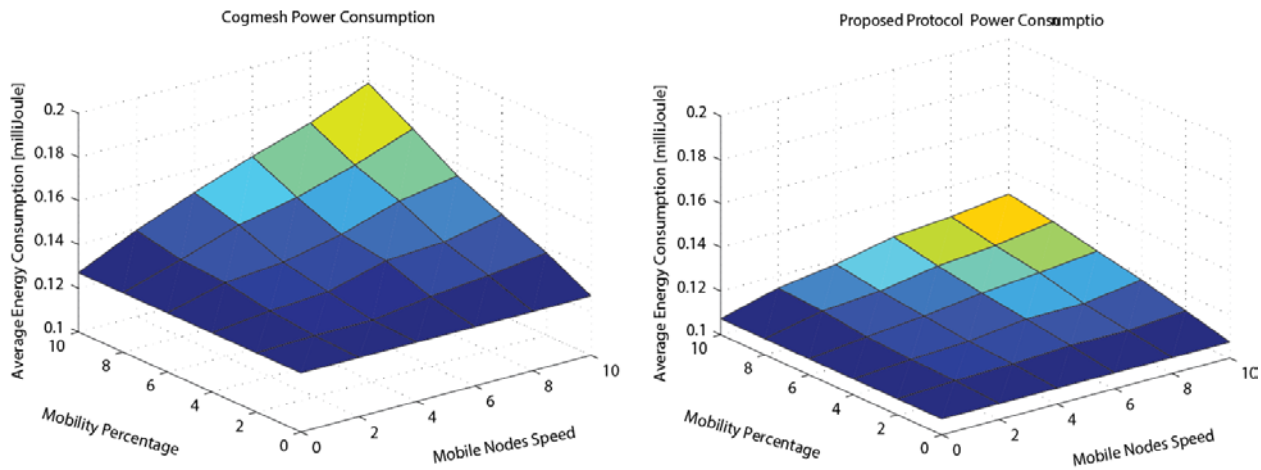


Fig. 3: Impact of nodes speed and network dynamics on the power consumption of protocols.

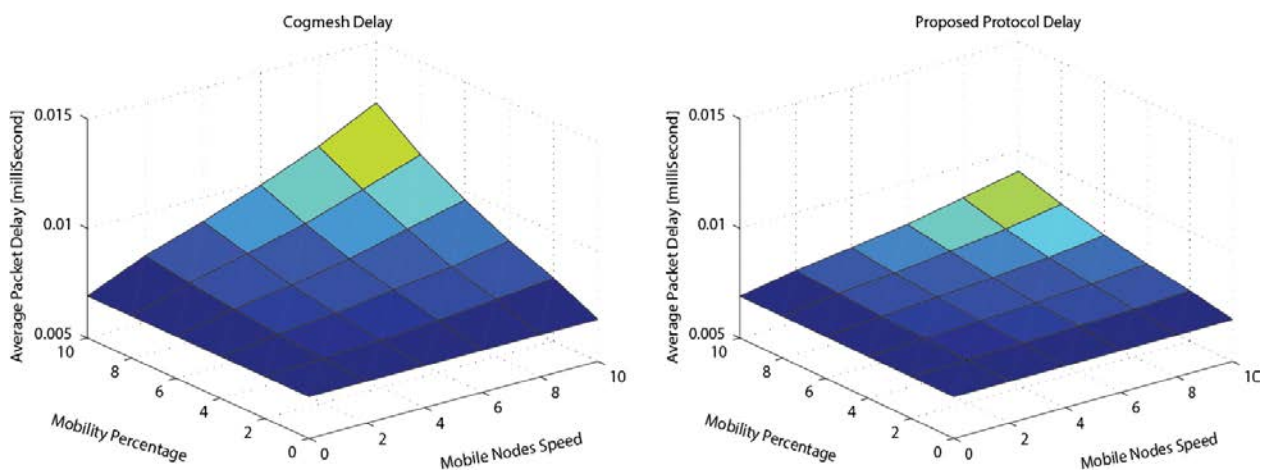


Fig. 4: Impact of nodes speed and network dynamics on the packet transmission delay of protocols

VI. CONCLUSION

This paper proposed a novel dynamic spectrum allocation method which is aimed to be more robust to PUs activity as well as node mobility or terminal mobility in the network by integrating the spectrum sensing at physical (PHY) layer with the packet scheduling at MAC layer. The proposed method uses a novel method to form the clusters in a very stable manner to reduce the number of re-clustering in the network and consequently consume less energy. The simulation study shows that the proposed method could outperform CogMesh in terms of packet delivery ratio, power consumption and delay, up to 5%, 30% and 25%, respectively. Considering cross-layer approach, appropriate routing protocol also can be integrated with the MAC protocol which remains as a future work.

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An Experimental Study of WSN Transmission Power Optimisation Using MICAz Motes

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Abstract—Adjusting nodes positioning to satisfy the needs of network connectivity and power conservation in wireless sensor networks (WSN) is not a trivial task. Too far node distance may result in unreachable transmission while too near node distance not necessarily guarantee nodes connection. By knowing an acceptable distance range to position nodes in the networks, it can greatly help researchers to better configure settings of WSN environment, while reducing precision of readings between real WSN testbed and the simulation counterpart. This paper demonstrates a real implementation of transmission power control to assist the decision-making on nodes positioning to be used in a sensor network communication. The implementation of the study is done using the MICAz motes with a nesC programming in TinyOS in an indoor environment. Our findings suggest that to assure network connectivity, motes must not be located more than 1.18 metres distance when they are placed on the floor building and not more than 2.37 metres distance when they are situated about one metre above the floor building. Our results also indicates that the maximum length the signals can travel using maximum power 0 dBm is merely about six metres long.

Keywords—WSNs; transmission power control; transmission range; topology control

I. INTRODUCTION

Several techniques and approaches have been proposed to tackle the issue of power control of WSN. The main idea of the techniques is to adaptively fine-tune the transmission power to be least energy consumption possible [1], [2], [3]. This can be done by measuring several parameters such as distance, position, and signal strength. In [4] and [5], every node adjusts the transmit power based on the local information it receives from its one hop neighbours. In other works by [6], the power adjustment is done based on distance between nodes. However, regardless the techniques used to solve the energy problem, power control is seen a vital strategy in doing so, and discretizing the transmission power level can substantially decrease the overall energy consumption while increasing the lifetime of the network [7].

In this paper, we properly study the nodes positioning to better support nodes configuration in testing environment of WSN such as simulation or emulation. The main aim is node

connectivity. We view the connectivity is a key issue as in any communication domain, without connection, no message can be delivered, no communication can be made. By knowing the distance in mind, nodes can strategise action to avoid an unwanted power loss, therefore, preserving residual energy, reducing energy waste, and improving lifetime.

The following sections describe the study in details. Section II defines transmission power control in sensor communication context. In Section III, we demonstrate the conducted experiment with results analysis. Finally, we envisage future works and some concluding remarks in Section IV.

II. TRANSMISSION POWER CONTROL AND MANAGEMENT

Transmission power poses a significant impact on energy consumption in WSN. A best reception of transmitted signals can be achieved if three physical conditions are carefully adjusted: low transmit power, good receiver sensitivity, and less multipath fading. Transmit power directly acquire battery power, where the more power used to transmit, the higher the battery depletes its energy.

The issues of power management had critically became a hot debate among the WSN communities. In addition, many researches have also demonstrated either by simulation [8], [9], [10] or emulation that transmission power is correlated with network coverage and connectivity, but less works are done in real implementation as though implementation results demonstrate a strong acceptance of output data while guaranteeing data validity. It is therefore, through the study we gather empirical information that is confidently acceptable to serve as an input parameters in simulating sensor networks communication.

III. EXPERIMENTAL ANALYSIS

In this section we present the transmission and reception power measurements conducted using MICAz motes [11]. The experiments are conducted related to distance and height without considering phenomenal conditions such as weather and temperature. In one occasion, we measure the reception power or the Received Signal Strength Indicator (RSSI) by varying the distance between transceivers when both motes positioned on the floor at the sixth level of a building, while

in another study, the same setup is replicated but with motes positioned on a computer table at certain height from the floor. The recipient mote behaves as a power recorder that displays the readings on a computer screen from a computer attached to the mote (the BS). The computer is running a power application programmed in nesC codes in TinyOS and a built-in java tools to display the readings on the screen.

MICAz motes use eight discrete transmission power levels of the CC2420 radio (as documented in [12]). The lowest power level 3 (equivalent to -25 dBm transmission (T_x) power) gives the very least battery power usage while the highest power level 31 (equivalent to 0 dBm of T_x power) consume the biggest percentage of battery power.

A. Description of Experiments

In its simplest form, RSSI can be used as indicator to adjust power in nodes communication. For instance, upon receiving packets and examining signal strength, the receiver determines what transmission power it should select for packet reply. This would benefit in energy reduction because transmission power significantly involved energy.

The signal strength of the received packets enable us to estimate distance among nodes. The RSSI not only imply the radius and coverage of the transmitter, but also enable us to interpolate the optimum distance between transmitter and receiver.

In brief, our main aim of the experiment is to find an acceptable distance between motes so we have a realistic idea how we can ideally position motes in simulation topology to investigate the energy factor of WSN. We are also interested to find the maximum transmission distance for certain power levels of the CC2420 radio.

B. Device Setup

In the study, we assume the MICAz monopole antenna radiates signals uniformly in all directions (omni-directional) as viewed by [13] and not losing more than thirty percent of energy as demonstrated by [14].

We conduct the experiment by considering the following motes setup: spatial motes communication where motes are positioned in an indoor computer laboratory environment, radio propagation among motes is in Line-of-Sight (LOS) and Non-Line-of-Sight (NLOS), fix and stationary mote behaviour, use IEEE 802.15.4 channel 26 with an assumption of no signal interference with any nearby 802.11 devices, have equal processing and communication capabilities, not equipped with Global Positioning Systems (GPS) capable antenna thus unaware of location, positioned randomly in the deployment area, capable of selecting a variable transmit power, and equipped with a new full power 3.0V AA zinc chloride batteries.

C. Node Position Measurement

In this experiment, we built a power application in TinyOS using nesC codes. The application is installed in two MICAz motes using a programming board. Both transmitter and receiver is programmed with lowest output power -25

dBm. Orientation of the antenna is vertical and motes positioned either on the floor or above the floor, specifically 0.72 m of the table height. We assume multipath interference exists and affecting the signal propagation.

We named the motes as Mote 0 and Mote 1 by setting the TOS_NODE_ID to 0 and 1 respectively. In our experiment, green LED on Mote 1 toggled upon transmitting each packet while red LED on Mote 0 toggled upon receiving packet. Typically, we design the motes with specific functionality, with Mote 1 functioned as a sender and Mote 0 functioned as a receiver. Mote 1 sends packets unicastly to Mote 0 in a simplex communication. Fig. 1 shows this configuration.



Fig. 1 Experiment setup of one-way-mote link: (a) Mote 1 sends packets to Mote 0; Mote 1 capable of sending and receiving packets while Mote 0 is sets to only receive packets, and (b) Perspective view of Mote 1 and Mote 0 radio components (T_x , R_x). Note that Mote 1 has the logic to send and receive packets but since Mote 0 are not sending any packets, thus no packets detected (on the R_x radio) at Mote 1

We placed Mote 1 at various angles between 45° to 360° from Mote 0. Due to multipath effects, each angle gives different readings on signal reception. For instance, at 45° angle from the receiver, both antennas may be blocked or experience a high multipath fading by chairs and computers. However, we can generalise the conclusion as we recorded many readings from all 48 tests run we conduct in all three labs and calculate the average.

For each node position, we also recorded ten readings in each three different nodes setup, namely Obstruction (OBS), Near Obstruction (NOBS), and Without Obstruction (WOBS). In the OBS setup, several opaque objects are placed between both nodes such as computer screen, computer casing, keyboard, cables, table stand, and chair where nodes unable to communicate in LOS. To introduce multipath effects, in the NOBS setup, nodes are positioned very near or next to these objects and able to communicate in LOS. In the WOBS setup, nodes positioned a distance from the objects, and can communicate directly with nothing blocks. The output power used by both motes is -25 dBm (current consumption equivalent to 8.5 mA) which is the lowest programming power output available in the CC2420 radio.

Comparing both modes, the ATF communications generates better readings which suggests height from the floor increase the range of transmission, where at the best position, they can be up to seven metres away from another motes. However, we may argue how high is high to achieve the best signals reception for an indoor motes communications. The ATF communication also produces a more stable and predictable range.

Based on Fig. 2 we conclude that the distance between any pair of motes is suggested to be between 0.37 and 1.18 metres when motes located on the floor, and between 1.07 to 2.37 metres when motes located about one metre from the

floor. Therefore, by considering a minimum values from each group, 0.37 metres for OTF and 1.07 metres for ATF are the best choice of distance among nodes to guarantee connectivity.

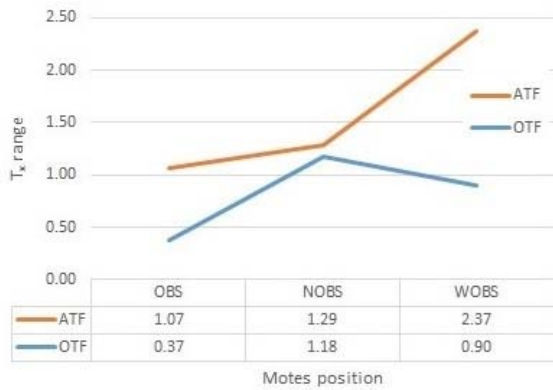


Fig. 2 Transmission range for above the floor and on the floor motes position at three different scenarios: with obstruction, near obstruction, and without obstruction

We also found that for the OTF mode of communication, multipath interference such as signal reflection, diffraction, and scattering not necessarily shorten the transmission range of the wireless signals; as being shown that OTF-NOBS in overall produces longer range compared to OTF-WOBS. This effect may due to antenna orientation between sender-receiver as also been suggested by [14], which is subject to future research in our study.

In addition, we conduct another experiment to measure the distance between motes when using maximum transmission power 0 dBm. We vary the distance of Mote 1 from Mote 0. The experiment is done in five settings: above the floor without obstruction (ATF-WOBS), above the floor with soft-partition obstruction (ATF-SP-OBS), above the floor with computer screen and keyboard obstruction (ATF-C-OBS), on the floor without obstruction (OTF-WOBS), and on the floor with computer table obstruction (OTF-T-OBS).

The graph in Fig. 3 demonstrates the strength of the signal received at Mote 0 is proportionally degraded with the increment of Mote 1 distance. The ATF in all settings gives better strength compared to the OTF settings, with the strongest RSSI detection measured at -63 dBm (ATF-WOBS) and -89 dBm (ATF-with obstacles).

The experiment indicates for any node positions in the laboratory setup, the detected RSSI falls between -63 to -89 dBm for ATF and between -73 to -93 dBm for OTF. For both of the setup, the results suggest that by using T_x power of 0 dBm, node placement between each other must not exceed six metres as no signal detected in the presence of obstacles. We also found that with the presence of obstacles, signal strength less than -63 dBm (or less than 1.07 metres) for ATF and less than -73 dBm (or less than 0.37 metres) for OTF can cause signal lost and undeliverable.

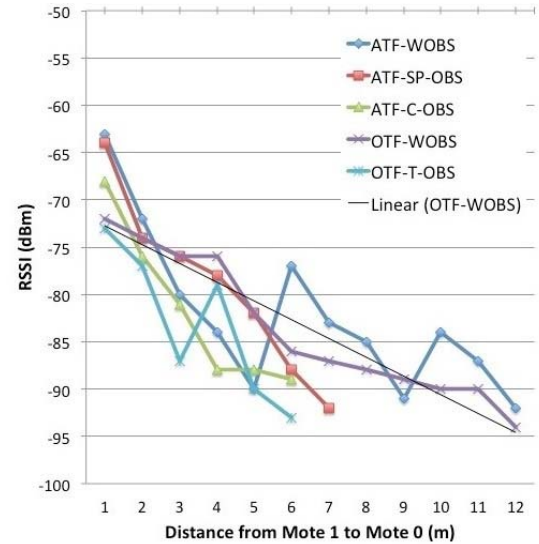


Fig. 3 The RSSI captured at Mote 0 at different Mote 1 distance when Mote 1 transmits packet with maximum power 0 dBm

D. Variable Power Range Measurements

In this experiment, we design a power range test to determine the maximum possible signal distance at certain transmission power of the CC2420 radio.

Node topology settings for this experiment has a similar setup with the experiments in Subsection C. With this setup, Mote 1 with functioned to send but not limited to receive packets, is staggerly programmed and flashed with eight different transmission power (as documented in [12], p. 51). In each power test, we adjust Mote 1's distance accordingly away from Mote 0 to a distance where Mote 0 lose its reception sensitivity, i.e., at a point where Mote 0 detects RSSI of -89 to -94 dBm. At this point, the distance is recorded.

Fig. 4 shows the transmission range for both ATF and OTF under different conditions at different power level. All recorded readings showed the expected results except for ATF-OBS and OTF-OBS where motes positioned on the floor give better range in the presence of obstacles. We again suggest the effects of multipath somehow helps the signals to travel further. Position granularity of motes may also plays a role to the result. We reserve to seek a better answer to this issue in future works.

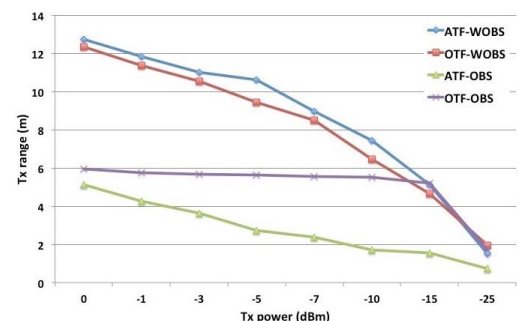


Fig. 4 Transmission range by using different transmission power in several nodes setup

Transmission range for all cases (except ATF-OBS) shows a big degradation at power transition from -15 dBm to -25 dBm. Therefore, we may conclude that by using transmission power of -15 dBm as a lowest power can better assure coverage compared to -25 dBm. Any power weaker than -15 dBm is considered unstable that signal may be unexpectedly lost. Finally, however at certain circumstances, these readings indicate an antenna positioned at the height of 0.72 metres seems not high enough to provide a strong and stable signal reception.

On the other hand, the readings may also help us to build a transmit power lookup table to be used in transmission power control. The lookup table is useful as a power reference to any node in power adjustment for communication with neighbours. In terms of sensor routing, a point of suggestion can be made that network coverage of six metres (according to the worst case scenario of OTF-T-OBS) can be guaranteed if a node receive packets with RSSI between -63 dBm (best of ATF) and -73 dBm (best of OTF).

IV. CONCLUSION AND FUTURE WORKS

In this study, we investigate the problem of finding the distance between nodes and transmission range in an indoor environment. The chosen distance must also be better enough to satisfy nodes connectivity. Our findings indicate for a case of motes positioning on the floor of a building, the distance between pair of motes achieve best signaling within 0.37 and 1.18 metres apart, while for a case of motes positioning about a metre high above the building floor, the best signaling can be achieved within 1.07 and 2.37 metres apart, even at some circumstances multipath effects are likely to affect the distance. The findings are beneficial for nodes positioning in simulation topology setup.

We also conduct two experiments to determine transmit power, thus realising power control in nodes positioning. By referring to specific motes configuration in an indoor environment, our findings indicate transmit power has to be controlled to achieve better power conservation and network connectivity.

We believe our contributions in this study particularly important to better understand the nature and behaviour of transmitted wireless signals while in other occasion assist a better improved accuracy and correctness formation of WSN simulation settings. Our study benefits not only many WSN researchers which deploy the idea of research on simulation, but also to WSN applications where energy conservation is an important issue.

Our future works may reserve to investigating power behaviour in the presence of mobile nodes in an inter-floor indoor environment.

ACKNOWLEDGMENT

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Adaptive Tree-Based Routing for Multi-Channel Wireless Sensor Network

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Abstract— In this work, a tree-based adaptive routing for multi-channel wireless sensor network is proposed where channel trees are used in routing and the trees are not static. The case for faulty nodes and energy depletion is considered where the topology of the network may change throughout the implementation to the network. In order to make the network to have a dynamic routing, the proposed channel assignment scheme is performed in a semi-dynamic manner. The proposed protocol consists of three phases; first, the neighbor classification phase, second phase is the channel initialization phase and the third phase is adaptive routing phase.

Keywords— *wireless sensor network; multi channel protocol; adaptive routing*

I. INTRODUCTION

Data gathering is one of many applications for WSN. In data gathering, data is relayed from the source nodes to the sink node. A proper architecture is useful to facilitate data gathering in WSN especially in multi-hop situation where data are relayed through transitional nodes until it is finally reaches the sink node. In critical applications, the network reliability is important to ensure data will reach to the destination successfully. In most of the existing protocol, power consumption for data transmission is disregarded. Power consumption for data transmission is equally important as it could changes the routing topology for the network. This work addresses the problem where the proposed algorithm offers maintenance solution for tree-based multi-channel WSNs.

Recent related studies is further discussed in section II followed by the presentation of the network model for the proposed adaptive tree based routing for multi-channel WSN in section III. In section IV, the simulation results were discussed. Finally, conclusion and future works is discussed in the section V.

II. RELATED STUDIES

The usage of availability of multiple channels is an ideal way to increase the packet delivery ratio. This is because by using a single channel the intra-interference is high and therefore the usage of multiple channels could enables message sending in neighbouring nodes simultaneously by using orthogonal

channels. There are 16 non-overlapping channels in the 2.4GHz ISM band that are available according to IEEE 802.15.4 standard.

However multi-channel WSNs face many practical issues. In multi-hop environment, without proper coordination one of the common problems is the deafness problem where message transmission between nodes failed because the nodes are tuned to different channel.

Since multi-channel comprises of complex channel assignment, simple channel coordination is needed. A common solution for this problem is the use of a dedicated control channel. The set back of using a control channel is the nodes still needs to tune to the control channel during coordination. Other simple solution is to use a centralized channel assignment to ensure the connectivity of the network. [1] proposed a centralized of tree structure to facilitate the routing tree for multi-channel wireless sensor network. However in a centralized setting, nodes need to share information extensively and it is not suitable for dynamic routing.

A dynamic routing is important for the case of energy depletion or node failure. The sudden change in the routing topology may disrupt the data collection because it may interfere with the network tree. There are few solution proposed for the single channel WSNs such as in [2]. A busy or congested channel would benefits from dynamic routing such as solutions proposed in [3].

An effective routing is often is a tree structure or a cluster-based structure that has been proven to show the better performance than a flat routing topology [1,2].

However, in multi-channel set up, a tree-based or a cluster-based is far more challenging to be applied than in a single-channel network. This is because a tree-based structure in multi-channel WSNs is usually developed in a centralized set up such as in [1]. A centralized set up is not suitable for a dynamic routing and it limits to fully exploit routing information as the nodes must share information through all the nodes in its routing tree.

In a situation where some of the parent node has depleted its energy, the path from the sender to the receiver will be lost in the initial fixed channel assignment. This is due to the

transceiver in both sender and receiver needs to be on the same channel for the communication to be possible. Another case to be considered is the case where there is high traffic in a certain channel. In this situation, the chance of collision is high and the allocated channel is not enough to cope with the high traffic and packet loss is likely to happen.

To solve the above mentioned problem, a multi-channel tree structure is designed and an efficient routing protocol is proposed using the improved structure to find an alternative path on the network graph. In this study, a maintenance algorithm is also proposed in the third phase which is called the Adaptive Phase algorithm.

III. PROPOSED ALGORITHM

The proposed protocol would adapt the tree-based approach by pre-determining the best channel to connect to. Due to changing requirements in the network, the channel assignments can be renewed by following the success and failure rates of the packet transfer. The proposed protocol consists of 3 phases; first, the neighbor classification phase, second is the channel initialization phase and the third is adaptive routing phase. Fig. 1 describes the flow for the proposed multi-channel protocol.

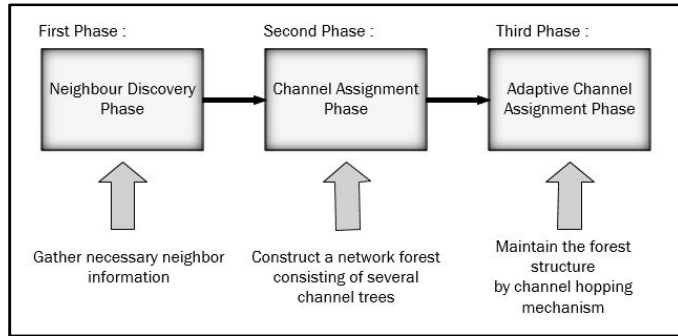


Fig. 1. Flow for the Proposed Multi-Channel Protocol

A. The First Phase : Neighbour Classification Phase

In this section, the model for the first phase construction is described as follows.

- w is defined as a neighbor of a node v if these nodes are interconnected with each other based on physical proximities, that is, the nodes are in the communication range with each other.
- There exist a root r where data is gathered
- This phase consists of two processes: getting number of hops for each node, and classifying neighbors into set of *children* C_v and set of parents P_v .
- In this phase, all nodes use one default channel as a control channel to ease the channel assignments.

The algorithm for the first phase is described step by step as follows:

Step 1 : The root sends a message to its nodes that are in the first level.

Step 2 : When a node receives the message, it records its number of hops as 1, and adds the root as its parent.

Step 3 : The node then sends notification messages about its number of hops to its neighboring node(s). When the neighboring nodes receiving it for the first time, it label the sender as the set of possible parents and records the number of hop of it parents and set the number of hops for itself where the $number_of_hops = number_of_hops + 1$. The node sends the notification message to the sender and the sender records it as a possible children.

Step 4 : After receiving subsequent BFS messages, each node updates its parent set and the children set. The process is repeated until the maximum height of the network.

After the implementation of this phase, the network is still considered as a flat network where it uses a single channel and has no network structure. In this phase, each node v in the network transmits one message and receives W_v messages, where $|w_v|$ is the degree of the node v in the tree. This phase consists of two processes: getting number of hops for each node, and classifying neighbors into set of *children* C_v and set of parents P_v . The classification of neighbors is important in the third phase in determining the path for the adaptive routing. Each node v has w neighbor that use the same channel in its communication range. Since all the nodes now in a single channel, the intra-tree interference for the nodes that is currently tuned to the default channel is the number of neighbors. Therefore, each node is assumed to have the knowledge of its number of intra-interference.

B. The Second Phase : The Channel Initialization Phase

After the implementation of the first phase, the second phase of the algorithm is initiated. In this phase, starting from the base station, the channel assignment will be done according to the least intra interference set.

- Starting from root r or the base station, a network forest consists of k channel trees, is built in this phase where the root is the control station for the second phase as it decides the channel assignment for each node.
- Each tree channel tree T_{ch} is the shortest path routing trees from the root to every node.
- Each channel tree T_{ch} starts from the root
- The maximum intra-interference allowed is set for the channel trees based on the network density.

Step 1 : Each node set its number of neighbour as the number of its intra-channel interference

Step 2 : Starting from $level = 1$ and channel $ch = 2$, the Base Station assigns a channel to each node by adding u into the channel tree T_{ch} for certain condition.

Step 3 : Node u is added to the channel tree T_{ch} such that by adding u makes the least interference to channel tree T_{ch} , and u is assigned channel ch such that it does not exceed the maximum intra-interference.

Step 4 : Node u then inform its neighbors about its current channel to be used in the third phase.

Step 5 : Once all of the nodes in this level are assigned with a channel, the process is repeated in the next level. The nodes at the next level choose one of the channels that are being used by its set of parents to ensure connectivity, and one parent that is using the same channel as the node is elected to be its main parent.

C. The Third Phase : The Adaptive Routing Phase

After the second phase, a dynamic routing is achieved by the channel hopping mechanism. Here, the proposed protocol suggests the concept of channel borrowing and channel reuse as it allows the network to respond to the requirement quickly. In this phase, each node is an autonomous node where decision for channel allocation is done locally.

Step 1 : A node forward the packet to its parent.

Step 2 : If the packet is received by the parent, do nothing.

Step 3 : Otherwise, update the probability of failure.

Step 4 : The probability of failure is updated until it reaches a certain threshold.

Step 5 : After the probability of failure reaches the threshold then the node will change its channel.

Step 6 : Eventually, the solution of the best channel of each node converges, and the node keeps using the same channel, until another occurrence of node failure or congested channel.

After the channel decision is done locally, the autonomous nodes would be able to divert its path whenever a channel is busy, to balance the network load, or to permanently divert the path to other candidate parent. This could ensure a clean end-to-end data transfer. Since each node now can make a decision within itself, the coordination among nodes is achieved without central authorization, scheduling or completely sharing information.

IV. SIMULATION RESULTS

The simulation experiments that have been carried in order to test the dynamic routing topology are described in this section. The performance of the proposed approaches is evaluated through extensive simulations.

A realistic communication system using the proposed multi-channel architecture for WSNs is developed using JAVA. In order to simulate it in a distributed setting, an object-oriented programming is used such that each node will have the properties of a sensor node. For a fair comparison the radio frequency model of the CC2420 radio is used where two nodes is assumed to be able to communicate with each other if the distance between them is less than the transmission range because the energy consumption for sending is relatively close to receiving [5].

First, the nodes are distributed randomly on the field. The base station is placed in the middle of the field. In this study, the performance of the proposed protocol is evaluated by comparing the end-to-end packet delivery ratio (PDR) that is the ratio of the number of packets delivered at the sink to the total number of generated packets.

In the simulation, the work is tested by simulating a realistic data gathering situation where a packet is generated

for each node and is relayed through the routing trees determined after the implementation of the first and the second phase. These packets will be forwarded along the channel trees until it is finally gathered to the root or a Base Station that is located in the center of the network.

In order to simulate the data transmission, time is discretized into a number of *rounds*. A reliable data transmission in each round is assumed. In each round, the transmission from the sender to the receiver is assumed to be successful according to a certain rules:

- When a sender s is sending message, node s changes its mode into sending mode. After each round, a node is assumed to be idle unless the node is still having some packets to forward to its main parent.
- Case 1: If the destination node t has depleted its energy or it is in offline mode, the data transmission is failed.
- Case 2: If the receiver is in idle mode, the destination node will change the status to receiving. In this case, data transfer is considered to be successful.
- Case 3: If the destination node t is in either sending or receiving mode from other sender node, the collision is assumed to happen and data transmission is failed.

In a tree-topology, a parent node may have a several children nodes, however, because of its half-duplex properties, only one winner could send or receive message at a time. Each node is given an ID, where the ID is given one by one based on the physical distance to the Base Station. The one that is closer to the Base Station is assigned a lower ID and vice versa. Each node can only either be in their sending or receiving mode because of its half-duplex properties.

A. Experiment 1 : Network Priority

In this experiment, the ID priority is investigated whether it affects the packet delivery ratio. For example, in ascending order id priority, among all nodes, nodes with lower id will be given the first priority to send message. The parameters that are used in the simulation is as shown in Table 1.

TABLE 1. PARAMETERS USED IN EXPERIMENT 1 AND 2

Properties	Network 1
Number of Nodes	250
Transmission Range	100 m
Field Size	1120 m × 650 m
Number of Channels	5

This experiment has been repeated in the same network coordinate. As it can be seen in Fig. 2 the ascending ID order achieves 100 percent of packet delivery ratio. By giving different type of priorities to the sending nodes, it is proven that the ascending order performs better than descending ID and random ID priority. Therefore, in the next experiment, ascending ID order is used to compare the result with similar work.

B. Experiment 2 : Comparison of the proposed protocol with existing tree-based protocol.

Using the same network such as in Experiment 1, the performance of the proposed protocol is compared to previous developed protocol [1]. The energy waste factor is measured as the average amount of collisions, overhearing, and idle listening happening in one node per round during the operation phase.

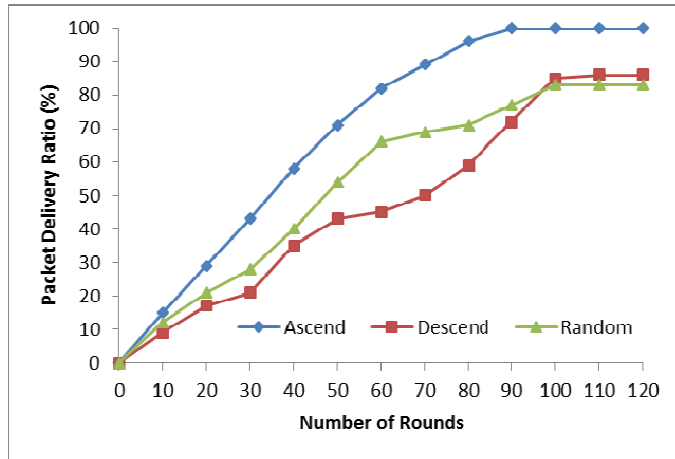


Fig. 2. Comparison of different type of priorities in network PDR

In this experiment, an ascending ID priority is used to generate the result to compare the proposed protocol with the existing tree-based protocol proposed in [1]. It can be seen in Fig. 3 that the proposed protocol reaches 100% packet delivery ratio and while PMIT protocol reaches 85%. Since the proposed protocol offers the maintenance solution, it can avoid the packet lost because it has been diverted to other channel trees. 15% of the packets lost in the transitional node(s) that has depleted its energy.

V. DISCUSSION AND CONCLUSION

In previous sections the proposed algorithm for tree-based multi-channel wireless sensor network has been presented. The conclusion that can be drawn from this research is that, although, consideration of route diversion is a simple approach but it is an effective way to address the node failure.

It is shown that the ID priority has given some impact in the overall network performance. It is also shown that the network density has much effect on the network. Implementation of the idea of this proposed protocol has shown the significant improvement of previous protocol where it offers more flexibility and fault tolerance form the one in [1].

By using the proposed protocol, the dedicated control channel can be fully eliminated and there is no need to do scheduling after the third phase implementation. Therefore, it can avoid time synchronization error as used in previous distributed protocols. It also uses less channel resource and could significantly increase PDR compare to a scheme with a fixed channel assignment in a high traffic rate. The exact

advantage depends mainly on other variations used in the network.

In this paper the performance of the proposed protocol is compared with existing literature in terms of packet delivery ratio. In the next experiment the end-to-end delay and the network lifetime will be further investigated.

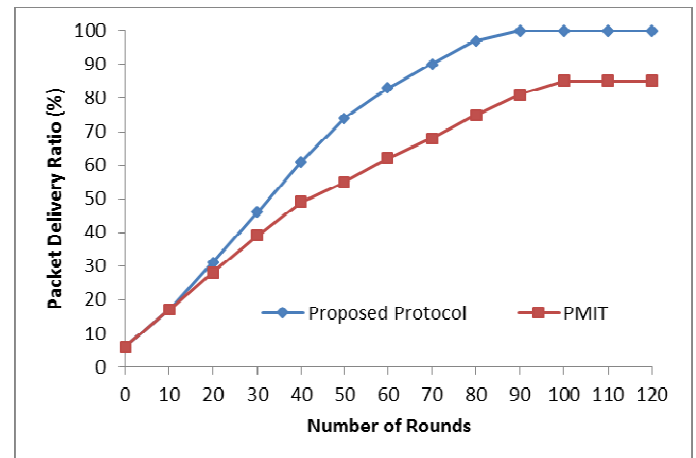


Fig. 3. Comparison of the proposed protocol with PMIT in terms of PDR

ACKNOWLEDGMENT

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An Analysis of Smart Grid Communication Infrastructure & Cyber Security in Smart Grid.

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Abstract—This paper investigates the smart grid communication infrastructure with an upgrade of existing electricity networks and advanced ICT technologies, intelligent devices & system. Considering the energy utilities play an increasingly important role in our daily life, smart grid technology introduces new security challenges that must be addressed. This paper focuses on the implementation of smart grid communication architecture to develop a real time situation and study the security challenges that is occurring on its communication infrastructure. A micro grid structure with particular generation and distribution specification along with its relevant communication infrastructure is designed to analyze the security challenges. The studies are performed based on a digital computer simulation approach using the Power Factory 14.1 software package. Some dynamic system events were formulated to link the designed micro grid architecture to the real time situation and to analyze the change in different system parameters due to cyber-attack and effect of these changes are examined on the smart grid generation and distribution network. This analysis is further used for identifying the cyberattack prone areas to ensure cyber security and determine the methods required for enabling cyber security in smart grid communication infrastructure.

Keywords—Smart Grid; Communication Infrastructure; Communication Technologies; Cyber Security; Security Challenges; Generation Power; Frequency Response.

I. INTRODUCTION

The smart grid is a next-generation power system and an upgraded electricity network that depends on two-way digital communications between supplier and consumer in an automated fashion to improve the efficiency, reliability, economics, and sustainability of the production and distribution of electricity [1]. The research on smart grid is mainly focused on three systems in smart grid-- the infrastructure system, the management system, and the protection system. Among these communication infrastructure underlying of the smart grid that supports advanced electricity generation, delivery, and consumption; advanced information metering, monitoring, and management and advanced communication technologies. The protection system in smart grid provides advanced grid reliability analysis, failure protection, and security and privacy protection services. A successful smart grid infrastructure requires integration of cyber system with the physical power system. Although adoption of cyber system has made the grid

more energy efficient and modernized, it has introduced cyber-attack issues which are critical for national infrastructure security and customer satisfaction [2]. Due to higher dependability of communication infrastructure on information and cyber technologies it is highly vulnerable to cyber intrusion and corruption. Objective of this paper is analyzing the communication infrastructure of a micro grid in order to identify the critical areas that are prone to cyber threats and also identifying the changes in the real time data in different parts of the system after various faults and analyzing the fault conditions and monitoring the values of these fault conditions with the help of graph based simulation so that proper cyber security can be ensured [3].

II. SMART GRID COMMUNICATION REQUIREMENTS

The communication infrastructure between energy generation, transmission, and distribution and consumption requires two-way communications, inter-operability between advanced applications and end-to-end reliable and secure communications with low-latencies and sufficient bandwidth [4]. Moreover, the system security should be robust enough to prevent cyber-attacks and provide system stability and reliability with advanced controls. Major smart grid communication is required to achieve security, System Reliability, Robustness and Availability, Scalability, Quality of Service (QoS), Network Latency, Critical timing requirements, Reliability, Time synchronization, Multicast support and Data delivery criticality. Smart power grid infrastructure is full of enhanced sensing and advanced communication and computing abilities. In the smart grid, reliable and real-time information becomes the key factor for reliable delivery of power from the generating units to the end-users. The impact of equipment failures, capacity constraints, and natural accidents and catastrophes, which cause power disturbances and outages, can be largely avoided by online power system condition monitoring, diagnostics and protection [5]. It is very critical for electric utilities to define the communications requirements and find the best communications infrastructure to handle the output data and deliver a reliable, secure and cost effective service throughout the total system. Basically, two types of information infrastructure are needed for information flow in a smart grid system. The first flow is from sensor and electrical appliances to smart meters, the second is between smart meters and the utility's data centers.

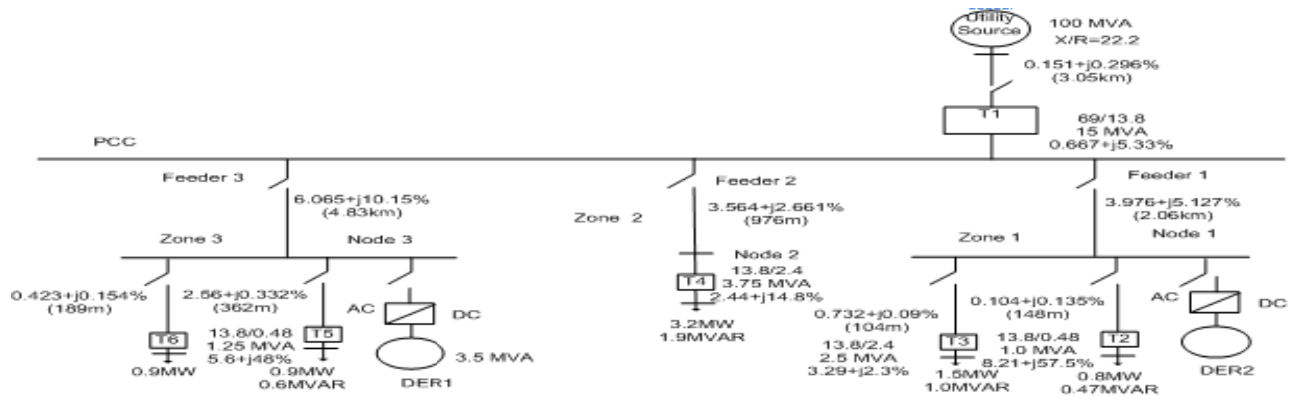


Fig.1. Single Line Diagram of Study System [9]

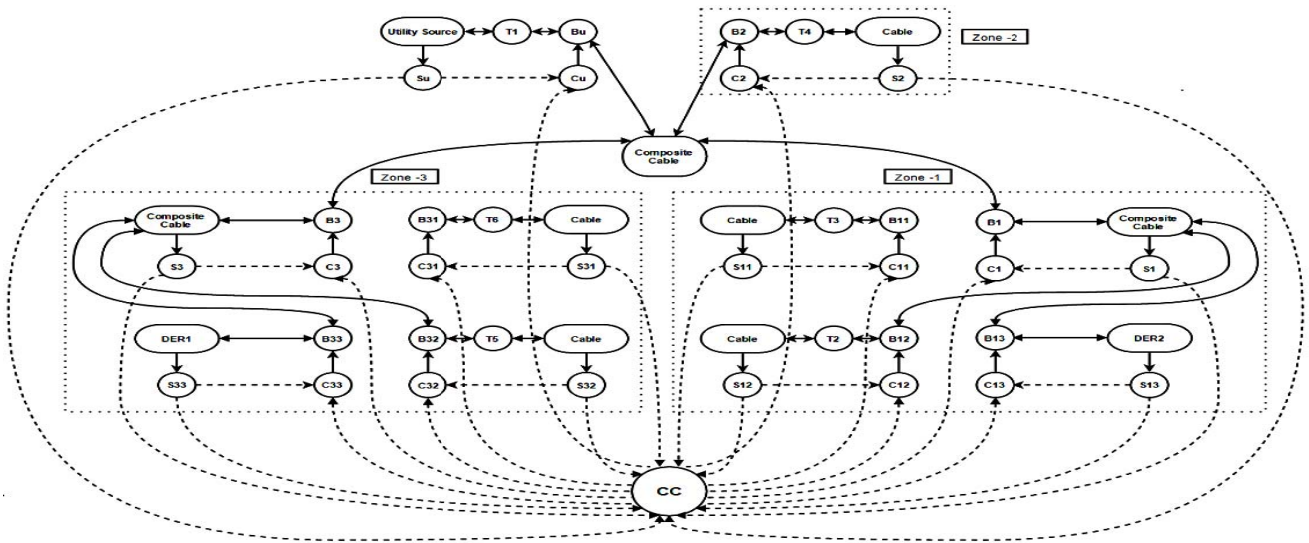


Fig.2. Electrical and cyber graphs for the test system of Fig.1 are in islanding mode. Nodes are comprised of DERs, circuit breakers Bi, cables, a transformer T of the electrical network and sensors si and actuator controls ci of the cyber network [7].

III. CYBER SECURITY OF SMART GRID

Rapidly-evolving threats require cyber security strategies that give manufacturers, utilities, and grid operators the flexibility to respond quickly and decisively. The basic concept of protecting a power grid from any means of cyber-attack is to ensure that the power grid has the ability to monitor, analysis, control and communication capabilities to the national electricity delivery system in order to maximize the output of the system while reducing energy consumption. Intelligent devices connected to smart grid can automatically adjust to changing conditions to prevent blackouts and increase capacity [6]. Seven major features must be applied in a smart grid to make it self-aware and self-protective from any kind of cyber-attacks and they are Enable active participation by consumers, Accommodate all generation and storage options, Enable new and upgraded products, technologies and services, Provide power quality for the range of needs in a digital economy, Optimize asset utilization and operating efficiency, Anticipate and responds to system disturbances in a self-healing manner and Operate resiliently against physical and kinds of natural disasters.

IV. CHALLENGES TO FACE IN IMPLEMENTATION

A. Security Strategies to Enhance Information Security in Smart Grid

The most common way to protect information is using a password to encrypt the network information. By using encryption of network information the access of the unauthorized people to the encrypted information is limited and only the people who have the encryption keys are able to open the encryption and use its information. Nowadays most of the information encryption methods and models are used about computers. Most of the information encryption systems for data networks are divided into two main groups: Symmetric key encryption and Public key encryption (asymmetric) [7].

B. Cyber Graph

Cyber graph has been used in this paper to assess the impact of cyber-attack. As a first modeling step, electrical and cyber graphs are formed such that each node represents associated grid elements; in this representation, nodes can be generators, transformers, loads or plug-in hybrids, circuit-breakers (electric), switches and control centers, sensors and

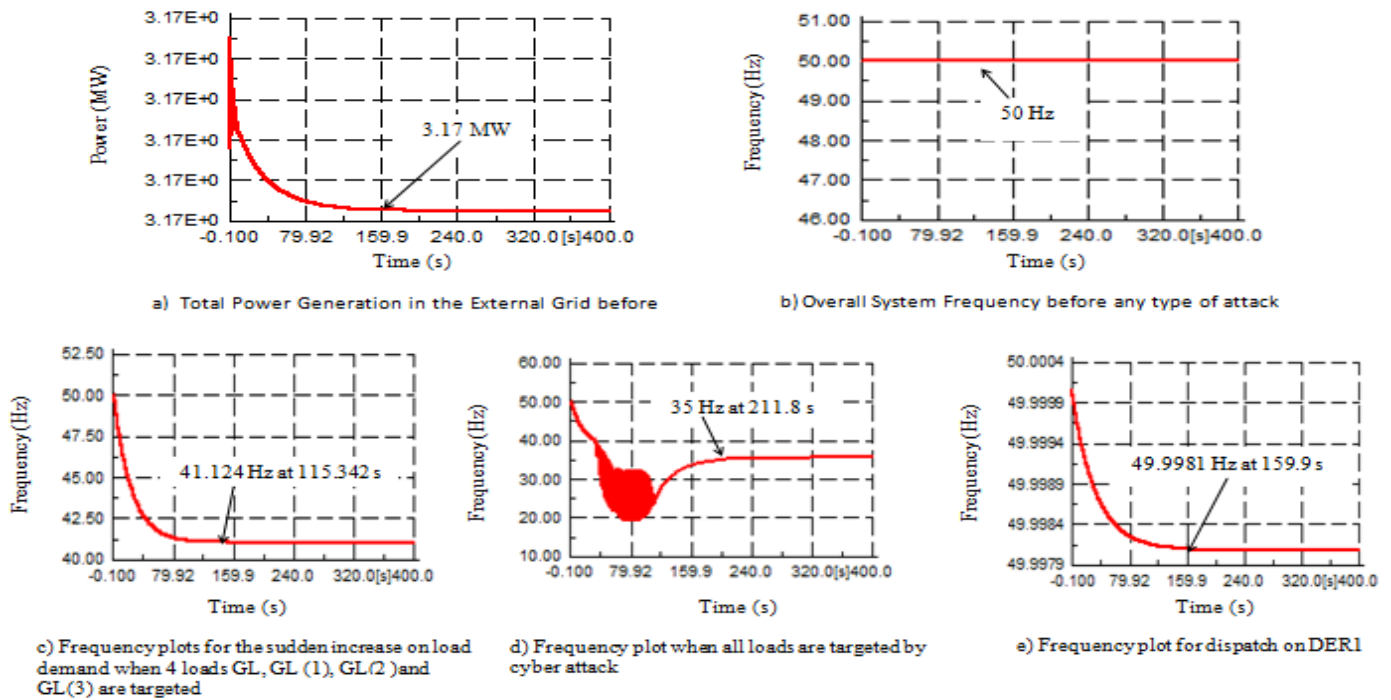


Fig.3. Simulation Results for Multi load event & Dispatch event performed on the test system.

breaker actuator controls (cyber). The electrical and cyber graphs are shown along with edges representing dependencies amongst components within the same network or at the cyber-physical bridge. The cyber infrastructure encompasses the sensors, SU, S1, S11, S12, S13, S2, S3, S31, S32, S33 that collect measurements from various system points, communication infrastructure, 10 breaker actuator controls and the control center as shown in Fig. 3. The sensor measurements can include active and reactive power, voltage and current phasors and on/off statuses of the switches. The control center acts as the overall system brain to connect or disconnect the loads to and from the grid based on their priorities and available generation capacity. The control center commands are sent to the switch actuators through communication links. We model each DER, switch, load, transformer, capacity bank and cable as nodes in the electrical graph. Each sensor, breaker actuator and control center are modeled as distinct nodes of the cyber grid [8].

V. RESULTS AND ANALYSIS

A. Test System and Simulation Tools

The test system in Fig 1 is composed of a 13.8-kV, three feeder distribution subsystem which is connected to a large network through a 69-kV radial line [9]. The 13.8-kV substation bus bar is radially connected to the main grid through the substation transformer and a 69-kV line. The network at the end of the 69-kV line is represented by a 69-kV, 100-MVA short-circuit capacity bus. A combination of linear and nonlinear loads (GL, GL1, GL2, GL3 and GL4) is supplied through three radial feeders of the subsystem. The system also includes two Distributed Energy Resource units, i.e., DER1 (3.5 MVA) and DER2 (3.5-MVA) on feeders F3 and F1 respectively. DER2 is a synchronous rotating machine

equipped with excitation and governor control systems. DER1 utilizes a voltage-sourced converter (VSC) as the interface medium between its source and the power system. DER1 represents a dispatch able source with adequate capacity to meet the real/reactive power commands, within pre-specified limits and also subsequent to disturbances. Such a dispatch able source may also include energy storage interfaced at the converter dc bus. DER1 also provides control on its output real and reactive power components independently.

B. Simulation Results

The micro grid test system we are working with is a dynamic system. Here we present results pertaining to the graph and dynamical systems modeling of cause-effect relationships of a cyber-attack on the test system. A graph theoretic formulation of distributed control is well-suited to this smart-grid representation. For this test system we performed two types of simulation events which are multi load event and dispatch event. The utility source of our test system has a generation capacity of 3.17 MW and 3.04 MVAR. Load shedding or trip off of generator will be occurred if an individual or the combined load demand exceeds generation. The impacts of cyber-attack on smart grid can be observed by creating events on different components of the system. Our system frequency is 50Hz. The threshold value for generator under-frequency is set at 48Hz (4%); that means when the system frequency drops under 48Hz, the generator will trip out. Preliminary results can be found in [10].

1) *Multiload Event:* When more than one load is targeted, sensors of corresponding loads are misinformed by the cyber attack which influences the control signals of corresponding loads to make a wrong decision by the control centre. For our test system as five loads are served at a time & multiple load

events are performed. Cyber attack execution for multiple loads is performed at $t=1\text{sec}$ where the total simulation is run for 400sec. Load event increases the load demand to 200% of its initial active power value and 5% of its initial reactive power value. When four loads i.e. GL, GL(1), GL(2) and GL(3) are attacked, the micro grid test system experience frequency deviation which is shown in Fig 3 (c). In the 400sec simulation run we observe that system frequency falls from its nominal value which is 50Hz to 41.124Hz at $t= 115.342\text{sec}$ and after that it becomes constant. From electrical and cyber graph we see that sensors S31, S32, S11 and S12 are respected sensors for these load which senses sudden changes on load demand. Control center, cc communicates all control signals at a particular time. Information of the attack on multiple load is collected by control center cc via corresponding control signals C31, C32, C11 and C12 respectively. Control center then perform necessary algorithm and make a quick decision for the whole system. As system frequency drops below the threshold value, generator may trip and frequent load shedding or blackout will take place. Fig 3(d) describes the frequency behavior of the total system when all loads are affected by cyber attack. Here we observe some unexpected fluctuations on the system frequency leading to system blackout.

2) *Dispatch Event*: Economic dispatch is the short-term determination of the optimal output of a number of electricity generation facilities, to meet the system load, at the lowest possible cost, subject to transmission and operational constraints. DER1 is a dispatchable source of 3.5MVA rated power with adequate capacity to meet the real and reactive power demands. The dispatch of DER1 is performed at $t= 1$ second after the total execution for the test system. Active power and reactive power change is considered as 200MW and 5Mvar of initial value respectively. Here we perform the simulation for 400 seconds. If load demand increases suddenly, we observe system frequency drops from 50Hz to certain range and after that it becomes constant. For DER1 sensor S33 has been targeted for the cyber attack. At any time t , for a switch i , the control centre, cc communicates with all control signal $C_i(t)$ where $C_i(t) = 0$ indicates open switch and $C_i(t) = 1$ indicates close switch. When sensor S33 is misinformed due to attack, corresponding control signal C33 helps control centre cc to collect this information so that it can perform necessary algorithm to make a quick decision for the whole system whether to keep corresponding breaker B33 open or close. Since, after dispatch of DER1 there are several more sources serving in the system at that time, frequency will fall as shown in Fig 3(e) but that will not affect the system severely. System frequency falls slightly due to the increased load on lesser number of sources compared to the time before dispatch event.

VI. CONCLUSION

Cyber-physical system security demands additional security requirements, such as continuity of power delivery and

accuracy of dynamic pricing, introduced by the physical system. Therefore, both information security and system-theory-based security are essential to securing cyber-physical systems, offering developing methodologies to overcome the effects of cyber-attacks, new regulatory approaches to encourage the adoption of innovative network technologies, improving the functioning and reliability of the grid by developing the sensing, enabling rapid response to faults and problems etc. Techniques to deal with the implementation and security challenges to enhance the overall security and efficiency of a smart grid, methodologies of dealing with vulnerabilities due to cyber-attacks and impacts of those attacks on the power system, observation of impact on power grid by employing intentional cyber-attack on the micro grid test system and effects of those attacks on different power system parameters, how the power system frequency, generation power and security components response to the sudden cyber-attacks to minimize the vulnerabilities due to the attacks are some major areas which should be given emphasis.

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Radio Network Optimization

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Abstract— This paper deals with the method of how to carry out radio network optimization for 2G and 3G systems with an overview of general GSM radio network optimization areas with regards to analysis and troubleshooting. The general steps and methods for wireless radio network optimization are first addressed. Then optimization methods are introduced. At last the issues of radio network optimization is discussed with special focus on the 2G and 3G networks as well as the required Quality of Service (QoS) to the subscribers. This study would help operators plan effective optimization of radio resources, enhance coverage, improve quality and increase capacity.

Keywords— radio network optimization; drive test; optimization methods; QoS; received signal level

I. INTRODUCTION

Optimization methods are used to solve cost functions reliably and rapidly. Four methods of optimization are frequently used: Random Walk, Simulated Annealing, Tabu search and Genetic Algorithms. GSM planning is the most essential part while working with radio network planning and optimization [1]. Network optimization has tradeoffs between quality, traffic/revenues and investments. Without fine-tuning the network, the customer complaints are increased and marketing becomes inefficient. The planning and optimization tools will assist the planner. All GSM operators find problems which are solved through Key Performance Indicator (KPI) and drive test analysis [2]. Parameters such as coverage, capacity, quality and cost for planning are considered during the planning and optimization process. Networks must be made in a way so that the network design can support the user demands [3].

RNP and optimization process play a very significant and vital role in optimizing a network in order to meet the increased demand from the customers. Coverage is the most important quality in a radio network [4]. Systems with good coverage will always be superior to a system with bad coverage. An area is referred to as being covered if signal strength received by MS in that area is higher than a certain minimum value. The aim of RNO process is to evaluate and maximize the quality of service in the network with the corresponding set of quality criteria [6].

A. Network Verification

The purpose of the Network Verification (NV) is to evaluate an independent and objective QoS inside a given service area. This is done with network measurement system. Some OMC (Optimization and Measurement Center) traffic measurements are done in parallel to provide a statistical data and to complete the network picture.

The field verification takes place after successful completion of site acceptance. It should be repeated before and changes to verify their effect on the network quality [1]. The service area, or the part of the network to be verified, is defined as a group of cells giving continuous coverage. It is always connected with a selection of test routes; the all verification and optimization activities are based in recurrent measurements over the same routes [5].

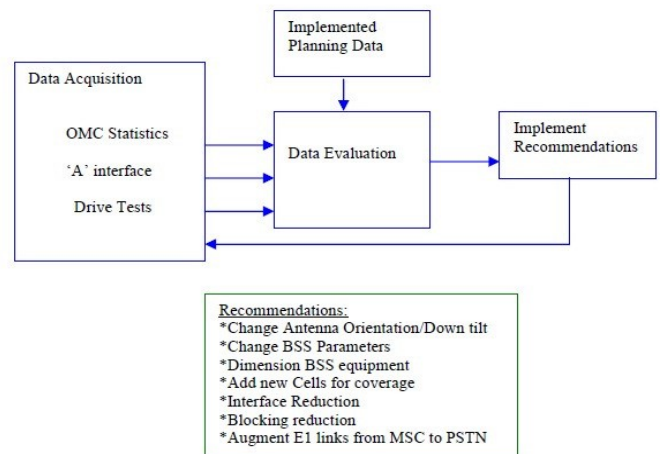


Fig. 1. Optimization Process

B. Network Optimization

Network Optimization can be defined as a continuous process of improving overall network quality [3]. Looking at network quality two different views should be considered. The customers (subscribers) view and the more comprehensive operators view overall network quality is illustrating this [4]. Usually a subscriber is not interested in site leasing or maintenance cost. As long as his service is not affected things like spectrum efficiency and network traffic are of no interest to him. Network optimization service and more general the Nokia Quality cycle service package are designed to support the operator in the most efficient way to improve all different aspects of network quality [6].

Radio network optimization means network performance improvement utilizing existing network resources [3]. Steps of optimization in general:

- Data collection and verification
- Data analysis
- Parameter and hardware adjustment
- Optimization result confirm and reporting

The purpose is increasing the utilization of network resources and solving the existing and potential problem on the network, identifying solutions for future network planning.

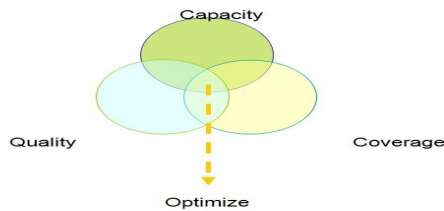


Fig. 2. Radio Network Optimization

C. Concept of RN Optimization

Due to the mobility of subscribers and complexity of radio wave propagation, most of network problems are caused by increasing subscribers and the changing environments. RN Optimization is a continuous process that is required as the network evolves [3].

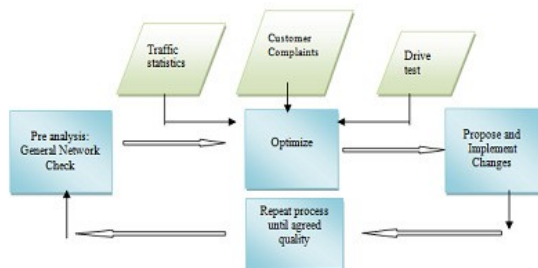


Fig. 3. Flow Diagram of RNO

In the following cases, the network should be optimized.

- New network or expansion on existing network.
- The network quality decreased seriously and there are many complaints from subscribers.
- An event occurs suddenly which affects the network performance seriously.
- The number of subscribers increased and affects the network performance gradually.

II. RADIO NETWORK OPTIMIZATION PROCESS

The process of Radio Network Optimization requires some steps. They are discussed here.

A. Pre-analysis

Pre-analysis basically involves the General Network Check. Following Tasks should be carried out:

- 1) Determine original network planning
- 2) Collect information about network
- 3) Determine functional network structure, e.g.
 - BTS/BSC locations, antenna direction / azimuth, tilting etc.
 - Services and features used
 - Network structure (macro cell, microcell etc.)
 - Determine the network element configuration, e.g.
 - Number of TRX per cell
 - Sector/Omni configuration
- 4) Visit selected sites
- 5) KPI analysis

B. Customer Complaints Analysis

The process of analyzing customer complaints requires some quality targets. Quality targets rely on two facts.

1) The customer's quality expectations are very simple:

- Availability of the service anywhere, anytime
- Call setup time within limits
- Good speech quality during the call
- Normal termination of the call

2) Quality problems are indicated by:

- Poor signal levels
- High locking rates
- High bit error rates
- Dropped calls/handover failures

C. Collect / Analyze OMC Statistics or KPI Analysis

This involves measuring the rate of some data.

- Handled traffic (congestion on TCH, SDCCH)
- dropped calls
- Interference
- Handover reason (due to UL_QUAL, Power budget, distance)

D. Collect / Analyze Drive Test Measurements

During drive test measurement, the following data are required for downlink as uplink is not required for test measurement.

- BER (RXQUAL)
- Serving signal level
- Channel Number
- CI and LAI
- Timing Advance
- Layer 3 messages
- BSICs
- Signal and power levels of neighboring cells

E. Propose / Implement Changes

The following changes are required to improve the network quality.

- Changes requested using standard forms
- Proper approval necessary (signatures)
- Physical change requests
 - ☐ Change antenna direction, tilt, height etc.
- Database change requests for:
 - ☐ Change frequency, and add neighbor etc.
 - ☐ Should be implemented quickly
- All involved parties must be informed

III. DRIVE TEST

In general, drive test is the most important method for radio network optimization. Drive Test can be defined as a method that used to verify the actual condition of RF (radio frequency) signals certain operator at certain place [5]. Drive test requires testing the network in particular area to give the real picture of the network's performance on the field, with a certain tool [3].

A. Functions of Drive Test

- 1) Analyzing customer complaint of certain operator home or office area
- 2) Finding problem in BTS (Timeslot Check, TRX Swap Feeder)
- 3) Analyzing the result of optimization process



Fig. 4. Equipment setup during drive test

B. Software Used for Drive Test

There are software packages can be used for drive test that installed on laptop. For the Drive test basically most used software's are TEMS Investigation (Ericsson), NEMO (Nokia) [3].

C. Drive test equipment

Some equipment used during drive test are listed below.

- 1) TEMS Handset (complete with Charger, Headset, Data Cable) and USB Hub
- 2) Laptop (installed TEMS Investigation) and Adapter
- 3) Inverter and Terminal
- 4) Battery and Charger
- 5) GPS

D. Maintenance requirement for Drive Test

- 1) Power supply needed, usually using inverter in the car from laptop, GPS and MS
- 2) GPS should use external antenna
- 3) MS can use external antenna or not use external antenna.

E. Mode of Drive test

There are two modes of drive test.

1) Dedicated / Continuous / Long Call Mode:

- Making continuous call along drive test activity
- Before starting the route, call the drive test number and only stop the call when the route is finished.

2) Idle Mode

- Along the drive test activity, the MS is "ON" but no call occur

IV. DRIVE TEST PARAMETERS

Several parameters are measured during the drive test. They are briefly described below.

A. Rx Level

Level of Received signal strength. In dBm or Steps. Value in the form of step subtracts 110 to the value to get dBm value. That is Receiving level in terms of dBm that mobile is receiving from the site. Range of -30 dBm to -110dBm [7].

B. Rx Quality

Received signal quality level, measured base on BER (bit error rate). The value is between 0-7, the lower the better.

C. FER

FER (Frame Error Rate) the percentage of frames being dropped due to high number of non-corrected bit errors in the frame. It is indication of voice quality in network.

D. BER Actual

Ratio of number of bit errors to total number of bits transmitted in given time interval [7]. BER is a measure of voice quality in network depends on BER Rx Qual is measured. e.g., BER 0 to 0.2% corresponds to Rx Qual 0. Max. BER countable and useful is up to 12.8 % which corresponds to Rx Qual of max. 7

	RXQUAL class	Mean BER (%)	BER range from... to
Good	0	0.14	< 0.2%
	1	0.28	0.2 ... 0.4 %
Fairly good	2	0.57	0.4 ... 0.8 %
	3	1.13	0.8 ... 1.6 %
Acceptable	4	2.26	1.6 ... 3.2 %
	5	4.53	3.2 ... 6.4 %
	6	9.05	6.4 ... 12.8 %
Intolerable	7	18.1	> 12.8 %

Fig. 5. SQI

E. SQI

The parameter used by TEMS measure Speech Quality. It has been designed to cover all factors that Rx Qual lack to measure. SQI is a more sophisticated measure is dedicated to reflect the quality of speech. SQI computation considers the following factors.

- a) The bit error rate (BER)
- b) the frame erasure rate (FER)
- c) data on handover events
- d) statistics on the distribution of these parameters

F. C/I

C/I (carrier-over-interference) is the ratio between the signal strength of current serving cell and signal strength of undesired (interfering) signal components. It should be at least ≥ 9 and $C/I \geq 12$ for practical purpose. It is useful in the verification and optimization of frequency plans [7].

V. COMMON PROBLEMS OF RN OPTIMIZATION

Some common problems of Radio Network Optimization are discussed here.

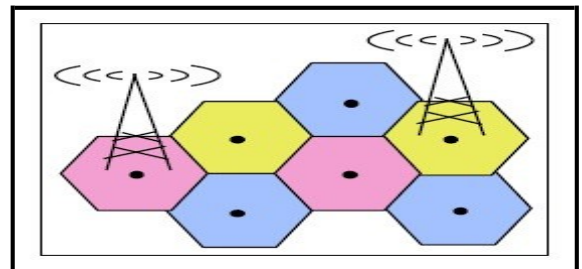


Fig 6. Coverage Problem

A. Coverage Problem

1) Coverage hole

Coverage holes may exist when coverage areas of two BTS do not overlap or there are some big obstacles in the coverage area. It leads to no signal or very poor signal [3].

2) Over shooting

In the actual network, the high BTS antenna can propagate far away along a road and serve in area which it is not suppose to serve, which results in the "Isolate Island" problem.

B. Capacity Problem

Capacity problem involves two items.

1) TCH congestion

Checking TCH congestion rate involves attempted TCH seizer's failure for all on OMC traffic statistics data, or comparing the busy hour traffic of each cell with the calculated capacity in different GoS (Grade of Service) to judge the situation of traffic congestion.

2) Traffic balance

- Adjust the antenna height, down tilt, and change the TX power of BTS and MS.
- Adjust parameters for cell selection, cell reselection and handover
- Enable load handover and direct retry functions.

C. Interference Problem:

Due to a limited spectrum bandwidth in a GSM network, interference of channels is unavoidable. There are two types of interference, co-channel interference and adjacent channel Interference [7].

1) How to locate the interference

- Interference band in TCH Measurement Function
- Receiving Level Measurement Function
- Rx Qual (Drive test)
- Ratio of handovers for uplink/downlink quality
- Call drop measurement function
- Too many handover failure and reestablishment failures Statistics

2) Check and solve the interference problems

- Check the place with interference and conversation quality by drive test.
- Use spectrum analyzer to find out the interference in the frequency band.
 - ☐ Enable frequency hopping, DTX and power control functions.
 - ☐ Check and eliminate equipment problems

3) Means to decrease the interference

- Increase the reuse distance between the co-frequency and adjacent frequencies.
- Decrease the transmitting power of BTS.
- Adjust the antenna height, azimuth and down-tilt.
- Use the anti-interference technologies, such as frequency hopping, power control, and so on.

D. Handover problem

Handover basically means changing the point of connection while communicating. When mobile station

switches from one set of radio resources to another set, handover is said to have taken place.

1) *Handover failure and handover delay lead to bad quality or call drop.*

2) *Frequent handover decreases the conversation quality and increases system signaling load.*

3) *Unreasonable proportion of outgoing inter-cell handover to incoming inter-cell handover results in imbalanced traffic.*

To check and solve handover problems, the following handover parameters (OMC) should be checked.

- Inter-cell Handover Measurement Function
- Undefined Adjacent Cell Measurement Function
- Outgoing Inter Cell Handover Measurement Function
- Incoming Inter Cell Handover Measurement Function
- TCH Measurement Function

E. Call Drop Problem

During the GSM network operation, call drop is the main reason the subscribers complain. Call Drops happen when the mobile is no longer has the signal strength that the tower can detect. Another reason is that bandwidths allocated for that particular call are being used elsewhere. There are many call drop causes, such as poor signal strength, interference and improper parameters.

VI. CONCLUSION

In this paper it has been shown that we can optimize an existing cellular network using optimization tools and fine parameter tuning. Some of the inputs for optimization will be needed from traffic statistics, customer complain, and drive test. This study helps plan operators to enhance coverage and quality and increase capacity of the network. In this analysis, the general process for radio network planning, optimization methods, GSM planning and UMTS (Universal Mobile Telecommunications System) planning have been discussed. Coverage planning and capacity planning are the most important ones in the radio network optimization. They can be independently implemented in the GSM system, while they are correlated in the UMTS system due to the significant role of interference in UMTS.

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Optimizing and Enhancing Performance of Database Engine Using Data Clustering Technique

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Abstract: The sizes of databases are increasing every day. Hence, now days, accessing data in an acceptable time is one of the biggest challenges in centralized database. In centralized databases, the records can be categorized according to the access frequencies; least accessed records (cold data) and most accessed records (hot data). In a study it shows that more than 90% cases query are requested for hot data, and in case of insertion operation, 99% are done on hot data. Thus categorizing of the data set may improve data accessibility. In this paper, we are proposing a data clustering mechanism based on data access frequency. We have considered only the hot data and the cold data. Here we divided the whole database into two separate files. The first file contains only hot data and the second file contains only the cold data. The time period of hot and cold data will vary for different application domains. The database engine will have direct access on the first database file and in case of unavailability of data; the database engine will look for the second database file. Finally, the experiment result shows how and why data accessibility time should outperform than other available data clustering techniques.

Keywords: Large Database Systems, access frequency, hot data, warm data, cold data.

I. INTRODUCTION

During the last two decades, Relational Database Management System (RDBM) has been established as the technology for data store and giving us the capabilities to handle databases size up to terabytes [1]. Our modern information technology allow us to produce even more powerful computers every year, which makes it possible to collect, store, transfer, and combine huge amounts of data at very low costs. Thus, companies and scientific research institutions can afford to build up large archives of documents and data like, numbers, tables, images, and sounds. However, it is still a challenge for efficient querying and exploiting data in an intelligent way from a gigantic database and to represent it in a structured way. It will be almost impossible to handle or access this large amount of data if database operation is not performed in an optimized way or data is not properly

clustered. When the data access routines of that application are not optimized, there is a 90% possibility is that a database based application performs slowly [2]. So far, scientific research works have been done on reducing overhead of database related to memory uses and storage management. And more scientific works have been performed on clustering database from different angle. However, clustering data based on hot/warm/cold (frequently accessed, least frequently accessed and least accessed) is somewhat missing.

From study on centralized databases, it was found that most data or records are stored but rarely or never accessed after a certain time frame. For example in a super shop the sell records are kept in the database but after calculating the year ending, when the reports are produced, these records are not used for any other purposes. Table 1 Shows some of this type of application areas where data are not accessed after a certain time (Time Frame). The data in table 1 is found from different renowned application fields.

Table 1: Data access frequency on hot cold data based on application area.

Application Fields	Time Frame (Days)	Hot Data (%)			Cold Data (%)		
		INSERT	UPDATE	QUERY	INSERT	UPDATE	QUERY
Education	1450	100	95	90	0	5	10
Sales	180	100	90	85	0	10	15
Finance	360	100	90	80	0	10	20
HR	360	100	90	85	0	10	15
E-Commerce	180	100	85	80	0	15	20
Digital Library & Digital Publishing	720	100	80	65	0	20	35
Banking	360	100	85	70	0	15	30
Credit card Transaction	180	100	90	90	0	10	10
Telecommunication	180	100	85	80	0	15	20
Airlines reservation	180	100	90	90	0	10	10
In Percent		100%	88%	81.50%	0%	12%	18.50 %

The percentage of average access of hot data on different application domains can be calculated as following:

$$OH_T = 1/N \sum_{i=1}^N (IO_i + UO_i + QO_i) / 3$$

Where OH_T is average access on hot data, and IO, UO and QO is the insert, update and query operation on hot data respectively. And N is the number of applications area.

Similarly average access of cold data on different applications can be calculated by the following formula:

$$OC_T = 1/N \sum_{i=1}^N (IO_i + UO_i + QO_i) / 3$$

OC_T is the average access on cold data.

From the above equation it can be shown that the average data access on hot data are 89.83% and on cold data are 10.17%. Based on this type of experiences, in this paper, we are proposing a framework where we will priorities the hot data and will give less priority to the cold data. We may also consider those data that resides between hot and cold data as well for future research. The technique we propose is to split the database file into different priority files where frequently used records will reside in highest priority file and the least accessed records will reside in the least priority file. For performance evaluation, we considered only the hot and cold data. The age of the data will be determined by the database admin from the global configuration part of the framework. For our experiment we considered hot data are those, which were accessed within six months and cold data are those which were not accessed within one year. This is because, any record that were accessed within six month will be considered as hot data so after another six month this data will be treated as cold data. The splitting will not affect the overall data access performance because we will keep an index about the data content of each database file.

We will also analyze the performance of a very large database with and without our suggested techniques. We have show a comparison of our proposed framework with existing system in SQL Server databases and found satisfactory performance. Although the technique is used for SQL Server databases but it could be used for most of the existed databases platforms.

II. RELATED WORK

Syedur et. al. [2] discusses the aspects of performance optimization related to data access in transactional databases. They have tried to prevent any possible performance issue that may take place later, by applying their knowledge before or along with their development activity. Their methodology

doesn't focus on preventing rather than trying to recover after the problem occurs.

Bin Li et. al. [3] proposed some optimization principles and methods of oracle9i database capability in the aspects of database structure, SQL Sentence and memory assignation, etc. And the system capability has been enhanced to comparatively great extent by using these above-mentioned methods of optimizing database during the course of system application. Their strategy gave some integrated and further conclusion about optimization of oracle9i database. It involves a lot of work and need often track diversified statistical targets and analyze the cause of capability change.

P. van Bommel et. al, [4] focus on optimization of database schema transformation by evolutionary search or genetic search. A framework was designed to transform conceptual data schemata into efficient internal schemata was presented. They have considered the problem from the viewpoint of searching through the space of all correct, but possibly incomplete internal representations of the conceptual schema at hand. A search strategy was established, based on the use of evolutionary operators.

III. PROPOSED FRAMEWORK

A. Basic Idea of Proposed Method

In this paper we are proposing one of the methods of optimizing data organization according to the age of the data in a database. The idea of this method is based upon the observation that, not all data are used all the time and unnecessarily hamper in the cost of a query. Our proposal is to separate data in a large database according to its age, means the most recent data and the least recently accessed data are kept in different database files. The first database file will contain most recent data and the second database will contain least used data and accordingly the database will be splitted into two separate database file. The working procedure of our proposed framework is very simple. Basically we will partition our database file into two database files, where each file will contain data according to frequency of access of a particular data. Database will be partitioned as DB1, DB2 where DB1 will act as primary database file. Any data may reside in file DB1, or DB2 depends on its use in main database (DB1). Though it is possible that sometime DB2 database files may work as primary database in case a data is not found in DB1 database file. The most recent used data (time frame will be configured by DB admin) will reside in main database (DB1) and according to data access date, data will transfer to

DB2. In our proposed framework (as show in figure 1) we have used multiple processes which will have different level of priority and will run simultaneously. Everything will be controlled from the Application layer. Inside framework engine we put some automatic processes whose job is to analyze the data used recently and according to the frequency of use of data, it will separate and send data to any of these two database files while the database engine is idle.

Our observation from table 1 shows that operation on database file depends on the type of SQL command, for example, insertion operation always hot data, update operation maximum time on hot data and query operation happens on both hot and cold data.

B. Framework Description

Since operation on database depends on the type of SQL commands, the proposed framework will have some components like SQL scanner, which in fact will be used to find out what type of operation it is and accordingly it will act on either of two database file. If it is an insert operation then it will directly access DB1 because 100% insertion operation will be in state hot data, and if it is an update, delete or a read operation, it will search DB1 first and if it is not found there, it will certainly be in DB2. Other components of the framework are shown below:

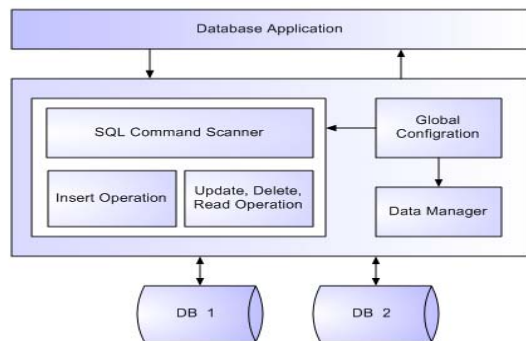


Figure 1: Proposed framework with all its components

1. Database Application:

Database application such as an ERP, Inventory, Airline Reservation System, University Management Application Software, and Industrial Application etc.

2. SQL Command Scanner

As described above the job of SQL scanner is to identify the type of SQL command it is (insert, update, delete or read operation). For insert it will directly hit on DB1 and for other operations it will hit to DB2.

3. Insert Operation:

Insert Operation directly access Primary Database (DB1).

4. Update, Delete, read Operation:

Update, Delete or Read Operation, it will search DB1 first and if it is not found there, it will certainly be in DB2.

5. Global Configuration:

Global configuration will have interface with application layer, where the application developer/user will have options to configure how database file will be splitted, the age classification of data etc. This system configuration will also be used for Database connectivity, clustering, data synchronization with others. The configuration includes:

- a) Database Server Name or IP Address
- b) Database Platform: MS SQL SERVER, MySQL, ORACLE, Others
- c) Database User
- d) Database Password
- e) Database Port (Optional)
- f) Application Deployment Date & Time
- g) Data Age Time Gap
- h) Clustering Mechanism Configuration, etc.

6. Database Manager:

Database manager will keep records of existed data on the database files. Any data exceeds the time frame given in the global configuration will be shifted from one database file to another (i.e. DB1 to DB2 or vice versa). Any data or record which was access from the DB2 will be converted as hot data, because according to our classification mechanism any data is accessed recently will be treated as hot data.

7. DB1 & DB2:

DB1 (Primary Database) & DB2 (Secondary Database) is a Relational database System.

IV. A PROGRAMMING IMPLEMENTATION

Our proposed model has been implemented in programming language by using C++ because increase execution time & reduce system resource. NetBeans IDE 7.4 has been used to write the source code of the persistence layer. NetBeans has been chosen because of its comfortable Integrated Development Environments. NetBeans is developed and maintained by Sun Microsystems. NetBeans has its own file format to maintain the source code. Oracle, MySQL, SQL Server and MS Access was used as the database. The tools used for our purpose are shown in table 2.

Table 2: Software used for developing the framework.

System Development Software	Specification
Operating System Windows XP	TM Professional
NetBeans IDE(C++)	7.4
Database: SQL SERVER	Version 2005
Database Management Studio	Version 2005

Our proposed model requires and certain minimum hardware. Though we have used much higher configured computers for our purpose. Hardware used for our experiments are shown in table 3.

Table 3: Hardware used for implementing the framework.

Hardware and software components	Specification
Processor	Intel Dual Core 2.00 GHz, or AMD Athelon 2800++
Cache	2048 MB
Primary Memory(RAM)	2048 MB
Secondary Memory(HDD)	160 GB
Chip Set	Intel 945

V. EXPERIMENTAL EVALUATION AND PERFORMANCE ANALYSIS

The functionality offered by our proposed model was for Microsoft's windows operating system. For the implementation of our model we have used windows XP operating system in a stand-alone system. In our experiment SQL server 2005 was used as the database server. We have tested our framework with different volume of data. For example we have configured our system to split our database after every three months (depends on global configuration) and each months we inserted 15, 00,000 (one and half million) numbers of data. The figure 2(in worst case) shows the time needed for a single query to execute in existing database and in our proposed framework. We considered the time needed to execute a query as cost for the system.

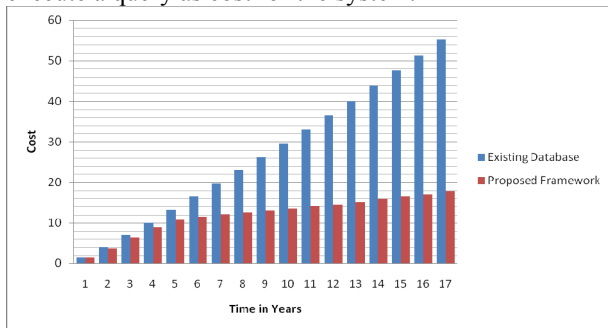


Figure 2: Shows costs in worst case

The cost metric is measured based on time in milliseconds. The clear observation is that the proposed algorithm outperforms on currently existing algorithms proportionally as time passes by. In some cases our proposed system increases database's performance up to 90%. From the above comparison we can clearly see that our framework performs better in all the cases and hence is ideal to be adopted.

VI. CONCLUSION

Idea of splitting an entire database is perhaps a new idea. Of course not all databases will be benefitted from our framework. In process industry, data change continuously with slight jitter [7]. Append operation is much more frequent than query operation, and update operation occurs rarely [6]. In cases where data are stored and after certain time those data are rarely or not accessed at all for long time, our frame will reduce overhead on database engine by reducing volume of data in the primary database. Hence it will increase performance of the database. The above experimental results showing a certain improvement in performance of our system.

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Behavior Analysis of Minimum Mean Square Error Equalizer of MIMO Wireless Channel

Comparison between 2×2 and 3×3 MIMO for BPSK Rayleigh channel

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Abstract— In present days, there has been a race among the communication systems in different schemes and generations. Day by day the data transmission rate is increasing and breaking the record of previous generations. Now, people are thinking of transmitting data beyond imagination. By employing multiple antennas both at transmitter and receiver side, MIMO communication has become most promising wireless technologies that can efficiently boost the data transmission rate, improve system coverage, enhance link reliability, provides very high spectral efficiency. When the data is transmitted at high bit rates, over mobile radio channels, the channel impulse response can extend over many symbol periods, which lead to Inter-symbol interference (ISI) and fading. Fading and interference effects can be combated with equalizer. This paper analyses the performance of Minimum mean square error method in terms of BER for MIMO (2×2) and (3×3) wireless channels. Binary Phase Shift Keying (BPSK) modulation technique was used in Rayleigh flat fading channel. Simulating in MATLAB results show that the equalizer based Minimum mean square error receivers is good for noise free channel and is successful in removing ISI. Successive interference method with optimal ordering outperforms the MMSE however their complexity is higher due to iterative nature of the algorithm.

Index Terms— MIMO, MMSE, BER, MRC, etc

I. INTRODUCTION

The next generation personal communication network will need to be support a wide range of services which will include high quality voice, data, facsimile, still pictures and streaming video. In this case, using multiple transmitting and receiving antennas or to form multiple-input multiple-output (MIMO) channels can resolve the problem [1]. This paper focused particularly on phenomenon of fading and inter-symbol interference (ISI) when data is transmitted at high bit rates. Equalization is a technique used to combat inter-symbol interference.[4] ML is better among ML and MMSE equalizers specifically in case of Binary Phase shift Keying. In [7] some of the detection algorithms are discussed with SIMO and MIMO systems and they are compared based on BER performance. It [5, 6] focuses on the comparison of equalization techniques for Rician Flat fading and Rayleigh frequency selective fading channel. This paper discuss the performances of MMSE equalization technique by considering

2 transmitting and 2 receiving antenna case (resulting in a 2×2 MIMO channel) and 3 transmitting and 3 receiving antenna case (resulting in a 3×3 MIMO channel for a flat fading Rayleigh multipath channel and the modulation is BPSK.

II. RELATED STUDY

A. MIMO Systems Overview

Multiple Input Multiple Output [12] technology is a wireless technology that uses multiple transmitters and receivers to transfer more data at the same time shown in Fig.1. MIMO technology takes advantage of a radio-wave phenomenon called multipath where transmitted information bounces off walls, ceilings, and other objects, reaching the receiving antenna multiple times via different angles and at slightly different times. MIMO technology leverages multipath by using multiple, “smart” transmitters and receivers with an added “spatial” dimension to dramatically increase performance and range [18]. MIMO makes antennas work smarter by enabling them to combine data streams arriving from different paths and at different times to effectively increase receiver signal-capturing power [19]. If there are more antennas than spatial streams, the Additional antennas can add receiver diversity and increase range. MIMO takes advantage of multiplexing to increase wireless bandwidth and range.

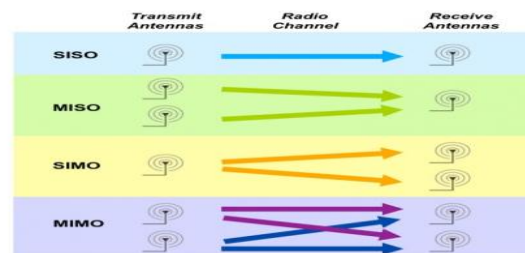


Figure1. Multiple path wireless communication technology to transfer data at the same time. [3]

MIMO algorithms send information out over two or more antennas and the information is received via multiple antennas as well. On normal radio, multiplexing would cause interference, but MIMO uses the additional pathways to transmit more information and then recombines the signal on

the receiving end. MIMO systems provide a significant capacity gain over conventional single antenna systems, along with more reliable communication.

MIMO system is represented by

$$Y=HX+N...(1)$$

Hence, X = Transmitting signal, Y = Received signal

H =Channel matrix, N = Noise vector.

B.2×2 MIMO Channel

A generalized ($M \times N$) channel MIMO wireless system as a block diagram is shown in Fig.2. In a 2×2 MIMO channel usage of the available 2 transmitting antennas can be as follows: Considered that a transmission sequence, for example $\{x_1, x_2, x_3, \dots, x_n\}$. In normal transmission, we will be sending x_1 in the first time slot, x_2 in the second time slot, x_3 and so on. However, as we now have 2 transmitting antennas, we may group the symbols into groups of two. In the first time slot, send x_1 and x_2 from the first and second antenna. In second time slot, send x_3 and x_4 from the first and second antenna send x_5 and x_6 in the third time slot and so on.

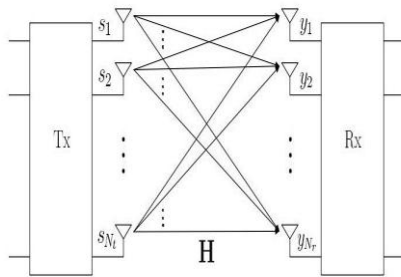


Figure 2.Block diagram of ($M \times N$) channel MIMO wireless system [4].

Notice that as two symbols were grouped and sent them in one time slot, which need only $n/2$ time slots to complete the transmission. Hence the data rate is doubled. This forms the simple explanation of a probable MIMO transmission scheme with 2 transmitting antennas and 2 receiving antennas. This system can be described in generalized beam widths [13].

C.3×3 MIMO Channel

In a 3×3 MIMO channel usage of the available 3 transmitting antennas can be as follows: Consider that we have a transmission sequence, for example $\{x_1, x_2, x_3, \dots, x_n\}$. In normal transmission, x_1 will be sent in the first time slot, x_2 in the second time slot x_3 and so on. However, we have 3 transmitting antennas; we may group the symbols into three groups. In the first time slot, send x_1, x_2 and x_3 from the first, second and third antenna. In second time slot, send x_4, x_5 and x_6 from the first, second and third antenna, send x_7, x_8 and x_9 in the third time slot and so on. Notice that as we are grouping three symbols and sending them in one time slot, we need only $n/3$ time slots to complete the transmission. Hence the data rate is tripled. This forms the simple explanation of a probable MIMO transmission scheme with 3 transmitting antennas and 3 receiving antennas. The three transmitted symbols interfered with each other and we will use minimum mean square error

equalizer to minimize the interference. Several approaches have been proposed for interference management [20, 21, 22].

D. Minimum Mean Square Error Equalizer

A more robust equalizer is the MMSE [10] equalizer where the criterion used is the minimization of the mean square error (MSE) between the desired equalizer output and the actual equalizer output. The MMSE equalizer maximizes the signal to distortion ratio at its output within the constraints of the equalizer filter length [14, 15, 16]. If an input signal has a time dispersion characteristic that is greater than the propagation delay through the equalizer, then the equalizer will be unable to reduce distortion. *Minimum mean square error (MMSE) equalizer for 2×2 MIMO channel*

Let us now try to understand the math for extracting the two symbols which interfered with each other. In the first time slot, the received signal on the first receive antenna is,

$$y_1 = h_{1,1}x_1 + h_{1,2}x_2 + n_1 = [h_{1,1} \ h_{1,2}] \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + n_1 \dots \dots \dots (2)$$

The received signal on the second receive antenna is,

$$y_2 = h_{2,1}x_1 + h_{2,2}x_2 + n_2 = [h_{2,1} \ h_{2,2}] \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + n_2 \dots \dots \dots (3)$$

Where, y_1, y_2 are the received symbol on the first and second Antenna respectively, $h_{1,1}$ is the channel from 1st transmit antenna to 1st receive antenna, $h_{1,2}$ is the channel from 2nd transmit antenna to 1st receive antenna, $h_{2,1}$ is the channel from 1st transmit antenna to 2nd receive antenna, $h_{2,2}$ is the channel from 2nd transmit antenna to 2nd receive antenna, x_1, x_2 are the transmitted symbols and n_1, n_2 is the noise on 1st and 2nd receiver antennas [23]. We assume that the receiver knows $h_{1,1}, h_{1,2}, h_{2,1}$ and $h_{2,2}$. The receiver also knows y_1 and y_2 . The unknowns are x_1 and x_2 . With two equations and two unknowns we can solve it. For convenience, the above equation can be represented in matrix notation as follows:

$$\begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} h_{1,1} & h_{1,2} \\ h_{2,1} & h_{2,2} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix} \dots \dots \dots (4)$$

$$\text{Equivalently, } y = Hx + n \dots \dots \dots (5)$$

The Minimum Mean Square Error (MMSE) approach tries to find a coefficient W which minimizes the criterion,

$$E\{[Wy - x][Wy - x]^H\} \dots \dots \dots (6)$$

Solving,

$$W = [H^H H + N_0 I]^{-1} H^H \dots \dots \dots (7)$$

Using the Minimum Mean Square Error (MMSE)

Equalization, the receiver can obtain an estimate of the two transmitted symbols x_1, x_2 , i.e.

$$\begin{bmatrix} \hat{x}_1 \\ \hat{x}_2 \end{bmatrix} = [H^H H + N_0 I]^{-1} H^H \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} \dots \dots \dots (8)$$

Minimum mean square error (MMSE) equalizer for 3×3 MIMO channel: Let us now try to understand the math for extracting the two symbols which interfered with Each other for 3×3 MIMO channel.

III. RESULTS & SIMULATION

The simulation has been done by Matlab for 2*2 and 3*3 MIMO in respect to Bit Error Rate (BER) and Signal to Noise Ratio (SNR).

BER and SNR for BPSK Modulation with 2x2 MIMO and ML Equalizer (Rayleigh channel)

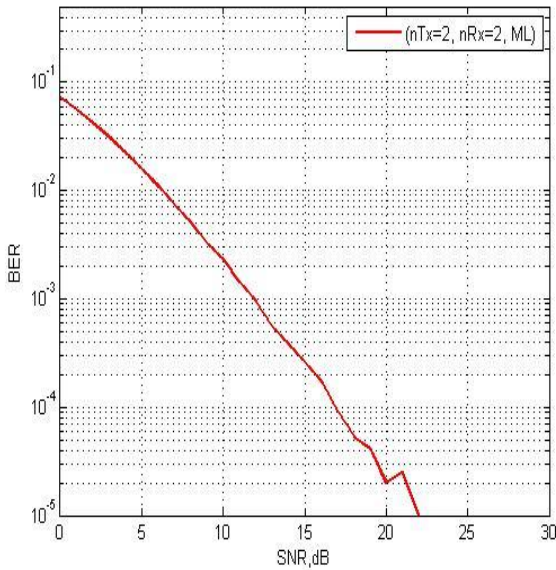


Figure 3: 2*2 MIMO and ML Equalizer

The MIMO and ML equalizer which gives the output of BER (Bit Error Rate) shown above with respect to SNR (Signal to Noise Ratio) (in dB). From the graph in figure 3, it is observed that for the SNR of 22 dB, the BER was close to 10^{-5} which is the best performance for 2x2 MIMO.

BER and SNR for BPSK Modulation with 2x2 MIMO and MMSE-SIC-Sort Equalizer (Rayleigh channel)

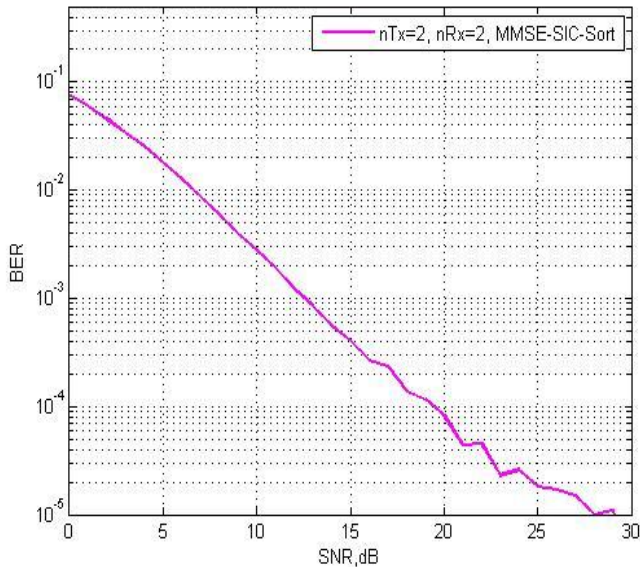


Figure 4: 2*2 MIMO and MMSE-SIC-Sort

From the figure 4, it is clear that the SNR of 28 dB, the BER was 10^{-5} and the SNR is quite higher than the 2x2 MIMO and ML Equalizer.

BER and SNR for BPSK Modulation with 2x2 MIMO and MMSE Equalizer (Rayleigh channel)

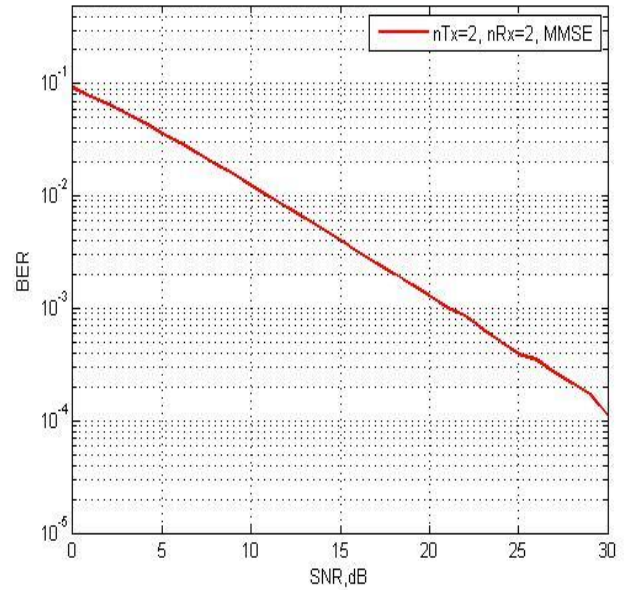


Figure 5: 2*2 MIMO and MMSE-SIC

From the figure 5, it is clear that the SNR of 30 dB, the BER was 10^{-4} and the BER is quite higher than the 2x2 MIMO and MMSE-Sort Equalizer.

BER and SNR for BPSK Modulation with 3x3 MIMO and MMSE Equalizer (Rayleigh channel)

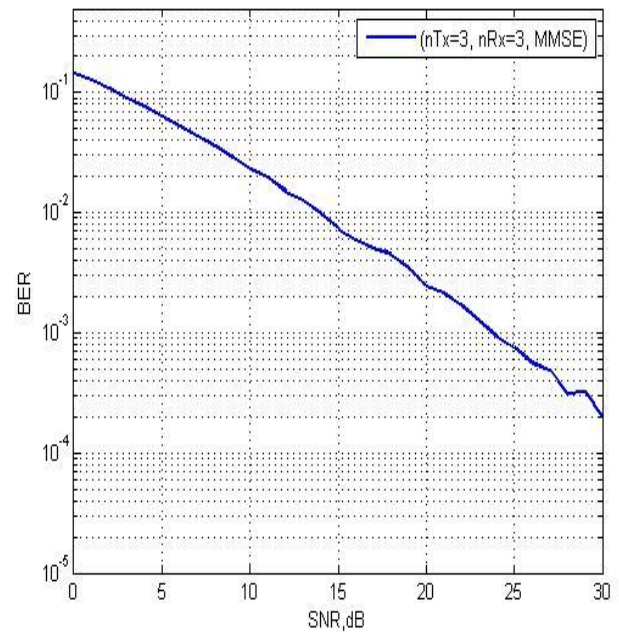


Figure 6: 3x3 MIMO and MMSE

From the figure 6, it is clearly observed that the SNR of 30 dB, the BER was $10^{-3.7}$ and the BER is quite higher than the 2x2 MIMO and MMSE Equalizer. However the BER increase in 3x3 MIMO than 2x2 MIMO.

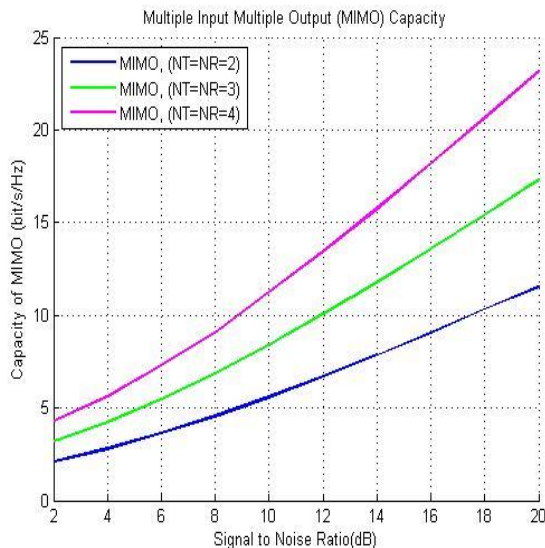


Figure 7: Shannon capacity for 2×2, 3×3, 4×4 MIMO

Figure 7, shows the capacity of MIMO for different NR (number of antenna in receiver) which are varies from 2 to 4. It is clear from the figure that the channel capacity increasing very rapidly with NR and NT (number of antenna in transmitter). For NR=2, the channel capacity is 12 bit/s/Hz when the channel capacity is 17 bit/s/Hz for NR= 3, and for NR= 4, the channel capacity is 23 bit/s/Hz.

IV. CONCLUSION

The BER of 3×3 MIMO increase in a very little amount respect to 2×2 MIMO because the number of bit transmitted in 3×3 MIMO is much more higher than the 2×2 which is quite clear from the Shannon capacity of MIMO and this capacity increase very rapidly for 4×4 MIMO. However the 3×3 MIMO has a very wide range of data transmission, the BER has not any significant effect on this. This rapidly development in MIMO will make the wireless communication more effective and the dream of 5G will not be a dream anymore. So the 3×3 MIMO is much more effective for high speed in wireless communication compare to 2×2 MIMO.

ACKNOWLEDGMENT

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Mobile Phone to Mobile Phone Wireless Power Transfer

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Abstract— Wireless power transfer (WPT) or wireless energy transmission is the transmission of electrical power from a power source to a consuming device without using solid wires or conductors. In this paper we describe the implementation of the concept of wireless power transfer to transmit power from mobile phones to mobile phones. In our design we use simple inductive coupling and basic electrical circuits to implement this technology.

Keywords—wireless power transfer; OTG; resonance; inductive coupling; Colpitts oscillator; voltage regulator

I. INTRODUCTION

In recent times, Wireless Power Transfer technology has opened a window towards a new revolution. It changed the general perception of transferring power. Commonly known as WPT, this technology can ease the way of energy transmission widely. Back in 2006, a group of MIT students came up with this innovative way to light a 60W bulb over two meter distance wirelessly. This amazed the concerning people with a hope to use the old theories in a new way. This noble concept was far foreseen by the “Father of Wireless” Nicola Tesla based Tesla theory, which would transmit the electrical energy across a large distance and receive it at the destination with negligible losses. We used this concept to make an efficient power transfer system to transmit a low voltage power over a short distance. We aim to charge a phone with 5V energy transferred through resonating coils from an On-The-Go (OTG) supported phone. An OTG supported phone can deliver 5V DC output which we intend to use to charge another phone [7].

II. RELATED WORK

Engineers have been striving to find better ways to recharge mobile phones. With the development of modern technologies, they were able to invent a new wireless charging technique. It includes a wireless charging pad which will recharge a device when kept over it. Besides recharging mobile phones engineers have used this device to wirelessly operate lights, laser etc. A group of researchers from, School of Electrical System Engineering, University Malaysia Perlis has published a study similar to our solution [3]. Although, the researchers tried to transfer the required voltage through coupling coils instead of implementing any more circuit on the receiver except rectifier. The study did not include the physical implementation of the idea to charge phones, rather it concentrated in wireless power transfer.

III. WIRELESS MOBILE CHARGING FRAMEWORK

We receive a 5V DC power from an OTG supported phone. This DC power is converted into a high frequency AC signal with the help of a Colpitts Oscillator. The high frequency AC signal is transmitted through a magnetic coil at the receiver end. The high frequency AC input from the coil is passed through a full-bridge rectifier circuit. This converts the AC signal into DC power which is then boosted to 5V using a step-up Voltage Regulator.

IV. THEORY

The basic idea of the system is to take the power from a charged phone and deliver it to another one wirelessly. The wireless power transfer of the system is based on inductive coupling between transmitting and receive coil. The transmitting phone needs to support USB OTG (On-The-Go) to draw the power into the system. USB OTG is a specification first used in late 2001 that allows USB devices such as mobile phones to act as a host, allowing other USB devices such as USB flash drives, digital cameras, mice or keyboards to be attached to them. This allows it to dissipate constant +5V voltage and up to 1.5mA current (depending on USB version and protocols) [6]. We can use this power to serve our purpose.

Fig. 1 shows the basic block diagram where the oscillator generates an AC power and transfers it wirelessly through coupling coils. The AC power is then rectified to DC with a simple full bridge rectifier. The DC power is then converted to 5V DC with a voltage regulator working as boost converter.

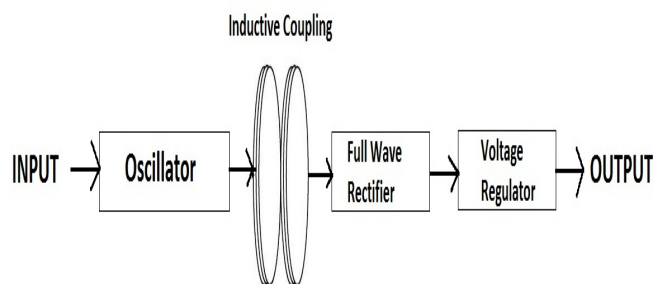


Fig. 1. Block diagram of the system

The idea is to convert the DC supply into a high frequency AC supply with the help of an oscillator, which we need in order to transfer wirelessly [7]. For the oscillator we used a simple Colpitts oscillator. The oscillator is a simple one to

implement and being a current oscillator it is helpful for the system [1]. The AC power is then transferred wirelessly through inductive coupling. In inductive coupling power is transferred between coils of wire by a magnetic field. Fig. 2 shows the block diagram for the wireless power transfer through inductive coupling. The transmitter and receiver coils together form a transformer. An alternating current (AC) through the transmitter coil (L1) creates an oscillating magnetic field (B) by Ampere's law. The transmitting coil (L1) is coupled with the receiving coil (L2) at the same frequency. This allows the magnetic field (B) to pass through the receiving coil (L2). The system induces an alternating EMF following Faraday's law of induction, while creating an AC current in the receiver. One important parameter for coupling is frequency, as the efficiency shows a proportional relationship with frequency for such systems [9].

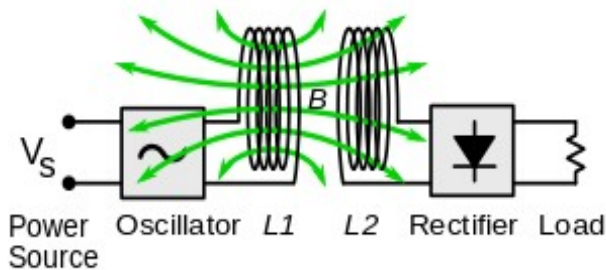


Fig. 2. Block diagram of WPT (Collected from [9])

Both of the coils act like a transformer, using which the power is transferred. The transferred power depends on the coupling frequency and mutual inductance (M) between the coils. As a matter of fact the output power increases with both the parameters, which can be manipulated since M depends on the geometry and the distance D_{range} between the coils. Another important parameter is the coupling coefficient

$$k = \frac{M}{\sqrt{L_1 L_2}} \quad (1)$$

When, the distance between the coils is zero, the parameter $k=1$, which is the maximum possible value for k. It theoretically means the link efficiency is 100%. The lower the value the less efficient our link becomes [9].

In the receiving side, the output of receiver coil is passed through a full bridge rectifier and converted to DC. The DC power is then passed through a voltage regulator to boost the expected low voltage into 5V DC power which is required for the load of the system.

V. SYSTEM SETUP

The setup of the system has been discussed below.

A. Transmitting End

The system will get a 5V DC input from a mobile phone. This power will consequently reach the receiving phone and recharge the phone. So, the DC voltage is passed through a Colpitts oscillator so that it is transformed into an AC signal. High frequency AC frequency is required to transmit the

power wirelessly through magnetic wires. The higher the frequency gets, the better the coupling occur. For the oscillator, NPN transistor is chosen. The resistances and capacitors are connected to a BC-548 (NPN) transistor in the design of the Colpitts oscillator. Multiple transistors were connected in parallel to improve the supply current.

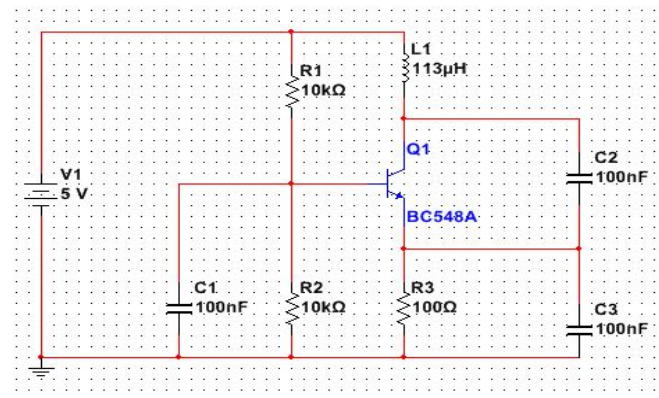


Fig. 3. Transmitting end

B. Coupled Coils

The transformed AC voltage is passed through a magnetic coil. For this system, a 24AWG Magnetic Enamelled Wire have been used. The diameters and number of turns of both the transmitting coil and receiving coil are chosen in such a way that they both resonate at the same frequency. The diameter and number of turns were arbitrarily chosen to be 15cm and 20 turns respectively. The coils were then separately connected to the transmitting end or Colpitts oscillator. The number of turns of both the coils were varied to get the expected outcome. When both the coils resonated at the same frequency, then the receiving part was accordingly. The chosen frequency was 66 KHz approximately. To find the inductance of the coils, the following formula were used:

$$freq = \frac{1}{2\pi \sqrt{ind \times \frac{49}{2}}} \quad (2)$$

$$ind = \frac{\left(\frac{1}{freq \times 2\pi}\right)^2}{\frac{49}{2}} \quad (3)$$

Thus the inductance was found to be 113 μH.

C. Receiving End

The receiving coil was connected in parallel to the coupling capacitors of 100nF. These capacitors helps the coils to improve coupling co-efficient.

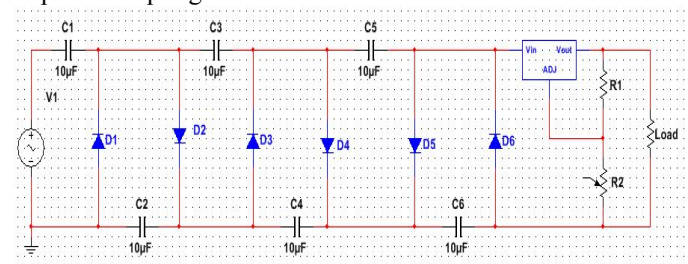


Fig. 4. The receiving end

In fig. 4, the AC signal from the coil is passed through a Cockcroft Walton (CW) voltage multiplier. A CW multiplier generates a high DC voltage from a low voltage AC. Hence, it can rectify the received AC voltage, as well as, step up the relatively low voltage. This is a classic multistage diode/capacitor voltage multiplier, which is economically efficient at the same time. The stages can be increased or decreased with ease to get desired output. The final voltage output of the stages is given by (4).

$$V_o = 2NV_p = NV_{pp} \quad (4)$$

where N is the number of stages.

This DC voltage is then regulated with a Voltage Regulator circuit. The idea of this regulator is to control the output voltage while changing the value of R1 and R2 from fig 4. The voltage can be controlled by using (5).

$$V_{out} = V_{ref} * \left(1 + \frac{R_2}{R_1}\right) + (I_{adj} * R_2) \quad (5)$$

where Vref is the voltage between Vout and adjustment pin and Iadj is the current through adjustment pin.

The voltage regulator has been chosen to keep the system simple and upon simulation results. Depending on the practical scenario which might vary, as there are lots of boost converter mechanisms to boost and control the DC power.

VI. RESULTS AND DISCUSSION

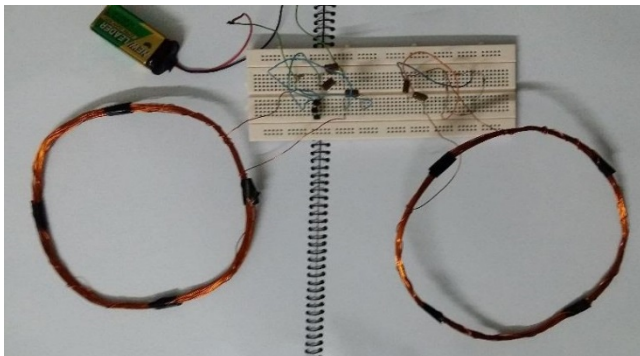
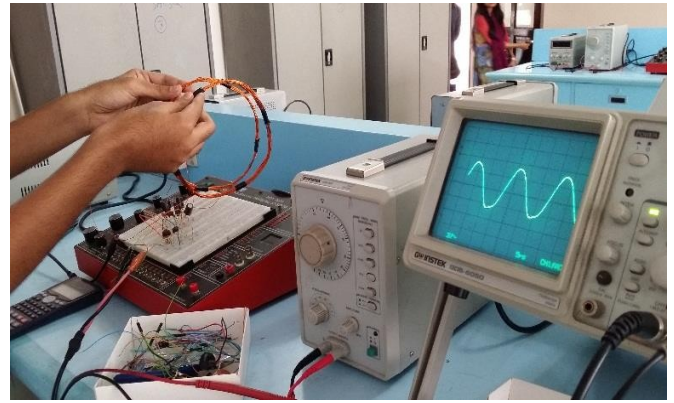


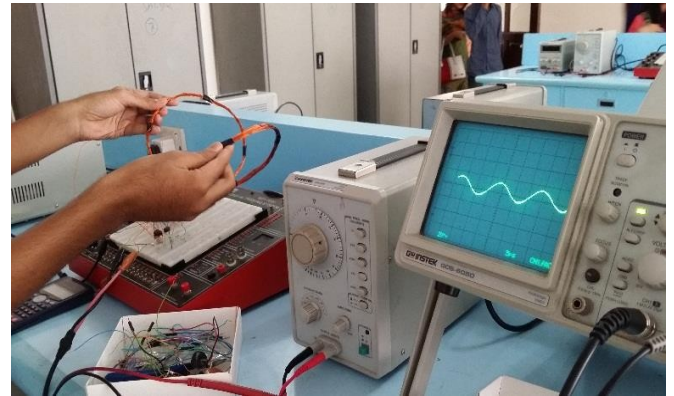
Fig. 5. Practical implementation of receiver and transmitter (Using 5V DC battery)

Fig. 5 shows the practical circuit design. The left portion of the design is the transmitting end and the right portion is the receiving end. Here, the voltage regulator circuit is not implemented.

The output of the receiving coil was measured in the lab in various circumstances. We took 5V DC from the power supply as our input power of the transmitting end.



(a)



(b)

Fig. 6. AC output across receiver (a) coils are near each other (b) Coils are taken further away

Fig. 6 shows how the change in distance between the coils changes the output AC. As we can see when the coils are closer the amplitude is better. The closer the coils, the better the efficiency. At approximately, 2 cm distance we got 3.4V AC output from the oscilloscope reading. Whereas, the RMS value measured was 3.26V.

This paper has been done as the project is ongoing and thus the results are only from the parts and experiments we have conducted so far. The voltage regulator has not been implemented here but we expect to have our desired output as it was simulated successfully.

VII. FUTURE WORK

There is a possibility that the Voltage Regulator used, might not meet the requirements of the system. Very high resistance value is required to attain the 5V DC output using this regulator. Although, similar topologies using MAX756 or MAX757 IC might be an optimum solution to this problem. It steps up voltage within 0.7V – 5.5V to 2.7V – 5.5V [11]. MAX757 can be adjustment according to the requirement, whereas MAX756 gives fixed 5V DC output.

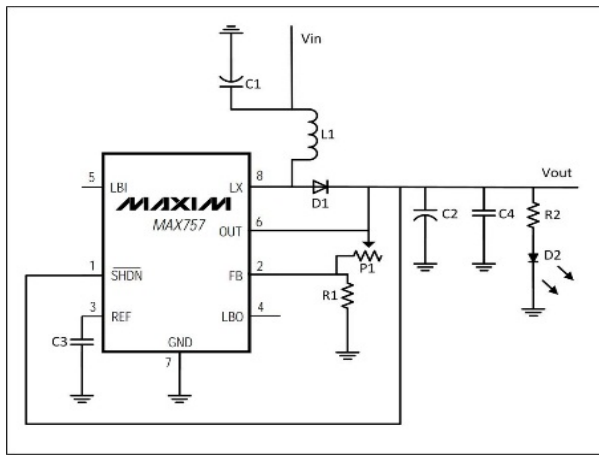


Fig 7 A 5V DC boost regulator using MAX757 IC

In future, this system can be built within the phones and thus no external device will be required to enjoy its service. The coil can be built using PCB boards which can be installed within the mobile phones. Two such phones will then be able to share the power within them using this technology.

VIII. CONCLUSION

In this paper, we have described a mechanism to transfer a low power DC voltage from one mobile to another for the purpose of charging the phone. We have dealt with coupling coils and their resonance. Colpitts Oscillator, Full-Bridge Rectifier and Voltage Regulator circuits have been implemented. This project gives us a new and convenient option of recharging our phones at times when charging from a phone remains the only option because of the lack of having proper charging outlets. This simple idea is a feasible and inexpensive one for mobile charging.

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This work is done as part of an undergraduate Capstone Design course in the Department of Electrical and Computer Engineering of North South University, Dhaka, Bangladesh. We cordially thank and gratefully acknowledge our supervisor Dr. Hasan Uz Zaman for his helpful suggestions and directions, which have improved the work. We also thank the Department of Electrical and Computer Engineering of North South University for giving us the opportunity to work on this project.

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Observation of the Common Phenomena from a MATLAB Model for M/M/1 Queuing System

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Abstract—Queuing theory was introduced for the first time about 100 years ago. Currently it is the key for internet trafficking and network modeling. For analyzing the viability of any specific model queuing theory is processed carefully while manipulating the different parameters to extract the best combination in real-time. A basic M/M/1 (a queuing system with Poisson arrivals, exponentially distributed service times and a single server) model explains how different entities react for variation of parameters. A MATLAB model for M/M/1 queue is taken to observe these changes.

Keywords—M/M/1 queue; packet scheduler; buffer; MATLAB model; network modelling

I. INTRODUCTION

For network packets trafficking at the switching centre, packets are stored in the input buffer if the link capacity is exhausted and they keep waiting for getting the service. Based upon this, various queues and buffers are classified. The scheduler governs the scheduling operations of the stored packets in the buffer and transmits them over the specified link in the network. The selection process of which flows to be transmitted must be taken within a short period of time. Per flow has the provided *QoS* (quality of service) parameter. For easier implementation the associated algorithms of the buffered scheduler must be simple.

The easiest way to understand the queuing concept is to start manipulating various parameters of M/M/1 system and observe the effects. The proposed MATLAB model is best fitted for analyzing these phenomena and relating the outcomes with theoretical features.

II. M/M/1 QUEUE

A. Theory

M/M/1 queue is basically the simplest queuing system. The Kendall's notation M/M/1 refers to a specific system [1]. First M refers to the arrival statistic which is Poisson; the second M refers to the departure statistic rate of service time which is exponentially distributed and the last 1 means that there is a single server in the system [2].

Sometimes the departure rate is [2] also Poisson distributed. This basic queue is of infinite length. The arrival rate is noted with λ and μ represents the departure rate. If the system has no

packet then the probability to have a single packet in the next time interval is $\lambda\Delta$ and this is applicable for all states transition.

If there is a single packet in the server then the probability that the packet is being served is $\mu\Delta$ and this is applicable for all states transition. This can be expressed by the state diagram in Fig. 1.

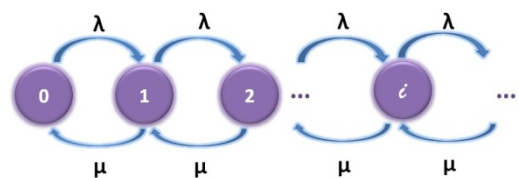


Fig. 1. State Diagram for M/M/1 Queue [3]

The arrival packets in the queue [3], the service rate of the server and the behavior of randomly selected packets are assessed in the system model. The single server does not work continuously. Sometimes it serves the packets and sometime it remains idle when there are no packets arriving in the queue. So the active working time [4] fraction is represented by ρ (traffic intensity that implies the single server utilization) where $\rho = \lambda/\mu$. The value of ρ is always less than 1; otherwise the length of the queue will be exploded.

B. MATLAB Model for M/M/1 Queue

The basic MATLAB model for M/M/1 queue [5] is presented in Fig. 2.

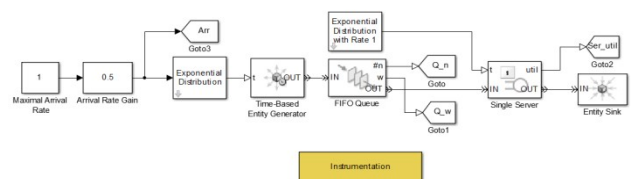


Fig. 2. MATLAB Model for M/M/1 Queuing System [5]

The *Instrumentation* block holds the background model of the system which is given in Fig. 3.

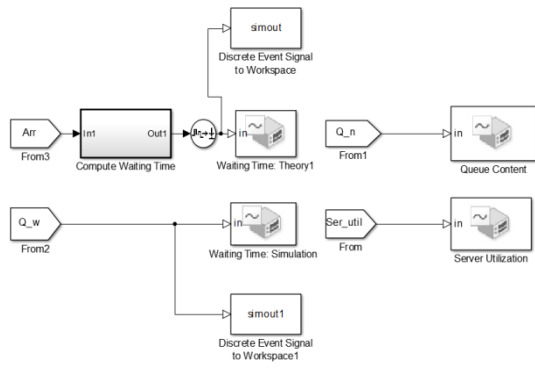


Fig. 3. Background Model for M/M/1 Queue System [5]

The background model represents the model for computing the built M/M/1 [5] MATLAB model simulation results with the theoretical values. The Computing Waiting Time block holds the mathematical operations for calculating the waiting time with service rate 1. The block is digressed below in Fig. 4.

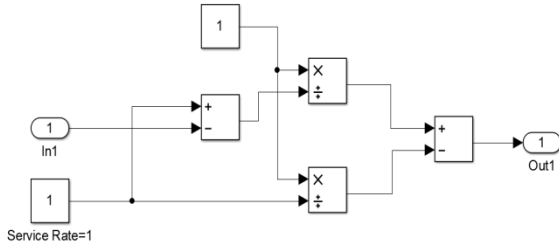


Fig. 4. Computing Waiting Time Operational Block

III. SIMULATION RESULTS OF THE BIULT MODEL

A. Comparison of Simulated Results for Different Arrival Rate Gains

This Section contains the graph that shows the comparison of simulated results using various Arrival Rate Gains. This graph gives the idea that how the outputs of the different mean packet waiting times evolves and settles down as the simulation time progresses.

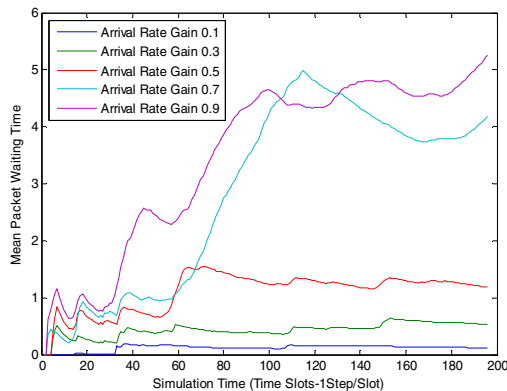


Fig. 5. Comparison of Simulated Results for Various Mean Packet Waiting Time

B. Probability Mass Function Plots of Queue Content for Different Arrival Rate Gains

This section contains the graph of probability mass function (*pmf*) of time varying queue content for various Arrival Rate Gains. The aim of all the *pmf* plots of time varying queue content for different mean packet waiting times is to show that how the density of data packets in the queue are increasing with larger values e.g. as load tends to 1 gradually.

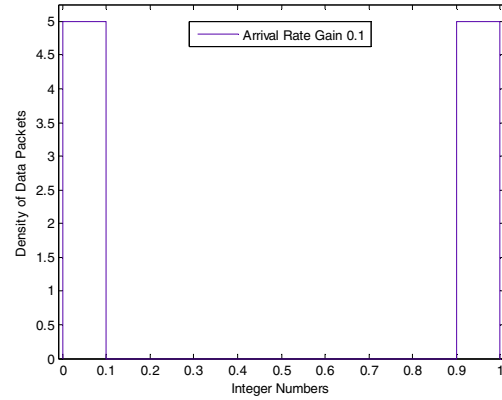


Fig. 6. *pmf* for 0.1 Mean Packet Waiting Time

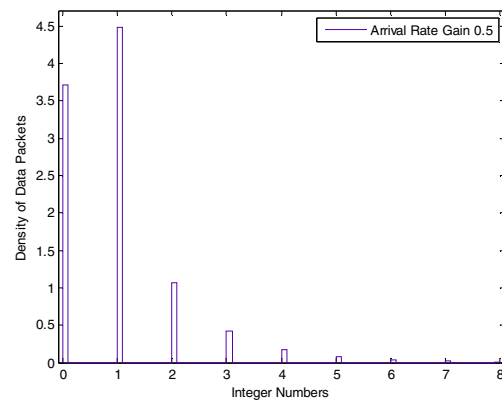


Fig. 7. *pmf* for 0.5 Mean Packet Waiting Time

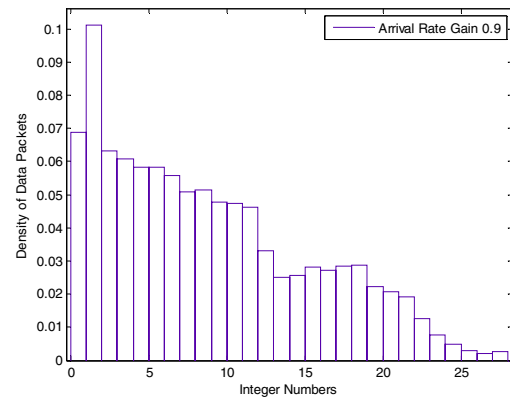


Fig. 8. *pmf* for 0.9 Mean Packet Waiting Time

IV. OBSERVATIONS

As the arrival rate is λ and the service rate/departure rate is μ for the M/M/1 queuing system model hence the theoretical value [6] gives that the single server utilization is λ/μ and the mean waiting time in the queue [7] is $[1/(\mu-\lambda)] - [1/\mu]$ where

First term $1/(\mu-\lambda)$ indicates the total waiting time in the system [8]

Second term $1/\mu$ [8] indicates the mean service time of the system

The built MATLAB model gives the empirical results that match with the theoretical results of M/M/1 queuing system. It is observed that:

- As the Arrival Rate Gain is being increased the time varying queue content, server utilization, mean waiting time in the queue [8] show much bigger variations.
- Increase in load also points out that more time is required for settling down the simulation results.
- The above mentioned issues are of basic importance for queuing theory.
- The reason for that is increase in load means the arrival rate λ tends to 1 and it gradually approaches to service rate μ that makes the active working time ρ bigger. This ρ is also called the offered load of the system. As ρ tends to become 1 then the single server always remains busy and a much longer queue is build.
- As the queue length tends to become infinite then the time varying queue content and mean waiting time of the M/M/1 model show much bigger variations for longer buffering time because then mean queue content becomes $N=\rho/(1-\rho)$.
- For the same reason settling time is also increased for as load tends to 1 e.g. longer runs are needed for higher load.

All these variances and changes are very common phenomena in queuing system.

V. CONCLUSION

The main purpose of this MATLAB model is to simulate the different random entities in M/M/1 queue and check whether the outcomes match with the theoretical assumptions or not. The aim is accomplished successfully.

It has been found that the common phenomenal findings exactly predict the distribution relations that are assumed in theoretical derivations.

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Implementation of Remote Laboratory for Engineering Education in the field of Power Electronics and Telecommunications

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Abstract— This study focuses on the implementation of web based low cost yet effective remote laboratory system for engineering education to utilize limited resources for improving engineering student's practical knowledge as well as self learning ability. The proposed system provides practical experience to the students in the field of power electronics and telecommunications. This lab allows the students to perform experiment by controlling various experimental apparatus on real time from their home or any preferred location using the internet connection. Several experiments related to Power electronics and Telecommunications have been designed for this remote lab. The system is flexible and the range of laboratory experiments can be expanded as required without any complexity. This initializes an openly integrated environment which facilitates the sharing of not only hardware resources, but also educational material for enhancing engineering education in developing countries like Bangladesh.

Index Terms— Pulse width modulation; Duty cycle; Virtual instrument; G-server; Remote Laboratory; Modulation Techniques; DAQ.

I. INTRODUCTION

Due to rapid growth of internet and software technologies, simulation based laboratories have been presented to the engineering field. As these simulation labs are incapable of providing practical experience, remote laboratories have been introduced which are based on controlling and operating real apparatus in real time and have the capabilities to provide practical experience to the students. The importance of remote laboratories has been discussed and presented in many journals [1-5].

A number of individual remote laboratories have been implemented in many universities around the world for power electronics and Telecommunications education. Although these remote lab systems seem less expensive compare to the iLab of MIT, the British Open University, the Swiss Federal Institute of Technology, NetLab of the University of South Australia [2-5]. But it is still highly expensive for developing countries. For power electronics, these systems have used expensive PWM signal generator, programmable DC power supply, fully assemble readymade or semi readymade experimental modules [6-8]. In order to control various input source an external module NI ELVIS which is quite expensive is used. For acquiring data, expensive data acquisition devices (DAQ) have been deployed. However, these systems do not have ability to change various parameters during experiment such as duty cycle. As a result,

users are unable to change different parameter during experiment. User has to stop the experiment to change the parameters. For Telecommunications, again these systems have used Emona DAtEx board with NI ELVIS [9] for acquiring and visualizing data. With addition of such an external instrument the overall cost of the remote lab system increases significantly.

Our proposed remote laboratory system is cost effective and is in integrated form, providing both Power electronics and Telecommunication related experiments. The system is low cost and aimed to provide students a real lab like environment where they are able to operate real instrument. This remote lab provides an opportunity for students to perform their experiment individually, understand what they are doing and relate the theoretical knowledge with practical experience. For Power electronics remote laboratory we have designed the entire experimental module by our self. These modules include different dc-dc converter and dc to ac inverter. These modules have been deployed in remote lab for various power electronics based experiments. For generating PWM signals with variable duty cycle, low cost PIC microcontroller has been used. General DC power supply coupled with servo motor is used to control and generate variable input voltages for power electronics modules. For Telecommunications remote lab, we have selected Emona 101 trainer board which has the same capability as Emona DateEX and we do not need an external module like NI ELVIS. In order to acquire and visualize data for both laboratories, we have not used any kind of expensive analog data acquisition cards or external module. Instead we have utilized our laboratory resources such as digital oscilloscope and microcontroller to acquire and visualize data. Our oscilloscopes have higher sampling rate and fast response time yet less expensive than those external modules or DAQ. Moreover, our designed remote laboratory systems have the ability to change various parameters on the fly, so students do not need to stop the experiments to change the parameters. LabVIEW [10] along with MATLAB [11] are selected as programming environment to design application interface of the remote laboratory system.

II. ARCHITECTURE OF THE REMOTE LAB

Our developed remote laboratory system is based on client/server architecture where all the experimental modules are connected on the server side through application interface. The general architecture of the remote laboratory system is

shown in figure 1. The whole system is divided into three parts:

- 1) *Authentication server*: This server contains all the necessary information of students such as ID, valid password and booked time slot. When the verification is done successfully it connects the clients to the application interface for conducting experiments.
- 2) *Application interface*: All the experimental hardware or module connected to it through controlling and acquiring module. Application interface helps the client to control, change different parameters of the experiments on real time through graphical user interface. It also contains a HD web camera so that clients can see the physical hardware and its response on real time on the web browser.

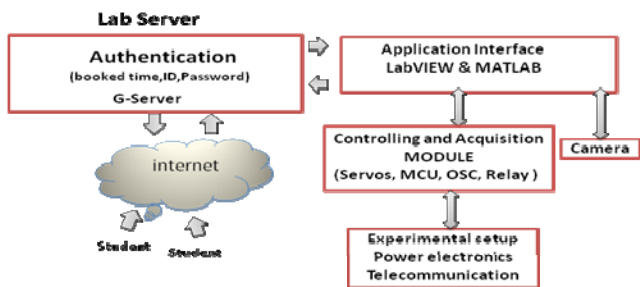


Fig 1. General architecture of the remote lab

- 3) *Controlling and acquisition module*: This module contains the necessary hardware to control and acquire the data through application interface by the client. This module includes low cost servos, relay matrix switch for connecting different points, microcontroller (MCU) and digital oscilloscope (OSC).

III. OPERATIONAL PRINCIPLE OF THE REMOTE LAB

Authorized students who booked time slot before and have valid ID and password can access the lab server through web browser using internet connection to perform the experiment. We adopt time division approach as our experimental module cannot allow parallel simultaneous connection for different users. A link is provided to the students to perform the experiment. When a student goes to the link a login window will be prompt for username and password. After login successfully a student sends a request to the G-server. The server processes the request and hand over the control to the student. On the client computer side, LabVIEW runtime engine which is a free web browser plug-in must be installed in order to control and view the VIs properly. After gaining the control students can perform the experiment and observe the result by changing different parameters from the front panel of the VI. Students can view the system by pressing the ‘camera view’ button. When a student finishes experiment he/she should save the acquired data by pressing the ‘Save data’ button and enter email address in the ‘email address’ box to get the acquired data by email. These buttons are found on the front panel. Data including graph will be sent in MS Excel format as shown in figure 2(a,b). Finally, at the end of allotted time student must release the control of the VI and log off.

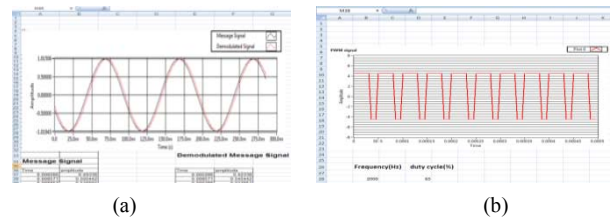


Fig 2. Data (a) FM and (b) PWM signal generation with variable duty cycle experiment received from the email

IV. POWER ELECTRONICS LAB SETUP

In this section, dc-dc power converter buck and boost are given as an example for power electronics remote lab experiment. The experimental setup is shown in figure 3.

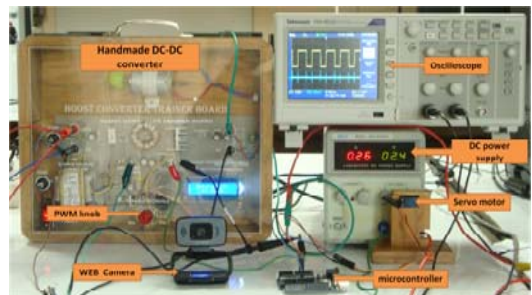


Fig 3. Experimental setup for Power Electronics Lab

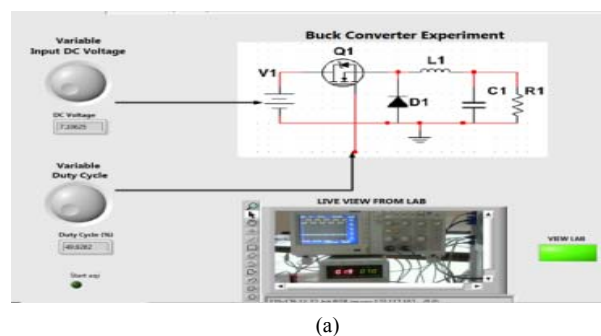
A. Buck Converter

As stated earlier we designed the main dc-dc buck converter by using our available resources. In order to drive the MOSFET of the converter we used low cost microcontroller to generate pulse width modulated (PWM) signal. The detail specifications of the converter are given in the table 1.

TABLE I
SPECIFICATION OF THE CONVERTERS

Converter Type	Buck	Boost
Input DC voltage	3v-15v	2v-5v
Switching frequency	20 KHz	20 KHz
Duty cycle	25%-75%	25%-75%
Filter Capacitor	200uF	200uF
Filter inductance	1mH	5mH
Load resistance	10 ohm	100 ohm

We designed and published the dc-dc buck converter for remote experiment. After login, the following panel will be shown on the browser for dc-dc buck converter experiment as shown in figure 4(a).



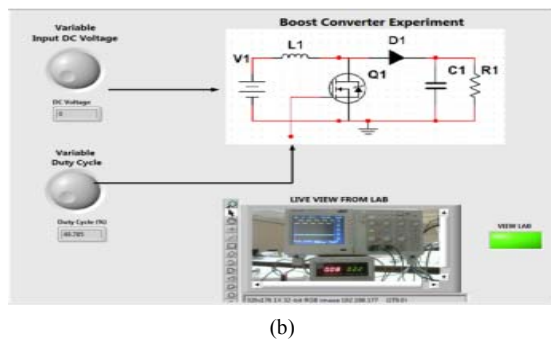


Fig 4: Front panel for (a) buck (b) boost converter experiment

From the front panel students can change different parameters and observe different numerical and graphical data as shown in figure 5(a,b).

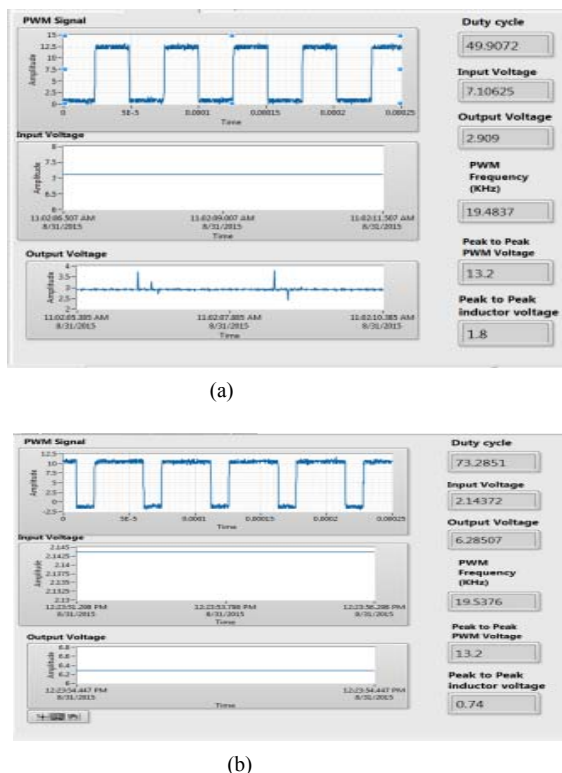


Fig 5. Data analysis for (a) buck (b) boost converter experiment

Some of these parameters are described below:

- Variable input dc voltage: Students can change dc input voltage from 3 volt to 15 volt with this option.
- Variable duty cycle: duty cycle can be changed from this option.
- PWM signal: From the observe tab student can visualize the generated PWM signal on the waveform graph.
- DC output: output voltage can be visualized with this option. This is important as student can see the output dc voltage has spikes on it as shown in figure 5(a). This spike occurs due to high speed switching of the MOSFET and improper filtering. This helps the students about practical situation which they cannot observe through simulation.
- Peak to peak inductor voltage: The voltage across the inductor can be seen with this option.

- Peak to peak PWM voltage: Generated PWM signal voltage can be measured with this option
- Data acquisition: From this tab students can acquire the experimental data and save it to database along with their id.
- Email: The saved data can be received by email from this option.

B. Boost Converter

We designed and published the dc-dc boost converter for remote experiment. After login, the following panel will be shown on the browser for dc-dc boost converter experiment as shown in figure 4(b). From the front panel students can change different parameters and observe different data as described earlier. Specification and analysis part of the boost is shown in table 1 and figure 5(b).

V. TELECOMMUNICATION LAB SETUP

In this section, Amplitude Modulation and Frequency Modulation are given as an example for telecommunication remote lab experiment. The experimental setup is shown in figure 6.



Fig 6. Experimental setup for Telecom Lab

A. Amplitude Modulation

After login the following panel will be shown on the browser for AM experiment as shown in figure 7(a).

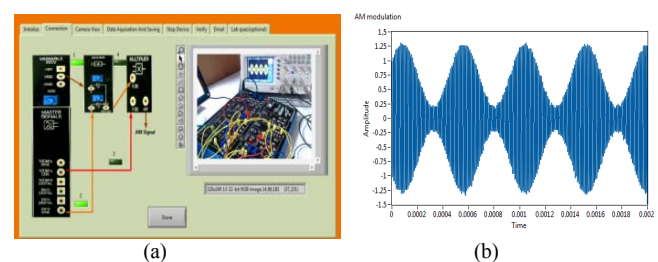


Fig 7. (a) Front panel for AM experiment. (b) Data acquisition for AM experiment

For AM experiment some of the parameters are same as mentioned earlier. A new parameter which is included for AM experiment is given below:

- Connect different point: Students can connect various points according to the experiment by simply turn on/off different buttons. These buttons send the signal the microcontroller and microcontroller drive the relay to make connection at different point. The data acquisition tab is shown in figure 7(b).

B. Frequency Modulation

We designed and published the FM for remote experiment. Students can observe and change parameter as stated earlier. The front panel and data acquisition tab is shown in figure 8(a,b).

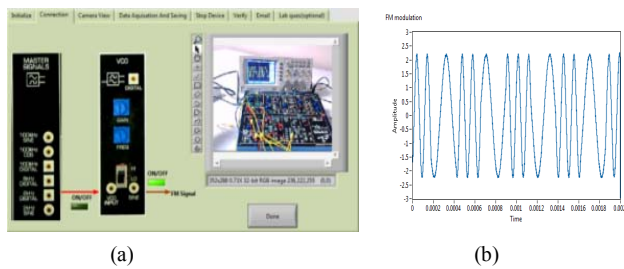


Fig 8. (a) Front panel for FM experiment (b) Data acquisition for FM experiment.

VI. SAFETY AND OTHER ISSUES

All kinds of safety and other issues are taken care of properly. If any kind of error or exception occurs then the server takes over the control and reset the experimental setup. For example if a student leaves in the middle of the experiment then the server detects this and terminates the program. Besides with server panel manager we can monitor who is currently performing the experiment and who is waiting and also the overall traffic. We can also disconnect student if any kind of emergency arises. In addition, a student cannot get more than allotted time to perform the experiment. If a student does not log out within the allotted time then the server will automatically disconnect the student and hand over the control to the other students waiting for the experiment.

VII. FEEDBACK

In order to study the feasibility and usefulness of the laboratory a group of 20 students have been asked to perform experiment using remote laboratory. Proper manual and instruction was also provided to the students. This survey was necessary to verify whether there is any problem or not and also overall evolution from the students. The following questions have been asked to students for the survey.

- Q1. Was the remote lab easy to use?
- Q2. How was the instruction manual provided with remote lab?
- Q3. How was the user interface?
- Q4. What is your opinion about perspectives of Remote lab in Bangladesh?
- Q5. Is the remote lab an effective way of teaching?
- Q6. Did you find any difficulties during experiment?
- Q7. Was the allotted time for experiment enough for you?
- Q8. Is the remote lab effective for you in studying practical engineering courses?
- Q9. Give an overall evaluation about remote lab.

From the figure 9 we can see it reflects a positive sign about remote lab from the students. We can see that allotted time was not enough for the students (Q7). Two students did find difficulties during experiment (Q6). When we asked them about the problem we found that they did not read the

manual properly. Students did find remote lab as an effective way for their study (Q8).

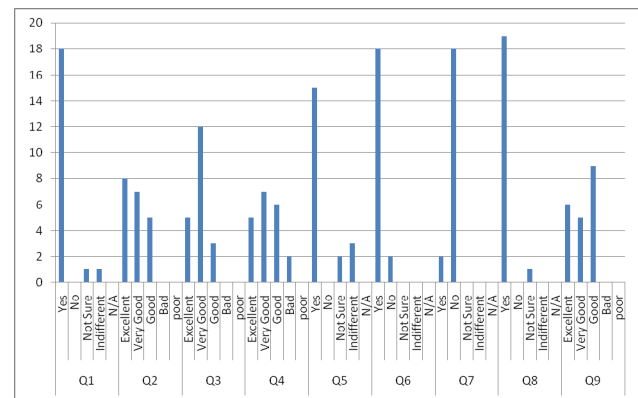


Fig 9 Students survey result

VIII. CONCLUSION

The model of a cost effective and feasible remote laboratory for power electronics and telecommunications was presented in this paper. Remote laboratories may not be a substitute of traditional laboratory system. It is an add-on with traditional lab system by which we can utilize and share our limited resources with each of the students. This can open opportunities to enhance class room lectures, create cooperation among other educational institution in sharing equipment or lab facilities and extend the learning process in the developing countries like Bangladesh.

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- [10] <http://www.ni.com/labview/>
- [11] <http://www.ni.com/labview/mathscript>

Simulation of DC to DC Converter with Embedded Snubber Circuit Maximum Power Point Tracking

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Abstract—This paper presents the proposed model and simulation of a DC to DC converter use an active snubber circuit and maximum power point tracking (MPPT) using fuzzy logic controller (FLC) for a standalone Photovoltaic (PV) System. The function of the circuit is to minimize power losses during switching operation and to improve efficiency of voltage conversion. FLC is used for tracking the maximum power point of the PV array. It will control the ON/OFF time by generating the pulse width modulation (PWM) to trigger the gate of the IGBT. The proposed DC to DC converter is designed by using the Multisim software while the MPPT controller is modelled using the Matlab Simulink software. The performance of the proposed model is tested using simulation and the result show that our proposed converter can produce 130% increasing on output voltage. By using the fuzzy logic method to track the maximum power of the PV array, it is faster and the voltage is stable.

Keywords— Photovoltaic (PV);DC-DC boost converter;Maximum power point tracking (MPPT);Fuzzy logic;Pulse width modulation (PWM)

I. INTRODUCTION

Photovoltaic (PV) is a renewable energy that is widely used. It is effective, harmless and reliable. However, two major setbacks which are high installation cost and low efficiency. The condition of weather and condition affect the stability of the system. Since PV system is exhibit nonlinear V-I characteristic, the power output changing based on the nature of the connected load. In PV system, DC-DC converter acts as front end power electronics which is connected between PV array and the load (1).

DC-DC boost converter is a step-up DC-DC converter which has a simple circuit design that consists of a few components. It is easy to design and develop just using only small and inexpensive components. Conventional boost up converter operates in hard switching where the components change dramatically from high value to zero or vice versa thus results in less efficient. Various soft switched methods has been implemented and the most of the circuit use passive or active snubber (2). Snubber circuit not only offer

the soft switching but it also offer a function to reduce the adverse effects of the reverse recovery characteristic of the boost rectifier on the conversion efficiency and electromagnetic compatibility (EMC) (3).

In a PV system, a technique called maximum power point tracking is applied to track the maximum power of PV array. As solar cell is nonlinear source, the characteristic of the system always changing depends on the temperature and irradiance. It is become a complicated problem to track the maximum power point of the system. To overcome this problem, a MPPT controller is needed to generate the pulse width modulation (PWM) signal to trigger the switching gate of DC-DC converter by specifying optimum duty cycle in order to locate the nearest maximum power point of the PV system (4).

Various techniques have been introduced and used to track the maximum power point of the system as presented in (5-8). Since these techniques determine the estimate values of load voltage and current by moving the functioning point along the I-V curve of the PV panel, it takes longer time to track the maximum point (6). Most simple and commonly used technique is P&O method because of its simplicity to implement but the main problem with this method is oscillation. Conductance incremental method requires a complex control circuit thus it is not an option to use this method for tracking the maximum point of the PV panel (5).

This paper presents an efficient MPPT controller for PV systems by incorporating soft switching high step up boost converter with fuzzy logic based rules for tracking the maximum power that generated by the PV array. The proposed model of DC-DC converter is designed using Multisim software by implementing active snubber circuit to reduce and minimize the switching losses thus increase the voltage conversion of the converter. The model of PV system is develop using Matlab/Simulink and fuzzy logic technique is use to track the maximum power point of the PV system where the output signal of the fuzzy logic controller set the duty cycle of pulse width modulation (PWM) boost converter.

II. SOFT SWITCHING DC-DC CONVERTER

The schematic diagram for the MPPT controller for the PV system showed in "Fig.1". From the left, the system contains a PV panel, DC-DC boost converter, MPPT Controller and DC load. The boost converter is the front-end power electronic device for the PV system. It is the middle system which connects the PV panel to the load. It consists of a switching component, an inductor as an energy storage component, an output rectifier diode and a filter capacitor element. Filter made of capacitor in combination with inductor are normally used to reduce the voltage ripple at the output voltage. The main problem in the converter is degradation during conversion cause by an inherent problem which is reverse recovery problem occur. Improvement in the converter design is made to improve the conversion efficiency.

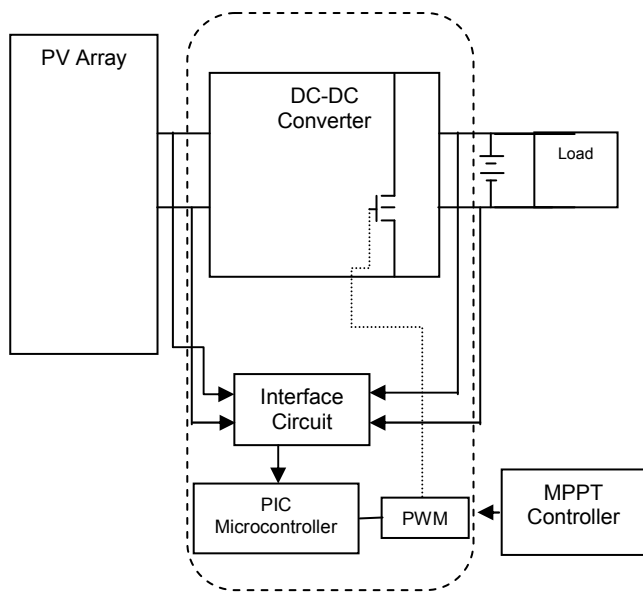


Fig. 1. Schematic diagram of the MPPT Controller for PV System

"Fig. 2" shows the proposed soft switching high step up boost converter design topology. This topology uses an active regenerative snubber to introduce the soft switching boost converter. The components use to develop this converter are capacitor input filter, inductor at the primary side (L_1), inductor at the secondary side (L_2), connecting capacitor (C_6), a switching device comprising of an IGBT (Q_1) with diode (D_1), an active regenerative snubber circuit, and output resistor (R_1). The circuit of active regenerative snubber consists of input diodes (D_2) and (D_3), an input capacitor, a snubber inductor (L_3), a switching device (Q_2) with diode (D_4), a rectifier diode (D_7), a snubber output capacitor (C_3) and a passive snubber contains a capacitor (C_2) and diodes (D_6) and (D_7). The coupled inductor in the primary side (L_1) and secondary side (L_2) can be regarded as an ideal transformer with turns ratio (n) and coupling coefficient (k). The turn ratio (n), coupling coefficient (k)

and the voltage gain (G) of the proposed converter are given as;

$$n = \frac{N_1}{N_2} = \frac{N_{L1}}{N_{L2}} \quad (1)$$

$$k = \frac{L_m}{(L_k + L_m)} \quad (2)$$

$$G = \frac{V_o}{V_{in}} = nk + \frac{1+D}{1-D} + \frac{1}{(1-D)^2} \quad (3)$$

V_{in} and V_o are the input and output voltages and D is the duty cycle of the switching.

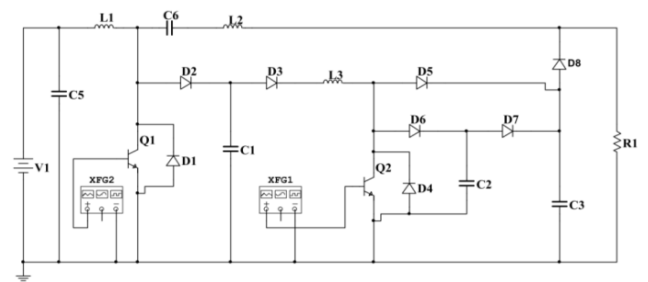


Fig. 2. Proposed High Step-up Boost Converter Circuit

III. IMPLEMENTATION OF FUZZY LOGIC METHOD

Fuzzy logic is well known for its simplicity. Other main advantages of the fuzzy logic controllers are can handle nonlinearity, can work with imprecise inputs, not needed an accurate mathematical model as (1). The inputs are normally an error E and a change of ΔE as given in (4) and (5) respectively.

$$E(n) = \frac{P(n) - P(n-1)}{P(n-1)} \quad (4)$$

$$\Delta E(n) = E(n) - E(n-1) \quad (5)$$

During the initial process which is fuzzification, E and ΔE are calculated and converted to the linguistic variable. Linguistic variables are non-precise variables that often convey a surprising amount of information. To simplify the control calculation, the values of error E and change of error ΔE can be normalized as (6) before fuzzification process.

$$E_f = \left\{ \frac{E}{E^*} \mid E \leq E^* \right\} \quad (6)$$

Where $E^* = V_{max}$, so the scopes of error E and change of error ΔE will be $[-1, 1]$.

The fuzzy logic output is typically a change in duty ratio ΔD of the power converter. The linguistic variables assigned

to ΔD for the different combinations of E and ΔE as shown in Tab. 1 which is based on a boost converter. If for example, the input E is PB, and ΔE is ZE, then we want to largely increase the output which is duty ratio, which is ΔD should be PB to reach the MPP.

TABLE 1. Fuzzy Rule Base

$\Delta E \backslash E$	NB	NS	ZE	PS	PB
NB	ZE	ZE	NB	NB	NB
NS	ZE	ZE	NS	NS	NS
ZE	NS	ZE	ZE	ZE	PS
PS	PS	PS	PS	ZE	ZE
PB	PB	PB	PB	ZE	ZE

The final step is to combine the fuzzy output into a crisp system output. The result of the defuzzification has to be numeric value which determines the change of duty cycle of the PWM signal to drive the IGBT gate. Various methods has been introduced and implemented to calculate the crisp output of the system. The most often method used which is give the better result is probably Centre of Gravity (COG) method. The COG can be expressed mathematically as Eq.(7).

$$\Delta D = \frac{\sum_{i=1}^n Y[i] \cdot F[i]}{\sum_{i=1}^n F[i]} \quad (7)$$

Where $Y[i]$ is the i^{th} members of the output vector and $F[i]$ is the multiplying coefficient of the output membership function as shown in Tab. 1, and ΔD is the change of duty cycle. The ΔD represents a signed number which is added or subtracted from the present duty cycle to generate the next system response for reaching the MPP.

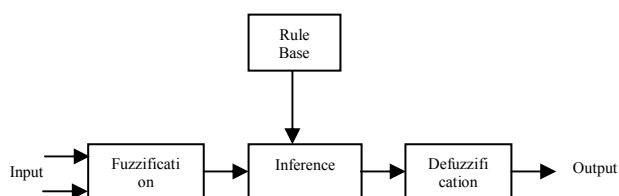


Fig. 3. Main components in fuzzy logic

IV. RESULTS AND DISCUSSION

Simulation is made to evaluate the performance of the boost converter by considering input voltage, $V_{in} = 30V$ and duty cycle, $D = 0.3$ with frequency, $f = 50kHz$. Fig. 4 shows the output voltage of the proposed boost converter obtained from the simulation using Multisim software. From the

figure, it shows that the proposed DC-DC boost converter has a very stable output signal with 70 V. The converter converts about 130 % of generated voltage from input voltage.

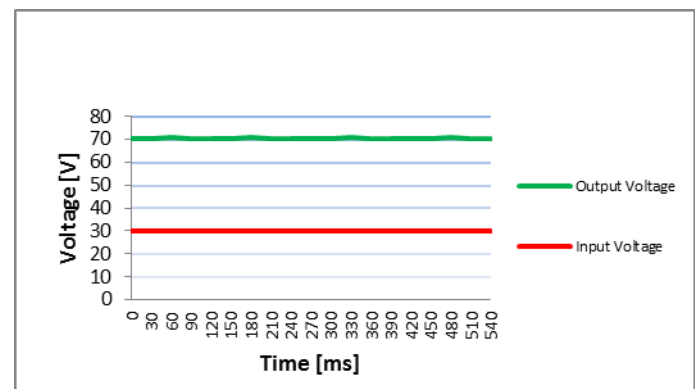


Fig. 4. Proposed DC to DC converter output voltage

The performance of the MPPT controller to reach the maximum power point is investigated using MATLAB/Simulink simulation. The response of the MPPT system is done under the standard condition with irradiation $1000W/m^2$ and ambient temperature $25^\circ C$. Fig. 5 shows the PWM and fig. 6 shows the duty cycle for the system. Fig. 7 shows the output of the system without applying the MPPT controller while fig. 8 is the result of using the MPPT controller. From fig. 8, it can be seen that the output power using the MPPT controller is characterized by a stable and small oscillation around the maximum power point with the tracking time is approximately 2 seconds. The result without using MPPT controller is having a bigger oscillation with dropping transient as the time to achieve the maximum point is approximately 5 seconds as in fig. 7. Therefore, the system using MPPT controller gives a better result for tracking the maximum power point (MPP).

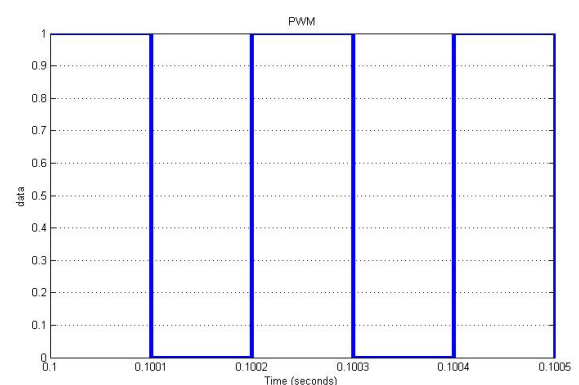


Fig. 5. Pulse Width Modulation (PWM)

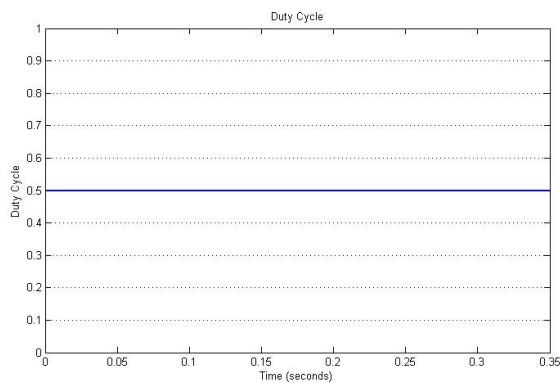


Fig. 6. Duty Cycle (D)

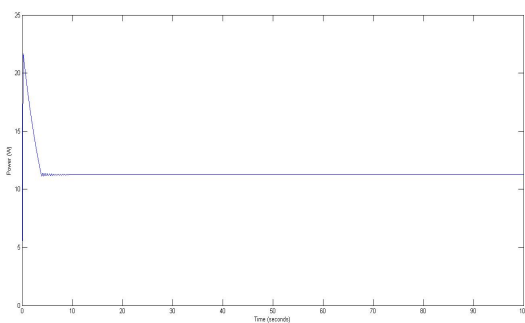


Fig. 7. Output power for without MPPT controller

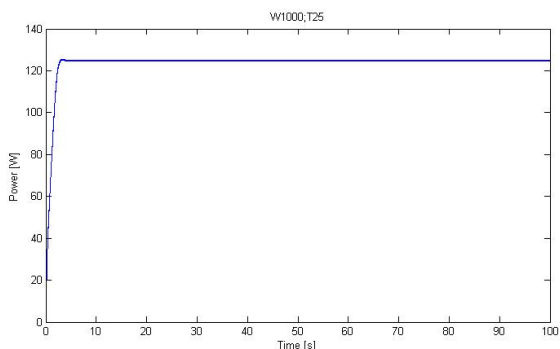


Fig. 8. Output power for with MPPT controller

V. CONCLUSION

The development of the DC-DC boost converter embedded with active snubber circuit applying MPPT technique for the PV system has been presented. The developed DC-DC boost circuit converter is design and simulates using Multisim software. A complete PV model is designed using MATLAB/SIMULINK. The simulation result of the develop DC-DC boost converter show that the converter can produce up to 130% voltage conversion with a stable output voltage. The result of the MPPT with the FL controller is compared with the system that operates without MPPT. It is shows that the FL based MPPT controller can operate better to extract maximum power from the system.

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Hardware Implementation of Grid connected transformer-less semi-Z-source inverter topology to mitigate common mode leakage current and THD

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Abstract— Transformer-less inverters are necessary parts for grid-connected renewable energy resources (RERs). Due to its cost effectiveness, downsize and less weight, great attention has been paid to these inverters development. However, these inverters have limitations like the flow of leakage current through PV arrays, high total harmonic distortion (THD) at inverter's output and DC current injection to the grid. This paper presents coupled inductor based single-phase transformer-less semi-Z-source inverter topology to lessen those limitations. For the purpose of handling the nonlinearity of the voltage gain of semi Z-source inverter, a nonlinear sinusoidal pulse width modulation (SPWM) technique has been employed. A laboratory prototype of the proposed inverter has been constructed. The feasibility of the configuration is ensured based on the mitigated common mode leakage current as well as the substantially lower THD at inverter's output voltage.

Keywords— Transformer-less Inverter, Semi-Z-Source, Total Harmonic Distortion (THD), SPWM.

I. INTRODUCTION

RERs, like wind turbine, solar photovoltaic (PV), and fuel cell are becoming more and more popular in recent era. These RERs are only capable of producing DC voltage at the output, while to connect these RER with grid, these require AC voltage at the output. Therefore, inverters are required to place in between RER and grid to regulate power conversion and control optimization [1-5].

Isolated and non-isolated inverters are widely used for connecting RERs with utility grid. The problems of using these isolated inverters are not only the high system cost and larger system size but also reduced overall efficiency of the system. The reason behind these disadvantages is the presence of transformers in the system with line or high frequency for electrical isolation [6]. In contrast, use of non-isolated inverters reduces both cost and size along with improved system efficiency. So, transformerless inverters have turned into the market mainstream [7]. However, transformerless systems require some safety issues to be considered, like minimization of connection effect between the input DC source and the grid as well as DC current injection to the utility grid. If same ground is not shared by both the PV cell and the grid, a variable common mode voltage is developed.

As a result, large common mode leakage current may flow through the parasitic capacitor between PV array and the ground which in turns, reduces the current quality of the grid, rises the system losses and induces electromagnetic interference (conducted and radiated) [7,8].

Conventionally, half or full bridge inverters have been used to mitigate the problem of common mode leakage current using bipolar SPWM. But, half bridge inverter requires approximately more than 700V of DC voltage to produce $220V_{ac}$ at the output. Hence, series connections of large number of PV arrays or high conversion ratio dc-dc converter are required. On the other hand, the full bridge inverter topology requires 50% of the input voltage than that of half bridge topology (approximately greater than 350V for $220V_{ac}$) [4]. But, the disadvantages of full bridge inverter topology are, high current ripple, lower efficiency and large filter inductor. On the contrary, some operational issues also need to be taken into consideration for transformer less grid connected systems in order to maintain power quality. Such operational issues contain THD, output current regulation, and DC current injection. The IEEE and the IEC standards have provided some limit on the maximum range of THD and injected DC current [9,10].

In order to improve the performance and lessen the cost, many transformerless inverter topologies based on traditional inverter and Z-source or quasi-Z-source inverters have been presented and analyzed for renewable energy distributed generators specially for PV application in [11-15]. But, the performance of the inverter under different loading conditions along with the analysis of THD and DC current injection has not included in [16].

A single stage transformer less semi-Z-source inverter topology for grid connected application is presented in this paper. The benefit of this semi-Z-source inverter topology over traditional single-phase full bridge inverters and Z source or quasi Z source inverters is, to generate sinusoidal voltage at the output of semi-Z-source inverter needs only two switches. Beside this, it contains Z source network in AC side of the semi-Z-source network which is different from conventional topology and results in size minimization. The feasibility of the inverter topology has been analyzed not only for R-L but also for utility grid. This inverter aims to minimize common

mode leakage current with its ground sharing features. It also ensures less THD and DC current injections to the grid by utilizing the coupled inductor techniques.

Section I contains an introduction specifying the rational of the project described in this paper. Basic principles of the transformerless semi-z-source inverters are briefly described in section II. Section III describes the important details of experimental setup as well as the results obtained and their discussion. At last, section IV includes the conclusion drawn from all the above discussed section.

II. BASIC PRINCIPLE OF A TRANFORMERLESS SEMI-Z-SOURCE INVERTER

Topologies of the single phase semi-Z-source inverters with coupled inductor are shown in figures 1(b), (c). From the duty cycle vs. voltage gain curve shown in Fig. 1(a), it is clear that when the duty cycle of the switch S_1 varies from 0 to 1/2, the inverters can provide positive voltage at the output, whereas, from 1/2 to 2/3 the output voltage is negative [12]. For the duty cycle of 1/2, the inverter produces zero voltage at output. Figures 1(b) and 1(c) show the two states of operation respectively. In state I, the switch S_1 conducts where input voltage source and capacitor C_1 charge the two inductors. For state II, switch S_2 conducts and two inductors have turned into sources.

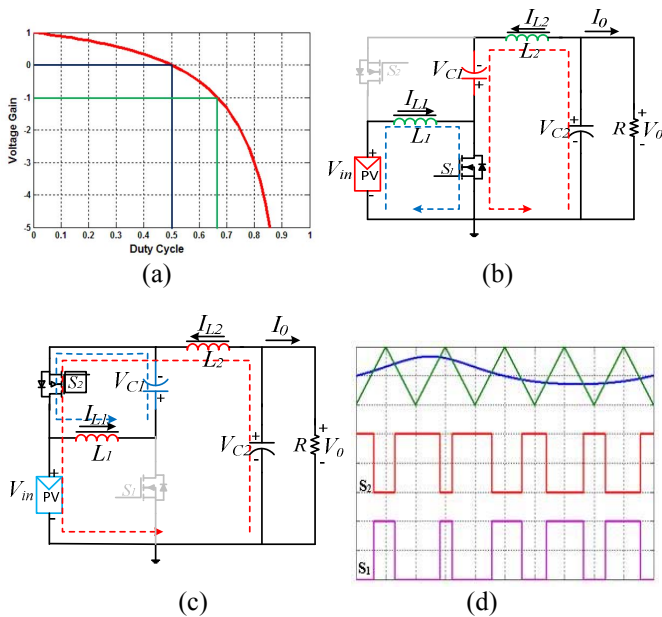


Fig. 1. (a) Voltage gain, (b) modes of operation at state I, (c) Modes of operation at state II, (d) Modulation methods, of semi-Z-source inverters.

The direction of current references of the inductor and the voltage references of the capacitor are mentioned in the figures 1(b) and 1(c) for the following steady state equations. The steady state equations can be derived based on inductor voltage-second balance and capacitor charge balance principle. The steady state equations are as follows:

$$\frac{V_o}{V_{in}} = \frac{1-2D}{1-D} \quad (1)$$

$$V_{C1} = \frac{D}{1-D} V_{in} \quad (2)$$

$$I_{L2} = -I_o \quad (3)$$

$$I_{L1} = -\frac{D}{1-D} I_o \quad (4)$$

If it is assumed that, inverter output voltage is (5) then the modulation index can be expressed as in (6). Equation (7) has been derived from (1), (5), and (6). $D' = 1-D$, the duty cycle of the switch S_2 which can be expressed as (8). Fig. 1(d) shows the switching signals of the switch S_1 and S_2 when the modulation index is 2/3.

$$V_o = V \sin \omega t \quad (5)$$

$$M = \frac{V}{V_{in}} \quad (6)$$

$$D = \frac{1 - M \sin \omega t}{2 - M \sin \omega t} \quad (7)$$

$$D' = \frac{1}{2 - M \sin \omega t} \quad (8)$$

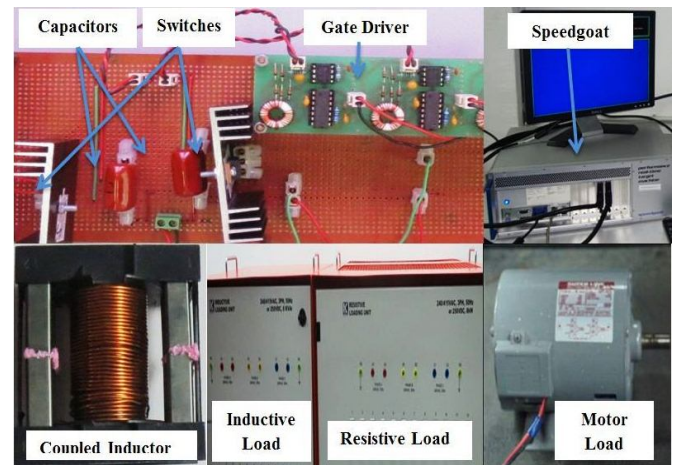


Fig. 2. Experimental setup of semi-Z-source inverter.

III. EXPERIMENTAL DESIGN

For the purpose of experimental validation, a prototype rated 48-W, 50Hz transformerless semi-Z-source inverter is constructed. The constructed laboratory prototype model of the semi-Z-source inverter system is shown in Fig. 2. Experimental results of the laboratory prototype model of 48-W transformerless semi-Z-source inverter are shown in figures 3 and 4. The prototype has been tested under R-L (27 Ω

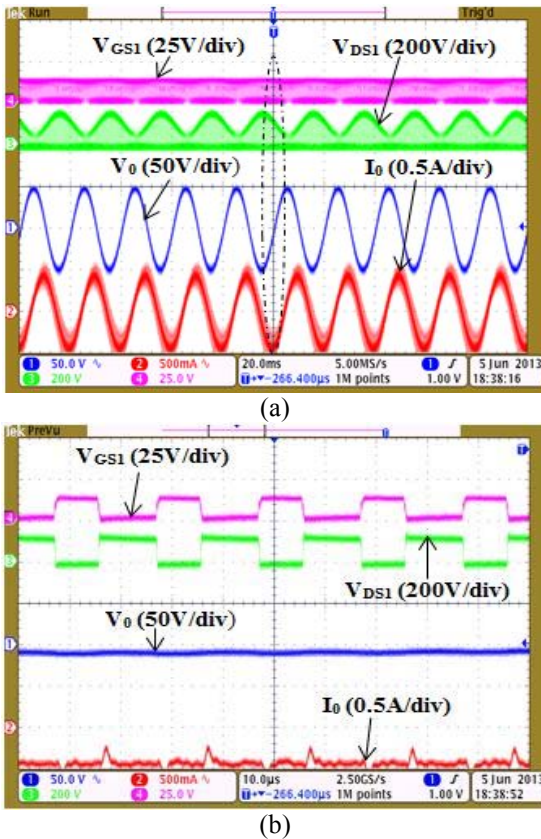


Fig. 3. Experimental waveforms for R-L load (a) Gate to source voltage, drain to source voltage, output voltage, output current, (b) Zoomed in waveform of gate to source voltage, drain to source voltage, output voltage, and output current

resistance and 285mH) load. During the laboratory experiment, around 50V input voltage has been applied and modulation index has been fixed to 0.95.

Fig. 3 illustrates the experimental results for R-L load. It is observed from Fig. 3 that the output current is lagging the output voltage and the magnitude of the output current decreases as the load increase. Waveforms of input voltage V_{in} , voltage across capacitor C_1 , output voltage V_o and output current I_o for R-L load are shown in Fig. 4. (a). These two figures illustrate that, peak capacitor voltage is twice the input voltage for R-L load. Moreover, both the positive and negative peak values of the output voltage are equal to input voltage. Fig. 4(a) display that for R-L load condition, the input voltage V_{in} is almost constant and contains no high frequency variation. As a consequence, the generation of common mode voltage is minimized which in turns results in reduction of common mode leakage current. The voltage across capacitor C_1 , current through inductor L_1 , output voltage and output current of the inverter for R-L loads are shown in Fig. 4(b). The figure notified, voltage developed across the capacitor C_1 followed by the inductor current I_{L1} and inductor current is approximately twice the output current. THD and harmonic spectrum of the output voltage and current for R-L are presented in Fig. 5. It can be depicted from Fig. 5 that, for R-L load the inverter produces RMS output voltage of 35.76V with

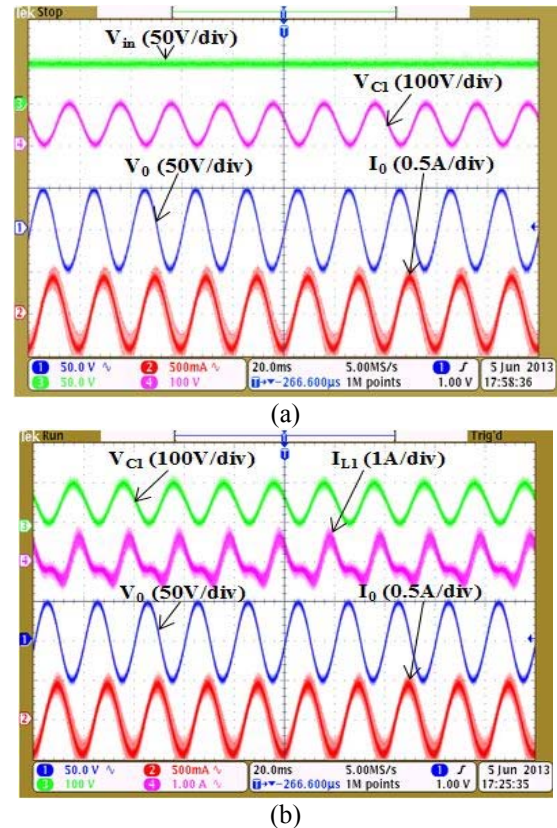


Fig. 4. (a) Input DC voltage, capacitor C_1 voltage, output voltage, output current for R-L load, (b) Capacitor C_1 voltage, inductor L_1 current, output voltage, output current for R-L load.

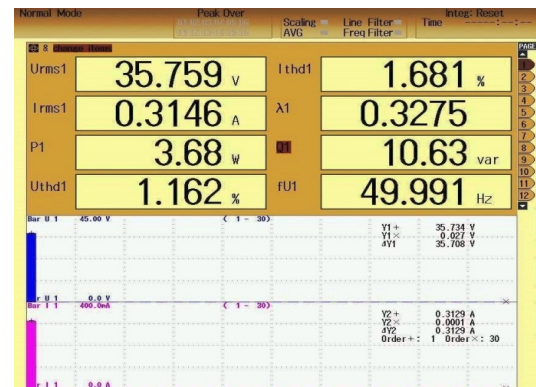
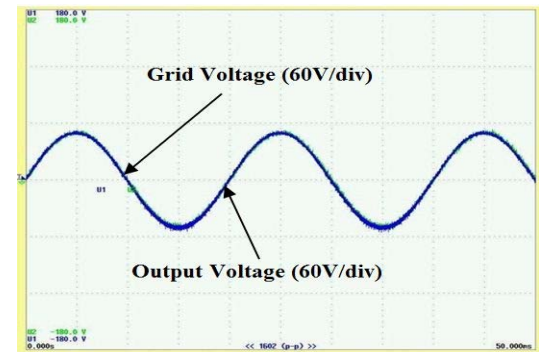


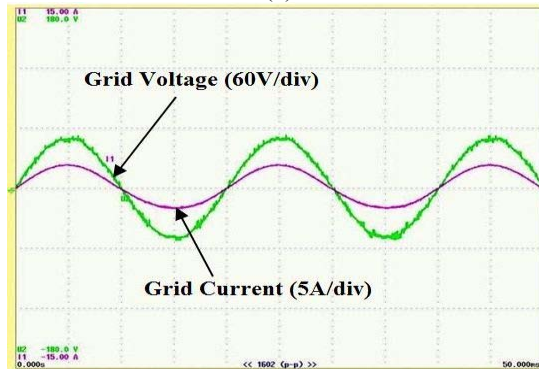
Fig. 5. THD and Harmonic spectrum of output voltage and current for R-L load

THD of 1.16% and RMS output current of 0.315A with THD of 1.68%. The inverter supplies real power of 3.68W and reactive power of 10.62var for R-L load.

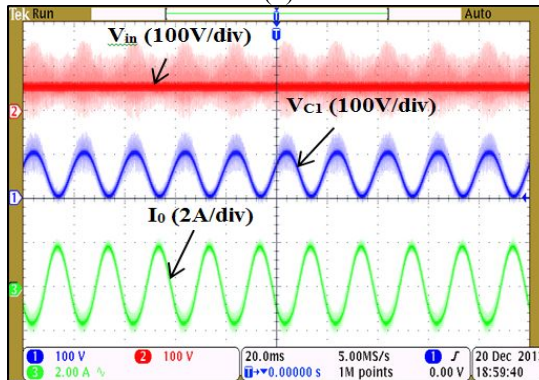
Finally, the prototype of the inverter has interfaced with the low voltage utility grid to observe the power transfer in synchronous mode. During interfacing with the grid in laboratory condition the utility grid voltage has been taken 50V_{ac} with frequency of 50Hz. Synchronous operation between grid voltage and inverter output voltage is shown in Fig. 6(a). Fig. 6(b) shows the experimental waveforms of grid



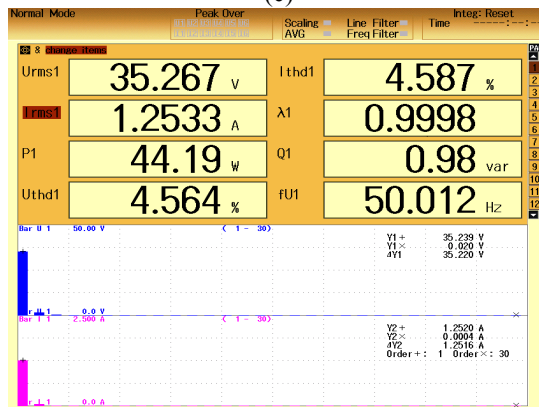
(a)



(b)



(c)



(d)

Fig. 6. (a) Experimental waveforms of inverter output voltage and grid voltage, (b) Experimental waveforms of grid voltage and grid current, (c) Input DC voltage, capacitor C_1 voltage, output current during interfacing with grid, (d) THD and Harmonic spectrum of output voltage and current during interfacing with grid.

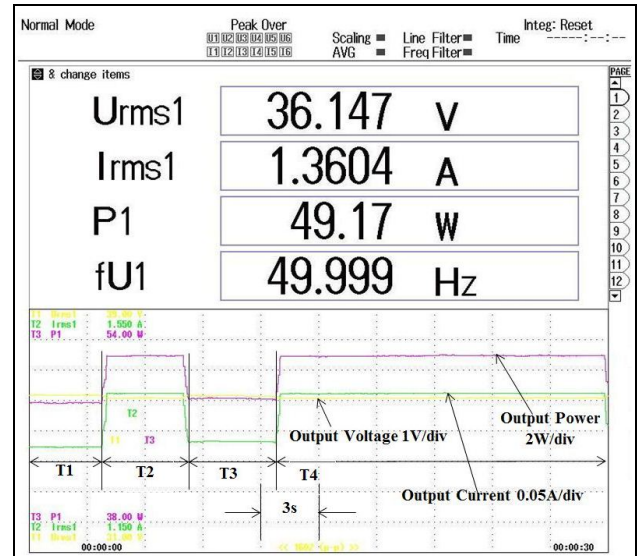


Fig. 7. Dynamic response of the inverter

voltage and grid current during full load condition. Waveforms of input voltage V_{in} , voltage across capacitor C_1 , and output current I_o of the inverter during grid interfacing are shown in Fig. 6 (c).

THD and harmonic spectrum of the inverter output voltage and injected grid current are shown in Fig. 6(d). It can be seen from this figure that, the inverter transfers power of 44.19W and injected grid current have THD of 4.58%.

Fig. 7 shows the dynamic response of the inverter for different loading conditions in terms of voltage, current and power. It is seen that, for the period of T1, the inverter supplies power of around 47.50W to 27.30 Ω resistive load with the RMS current of 1.32A. At the period of T2, suddenly the values of load resistance has been changed to 26.60 Ω and the inverter response within very short time by supplying 49.14W power with the RMS current of 1.36A. Again, at the period of T3 and T4 the values of load resistance have been changed to around 27.15 Ω and 26.60 Ω respectively and the inverter response very quickly by changing the output current levels to 1.33A and 1.36A respectively. For all the loading conditions, the value of output RMS voltage is constant to 36.15V.

IV. CONCLUSION

This paper describes a transformerless semi-Z-source inverter topology. By utilizing semi Z-source network; this topology employs only two active switches and a nonlinear SPWM technique to generate required sinusoidal voltage at the output. As the input and output terminals share the same ground, the leakage current flows through both grid and RERs get minimized. In addition to those, this low cost-compact inverter configuration has ensured the reduced

THD of output voltage and current without any filter and can response to the load very fast.

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Input Switched High Performance Single Phase Single Switch C^{uk} AC-DC converter

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Abstract— Single phase full wave bridge rectifier draws non sinusoidal current from a sinusoidal voltage source which causes low performance of the rectifier. Switching mode regulators are used to improve the shape of input current and power factor. In this paper a new single phase single switch AC-DC converter using C^{uk} topology is proposed. In this converter input current is chopped at high frequency during positive and negative cycle by a single switch to get step up/step down conversion. Comparison is made between proposed and conventional circuit under same load and operating conditions. The proposed circuit shows better performance. The efficiency of these converters is also high. No additional control scheme is required to achieve the advantages.

Keywords—AC-DC converter; THD; Switching mode regulators;

I. INTRODUCTION

Single phase full wave bridge rectifier draws non sinusoidal currents from sinusoidal voltage source due to its nonlinear characteristics. Several problems related to power quality arise at the input side of the rectifier due to this nonlinear characteristic. These problems include low input power factor, high input current THD [1- 3] etc. Due to high input current THD power supply is interrupted, which cause malfunction or damage of sensitive electric device. There are several methods to solve the problem. One of these methods is to use a passive filter which improves THD but input power factor remains low. Another method employs a single phase rectifier followed by a DC-DC converter. This method is also known as active filtering [4-9]. In this case the input filter required is relatively smaller than passive filter and improved input power factor is achieved in this method.

The Boost DC-DC converter is commonly used after rectifier for power factor correction. Buck, Buck-Boost etc. are also used to get different input/output voltage gain relationship [9-13]. The DC-DC converter chopped the input current at high frequency which can be filtered with a low pass filter to get near sinusoidal input current with high input power factor and low THD.

In this paper a new single phase single switch AC-DC converters based on C^{uk} topology of high performance is

reported. C^{uk} topology has several advantages over Buck-Boost topology like continuous input and output current. Moreover C^{uk} converter supplies energy to load both when switch is open and close. The reported converters chop input current by a single switch. This converter is different from conventional DC-DC converter regulated rectifier because input current switching ensures input AC current to be almost in phase with input voltage without any control scheme. This would result almost sinusoidal input current with using of small filter and also ensure good input power factor.

II. Proposed Circuit Configuration

Figure 1 shows the circuit diagram of proposed single phase single switch C^{uk} AC-DC converter. The circuit consists of inductors (L1, L2), capacitors (C1, C2), diodes and a switch (M1). As the given supply voltage is AC, the conversion on both positive and negative cycle of the input voltage is required. L1 and C1 are used for positive half cycle and L2, and C2 are for negative half cycle. The inductors L1_pos, L1_Neg, L2_pos, L2_Neg and capacitors C_pos and C_Neg form the input filter. C_Load and R_Load are the output capacitor and load respectively.

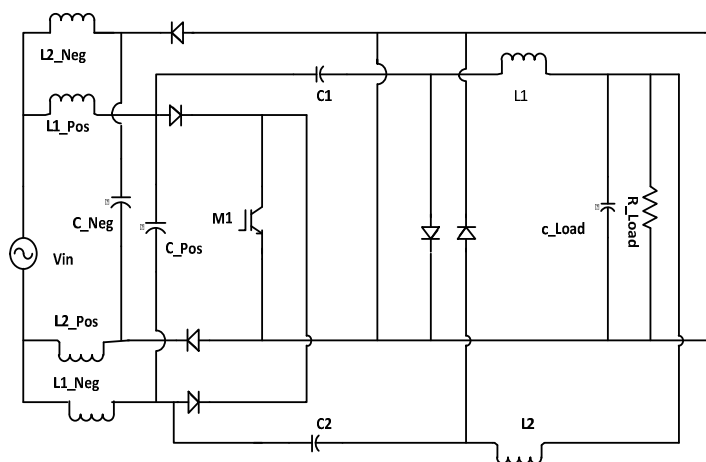


Fig. 1. Proposed AC-DC C^{uk} Converter

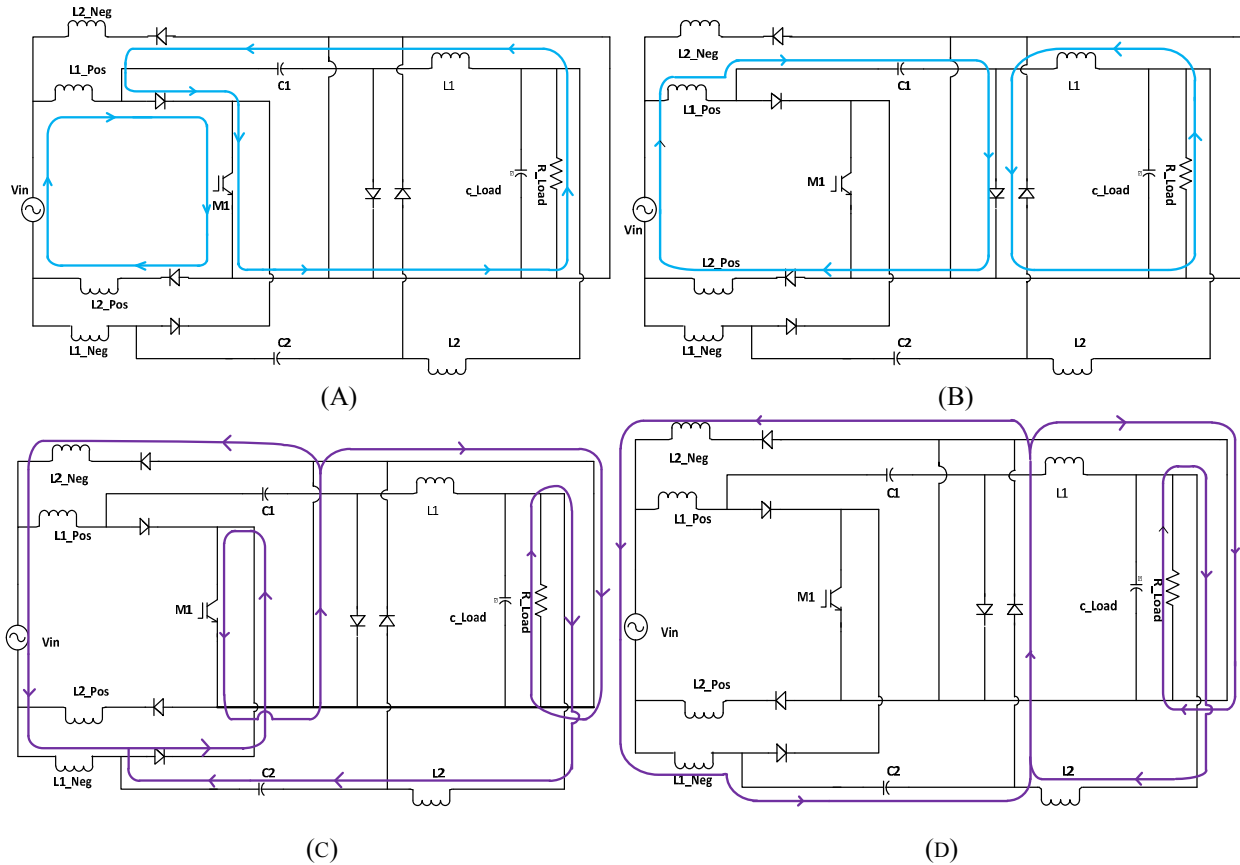


Fig. 2. Four states of operation of proposed AC-DC converter (A) State 1, circuit when switch is ON during positive half cycle of the AC supply, (B) State 2, circuit when switch is OFF during positive half cycle of the AC supply, (C) State 3, circuit when switch is ON during negative half cycle of the AC supply, (D) State 4, circuit when switch is OFF during negative half cycle of the AC supply.

III. Principle of Operation

Input AC current chopping at high frequency, provides switched AC current that requires small filter to make it nearly sinusoidal. As a result, the input current THD reduces and the input power factor improves. The proposed C^{uk} AC to DC converter has four operating states – positive and negative half cycles each with switch ON and OFF condition, as shown in Fig. 2. States 1 and 2 represent the positive half cycle operation with switch ON and OFF position respectively. Similarly, states 3 and 4 correspond to the negative half cycle operation of the converter. In this proposed scheme energy transfer is associated with inductors and capacitors L1, C1 and L2, C2 for positive and negative half cycle respectively

IV. Simulation Results

Simulation of the proposed circuit is performed with PSIM professional version 9.0.3.400 and with OrCAD Capture CIS 9.2 version.

A. Circuit Parameters

For simulation of C^{uk} configuration, an input ac source of 300V amplitude with frequency of 50 Hz is employed. An IGBT is used for switching purpose. For C^{uk} scheme, the inductors L1 and L2 have the values of 1.5mH each. The filter inductors L1_pos, L1_Neg, L2_pos, L2_Neg have the values of 4 mH and capacitors C1 and C2 have the values of 0.5μF each. The filter capacitors C_pos and C_neg have the values of 1μF each. The output capacitor C_Load has the value of 100 μF and a resistor R_Load of 100 Ω is used as load in the proposed circuit. The proposed single-phase converter has been compared with the conventional single-phase diode bridge rectifier followed by a DC-DC C^{uk} converter.

B. Results from Simulation

Typical input voltage, input current and output voltage waveforms of the proposed single-phase AC-DC converter circuit at a switching frequency (f_s) of 6 kHz, duty cycle (D) of 0.6 and load resistance (R_Load) of 100 Ω are given in Fig. 3. From this figure it is observed that, the input current is nearly sinusoidal. Fig. 4 shows the corresponding input current spectrum of the proposed circuit.

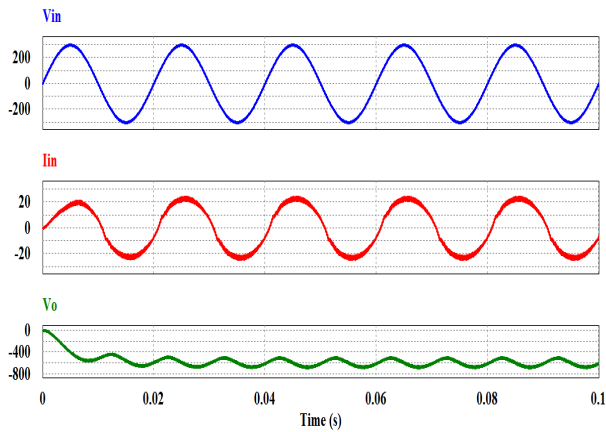


Fig. 3. Input voltage, input current and output voltage waveforms of the proposed converter at $f_s=6$ kHz, $D=0.6$, and $R_{Load} = 100\Omega$.

Fig. 5 and Fig.6 show the waveforms and input current spectrum of conventional circuit at $f_s=6$ kHz, $D=0.6$, and $R_{Load}= 100 \Omega$, under same operating condition to show the typical improvement.

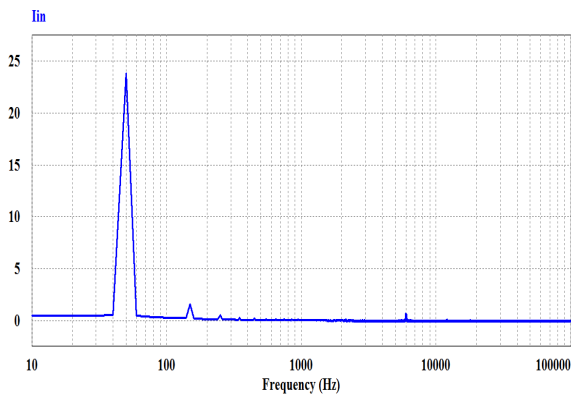


Fig. 4. Input Current spectrum of the proposed converter at $f_s=6$ kHz, $D=0.6$, and $R_{Load}=100\Omega$.

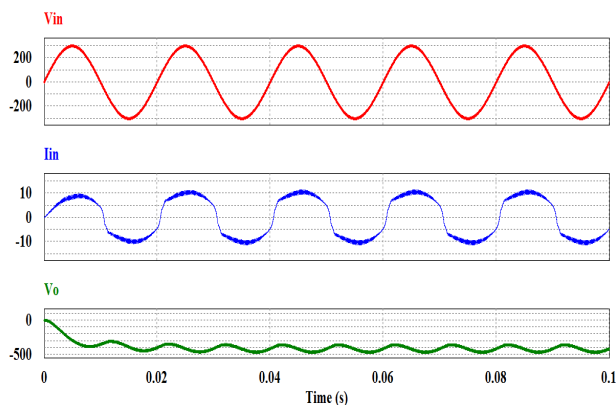


Fig. 5. Input voltage, input current and output voltage waveforms of the conventional converter at $f_s=6$ kHz, $D=0.6$, and $R_{Load} = 100\Omega$.

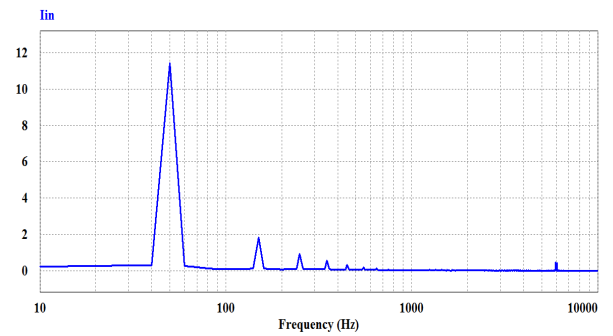


Fig. 6. Input Current spectrum of the conventional converter at $f_s=6$ kHz, $D=0.6$, and $R_{Load} = 100 \Omega$.

C. Quantitative Comparison

The performances of the proposed single phase C^{uk} AC-DC converter scheme is compared with conventional one having DC-DC converter at the output. For comparison, results are evaluated in terms of THD (%) of input current, input power factor and efficiency (%). The results of comparisons are presented in Fig. 7 to 9.

The bar chart shown in Fig. 7 indicates that the THD (%) of input current of the proposed converter is better than the conventional one for all duty cycles. The proposed converter exhibits high input power factor at all duty cycles compared to conventional one (Fig. 8). In terms of efficiency, the proposed C^{uk} converter has almost same efficiency as the conventional one (Fig 9).

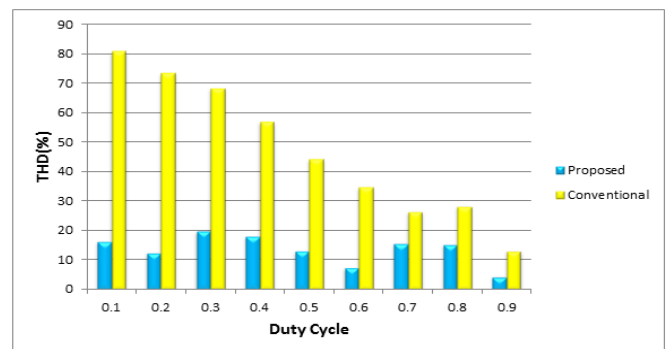


Fig. 7. Comparison of input current THD (%) between conventional and proposed scheme at $f_s=4$ kHz and $R_{Load} = 100\Omega$.

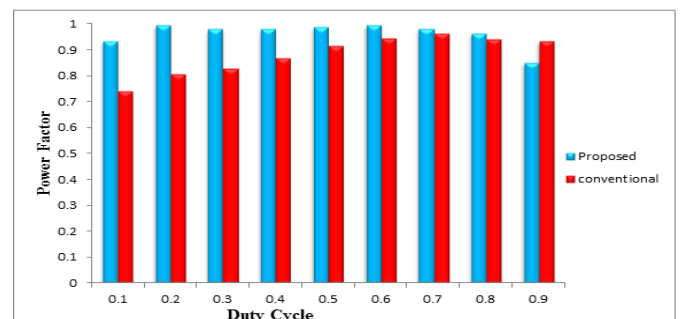


Fig. 8. Comparison of input power factor between conventional and proposed scheme at $f_s=4$ kHz, and $R_{Load} = 100 \Omega$.

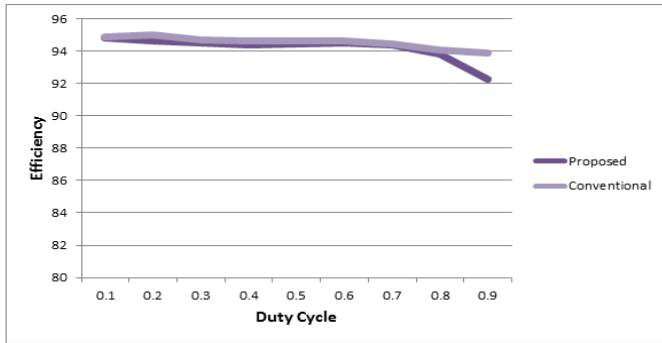


Fig. 9. Comparison of efficiency (%) between conventional and proposed scheme at $f_s=8$ kHz, and $R_{Load}=100 \Omega$.

D. Performances under Load Variation

Both circuits are subjected to load variation at frequency 8 KHz and 40 % duty cycle. The load resistance was varied from 50Ω to 500Ω and the performance was monitored in terms of input power factor and input current THD (%).

The comparisons of input current THD (%) and input power factor between the proposed and the conventional converter are presented in Fig. 10 to Fig. 11. The proposed converter gives low input current THD (%) with respect to the conventional circuit. The proposed circuit also has higher power factor than the conventional single-phase rectifier for all load resistance.

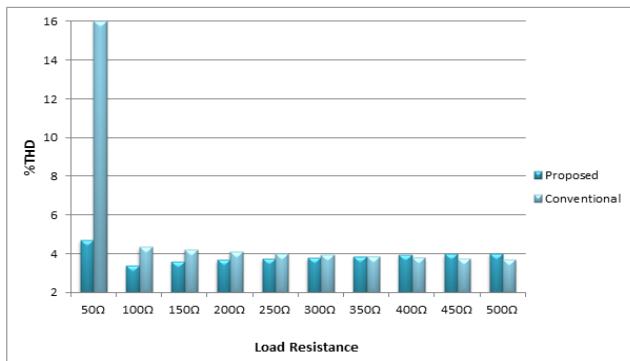


Fig. 10. Comparison of input current THD (%) between conventional and proposed scheme with load variation at $f_s=8$ kHz, and $D=0.4$.

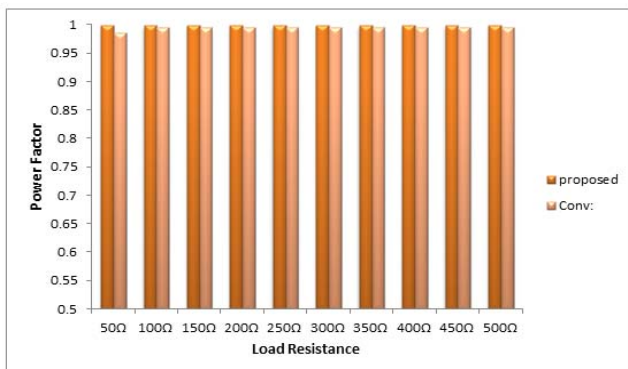


Fig. 11. Comparison of input power factor between conventional and proposed scheme with load variation at $f_s=8$ kHz, and $D=0.4$.

VI. CONCLUSION

A new scheme for single-phase C^{uk} AC-DC converter has been proposed with low input current THD and high input power factor. The circuit chops the input current at the AC side in contrast to the conventional scheme, which chops the rectified output. In this paper, simulation results are compared between proposed converter and conventional one. The main advantage of the proposed circuit is that smaller input filter is required to achieve better performances in terms of low input current THD and high input power factor. The efficiency of the proposed circuit is also high with small filter inductor and capacitor.

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An Efficient Wind Speed Sensor-less MPPT Controller Using Adaptive Neuro-Fuzzy Inference System

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Abstract: An adaptive neuro-fuzzy inference system (ANFIS) based maximum power point tracking (MPPT) algorithm has been proposed. The ANFIS based controller has the ability to track the maximum power point (MPP) and the corresponding rotor speed of the wind generator by estimating wind speed with very little error compared to the conventional ANN based MPPT techniques. The algorithm developed is based on two series ANFIS networks, one for wind speed estimation and the other to determine the maximum power point and the corresponding rotor speed. The method demonstrates remarkable performance in estimating wind speed and to predict MPP accurately without undesired oscillations around maximum power point. The algorithm does not require any mechanical sensor for wind speed measurement. Nonlinear time domain simulations have been carried out to validate the effectiveness of the proposed controllers under different operating conditions. Simulation results confirm the effectiveness of the proposed MPPT controller in tracking the maximum power point under rapidly changing wind conditions.

Keywords: ANFIS, MPPT, renewable energy, optimal rotor speed determination, Wind energy.

I. INTRODUCTION

Electric power generation using renewable energy systems have gained attention over the past decade due to the high costs, probable depletion, and harmful environmental impact of the conventional energy sources. Wind, solar, tidal, wave, geothermal and bio-fuels are some of the renewable energy sources which have received attention in the present days. Among the renewable energy sources (RES), wind energy is quite promising; wind turbine systems provide most direct method to convert wind energy into clean electric energy. To extract maximum power the rotor should always rotate at certain desired speed corresponding to a particular wind speed. The variable speed wind turbines are becoming more popular due to its capability to extract maximum power at different velocities. Anemometers are generally used to determine the wind speed. Generally, a number of anemometers are required at different locations for accurate measurement of wind speed, this makes it costly.

Since the efficiency of the wind turbine system depends on how accurately the maximum power point can be tracked, the determination of the optimum speed for varying wind conditions have been extensively researched over the past twenty five years. Various algorithms such as tip speed ratio control [1]-[3], power signal feedback control [4]-[6], hill-climb search control [7],[8], etc. are reported in the literature. Perturb and observe technique is a simple and well used

method, but the tracking performance is slow and it may give rise to continuous oscillation around operating point [9]-[10]. Most of these methods have drawbacks associated with them like large convergence time, fluctuations in the steady state and possible failure to track MPP in rapidly changing wind conditions. Neural networks can be used for MPPT problem [11] based on the ratio of mechanical power and turbine rotation speed. This will take care of uncertainty in parameters and also avoid oscillation. A neural network based control can be used to track MPP for both the dynamic and steady states, and to estimate wind velocity fast and accurately without using anemometers [12]. ANN based MPPT technique is proposed, in which two series networks are considered for wind speed estimation and maximum power tracking [14]. Combining fuzzy interface system (FIS) with ANN results in a powerful artificial intelligence technique known as adaptive network-based fuzzy inference system (ANFIS) [13]. Many methods have been proposed for ANFIS based MPPT determination. ANFIS programs developed has only one output. That's why most of the researchers used ANFIS for wind speed estimation and some evolutionary programming techniques (like Tabu search, differential evolution, evolutionary programming, particle swarm optimization, and genetic algorithm) for maximum power point tracking. This makes the procedure computationally burdensome. This study has developed an improved tracking algorithm using adaptive network-based fuzzy inference system principles. It uses two ANFIS networks, one for wind speed estimation and the other for maximum power point. This computationally efficient intelligent MPPT controller is implemented using MATLAB/Simulink and its performance has been compared with reported results.

II. MAXIMUM POWER POINT TRACKING

The wind turbine converts the kinetic energy of wind to mechanical energy. The mechanical power output of the turbine in terms of wind speed (V_w) can be expressed as,

$$P_m = \frac{1}{2} \rho A C_p V_w^3 \quad (1)$$

where, ρ is the air density and A is the swept area by the turbine blades. The wind power captured by wind turbine depends on its power co-efficient C_p written as,

$$C_p(\lambda, \beta) = 0.5176 \left(\frac{116}{\lambda_i} - 0.4\beta - 5 \right) e^{\frac{-21}{\lambda_i}} + 0.0068\lambda \quad (2)$$

$$\frac{1}{\lambda_i} = \frac{1}{\lambda + 0.08\beta} - \frac{0.035}{\beta^3 + 1} \quad (3)$$

Here, λ is the tip speed ratio and β is the blade pitch angle.

The value of λ can be calculated using the following formula:

$$\lambda = \frac{\Omega R}{V_w} \quad (4)$$

The wind power vs. turbine rotor speed characteristics are plotted for different wind speeds shown in Fig.1.

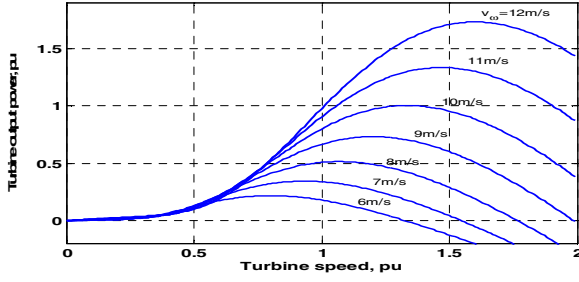


Figure 1: Speed vs. power output characteristics of a wind turbine

It can be observed from Fig.1 that for each wind speed, there is a maximum power point corresponding to a particular generator rotor speed. Maximum power point tracking problem involves identifying the maximum power and corresponding optimum generator speed for a particular wind speed. The maximum power point tracking concept with the change in wind velocity is illustrated in Fig.2. With the change in wind velocity, the proposed ANFIS will identify the maximum power corresponding to the wind velocity and will also specify the turbine rotor speed for which we will get maximum power. In Fig.2, point A is turbine power at a particular speed and point B is maximum power point at that wind speed.

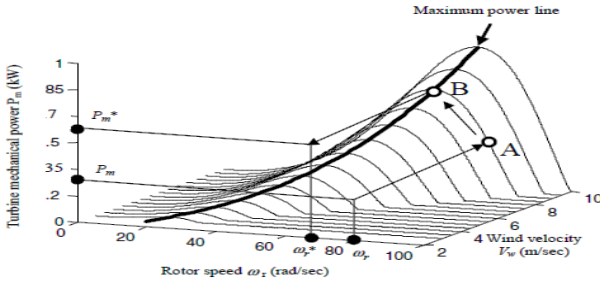


Figure 2: Wind turbine power curves [12]

III. ADAPTIVE NETWORK-BASED FUZZY INFERENCE SYSTEM (ANFIS)

In comparison to the neural network, ANFIS also has similar type network structure and maps for the input-output data set using the parameters of fuzzy membership functions. A simple ANFIS architecture, based on the two rule Sugeno system[13] with two inputs (X and Y) and single output (F) is shown in Fig. 3, where A1, A2 and B1, B2 are fuzzy input memberships for input X and Y, respectively and are used to fuzzify the inputs.

For a Sugeno ANFIS model the typical rule set with two fuzzy if-then rules can be defined as in (5) and (6):

If X is A₁ and y is B₁ THEN $f_1 = p_1x + q_1y + r_1$ (5)

If X is A₂ and y is B₂ THEN $f_2 = p_2x + q_2y + r_2$ (6)

As shown in Fig.3, ANFIS consists of five layers; the function of each layer is described below:

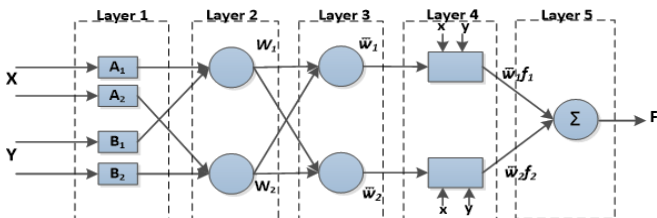


Figure 3: ANFIS architecture

a) Layer-1:

In layer-1 every node is adaptive node. The number of adaptive node in layer-1 depends on the number of input membership functions. Each input is assigned a certain number of membership functions and each membership function corresponds to a node. The main task of this layer is to fuzzify input parameters using some variables named as large, medium, small etc. Their output is given by:

$$O_{1,j} = \mu_{A_j}(x) \quad \text{for } j = 1,2 \quad (7)$$

$$O_{1,j} = \mu_{B_{j-2}}(y) \quad \text{for } j = 3,4 \quad (8)$$

Here, μ is the membership function and $O_{1,j}$ is the membership value for inputs X and Y. Membership functions μ_A can be different shaped function like trapezoidal, triangle, Gaussian. The most commonly used membership function is generalized bell (g-bell) and can be defining as (9).

$$\mu_A(x) = \frac{1}{1 + \left| \frac{x - c_j}{a_j} \right|^{2b_j}} \quad (9)$$

where a_j is the standard deviation, b_j is a positive number and c_j is the mean. These parameters are also termed as premise parameter and will be optimized during the training period. The value of a, b, c used for this plot are 2, 4, 6 respectively.

It has been observed that the wind speed estimation from turbine power vs rotor speed curve follows approximately bell shaped behavior. Though the turbine power vs rotor speed curve for different wind speed is not precisely follow the normal distribution characteristics but for large sample size (more than 30 samples) would guarantee that means will be bell shaped.

b) Layer-2:

All the nodes in this layer is fixed node and admits the output (membership values) from layer-1, where t-norm is used to “AND” these values, given by (10). The following four properties are to be satisfied for all a, b, c $\in [0, 1]$:

Commutativity: $T_1(a, b) = T(b, a)$

Associativity: $T_1(a, T(b, c)) = T(T(a, b), c)$

Monotonicity: $T_1(a, b) \leq T_1(a, c)$ whenever $b \leq c$ and

Boundary condition: $T_1(a, 1) = 1$

$$O_{2,j} = w_j = \mu_{A_j}(x) \mu_{B_j}(y) \quad j = 1,2 \quad (10)$$

Output of all the nodes correspond to the firing strength of the rule. Firing strength is a threshold limit above which a rule/ nodes get active and pass the output to other nodes.

c) Layer-3:

In this layer-3, each node is fixed node and used to perform normalization operation. The j-th node of this layer calculates the ratio between the j-th node activation level and the sum of all activation level, which is known as normalized firing strength. The output of every node can be defined by (11).

$$O_{3,j} = \bar{w}_j = \frac{w_j}{w_1 + w_2} \quad (11)$$

Output of every node interprets the normalized firing strength of the rule.

d) Layer-4:

In layer-4, every node is adaptive and the function can be defined as (12).

$$O_{4,j} = \bar{w}_j f_j = \bar{w}_j (p_j x + q_j y + r_j) \quad (12)$$

where, p_j , q_j and r_j are the resultant/ consequence parameters, and these need to be minimized during the training period.

e) Layer-5:

This layer is the output layer. The output signals from layer-4 are summed together to obtain the result at the layer 5. This layer can be described as (13).

$$O_{5,j} = \sum_j \bar{w}_j f_j = \frac{\sum_j w_j f_j}{\sum_j w_j} \quad (13)$$

IV. RESULTS

The ANFIS network is trained with about 1000 data sets created from the wind power-rotor speed-wind speed plot in Fig.1. The input to the first network is power and rotor speed and output is wind speed. The second network uses this wind speed as the input to give the corresponding optimum rotor speed (and maximum power). The test results for the given wind speed, power and rotor speed are compared with network outputs. The comparisons of the above 3 are given in Figs. 4, 5 and 6 respectively. Sample error plots for the actual and predicted wind speed, maximum power and optimal rotor speed are shown in Figs. 7, 8 and 9 respectively.

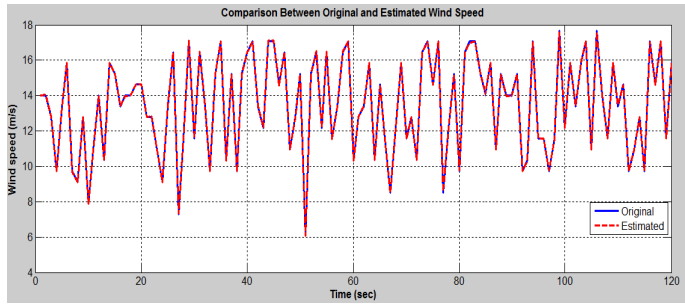


Figure 4: Original and estimated wind speed

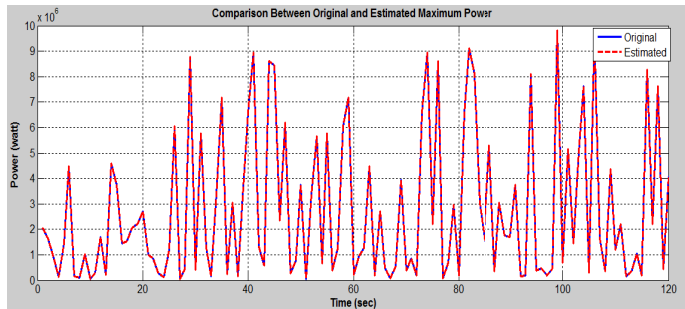


Figure 5: Original and estimated maximum power

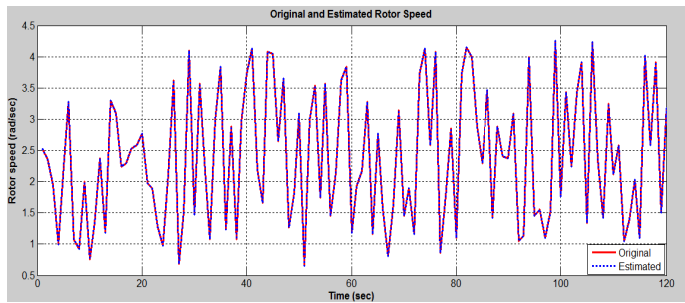


Figure 6: Original and estimated optimal rotor speed

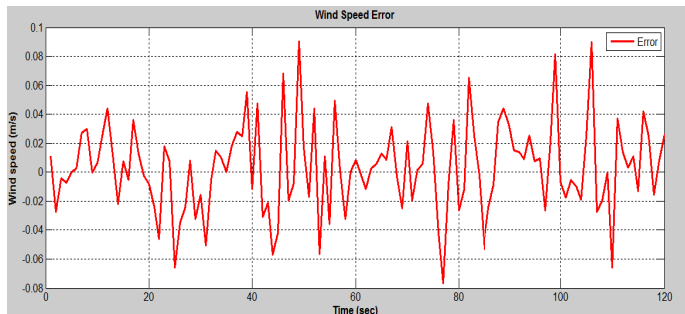


Figure 7: Wind speed estimation error
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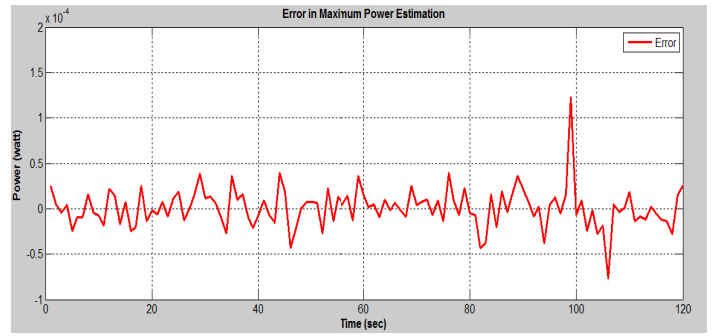


Figure 8: Maximum power estimation error



Figure 9: Rotor speed estimation error

The given (original) quantities in Figs. 4-6 are shown by the solid lines and the ANFIS network output is represented by the dotted lines. It can be seen that the predicted wind speed, maximum power, and optimal rotor speed variations completely overlap the original ones. Fig.7 shows the difference between the given and estimated values of wind speed, the maximum difference being 0.09 m/s. The maximum power and rotor speed estimation errors are 1.3×10^{-4} watt and 1.487×10^{-4} rad/sec respectively as shown in Fig.8 and Fig.9.

From the results presented, it can be observed that the proposed MPPT controller has the ability to track maximum power point and corresponding rotor speed under rapid changing wind conditions with very little error.

A. Comparison with reported results

The proposed ANFIS based MPPT algorithm is compared with the results presented in reference [12]. A variable speed small wind generation system is considered. The rating of the wind generator is 1.4 kW and radius of turbine blades is 1m. The article considers a random speed variation, the maximum and minimum values of which are similar to the values considered in the present study. Fig.10 shows the wind speed estimation error as reported, the maximum error being 0.25m/s. Note that this value is 0.09 m/s in this study as presented in Fig.7.

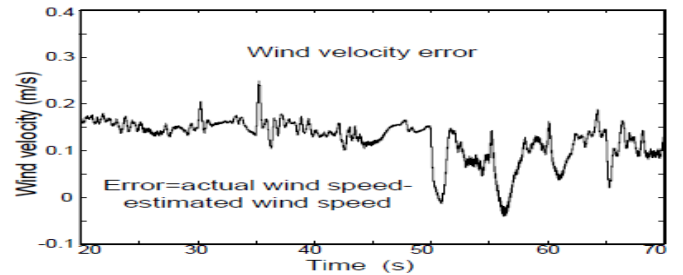


Figure 10: Reported wind speed estimation error [12]

From Fig.9 it is evident that maximum rotor speed error is around 1.487×10^{-4} rad/sec with the proposed controller, while this is about 1rad/sec in the study conducted in reference [12]. Record of the original and estimated rotor speeds of the reported study is reproduced in Fig.11.

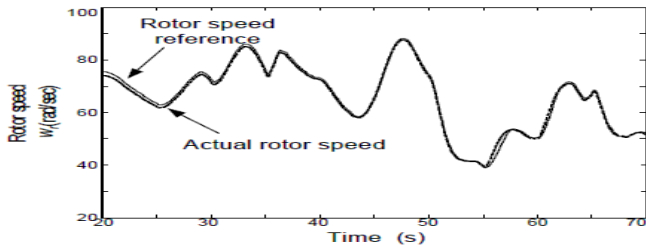


Figure 11: Original and estimated optimal rotor speed [12]

The maximum wind speed estimation error from the reported controller is about 0.25 m/s, while it is about 0.09 m/s with the proposed ANFIS controller. On the other hand the maximum error in optimal rotor speed tracking is about 1.4875×10^{-4} rad/sec and 1 rad/sec for proposed and reported controller respectively as shown in Table 1.

Table 1: COMPARISON BETWEEN PROPOSED AND REPORTED CONTROLLER

MPPT Methods	Minimum wind speed error (m/s)	Maximum wind speed error (m/s)	Minimum rotor speed error (rad/sec)	Maximum rotor speed error (rad/sec)
Reported Method	0.1	0.25	1.502×10^{-4}	1.00
Proposed Method	4.2×10^{-5}	0.09	1.130×10^{-6}	1.487×10^{-4}

From Table 1, we can observe that the proposed ANFIS based controller provides more precise tracking of the maximum power point.

The variation of power coefficient (C_p) for a sample wind speed variation is shown in Fig.13. For staircase changes in wind speed pattern, given in Fig 12, it can be observed that there is virtually no variation in C_p . This shows that the proposed controller can track the optimum values guaranteeing maximum transfer of wind energy to the turbine-generator system.

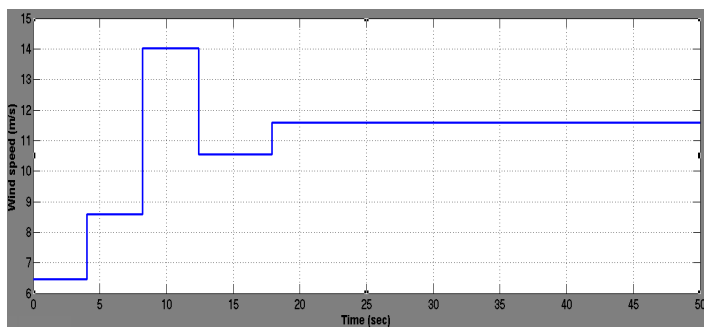


Figure 12: Random wind speed variation

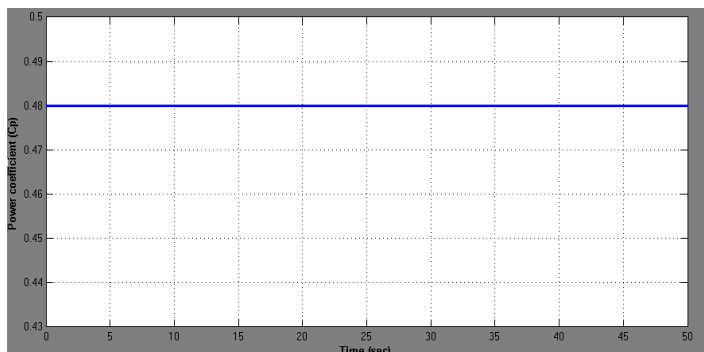


Figure 13: Change in power coefficient for wind speed variation

An adaptive neuro-fuzzy inference system based maximum power point tracking algorithm is presented. The method involves operating two ANFIS networks for estimation of wind speed and maximum power. The proposed MPPT algorithm does not require any mechanical sensor for wind speed estimation. Tests under rapidly changing wind conditions have shown that MPPT controller is very accurate in maximum power point tracking and yields optimum value of power coefficient (C_p). The feasibility of the proposed controller is validated and the simulation results exhibit the robustness, fast response, and exact wind speed estimation. Comparison with other reported results indicate that the proposed method gives minimum tracking error. The method is also computationally more efficient compared to other methods employing ANFIS.

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Generation of Switching Pulses for Operating a Double Purpose Converter Using Microcontroller

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Abstract—This research is on generating switching pulses for operating a double purpose converter using microcontroller. This paper describes the operating principles of a common circuit which contains two power electronic converter circuits. This common circuit structure can perform controlled rectification and dc chopper actions. Each power electronic conversion circuit requires a particular switching pattern for their desired output. Microcontroller based pulse generation is completely described and simulated in this paper.

Keywords—common circuit structure; pulse generation; double purpose converter; microcontroller.

I. INTRODUCTION

AC to DC converter is known as rectifier [1-3]. Controlled switching is useful to vary and control the output voltage of a rectifier. Rectifier is necessary to get DC power from general power system which is AC in nature [3]. Generally, the applications of rectifier are seen in DC drives, battery chargers and DC power supplies. A low pass filter is used to get pure DC output [1-3].

A buck converter also known as DC chopper is used to convert high DC voltage to low DC voltage [4]. The desired voltage is obtained here by changing width of the switching pulses or duty ratio [4]. The chopper is also used in DC drives, battery charger and DC power supplies like rectifiers.

These widely used two converters can be combined in a common circuit structure [1]. The pattern of conversion can be changed using different switching pulses and techniques [1]. The controlling of switching pulses is very important to operate this converter. So, a controller can be designed using different digital logic equations for the proposed research.

The control signals for operating the double purpose converter are generated using microcontroller in this research. This converter can be termed as a double purpose converter, because it operates both as controlled rectifier and DC chopper. The operating principle of the double purpose converter and the pulse generation by microcontroller are explained here. All necessary simulations were performed using Arduino, Protues and PSIM.

II. OPERATING PRINCIPLES AND SWITCHING PULSES

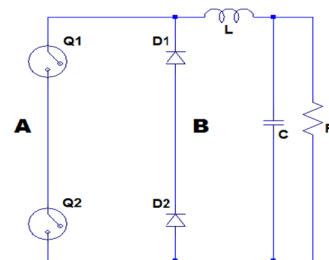


Fig. 1. Common circuit structure.

A common circuit structure shown in Fig.1 can be used as controlled rectifier and DC chopper [1]. For controlled rectification, an AC source is connected between A and B terminals. The switches and diodes are operated for controlled rectification.

Same structure is used as a DC chopper in which the DC source is connected between A and B terminals. Q1, D2 and D1 operate in case of DC to DC conversion. Switch Q1 will be operated by PWM signals.

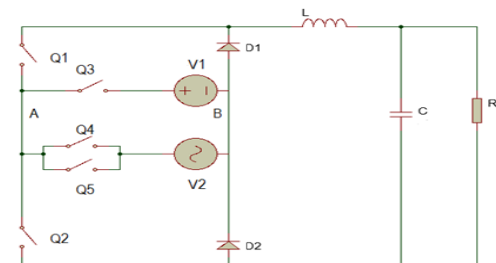


Fig. 2. Circuit diagram of the double purpose converter.

The circuit diagram of double purpose converter is shown in Fig.2. Q3, Q4 and Q5 are isolation switches. These switches are used to isolate the two conversion processes. Q1, Q2, Q4, Q5, D1 and D2 will be operated for controlled rectification and for DC chopping circuitry Q1, Q3, D1 and D2 will be in operation.

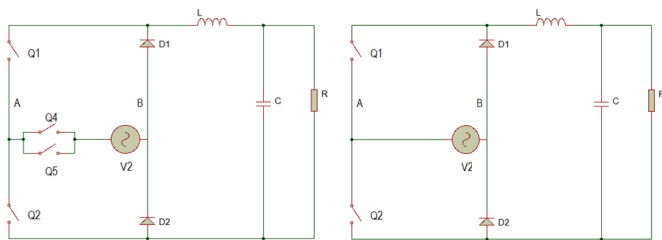


Fig.3. (a) Switch controlled rectifier circuit and (b) The equivalent switch controlled rectifier circuit.

If Q3 of Fig. 2 is out of operation then, the equivalent circuit acts as a controlled rectifier with isolation switches Q4 and Q5 (Fig.3(a)). The isolation switches will be in action for each half cycle of the sinusoidal input. So, they can be treated as short circuit at conduction state. The equivalent circuit indicating exact rectifier is shown in Fig.3(b). Q1 and Q2 are operated by a particular PWM. In rectification purpose, an AC source is placed between A and B terminals. In positive cycle of AC input Q1 and D2 are operated and in negative cycle of input D1 and Q2 are operated. The rectified voltage is available across the load (R). The LC filter is used to reduce the ripple of output voltage [5-7].

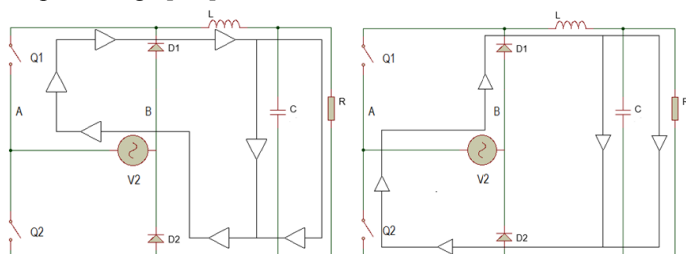


Fig.4. (a) Rectifier circuit in positive cycle of input and (b) Rectifier circuit in negative cycle of input.

Q1 and D2 are operated in positive cycle of the input. Fig.4(a) shows the direction of current flow. At negative cycle Q2 and D1 are operated. Fig.4(b) shows the direction of current flow in negative cycle. The output is unidirectional in every case. So, this action proves the rectifying action.

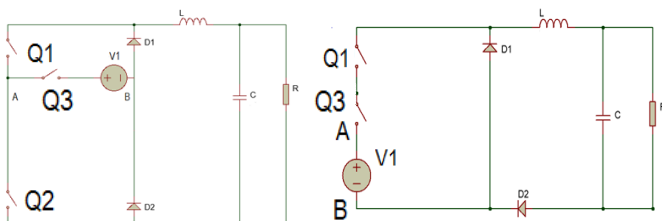


Fig.5. DC chopper in the common circuit structure.

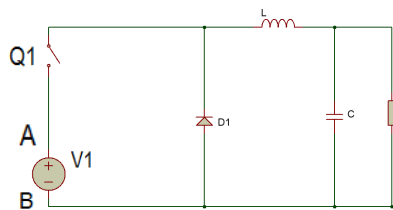


Fig.6. The equivalent buck chopper circuit.

If Q2, Q4 and Q5 of Fig. 2 are out of operation then, the equivalent circuit behaves as DC chopper with isolation switch

Q3 (Fig.5). Q3 will be in continuous operation in this mode. Fig.5 shows the DC chopper with an extra isolation switch Q3 and diode D2. In conduction state, diode D2 and switch Q3 can be replaced by short circuit. The corresponding equivalent circuit will be exactly like a buck converter (Fig.6) [4].

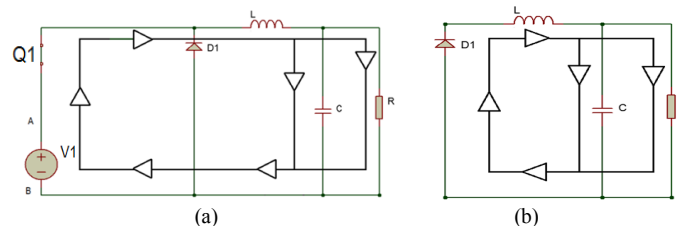


Fig.7. The operation of buck converter (a) ON switch and (b) OFF switch.

In case of DC chopper mode of operation the current flows from source (V1) to the load via Q1 (ON), filter inductor L, filter capacitor C. During this ON switch mode of operation the inductor stores energy [4]. The ON state mode of operation is shown in Fig.7. At OFF switch state of operation the stored energy in the inductor flows via L, C, R and D1 (freewheeling diode). This is the operation of a buck converter [4].

After proving all equivalencies, it is important to analyze the desired switching signals.

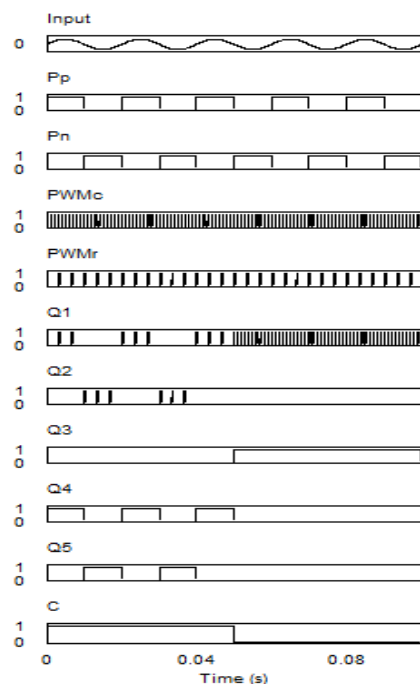


Fig.8(a). The switching signals.

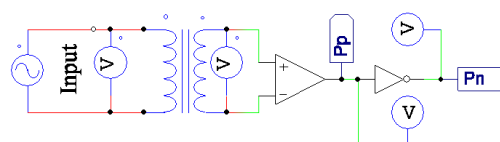


Fig.8 (b). Isolation pulse (Pp and Pn) generation circuit.

Fig. 8(a) shows the desired switching signals for the double purpose converter. It is previously mentioned that, while operating the common circuit structure as a rectifier, synchronization of switching pulses (for Q4 and Q5) with input

Sinusoidal wave is necessary for proper operation of the converter.

A transformer is connected with the input sinusoidal wave so that, a sine wave of small magnitude can be generated for operating a comparator (Fig. 8(b)). From Fig.8(a) it is seen that, the comparator provides positive pulse (Pp) during positive cycle of input. Correspondingly a negative cycle detector signal (Pn) is produced using a NOT gate at the output terminal of comparator. Pn provides high output during negative cycle of AC input.

While operating the converter as a rectifier, Q4 and Q5 act as isolation switches. Isolation of two modes of operation (rectifier and DC chopper) is very important for perfect operation. In this mode of operation control signal C is high and Q3 is out of operation which are shown in Fig.8(a). Q4 operates during positive half cycle of input and Q5 during negative half cycle of input providing isolation. During rectification, Q1 and Q2 are operated with PWMr.

The control signal is low (0) for operating the DC chopper. For chopper action, Q1 and Q3 (isolation switch) are operated and all other switches are at OFF state. Q1 is operated with PWMc. This is shown in Fig. 8(a).

The patterns of switching pulse are explained for the desired operations. Now these switching pulses will be generated using microcontroller.

III. GENERATION OF SWITCHING PULSES FROM MICROCONTROLLER

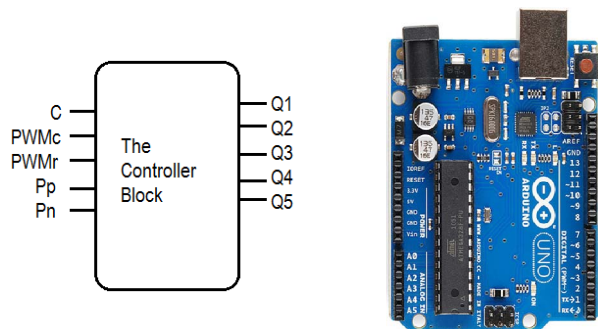


Fig.9. The controller block and microcontroller.

Fig. 9 shows a controller block and microcontroller [8]. In this research the pulse generation will be performed by the microcontroller unit (ARDUINO).

TABLE. I. INPUT AND OUTPUT OF THE CONTROLLER

Inputs		Outputs	
C	Control Signal	Q1	Pulse for Q1
PWMc	PWM signal for chopper	Q2	Pulse for Q2
PWMr	PWM signal for controlled rectification	Q3	Pulse for Q3
Pp	Positive cycle detector	Q4	Pulse for Q4
Pn	Negative cycle detector	Q5	Pulse for Q5

Table.I shows the inputs and outputs of the controller which will operate the converter of Fig. 2. Some digital logic

equations can be developed to design the controller from Fig. 8(a) and Table.I,

$$Q1 = C.PWM_r.Pp + \overline{C}.PWM_c \tag{1}$$

$$Q2 = C.PWM_r.Pn \tag{2}$$

$$Q3 = \overline{C} \tag{3}$$

$$Q4 = C.Pp \tag{4}$$

$$Q5 = C.Pn \tag{5}$$

The necessary code for microcontroller can be written using (1) to (5). The code will determine the pulses for the converter circuit. The necessary code is given below,

```

int q1 = 1;
int q2 = 2;
int q3 = 3;
int q4 = 4;
int q5 = 5;
int c = 6;
int pwmr = 7;
int pwmc = 8;
int pp = 9;
int pn = 10;

void setup()
{
  pinMode(q1, OUTPUT);
  pinMode(q2, OUTPUT);
  pinMode(q3, OUTPUT);
  pinMode(q4, OUTPUT);
  pinMode(q5, OUTPUT);
  pinMode(c, INPUT);
  pinMode(pwmr, INPUT);
  pinMode(pwmc, INPUT);
  pinMode(pp, INPUT);
  pinMode(pn, INPUT);
}

void loop()
{
  boolean cState = digitalRead(c);
  boolean pwmrState = digitalRead(pwmr);
  boolean pwmcState = digitalRead(pwmc);
  boolean ppState = digitalRead(pp);
  boolean pnState = digitalRead(pn);
  boolean q1State;
  boolean q2State;
  boolean q3State;
  boolean q4State;
  boolean q5State;
  // and
  q1State = (cState & pwmrState & ppState) | ((!cState) & pwmcState) ;
  q2State = cState & pwmrState & pnState ;
  q3State = !cState;
  q4State = cState & ppState;
  q5State = cState & pnState;
  digitalWrite(q1, q1State);
  digitalWrite(q2, q2State);
  digitalWrite(q3, q3State);
  digitalWrite(q4, q4State);
  digitalWrite(q5, q5State);
  //digitalWrite(pinOut, pinOutState);
}

```

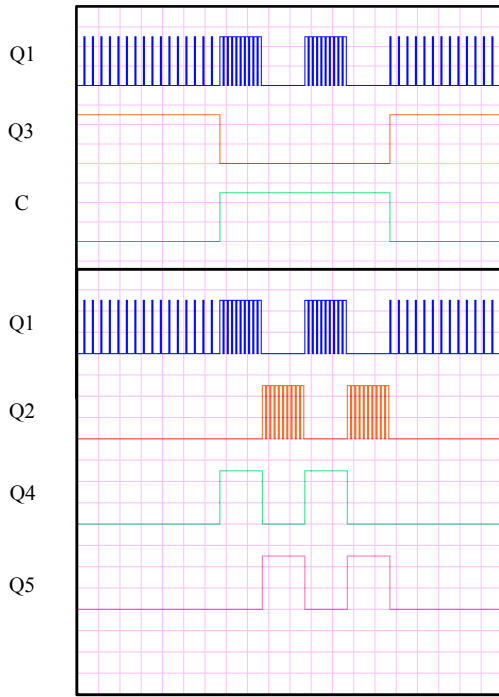


Fig.10. Generated switching pulses by Microcontroller.

Fig. 10 shows the generated switching pulses from microcontroller. In Fig.10 one unit of vertical axis is 2 V. The pulse generation is similar with the desired pulses of Fig. 8(a).

It is previously stated in the explanation of Fig. 8(a), High control signal C operates the circuit of Fig.2 as a switch controlled rectifier. In this mode of operation all switches (except Q3) operate. This is also seen in Fig. 10.

For low control signal C , the circuit of Fig.2 behaves like a DC chopper. According to previous discussion only switch Q1 and Q3 operate. The pulse generation for Q1 and Q3 is justified in Fig.10 under low state of control signal C .

So, all necessary pulse generations using microcontroller are justified in this section. The simulation is performed in Proteus. Although, this research is on pulse generation for the double purpose converter, the output voltage is shown in next part for complete analysis.

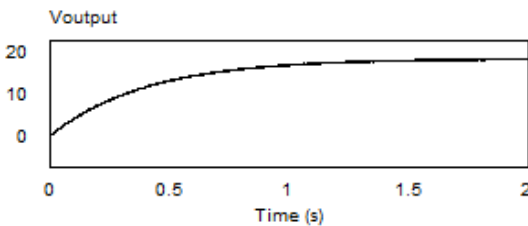


Fig.11. Output voltage of rectifier (C=1).

Fig.11 shows the output voltage of rectifier. The output is 18 V. The equation for calculating output voltage is given below [1-2],

$$V(average) = \frac{2V_p}{\pi} \cdot \sin\left(\frac{\pi D}{2N}\right) \sum_{n=1}^N \sin\left(\frac{\pi}{N}(n-1 + \frac{D}{2})\right) \quad (6)$$

In (6), D (0.10) is the duty cycle, V_p (311 V) is the peak value of input voltage. N (3) is the number of switching pulses and n is integer. The frequency of input is 50 Hz and the switching frequency is 300 Hz (frequency of PWM_r). Output power is 324 W.

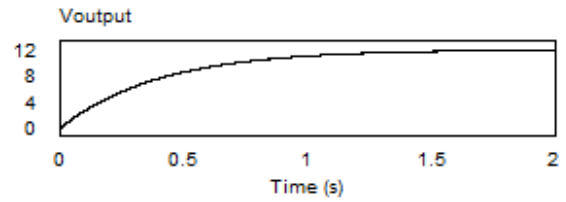


Fig.12. Output voltage of DC chopper (C=0).

Fig.12 shows the output voltage of DC chopper. According to design, the control signal is low. The output voltage is 12 V. The equation for output voltage is given as [4],

$$V(average) = D \cdot V_1 \quad (7)$$

Where, D (duty cycle) is 12% and V_1 (DC input) is 100 V. Switching frequency is 1 KHz (frequency of PWM_c). The output for Fig.2 is obtained from PSIM simulation.

IV. CONCLUSION

Microcontroller is widely used versatile device. This research demonstrates generation of switching pulses for operating a double purpose converter utilizing microcontroller. This unique circuit can be operated both as rectifier and DC chopper depending on the generated control signal from microcontroller. The proposed compact design can be used in both DC and AC systems to get controlled DC output voltage.

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Design & Simulation of Grid Connected Photovoltaic System using Simulink

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Abstract-

The paper proposes an up to date design and simulation of a grid connected photovoltaic system using Simulink. A Photovoltaic (PV) cell, a DC/DC boost converter and a DC/AC inverter constitutes the system. The internal mechanism of solar cell with diagram & approximation of PV cell are described. The PV cell model is simple, precise, and takes external temperature as well as solar radiation into consideration. It also analyzes different types of maximum power point tracking (MPPT) methods with algorithm. To track the maximum power of PV cell, the algorithm conjoined in a DC/DC converter is used. Finally, by using DC/AC inverter, the output voltage of DC/DC converter is regulated and synchronized with the grid. Simulation results show that the proposed model can effectively realize the actual physical characteristics of a grid-connected PV system by matching the phase of grid voltage to generated photovoltaic current.

Keywords:

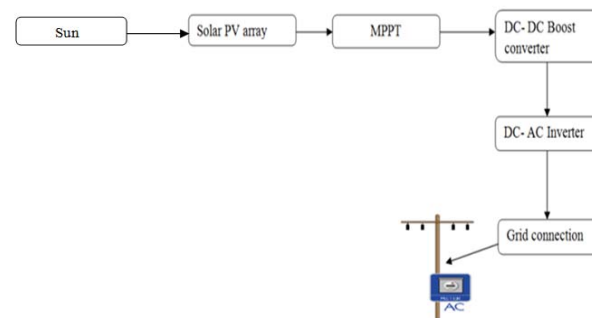
Photovoltaic system, MPPT, DC/DC Converter, DC/AC inverter, Grid connection, Simulink.

I. Introduction

Comparing with others renewable energy sources, solar energy is the largest contributors to world energy for its clean and unlimited supply characteristic. Historically, though the United States has been the leader of installed photovoltaics for many years but China is expected to keep on its rapid growth and to triple its capacity to 70,000 megawatts by 2017, becoming the world's leading producer of photovoltaic power any time soon. By the end of 2014, cumulative capacity of photovoltaic reached at 178 Giga Watts (GW), which was enough to meet 1 percent of global electricity demands. In Italy and Germany, the contribution of solar energy is reached to 7.9 percent and 7.0 percent of their annual domestic consumption respectively. By the end of 2015, forecasted worldwide expansion is about 500 GW, and installed capacity is predicted to more than double or even triple beyond 500 GW between now and 2020. By 2050, solar power is predicted to be become the world's leading source of electricity, with solar photovoltaic's along with concentrated solar Power contributing 16 and 11 percent respectively [1].

II. Photovoltaic System

Flow Diagram



Photovoltaic system consists of solar PV array where sun light will be converting into electrical energy. From the PV array by the use of a MPPT device try to reach MPP, which will produce maximum power. In order to connect it with the grid we need to raise the magnitude of the voltage which will be done by the use of a boost converter, from which the output of DC-DC converter will be invert into AC by the use of DC-AC inverter for grid connection [2].

Modeling of PV cell, MPPT algorithm, DC/DC converter and DC-AC conversion for grid connection organized as follows:

III. Photovoltaic Cell

PV cell is a semiconductor device that absorbs and converts the energy of photon comes from sunlight into electricity.

In the ideal PV Cell model, the parameter related to cells temperature and radiation intensity such as I_{LG} , I_{OS} , R_{sh} , and R_s are very difficult to be determined. Due to this engineering application become very challenging. Some experimental technical parameters such as open-circuit

Voltage V_{OC} , short-circuit current I_{SC} , the maximum power point current I_m , the maximum power point voltage V_m and the maximum power point power P_m are provided by Manufacturers of PV arrays [3].

The equivalent circuit of the ideal PV cell is shown in figure 1.[5]

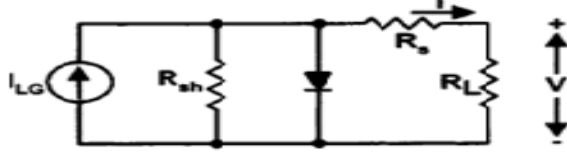


Fig 1: The Equivalent Circuit of the Ideal PV cell

$(V+IR_s)/R_{sh}$ can be ignored compared with Photon current cause R_{sh} is between 100Ω and 10000Ω .

$$I = I_{LG} - I_d - I_{R_{sh}} \\ = I_{LG} - I_{os} \left\{ \exp \left[\frac{q}{AkT} (V + IR_s) \right] - 1 \right\} - \frac{V + IR_s}{R_{sh}} \quad (1)$$

Where,

$$I_{os} = I_{or} \left[\frac{T}{T_r} \right]^3 \exp \left[\frac{qE_{Go}}{Bk} \left(\frac{1}{T_r} - \frac{1}{T} \right) \right] \quad (2)$$

$$I_{LG} = [I_{SCR} + K_l(T - 298)] \frac{S}{1000} \quad (3)$$

$$I = I_{SC} \left\{ 1 - C_1 \left[\exp \left(\frac{V}{C_2 V_{OC}} \right) - 1 \right] \right\} \quad (4)$$

At the maximum power point, when $V = V_m$, $I = I_m$,

$$I_m = I_{SC} \left\{ 1 - C_1 \left[\exp \left(\frac{V_m}{C_2 V_{OC}} \right) - 1 \right] \right\} \quad (5)$$

Under normal temperature,

$$\exp \left(\frac{V_m}{C_2 V_{OC}} \right) \gg 1, \quad C_1 = \left(1 - \frac{I_m}{I_{SC}} \right) \exp \left(\frac{-V_m}{C_2 V_{OC}} \right) \quad (6)$$

$$0 = I_{SC} \left\{ 1 - \left(1 - \frac{I_m}{I_{SC}} \right) \exp \left(\frac{-V_m}{C_2 V_{OC}} \right) \left[\exp \left(\frac{1}{C_2} \right) - 1 \right] \right\} \quad (7)$$

Under normal temperature,

$$\exp \left(\frac{1}{C_2} \right) \gg 1, \quad C_2 = \left(\frac{V_m}{V_{OC}} - 1 \right) / \ln \left(1 - \frac{I_m}{I_{SC}} \right) \quad (8)$$

We took the standard condition as temperate 25°C , 1000W/m^2

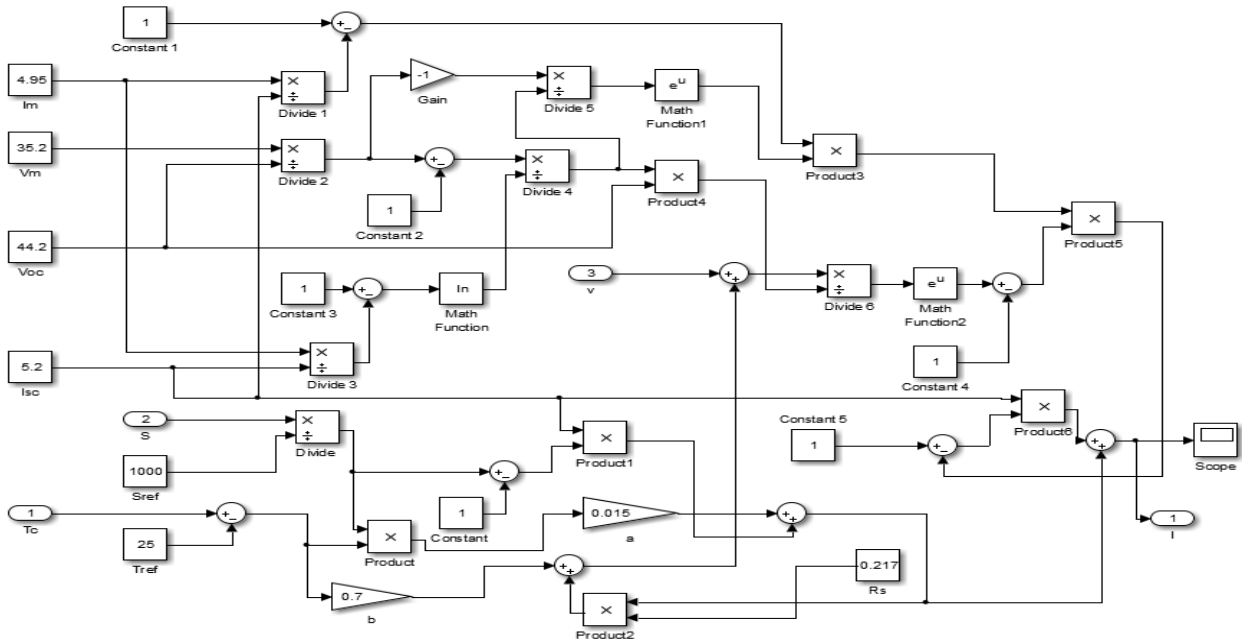


Fig 2: Simulation structure of PV model in Simulink

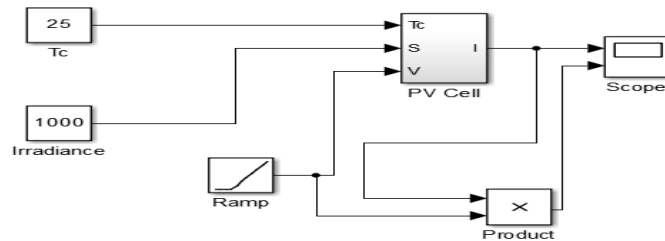


Fig 3: Packaged model of the PV cell

Table I: Symbol related to PV Cell with meaning

Symbol	Meaning of the symbol
V	PV Cell output voltage (V)
I	PV Cell output current (A)
I_{LG}	Photon current (A)
I_{OS}	Reverse saturation current of PV cell's (A)
T	Operating temperature of PV cell ($^{\circ}$ K)
Q	Charge of an Electron = 1.6×10^{-19} C
K	Boltzmann's constant = 1.38×10^{-23} J/ $^{\circ}$ K
K_1	Temperature coefficient
I_{SC}	Short circuit current (A/ $^{\circ}$ K)
S	Operating solar radiation (W/m ²) Short circuit current at STC (A)
E_{GO}	Band gape energy of the semiconductor (J)
A, B	Ideality constant between 1 and 2
Tr	Absolute temperature at Standard Test Condition of PV cell = 301.18 $^{\circ}$ K
I_{OR}	Reverse saturation current of PV cell's at temperature Tr (A)
R_{Sh}	Intrinsic parallel resistance of PV cell (Ω)
R_S	Intrinsic series resistance of PV cell (Ω)

Table II: Parameters of PV cell

Parameter	Symbol	Value
Voltage at maximum power	V_m	35.20 V
Current at maximum power	I_m	4.95 A
Open circuit voltage	V_{OC}	44.20 V
Short circuit voltage	I_{SC}	5.20 A
Temperature coefficient of short circuit current	a	0.015 A/ $^{\circ}$ K
Temperature coefficient of open circuit voltage	b	0.70 V/ $^{\circ}$ K
Internal series resistance	R_s	0.217 Ω
Reference solar radiation	S_{ref}	1000 W/m ²
Reference temperature	T_{ref}	25 $^{\circ}$ C

The simulation result in Figure 4 shows the I-V characteristic & Figure 5 shows the P-V characteristic for the PV cell.

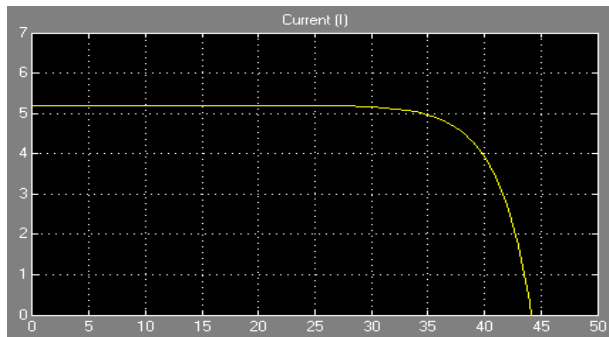


Fig 4: I-V Characteristic curve for PV cell

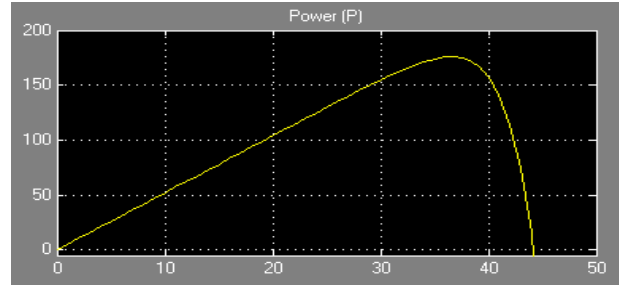


Fig 5: P-V Characteristic curve for PV cell

IV. MPPT Algorithm Simulation in Simulink

In each solar cell there is a point where it can produce maximum power and MPPT utilizes several methods to extract that point. In our paper we applied the perturb & observe (P&O) method [4,6] to extract that point, The simulation model of P&O MPPT algorithm is given in Figure 7 .

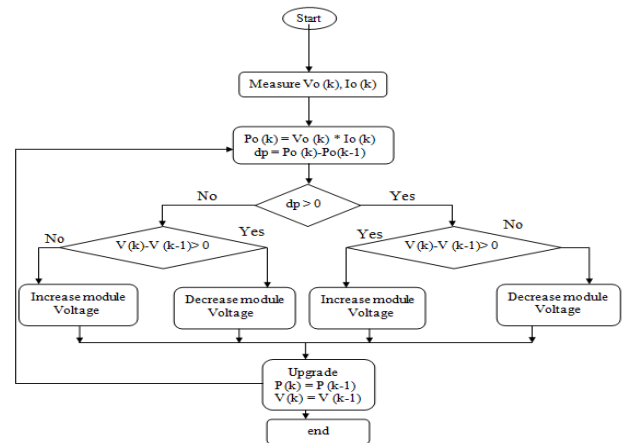


Fig 6: Flowchart of P&O Method

Simulink model of this Perturb & Observe method is given below:

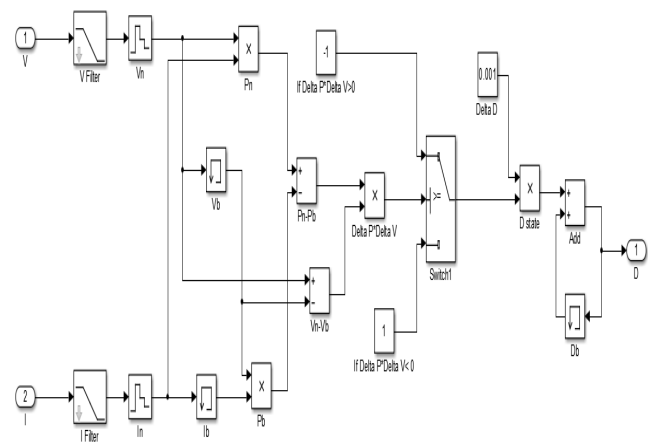


Fig 7: Simulation Model for P&O MPPT

V. DC-DC Converter

The voltage we that have generated using the PV panel, in order to connect it to the Grid we have to increase the magnitude of the voltage. Which can be done with the use of a DC-DC converter, for our purpose we have chosen Boost converter which will help us to increase the level of voltage that we have generated

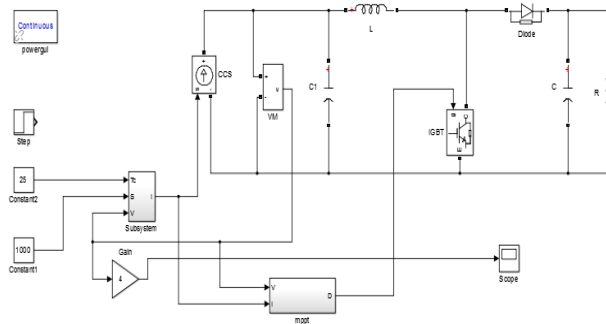


Fig 8: Simulink Model for DC/DC Converter

Table III: Parameters for DC/DC Converter model

Parameter	Symbol	Value
Inductor	L	0.01 H
Capacitor	C1	0.002 F
Capacitor	C	0.002 F
Resistor	R	500 Ω

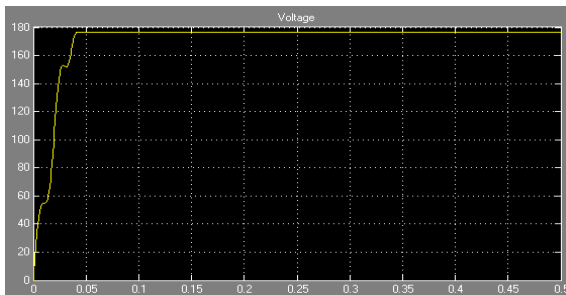


Fig 9: Boost Converted Voltage

VI. DC/AC Inverter

To supply the generated DC output voltage to Grid we need to convert the DC output to AC. To do so we have to use a DC-AC inverter. Before connecting it to the Grid-

- The output voltage from DC-AC inverter is must be sinusoidal and in phase with the grid voltage.
- Current-mode control is used to control the DC/AC inverter. The DC/AC inverter output current is compared with reference current i^* .
- To generate signals for a PWM generator, the compared result is sent to a discrete PI controller.

In PI controller

T	5×10^{-7}
K_p	6.00
K_i	256

In Grid we have

Peak amplitude	220 V
Phase	90
Frequency	50 Hz

In the LR filter

L1	10 mH
R	0.01 Ω

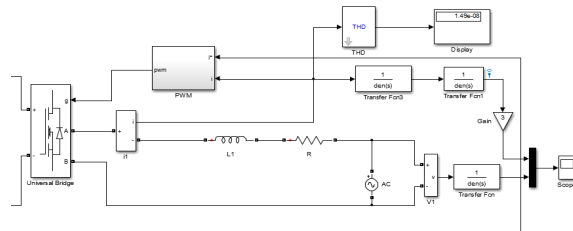


Fig 10: DC-AC Inverter with Grid

The structure of single phase controller used in fig 10 is given in fig 11

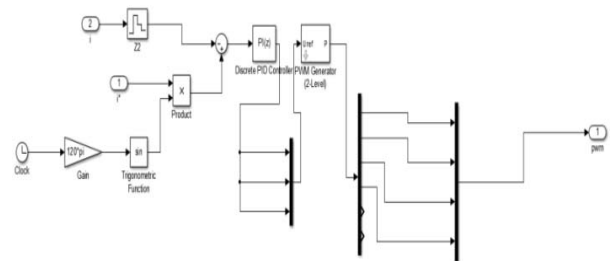


Fig 11: Single Phase Controller

When the temperature and solar radiation keeps unchanged, $T=25^\circ$, $S=1000 \text{ W/m}^2$, which is the standard condition, we get the output as below

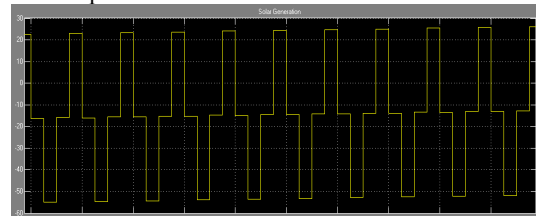


Fig 12: i-t PV Generated Output Wave Form

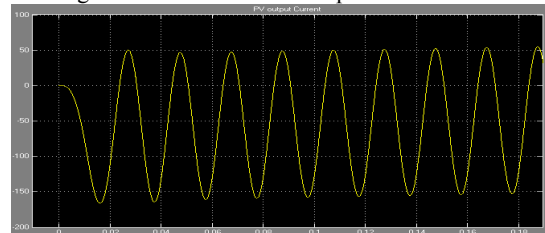


Fig 13: Modified i-t PV Generated Output Wave Form

VII. Grid Connection

The Grid connected PV System Simulink diagram is given below

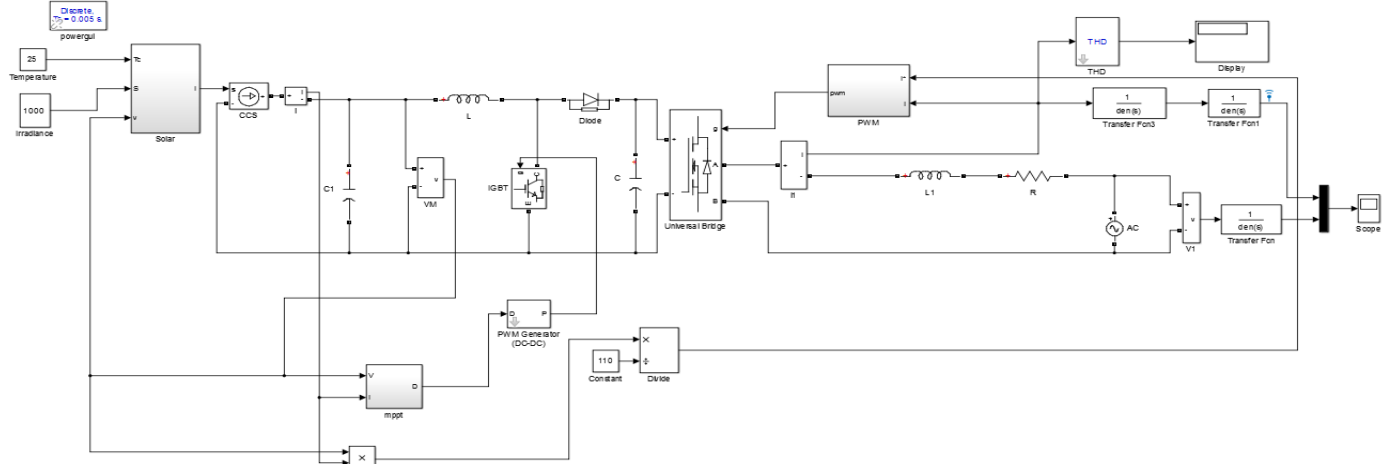


Figure 14: Whole PV Grid-Connected System

In order to convert PV generated output waveform into Sine waveform we have used a low pass filter

A. Low Pass Filter Calculations

Using RLC circuit we can convert rectangular waveform to sinusoidal waveform [7]

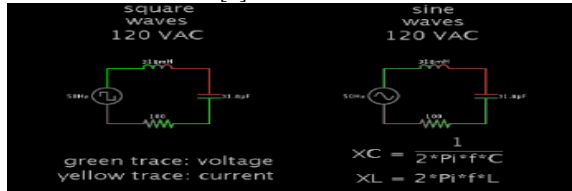


Fig 15: RLC circuit for sinusoidal conversion

Using this we get the Transfer function as

$$TF = 1/(LCS^2 + RLS + 1)$$

$$= 1/[0.000010112 S^2 + 0.00318S + 1]$$

Finally when we compare the PV Generated Current with the grid Voltage we get,

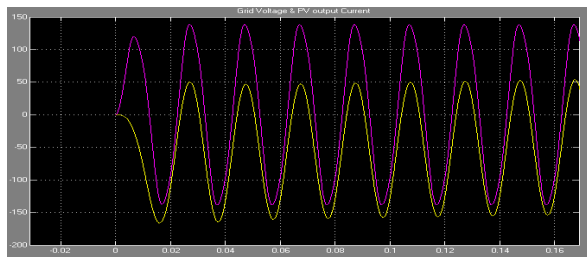


Fig 16: PV Current with the Grid Voltage

Results

Form the fig 16 which is the comparison between PV Current & Grid Voltage, we can see the generated PV current is leading to that of the Grid voltage by Φ for a minor fraction

of time (0.03s), for this period of time reduce its current waveform equation to $\sin(\omega t - \Phi)$ where as the grid voltage waveform is $\sin(\omega t)$. As soon as the time reaches 0.03s the generated PV currents gets 100% in phase with the Grid Voltage, producing an accurate & efficient results.

VIII. Conclusion

Mathematical design & simulation of photovoltaic system is presented, which can perform its required task under various temperature & irradiance. For a small fraction of time this generated current hence power was not in phase with the grid voltage but after 0.03s it was 100% in phase with the Grid & continued thereafter which was our ultimate aim of this thesis paper.

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Performance Improvement of PID Controller for AVR System Using Particle Swarm Optimization

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Abstract—A novel time-domain fitness function for determining the parameters of an optimum proportional–integral–derivative (PID) controller for an automatic voltage regulator (AVR) system using Particle Swarm Optimization (PSO) Algorithm approach is presented. An algorithm of employing proposed approach to minimize the maximum percentage overshoot, the rise time, the settling time, and the steady-state error of the terminal voltage of the synchronous generator is presented. From the MATLAB simulation results, it can be settled that the proposed approach has superior features, including fast tuning of PID parameters, stable convergence characteristic, and good computational efficiency. The results obtained from proposed PSO-PID controller are compared with that of estimated by genetic algorithm (GA) and PSO with different fitness functions and is found to have better performance. The outcome of this paper would be a boost to design robust AVR system with improved step-response.

Keywords—AVR system, Particle swarm optimization, PID controller, Design optimization, Step response

I. INTRODUCTION

Synchronous generators have a significant role in providing energy for electrical networks. Their efficient control is crucial as they are mostly responsible for maintaining stability and security of the power systems. An Automatic Voltage Regulator (AVR) is used in conjunction with synchronous generator to regulate reactive power and voltage level under normal operating conditions at various load levels [1]. The AVR controls the terminal voltage by adjusting the exciter voltage of the generator. During the past decades, numerous control methods such as adaptive control, fuzzy logic and neural network have been developed to improve the efficiency of process control in the industries. In despite of many efforts, the proportional–integral–derivative (PID) controller continues to be the main component for the AVR system because of its simple structure and robust performance within a wide range of operating condition [2]–[4].

Though it is popular, it has been quite difficult to tune properly the gains of PID controllers because many industrial plants are often burdened with problems such as high order, time delays, and nonlinearities. Several methods have been developed in literatures for determining the PID parameters. The first method is the Ziegler Nichols method [3–4]. In general, it is difficult to determine optimal PID parameters with the Ziegler-Nichols formula in many industrial plants and the designer has to depend on his experience for obtaining the best performance. Artificial intelligence (AI) techniques such as neural network, fuzzy system, and neural-fuzzy logic have been used also for proper tuning. [3], [5], [6]. But, the artificial neural networks suffer from the convergence time and the length of the training process. Also, the fuzzy logic systems depend on the experience of the designer. Furthermore, evaluation and tuning of robust PID controllers have been proposed by using a lambda tuning approach [7]. Then, different random search methods, for example, genetic algorithm (GA), simulated annealing (SA) and Chaotic

algorithm have been introduced to get the optimal design of the PID controller [8]–[10]. Though the GA methods have been implemented successfully to solve complex optimization problems, the crossover and mutation operations are unable to ensure better fitness. This is because chromosomes in the population have similar structures and their average fitness is high toward the end of the evolutionary process [11].

The PSO algorithm, proposed by Kennedy and Eberhart [10] in 1995, is another popular optimal algorithm. [12]. It was developed through simulation of a simplified social system, and has been found to be robust in solving continuous nonlinear optimization problems. The PSO technique can generate a high-quality solution within shorter calculation time and stable convergence characteristic than other stochastic methods [11]–[15]. In 2004, Gaing presented the PSO algorithm to optimize PID controller parameters of an AVR system [16]. Many researchers provided many improvement methods to improve the search performance of the PSO-PID controller where different fitness function has been used [17–20]. Unfortunately; those have the disadvantage of slow convergence and poor performance.

In this paper a new fitness function of the PSO-PID controller for high-order AVR system in time domain is introduced to overcome the disadvantages of previous studies. Parameters of proposed approach can yield a good step response that ensures the optimization of performance criteria such as rise time, settling time, overshoot and steady-state error.

This paper has the following discussions. Section II discusses the algorithm of particle swarm optimization and the proposed fitness function. The AVR model and the technique used in the optimal design of PID parameters are discussed, and the results are presented in Section III. This section also focuses on the advantages of the proposed function in comparison to the previous similar work. Conclusively, a summary is depicted in Section IV.

II. PSO METHOD

A. PSO Algorithm

A PID controller using the PSO algorithm is developed to improve the step transient response of AVR of a generator. The PSO algorithm is used to determine three optimal controller parameters k_p , k_i and k_d to obtain a good step response output of the system. Each individual k contains three members k_p , k_i and k_d . The matrix representation of the initial population with its dimension is $n \times 3$ [16].

The optimization method of the PSO-PID controller is shown below:

1. Specify the boundary of the three controller parameters and initialize randomly the individuals of the population including searching points, velocities, pbests and gbest.
2. Find out the values of the four performance criteria in the time domain, namely M_p , E_{ss} , t_s and t_r .

3. Calculate the evaluation value of each individual and compare each individual's evaluation value with its pbest. The best evaluation value among the pbest is denoted as gbest.

4. Modify the member velocity v of each individual k according to the equation (1):

$$v_{j,g}(t+1) = w \cdot v_j(t) + C_1 \cdot \text{rand}() (pbest_{j,g} - k_{j,g}(t)) + C_2 \cdot \text{rand}() (gbest - k_{j,g}(t)) \quad (1)$$

Where,

$j=1,2,\dots,n$ and $g=1,2,3$

$C_1=C_2=\text{constant}$

$\text{rand}()$ is the random number.

$$\text{And } w = \text{inertia weight} = w_{\max} - \frac{w_{\max} - w_{\min}}{\text{iter}_{\max}} \times \text{iter} \quad (2)$$

Where $\text{iter}_{\max}$ is the maximum number of iterations (generations), and iter is the current number of iterations [16].

5. If $v_{j,g}(t+1) > V_g^{\max}$ then $v_{j,g}(t+1) = V_g^{\max}$

If $v_{j,g}(t+1) < V_g^{\min}$ then $v_{j,g}(t+1) = V_g^{\min}$

6. Find the updated position of each individual k according to (3)

$$k_{j,g}(t+1) = k_{j,g}(t) + v_{j,g}(t+1) \quad (3)$$

$$k_g^{\min} \leq k_{j,g}(t+1) \leq k_g^{\max}$$

7. This process continues and the individual that generates the latest $gbest$ is an optimal controller parameter.

B. Proposed Fitness Function

The proposed fitness function is defined as follows:

$$J(k) = p(M_p + e) + q(t_s - t_r) \quad (4)$$

Where p & q are weighting factor and

$$e = \sqrt{\frac{\sum_0^n E_{ss}^2}{n}} \quad (5)$$

Here, e is the root-mean-square of steady-state error over the generation (n). Using the fitness function we can control the values of M_p , t_r and t_s . The most important parameter in this fitness function is e . Here, this parameter influences the other parameters to be optimum. For example, to have a minimum e the other parameters have to be in their optimal values. *i.e.* We would prefer to minimize M_p , t_r and t_s directly in the function and indirectly by minimizing e . Firstly, M_p , t_r and t_s are initialized randomly and e as zero. Then we get a value of the fitness function $J(k)$. This value uses in the AVR model introduces the steady state error. But we have to make the error zero. In each step, we take root mean square of the error and compare with the previous one to go towards the zero. If we take average or squared value of error there is a possibility of divergence of the error from the zero. It is required to modify the value of M_p , t_r & t_s which in turn modify the $J(k)$. On the other hand, $J(k)$ specifies the output of the controller. So, the convergence or divergence of e is strictly bound with the output performance of the PID controller. Here the divergence of error from zero is rejected and only the convergence is taken in account. When the steady state error of the AVR becomes zero we can consider the parameters of the PID controller optimal.

III. SIMULATION RESULTS

A. AVR Model

A simple AVR system consists of four main components, *viz.*, amplifier, exciter, generator, and sensor. The block diagram of an AVR system with PID controller is illustrated in Figure 1.

Mathematical modelling of these components needs to consider their transfer functions linearized. The transfer function of these components and the range of the time constants and gains of all components are given in Table I.

TABLE I
Transfer Function of AVR Components

Component	Transfer Function	Parameter Range
Amplifier	$K_A/(1 + \tau_a s)$	$10 < K_A < 400$, $0.02 < \tau_a < 0.1$ s
Exciter	$K_E/(1 + \tau_e s)$	$10 < K_E < 400$, $0.5 < \tau_e < 1$ s
Generator	$K_G/(1 + \tau_g s)$	$0.7 < K_G < 1$, $1 < \tau_g < 2$ s from full load to no load
Sensor	$K_R/(1 + \tau_s s)$	$0.001 < \tau_s < 0.06$ s

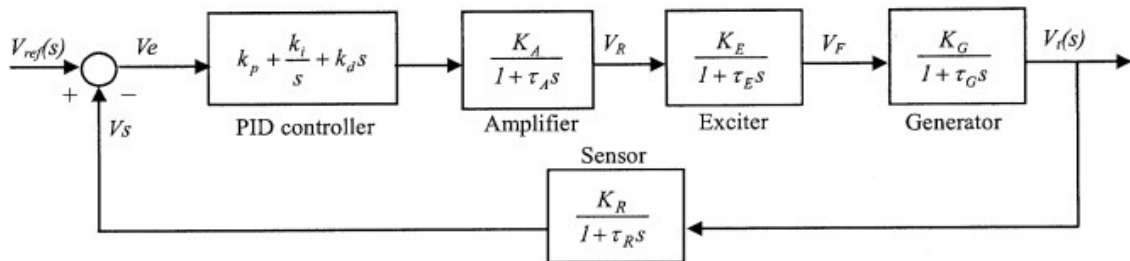


Figure 1: Block Diagram of AVR system with PID controller

The terminal voltage step response of this AVR system without PID controller is shown in Figure 2. By simulating this case with the help of MATLAB, we found that $M_p=65.7\%$, $E_{ss}=0.091$, $t_r=0.261(s)$, $t_s=6.99(s)$.

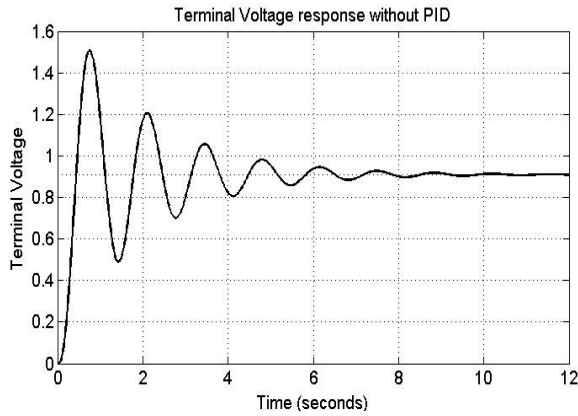


Figure 2: Terminal voltage step response of an AVR system without PID controller

This kind of voltage oscillation and large overshoot is unsafe for power systems stability. So, the terminal voltage should be controlled.

C. Case Study

Consider a system with the following PSO parameters:

- ✓ Population Size= 50
- ✓ Number of generation= 100
- ✓ w can be measured from Equation (2) with the values $w_{max}=0.9$ and $w_{min}=0.4$
- ✓ $C_1=C_2=2$
- ✓ Weight factors $p=1$ & $q=0.25$

Using the above parameters the simulation result that shows the best solution is $[k_p, k_i, k_d] = [0.6478, 0.5245, 0.2388]$. The terminal voltage response of the AVR system i.e. the optimum response is presented in figure 3. This figure shows that $M_p=1.0759\%$, $E_{ss}=0$, $t_r=0.2596(s)$ and $t_s=0.3933(s)$. This result shows that the steady state error becomes zero and the other parameters are improved.

Figure 4 represents the convergence characteristic of the PSO-PID controller. In this figure we can see that through 10 steps, the PSO algorithm can prompt convergence and obtain good evaluation value. This result shows that this PSO-PID controller needs less execution time and can obtain optimal PID controller parameters quickly and efficiently.

The terminal voltage step response of the AVR with PID controller with different weight factors is shown in figure 5. This is a comparison among different system with different parameters.

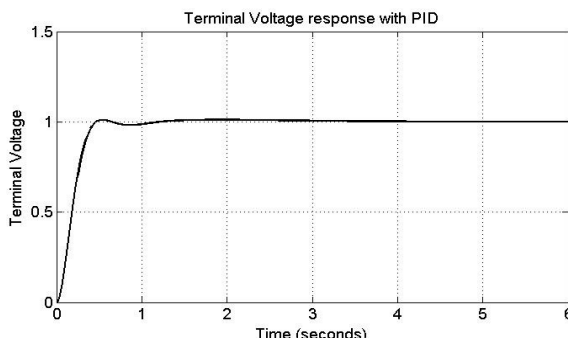


Figure 3: Terminal voltage step response with PID controller

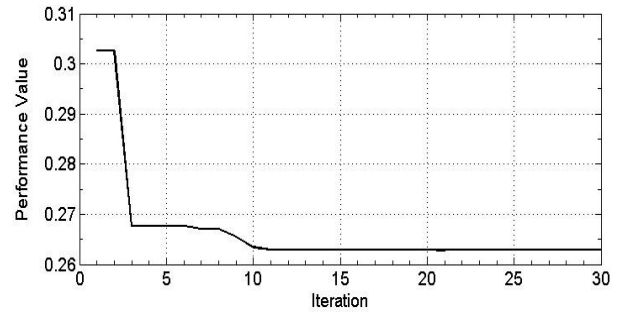


Figure 4: Convergence Tendency of PSO-PID controller

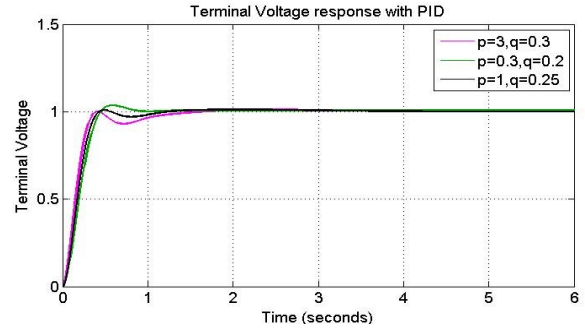


Figure 5: Terminal voltage step response of the AVR with PID controller for different weight factors.

D. Performance Analysis

To observe the efficiency of the proposed PSO-PID controller, the obtained result is compared with the GA-PID controller. In figure 6 the terminal voltage step response of the AVR system for the two different algorithms are shown. We can see from the figure that the PSO-PID controller has a better step response than the GA-PID controller. Table II shows further proof of improved performance of PSO-PID in which PSO shows better percentage overshoot, settling and rise time than that of GA.

In order to emphasize the efficiency of proposed fitness function, we should compare the performance of the fitness functions described in previous studies [16-18]. The proposed function shows better performance than each of the previous ones. The comparison among the function in [16] (Let PSO1) & the function in [18] (Let PSO2) with the proposed function (Let PSO3) is shown in Figure 7 and Table III. Both the figure and the table show that the proposed fitness function has better performance than that of others.

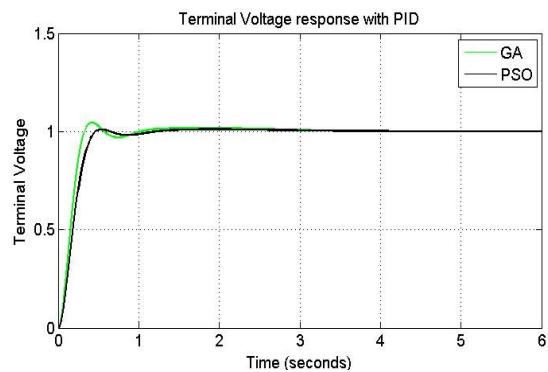


Figure 6: Terminal voltage step response of the AVR with GA & PSO Method (population= 50, number of generation= 100).

TABLE III
Comparison of the performance of GA-PID and PSO-PID

Type	k_p	k_i	k_d	M_p (%)	E_{ss}	t_s (s)	t_r (s)
GA-PID	0.7722	0.7201	0.3196	4.54	0	0.865	0.214
PSO-PID	0.6751	0.5980	0.2630	1.7	0	0.358	0.157

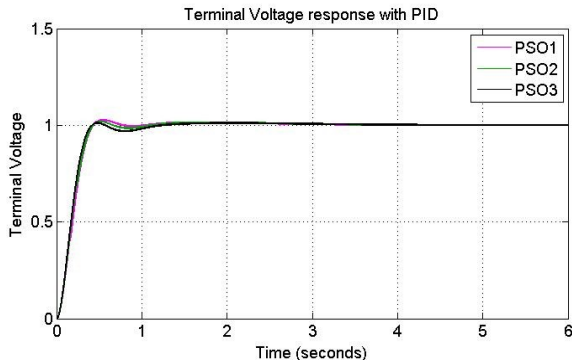


Figure 7: Terminal voltage step response with different fitness

TABLE IIIII
Performance comparison of each function for different system

Parameter	PSO1	PSO2	PSO3
k_p	0.6721	0.6721	0.6778
k_i	0.5709	0.5709	0.5545
k_d	0.2360	0.2497	0.2678
Evaluation Value	4.1843	4.0046	3.7344
Rise Time(s)	0.2718	0.2657	0.2552
Settling Time(s)	0.6060	0.3973	0.3490
Over shoot (%)	2.5432	1.9225	1.3800
Steady State Error((%))	0	0	0

The optimum result is obtained with proposed fitness function $J(k)$ after 10 iterations. On the other hand, the fitness function at [16] and [18] reach their optimum result after 34 and 18 iterations, respectively. The comparison result shows that the proposed fitness function can find an optimal PID control parameters faster than the other fitness functions.

IV. CONCLUSION

In this paper, a PSO algorithm with a novel fitness function are proposed to optimally design a PID controller for the AVR system to improve the step response of terminal voltage. The proportional gain, the integral gain, the derivative gain, and the saturation limit were chosen to define the search space for the optimization problem. The simulation result depicts that the proposed fitness function can let PSO algorithm find a high-quality PID control parameter set with better computation efficiency so that the controlled AVR system has more robust stability and can solve the searching and tuning problems of PID controller parameters more easily and quickly than the other methods. This result regards that the proposed approach would be a boost in raising the precision of optimization of PSO-PID controller for an AVR system.

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A Comparative Analysis of Two Different PCF Structures for Gas Sensing Application

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Abstract— Toxic or Flammable gases and leakage of the gas lines can be detected by photonic crystal fibers for avoiding pollution, explosion or fire. Two different structures of hollow core based Photonic crystal fiber (HC-PCF) has been presented in this paper. The sensitivity of the proposed structures was numerically investigated by finite element method (FEM) varying the optical and geometrical properties. The numerical results show that the proposed PCF with six air holes shows 2.22 times higher sensitivity than the PCF with eight air holes in the first layer. Both of the PCFs exhibit reduced confinement loss also. The proposed PCFs can sense the lower refractive index based gases (toxic/ flammable) at a wide range of wavelength 0.8 μm to 2 μm .

Keywords— Gas sensor; Sensitivity; Confinement loss; Finite Element Method; Toxic and flammable gas sensor

I. INTRODUCTION

In recent years, a wide range of applications of photonic crystal fibers (PCFs) have drawn much more attention to the researchers than conventional optical fibers for its versatile optical properties. Multiple air holes are arranged in the cladding part, which help to confine light in the core by the mechanism of total internal reflection (TIR). PCF attracted much attention for its incredible performance and tremendous applications. PCF gas sensors are smart application in the field of fiber optic technology which already proved it's potentiality to detect flammable and toxic gases in industry. Compared with electric sensor PCF sensors have many advantages like high sensitivity, high safety and low cost [1]. Researchers are working to develop the performance of PCF. For better guiding properties, proper selection of geometric parameters of PCF have great impact. A large number of studies have been reported to enhance sensitivity of PCF on the basis of geometrical parameters.

Different geometries of PCF structures like Hexagonal [2], octagonal [3], decagonal [4], elliptical [5], circular honey comb cladding [6] and so on have been analyzed to achieve better performance. Cregan et al suggested a hollow core photonic crystal fiber (HC-PCF) for gas detection which allows a better confinement of optical mode and gaseous materials simultaneously [7]. HC-PCF is a novel transmission fiber. The guiding mechanism of HC-PCF is photonic band gap (PBG). G.Xin et al. improved sensitivity of PCF for gas detection using a hollow core [8]. Park et al. proposed an index guiding PCF with a hollow high indexed ring defect in the core to enhance the sensing capabilities [9]. M. Morshed et

al. proposed a simple three layers of air holes with a hollow core which enhances the sensitivity and also reduces confinement loss [10]. All of the above discussion proves the potentiality of hollow core based photonic crystal fiber for sensing applications. For sensing gas or chemicals PCF sensors should have good relative sensitivity and also need the ability to confine the propagated signal in the core region properly. To resolve this challenge, we developed and investigated good sensing properties of hollow-core based PCF for gas sensing which will help to further theoretical inquiries.

In this paper, two hollow core based photonic crystal fibers (PCFs) have been presented, where the only difference between the two PCFs is the innermost layer of the cladding part. For the PCF₁ the innermost layer is hexagonal (Fig. 1(a)) and it is octagonal for the PCF₂ (Fig. 1(b)). Both of the PCFs have been analyzed by changing the geometrical as well as optical properties. The proposed structures depicts that the innermost layer (first layer) with six air holes (hexagonal in shape) shows higher sensitivity and lower confinement loss than the another PCF where first layer consists of eight air holes (octagonal in shape). The proposed PCFs can sense gas at a wide range of wavelength 0.8 μm to 2 μm . The background material of the PCFs are pure silica. The simplicity of the structures seems that the fabrication process will be easier as we have not used any material defect at core as the previous gas sensor presented in [11]. These structures were numerically investigated by finite element method (FEM). As well as sol-gel technique [12], the traditional stack and draw technique can be applied to fabricate those PCFs [13].

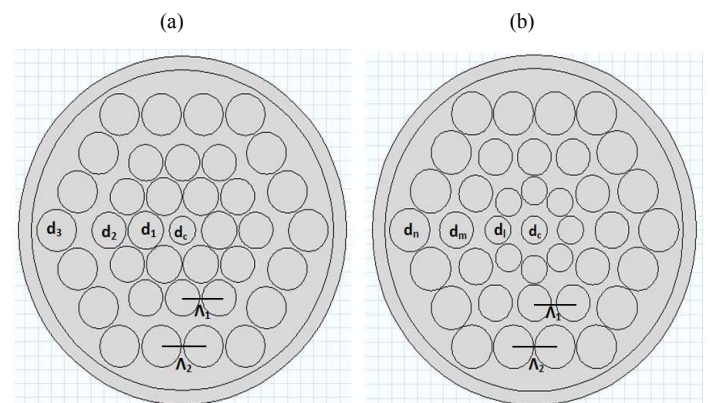


Fig. 1 Cross sectional view of proposed (a) PCF₁ and (b) PCF₂

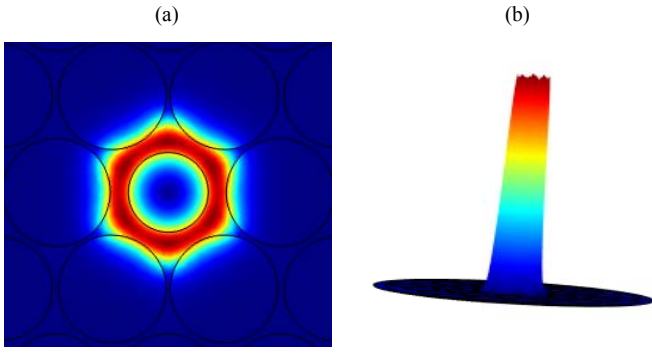


Fig. 2 (a) Modal intensity distribution for PCF₁ with six air holes at first layer (b) 3-D view of modal intensity of PCF₁.

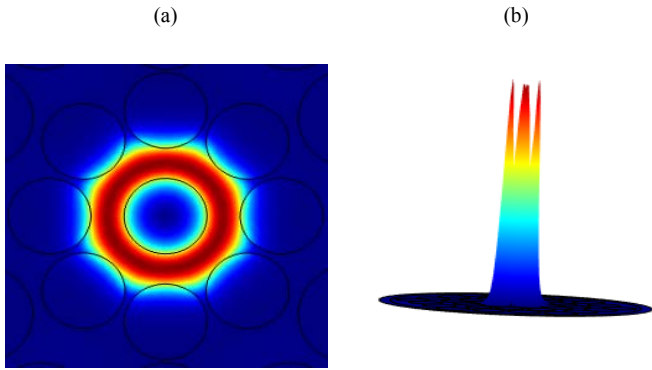


Fig. 3 (a) Modal intensity distribution for PCF₂ with eight air holes at first layer (b) 3-D view of modal intensity of PCF₂.

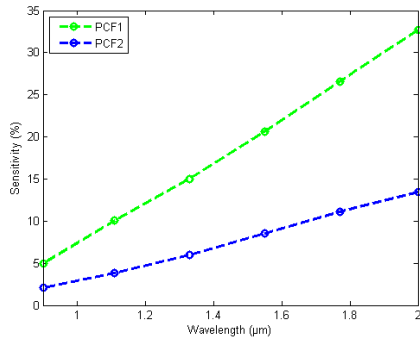


Fig. 4 Comparison of sensitivity of the PCF₁ and PCF₂ for optimized value $d_1=2.7 \mu\text{m}$, $d_3=3.1 \mu\text{m}$, $\Lambda_1=2.8 \mu\text{m}$.

II. GEOMETRICAL STRUCTURE

Fig. 1 (a) and Fig. 1 (b) present the cross sectional view of the proposed two PCF based gas sensors. Both PCFs have three rings of air holes. In PCF₁ all the three rings are hexagonal in shape and in the PCF₂ the innermost ring is octagonal and rest of the two rings are hexagonal in shape. Both of the PCFs have a central hole of $d_c=2 \mu\text{m}$ diameter filled with air in the core. The first ring air holes are arranged in the same distance considered as pitch1 $\Lambda_1=2.8 \mu\text{m}$ from the central hole for both of the configurations where the Λ_1 implies the distance between the two adjacent air holes of the first layer also. Pitch2

(Λ_2) is the distance between the center of two adjacent air holes of outer layer. In the PCF₁ diameter of all holes of the three rings are considered as d_1 , d_2 and d_3 respectively. In the PCF₂ the air hole diameter of the first ring is d_1 , d_m and d_n respectively where $d_m = d_2$ and $d_n = d_3$. The background material of both of the configurations is made of pure silica whose refractive index is selected using Sellmeier equation [14]. COMSOL Multiphysics 4.2 is used here for simulating the proposed PCFs by selecting Fine mode of Mesh size.

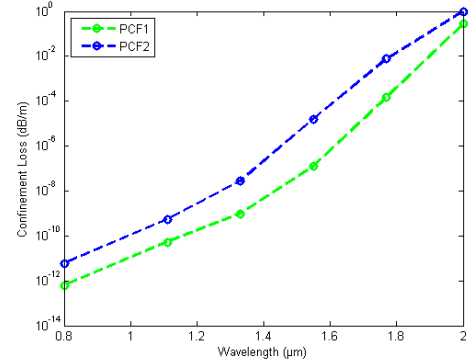


Fig. 5 Comparison of confinement loss of the PCF₁ and PCF₂ for optimized value $d_1=2.7 \mu\text{m}$, $d_3=3.1 \mu\text{m}$, $\Lambda_1=2.8 \mu\text{m}$.

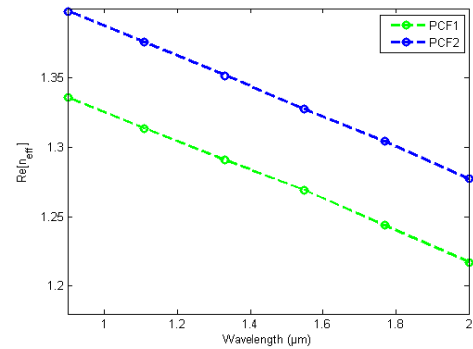


Fig. 6 Comparison of effective refractive index for the PCF₁ and PCF₂.

III. NUMERICAL ANALYSIS

We solved Maxwell's equation because of its high accuracy of analyzing the PCF [15] utilizing Finite Element Method to the perfectly matched boundary layer (PML).

Leakage of light energy from core to exterior matrix material happens when light propagate through the PCF is also known as confinement loss which is an important parameter of the PCF. The confinement loss (dB/m) L_c can be calculated through the following equation.

$$L_c = 8.686 K_0 \text{Im}(n_{eff}) \times 10^6 \text{ (dB/m)} \quad (1)$$

Where, $k_0=2\pi/\lambda$; λ wave length of light and $\text{Im}[n_{eff}]$ is the imaginary part of the effective refractive index.

Sensitivity depends on the relationship between gas concentration and optical intensity. According to the Lambert Beers Law the relationship between gas concentration and optical intensity can be expressed as-

$$C = \frac{1}{r\alpha(\lambda)l} \ln \frac{I(\lambda)}{I_0(\lambda)} \quad (2)$$

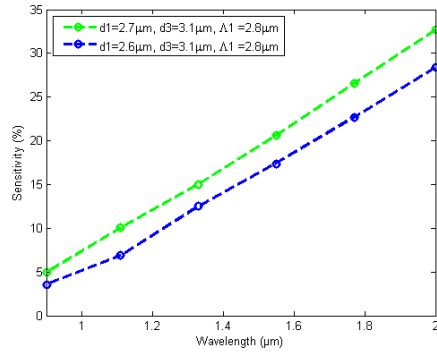


Fig. 7 Comparison of sensitivity varying d_1 keeping other parameters fixed for $d_3 = 3.1 \mu\text{m}$, $\Lambda_1 = 2.8 \mu\text{m}$ for PCF₁.

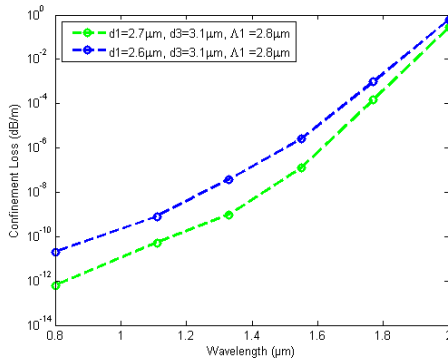


Fig. 8 Comparison of confinement loss varying d_1 keeping other parameters fixed for $d_3 = 3.1 \mu\text{m}$, $\Lambda_1 = 2.8 \mu\text{m}$ for PCF₁.

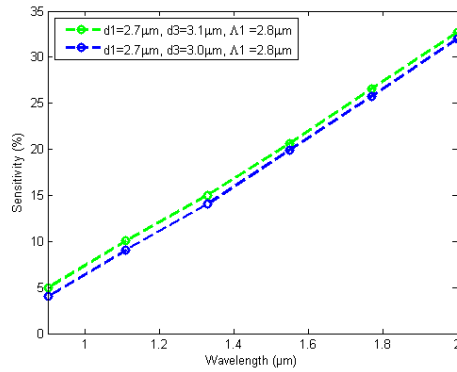


Fig. 9 Comparison of sensitivity varying d_3 keeping other parameters fixed for $d_1 = 2.7 \mu\text{m}$, $\Lambda_1 = 2.8 \mu\text{m}$ for PCF₁.

Where, C is the gas concentration, α is the gas absorption coefficient, l is the length of the PCF, I and I_0 present input and the output light intensity. r is the relative sensitivity coefficient which can be defined as-

$$r = \frac{n_r}{Re[n_{eff}]} f \quad (3)$$

$Re[n_{eff}]$ Is the real part of the effective refractive index of the guided mode; n_r is the refractive index of absorbing material and f is the ratio of the air hole power and the total

power where f is also defined as optical power distribution function.

$$f = \int_{sample} Re(E_x H_y - E_y H_x) dx dy / \int_{total} Re(E_x H_y - E_y H_x) dx dy \quad (4)$$

Where E_x , E_y and H_x , H_y are the transverse electric fields and magnetic fields of the mode. The higher value of f will show the better sensitivity of the PCF. The mode field pattern E_x , E_y and H_x , H_y can be obtained by the COMSOL Multiphysics 4.2 utilizing finite-element method (FEM) [16].

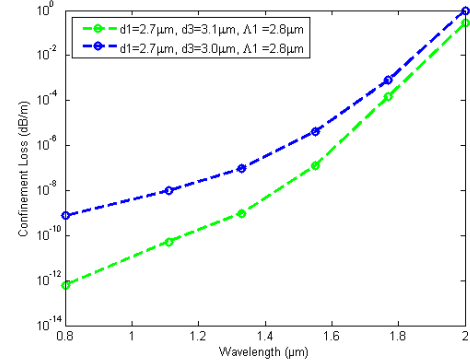


Fig. 10 Comparison of confinement loss varying d_3 keeping other parameters fixed for $d_1 = 2.7 \mu\text{m}$, $\Lambda_1 = 2.8 \mu\text{m}$ for PCF₁.

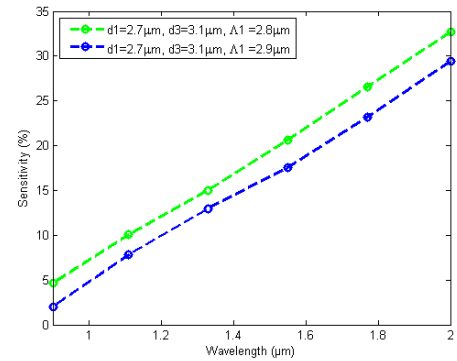


Fig. 11 Comparison of sensitivity varying Λ_1 keeping other parameters fixed for $d_1 = 2.7 \mu\text{m}$, $d_3 = 3.1 \mu\text{m}$ for PCF₁.

IV. RESULTS AND DISCUSSION

In this section the comparison between PCF₁ (first layer with six air holes) and PCF₂ (first layer with eight air holes) has been analyzed. The geometrical properties like first layer diameters second layer diameters has been numerically analyzed for the PCF₁ as we got better results for sensitivity and confinement loss for PCF₁ over PCF₂.

The modal intensity distribution (Electric field) of the PCF₁ and PCF₂ has been shown in Fig. 2 and Fig. 3 which demonstrates that at the core region the modal intensity distribution is tightly confined.

Fig. 4 and Fig. 5 show the comparison between PCF₁ and PCF₂ for optimized value $d_1=2.7 \mu\text{m}$, $d_2=2.6 \mu\text{m}$, $d_3=3.1 \mu\text{m}$ and $\Lambda_1=2.8 \mu\text{m}$. From Fig. 4 it is clear that PCF₁ shows

higher sensitivity than PCF₂. Fig. 5 shows that the PCF₂ shows higher confinement loss than PCF₁ which is not better. So it can be demonstrated that the PCF₁ is better than PCF₂.

So further analysis is focused on PCF₁. The PCF₁ also shows lower effective refractive index than PCF₂ which has been shown in Fig. 6. From the figure it can be illustrated that the effective refractive index decreases with the increase of wavelength.

By varying the diameters of d_1 keeping other parameters (d_3 and Λ_1) constant the relative sensitivity is calculated which is shown in Fig. 7. It is clear that the sensitivity increases linearly with the increase of d_1 . From the Fig. 8 it is also proved that the larger diameter shows lower confinement loss. So for further investigation $d_1 = 2.7 \mu\text{m}$ has been selected.

Fig. 9 and Fig. 10 show the comparison of sensitivity and confinement loss by varying d_3 . The outer layer has been varied for $d_3 = 3.1 \mu\text{m}$ and $d_3 = 3.0 \mu\text{m}$ where the other geometrical structure are constant ($d_1 = 2.7 \mu\text{m}$ and $\Lambda_1 = 2.8 \mu\text{m}$). From Fig. 9 it is clearly visualized that the sensitivity does not significantly but from Fig. 10 it is proved that the larger diameter of air holes at the outer layer is responsible for lower confinement loss. So for the next investigation $d_3 = 3.1 \mu\text{m}$ has been optimized.

Sensitivity versus wavelength for pitch (Λ_1) variation is shown in Fig. 11 where d_1 and d_3 are kept constant. It has been observed that $\Lambda_1 = 2.8 \mu\text{m}$ shows better sensitivity for each case and it has been optimized. For every cases optimized values show better performance in proposed PCF.

V. CONCLUSION

The detection of gases and liquids using PCF has already proved its potentiality. Detection of toxic and flammable gases in the industrial process is an important part of safety issues. Photonic crystal fibers are used for gas detection at a wide range of wavelengths. In this research, two PCF structures have been proposed and also described a numerical study of comparative performance analysis between the two proposed PCF. The PCF with six air holes at the first layer shows much better results than the structure where there were eight air holes at the first layer. Both of the PCFs show lower confinement loss. The analysis were performed for methane gas sensing but other gases can also be detected with the similar expected performance.

ACKNOWLEDGMENT

The authors are grateful to those who performed in this research

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Home Appliances Control Using Mobile Phone

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Abstract—Now a days mobile phone has become a part of our daily life. Due to low cost of mobile phones, mobile phones are widely used for home automation. In this paper a remotely operated mobile phone controlled home appliances system is proposed. It is a DTMF (dual-tone multiple-frequency) based system consists of two mobile phones, DTMF decoder and ATmega8 microcontroller. One mobile phone is used as remote which may locate at far distance from home and another mobile phone is located at the home which acts as a receiver. The control information are sent via the remote mobile phone as DTMF tone, this DTMF tone is received by the mobile phone located at home, the received DTMF tone is then decoded by DTMF decoder MT8870 IC. The output logic signal of the decoder is used as input to the ATmega8 microcontroller. The ATmega8 microcontroller is previously programmed to control home appliances according to output of the DTMF decoder.

Keywords—Mobile phone; home automation; DTMF; decoder; microcontroller.

I. INTRODUCTION

A mobile phone has become an essential component in our day to day life. Home appliances control using mobile phone can give us flexibility as cellular networks has wide range of network coverage. So it is possible to control home appliances from a far distance. Conventional wireless control system such as radio frequency based systems, infrared remote based systems has limited working range. A mobile phone based control system can overcome this drawback, it also provides some advantages like no interference with other controllers.

In order to obtain a secure remote control operation considerable efforts should be given for the development of remote control for home appliances. A huge number of research work has been carried out to obtain a fast, secure remote control system. There are a number of telephone line based control system for home automation has been carried out, such as a telephone based control system for home automation using a hardware based remote controller [1][2] and telephone line with personal computer based systems [3] also developed. In telephone based system there was several drawbacks such as mobility problem, less flexibility and less friendly. With rapid growth of internet technology various internet based control systems for home automation has been proposed [4]-[7]. Internet based system is costly as it requires home server and personal computer. Again in an internet based system it is required to run the home server and computer all time. Internet based system gives worldwide coverage, one can control home appliances from anywhere. In internet based

system there is a considerable energy loss in home server and personal computer. Again it is not required to control a home appliances from one country to another country. Bluetooth based control systems for home appliances control has been proposed [8][9]. Bluetooth based system has the limitation of short working range. Long range control is not possible using Bluetooth based system. A mobile phone based home automation system that consists of a mobile phone with Java capabilities, a cellular modem, and a home server has been presented [10]. This system requires a mobile phone that should have Java capabilities to control home appliances.

A DTMF based home appliances control using mobile phone is shown in Fig. 1. The control system consists of two mobile phones, one DTMF decoder IC (MT8870), electromagnetic relays and a ATmega8 microcontroller. The system consists of two sub-system Unit-I and Unit-II. Unit-I consists of a mobile phone which may locate inside or outside the home. Unit-II consists of a mobile phone which is connected to decoder circuit via a headphone, the output of decoder circuit is connected to ATmega8 microcontroller. ATmega8 microcontroller controls the electromagnetic relays in order to control home appliances. The ATmega8 microcontroller works according to the output logic combination of the decoder circuit. The decoder circuit has 4-bit output logic signal. Unit-II resides at home. Two mobile phones communicates with each other via cellular networks.

This paper is organized into four sections. Section II describes working principle. Hardware implementation is described in section III. System evaluation is described in section IV. Finally, Section V concludes the paper.

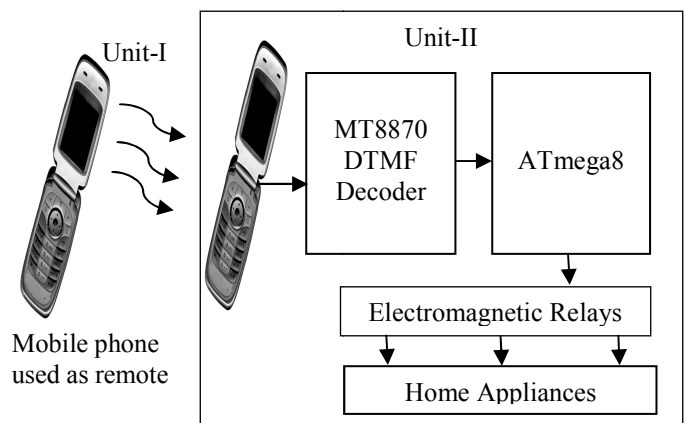


Fig. 1. Block-diagram of home appliance control system using mobile phone.

II. WORKING PRINCIPLE

Communication is established between Unit-I and Unit-II by initiating a call from Unit-I to the Unit-II mobile phone that resides at home. Unit-II mobile phone is kept in auto answer mode so that the incoming call can be received automatically. When the incoming call is received by Unit-II mobile phone, in the course of a call, if any button is pressed by Unit-I mobile phone, a tone corresponding to the button pressed is heard at the other end of the call. This tone is called dual-tone multiple-frequency (DTMF) tone. The keypad tone of the Unit-I mobile phone should be on and should be in DTMF mode. The DTMF signal is then sent to the DTMF decoder circuit via a head phone attached to the mobile phone of Unit-II. The DTMF decoder IC MT8870 is a full DTMF receiver that integrates both band split filter and decoder functions into a single 18-pin DIP or SOIC package. Its filter section uses switched capacitor technology for both the high and low group filters and for dial tone rejection. Its decoder uses digital counting techniques to detect and decode all 16 DTMF tone pairs into a 4-bit code. The received DTMF tone is processed by the MT8870 decoder IC and a set of digital output is produced corresponding to the keypad button pressed by Unit-I mobile phone. The output digital output of the DTMF decoder circuit is compatible to ATmega8 microcontroller. The output of the decoder circuit is used to tell the preprogrammed microcontroller which load to be turned on or off. Most of the household loads are alternating current (AC) load and microcontroller cannot control AC load directly. In order to control AC load electromagnetic relays are used. Again a microcontroller cannot drive a relay as microcontroller has low output current, so transistors are used to energize the relay coil for the purpose of load control. Electromagnetic relay driving circuit is shown in Fig. 2. The output of DTMF decoder circuit is 4-bit digital value with 12 distinct combination, so 12 different control signals can be obtained from the decoder circuit.

The mobile that makes a call to the mobile phone stacked in the home acts as a remote. So the proposed system does not require the construction of receiver and transmitter units. DTMF assigns a specific frequency, consisting of two separate tones to each key so that it can easily be identified by the electronic circuit. The signal generated by the DTMF encoder is a direct algebraic summation, in real time, of the amplitudes of two sine or cosine waves of different frequencies, "i.e." pressing '5' will send a tone made by adding 1336 Hz and 770 Hz to the other end of the line. DTMF keypad frequencies are

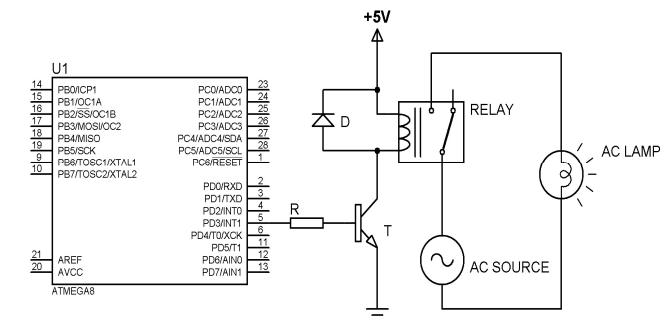


Fig. 2. AC load control using ATmega8 microcontroller and relay.

TABLE I. DTMF KEYPAD FREQUENCIES

Frequency	1209Hz	1336Hz	1477Hz
697Hz	1	2	3
770Hz	4	5	6
852Hz	7	8	9
941Hz	8	0	#

shown in Table I.

III. HARDWARE IMPLEMENTATION

Hardware implementation requires design of DTMF decoder circuit and algorithm development for the control system.

A. Design of DTMF Decoder Circuit

The DTMF decoder circuit is designed using MT8870DE IC. The complete circuit diagram [11] is shown in Fig. 3.

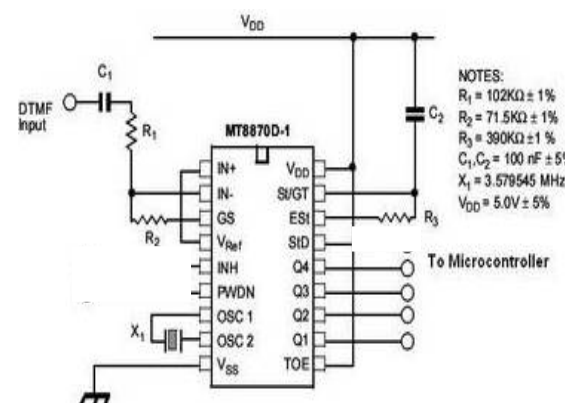


Fig. 3. DTMF decoder circuit using MT8870DE IC [11].

The output bit pattern of DTMF decoder circuit is shown in Table II.

TABLE II. OUTPUT OF THE DECODER CIRCUIT [11]

Digit (DTMF Tone)	Output of decoder circuit			
	Q1	Q2	Q3	Q4
1	0	0	0	1
2	0	0	1	0
3	0	0	1	1
4	0	1	0	0
5	0	1	0	1
6	0	1	1	0
7	0	1	1	1
8	1	0	0	0
9	1	0	0	1
0	1	0	1	0
*	1	0	1	1
#	1	1	0	0

B. Algorithm Development

Depending on the bit patterns of the DTMF decoder circuit a suitable algorithm is developed and written for the purpose of controlling home appliances. The algorithm is written in Atmel Studio 6.0. Internal EEPROM of ATmega8 is used to store information of the load status. If a power failure occurs the microcontroller stores the present status of home appliances in its EEPROM. When the power is back the microcontroller takes the status of home appliances from its EEPROM and maintains the load status. The control algorithm for one home appliance control is given in Fig. 4. The algorithm can be further extended for all home appliances.

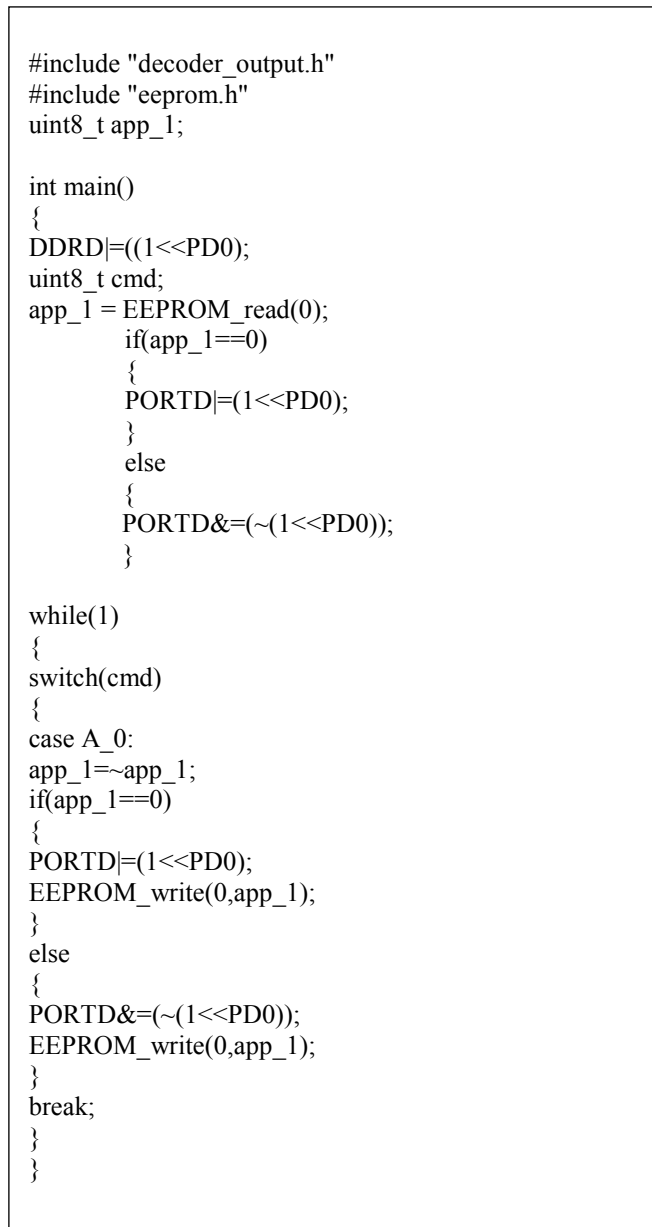


Fig. 4. Algorithm for home appliances control system.

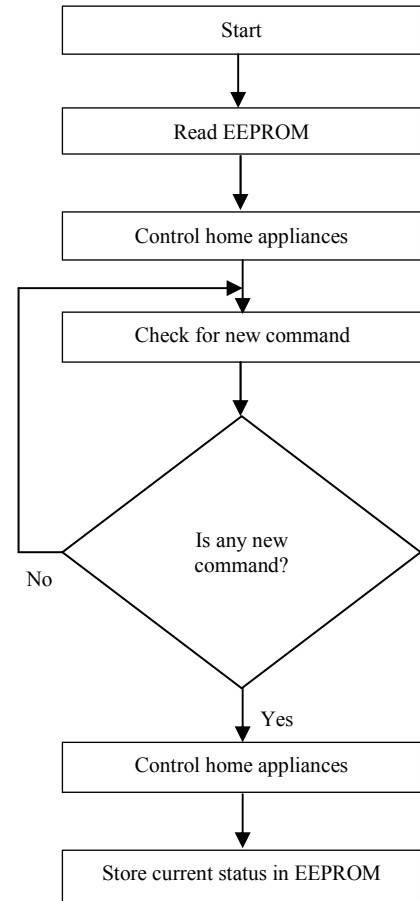


Fig. 5. Flow chart of the developed system.

The flow chart for the mobile controlled home appliances is shown in Fig. 5. The complete circuit diagram is shown in Fig. 6. In Fig. 6, the decoder circuit and relay driving circuit are not shown because the decoder circuit is shown in Fig. 3 and the relay driving circuit is shown in Fig. 2.

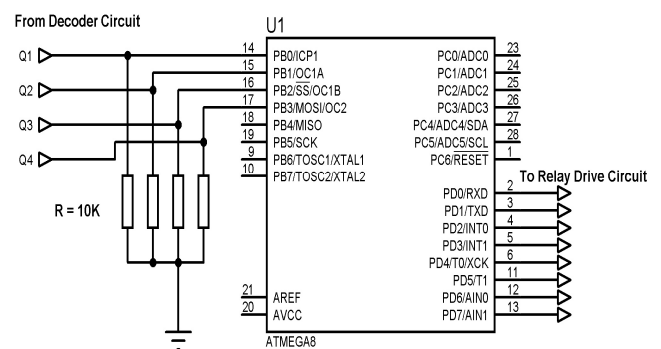


Fig. 6. Complete circuit diagram of the proposed system.

IV. SYSTEM EVALUATION

The proposed system can be used practically because of its flexibility, cost effectiveness and greater security. Due to simplicity of the design, use of less components makes the proposed home appliances control system less costly than other same types of system. Only the mobile phone is the costly parts of the proposed system. A mobile phone of very simple configuration can be used for this purpose. Now a days simple mobile phone is very cheap (available within US\$15) in Bangladesh. Due to low cost of mobile phone the proposed control system is cost effective. This system requires only one mobile phone which locates at home, the other mobile phone can be the personal mobile phone of the user. So within US\$20 the proposed control system can be implemented.

The proposed system also eliminates security threats completely by assigning a specific phone number to the mobile phone to operate which acts as a receiver and located at home. Now a days mobile phones have call blocking facilities, so one can assign a specific phone number as he wants and can block all other incoming phone calls. By assigning a specific phone number to the receiver mobile phone prevents establishing a phone call except himself. As call establishing is mandatory, so no one can be harmful for the proposed control system.

Another fact is, the proposed system can be used for some other purposes like main switch control of industries. If fire hazard occurs, this control system can be used to turn off the main switch automatically and immediately. It can reduce fire hazard subsequently.

The proposed control system has some drawbacks. The main disadvantage is cellular phone call establishing is mandatory for the proposed control system, so the proposed control system can only be used within cellular coverage area, another disadvantage is that every time, when one wants to control home appliances it is necessary to make a call to the mobile phone that resides at home, there may be a time delay to establish a phone call and for each call charge is applicable.

V. CONCLUSION

In this paper a low cost DTMF based mobile phone controlled home appliances system was developed. The developed system uses DTMF tone as control signal. The DTMF tone is received by the mobile phone that resides at home. The received DTMF tone is further decoded to digital values using DTMF decoder. A microcontroller based control algorithm was developed according to the output logic of the decoder for the purpose of controlling home appliances from a remote place. The proposed system is not fixed for home appliances control, it can be used anywhere for the purpose of controlling any types of electronic devices remotely. The developed system requires two mobile phones of any configuration for remote communication. Within cellular network coverage the home appliances can be controlled from anywhere, anytime. Again the proposed system does not suffer from security

threats because mobile phones have call blocking facilities, so the user can block all incoming calls except specified one.

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Remotely Operated Underwater Vehicle With Surveillance System

Abstract—Remotely operated underwater vehicles or ROVs are underwater robots that are used in science, entertainment, military and offshore oil industries. Their main function is to interact with the environment under the water in various ways. It is a very complicated system and uncommon in the developing or the underdeveloped countries around the world. Countries consisting many water bodies and prone to maritime incidents ROVs can be very useful in rescue missions. In this research work we built a ROV and equipped it with a surveillance system. Our ROV will be quite useful beside rescuers by monitoring the underwater and sending the video. It can also be used in military, scientific research, film making under the water and monitoring underwater industrial structures and underwater network devices.

Keywords—ROV, Underwater Robots, Underwater Surveillance System.

I. INTRODUCTION

About three fourth of the world is covered with water. To work and in this vast region of water the current methods being used are scuba, ROV, submarine, autonomous underwater vehicle(AUV) etc [1,2] . ROVs and AUVs are the two types of underwater robotic vehicles used to explore the underwater. Our work presented here is about ROVs. ROVs are teleoperated vehicles employing an umbilical cable to carry both power and telemetry from a mother ship to the vehicle where a crew controls the vehicle[3]. ROVs often work in closer quarters and deploy robot arms or other tools to interact with objects nearby [2]. ROVs are becoming more and more important in exploration and exploitation of the underwater environment because it is a necessary underwater technical system for observation [4]. Underwater robotic vehicles such as a ROV is also becoming important because of having greater speed, endurance and depth capability, as well as a higher factor of safety than human divers [2,5].

The first ever developed ROV was known as POODLE. The vehicle was mainly used for archaeological research [6]. Since then till now the ROVs are being widely used both in industry and science for a variety of purposes including exploring hydrothermal vents, inspection of underwater oil derricks, inspection of high radiation area, inspection of subsea phenomena, surveying archaeological sites and fixing underwater infrastructure such as cabling and piping. One of their most challenging uses is in deepwater search and rescue [5,7-9]. The rate of development regarding the control of ROV and the rate of the development of new sensors and environmental perception systems is on the rise. This caused an outstanding rise in cost [10].

ROV control system went through numerous developments. Yoerger and Slotine proposed a sliding mode controller for a ROV[11]. Goheen and Jefferys designed multivariable self-tuning autopilots for underwater vehicles[12]. Dougherty and Woolweaver showed that a sliding mode control on the MUST vehicle provides satisfactory behavior[13]. Yuh described the vehicle modeling and proposed a discrete-time adaptive control for underwater vehicles[1]. Yuh proposed a neural net control system using a reinforcement learning approach (critic function) [14,15]. Mahesh et al. proposed a coordinated motion control system for the vehicle and manipulator using the adaptive control approach[16]. Brock et al. illustrated collision avoidance control of multiple autonomous vehicles[17]. Healey and Marco described a slow speed flight control of autonomous underwater vehicles using traditional and sliding mode controllers and presented experimental results[18]. Kato et al. presented a fuzzy control approach to an autonomous underwater vehicle, the Aqua Explorer 1000[19].

With the development of technology the underwater robot system is getting expensive. As a result, in developing and under developed countries the production and deployment of ROVs are limited. Many of these countries consist of a large number of water bodies and are prone to maritime incidents. Some researchers have been involved in the design and development of low-cost underwater robots [10,20]. Our work focuses on to provide an alternative to the expensive ROVs. Our ROV can be produced in a low cost and can be deployed to assist in rescue missions, industrial and scientific works.

II. DESIGN OF THE SYSTEM

While designing the system our main target was to achieve video footage from underwater by moving our vehicles towards various directions. Our system functions according to the block diagram given in Fig.1.

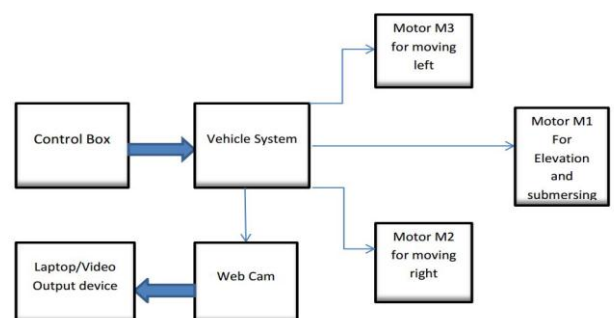


Fig. 1. Block Diagram of the System

III. CONSTITUENTS AND SUBMODULES

A. Control Box

The control box consists of 3 DPDT center off toggle switches within it. The switches are connected to the ROV by the tether cable and to the motors by a UTP cable. The switches provide the power from a 12volt battery. Fig. 2 shows the internal view of the control box. The DPDT center off toggle switch can control two separate circuits, but are always switched together by a single actuator.

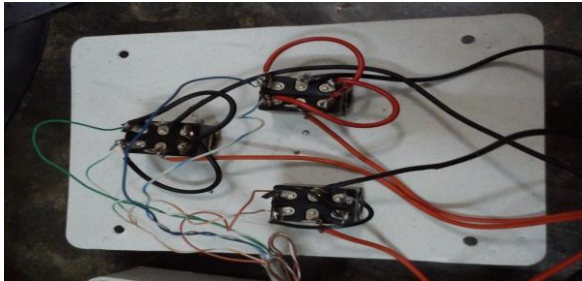


Fig. 2. Control Box(Internal View)

B. Vehicle System and motors

The vehicle system is mainly constructed using UPVC pipes. A 5 feet (1.5m) long UPVC pipe is cut according to the requirement which is given here:

- Two pieces – 2 1/2" (6.4 cm) long
- Two pieces – 4" (10.2 cm) long
- Two pieces – 4 1/2" (11.4 cm) long
- Four pieces – 1 1/2" (3.8 cm) long PVC pipes.
- Four pieces – 5" or (12.7 cm) long PVC pipes.
- 4 pieces of 1/2" PVC T's.
- 10 pieces of 1/2" PVC elbows

Waterproofed motors with the propellers are attached to the UPVC frame. The frame of the vehicle is light and can easily be transported to any area. The choice of using UPVC pipes was done because of availability and its lower mass. Being light weight in nature the vehicle is easily maneuverable in the water. The light vehicle frame allows it to be easily deployed in any conditions. Fig.3 shows the assembly of the vehicle frame.

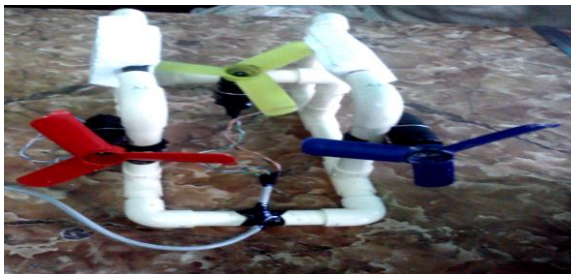


Fig. 3. Vehicle frame and the motors with the propellers

C. Web Camera

The surveillance unit consists of a web camera and a flashlight. The final assembly of the ROV is given below in the Fig. 4.

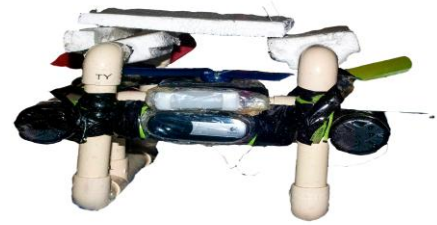


Fig. 4. The Final Assembly of the ROV

D. Video Output device

Video output device is connected to the web camera. Where the live video footage from the underwater can be displayed and stored for further analysis.

IV. CONTROL OF THE ROV

Our ROV's is controlled by thrusters. The dynamics of our ROV depends upon the thruster dynamics. Using the propeller angular velocity Ω as the state, the thruster dynamic model and the thrust output equation is given by

$$\dot{\Omega} = \beta_{th} \tau_{th} - \alpha_{th} \Omega |\Omega| \quad (1)$$

$$\text{Thrust} = C_{th} \Omega |\Omega| \quad (2)$$

τ_{th} is the input torque for the thruster. α_{th} and β_{th} are constant model parameters and C_{th} is a proportionality constant [21]. There will be thruster efficiency lost due to disturbances in the water inflow to the thruster propellers caused by thrust to thrust, thrust to hull interactions, currents and vehicle velocity and disturbances by the water inflow [22]. ROVs mainly have low relative velocities. The hydrodynamic forces effecting the ROV can be calculated by the Morison equation [23]

$$F_h = C_d \frac{1}{2} \rho_w A (V|V|) + C_m \dot{V} \quad (3)$$

F_h is the hydrodynamic force itself, ρ_w is the water density, A expresses a reference area and V and $\dot{V}$ are the vectors of the relative velocity between fluid and vehicle and of the relative acceleration, respectively. The coefficients C_d and C_m depend on the direction of the relative velocity. Since the acceleration of the water flow is normally very small, the second term on the right-hand side of (3) can be added to the inertia terms of the mass matrix. The coefficient C_m can then be called hydrodynamic added mass. The first term describes the quadratic damping of the water and represents the nonlinear effect of hydrodynamic forces [24].

We are using three 9v DC motors as thrusters. The controller controls the thrusters and moves the vehicle in 4 directions-up, down, right and left.

A. Switching System of the Vehicle

We are using three switches to three motors. In Fig. 5 the switching system has been visualized. The top motor M1 is controlled by switch S1 to elevate and submerge the vehicle by rotating it clockwise and counterclockwise. The

clockwise motion aids the submersing action. The counter-clockwise motion does the opposite. Right Switch S2 is used for moving the vehicle to the right by controlling the motor M2. Left Switch S3 is used for moving the vehicle to the left by controlling the motor M3. To move the vehicle forward both the M2 & M3 will be turned on and to move the vehicle backwards both the motors rotate in the opposite direction using both the switches S2 & S3. The motors can be turned off moving the switch into the center position.

The power chords provide the necessary power to conduct the switching operations.

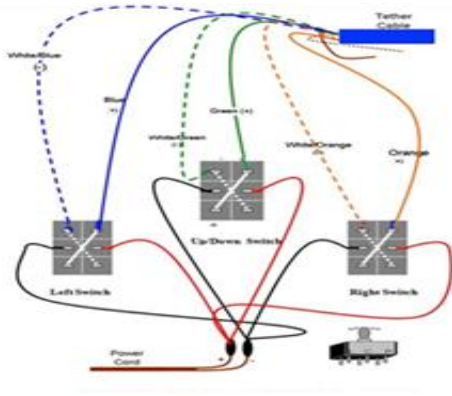


Fig. 5. The Switching System

B. Complete Overview of the System

The total overview of the control system given here shows the connections between the switches and the motors.

- Switch S1 is connected to motor M1 by the UTP cable's Green and Green-White wires.
- Switch S2 is connected to motor M2 by the UTP cable's Orange and Orange-White wires.
- Switch S3 is connected to motor M3 by the UTP cable's Blue and Blue-White wires.
- The dotted lines denote that Green-White, Blue-White and Orange-White wires.

Fig.6 shows every connection used in the vehicle. The motors are connected to the batteries through switches. The UTP cables are used here for sending signals of both polarities. The dotted wires are used to transmit the signal of the opposite polarity from the signal of the colored wires.

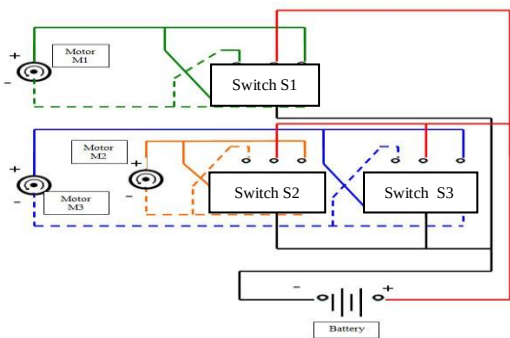


Fig. 6. Total Overview of the Control System

TABLE I shows the operation of the switching system. Position 1 of our DPDT center off toggle switch turns the motor in a clockwise direction and position 2 turns the motor in a counter clockwise direction.

TABLE I Switching System Index

Switch	Position	Motor	Propeller Direction
S1	1	M1	Clockwise
S1	Center	M1	Off
S1	2	M1	Counter-clockwise
S2	1	M2	Clockwise
S2	Center	M2	Off
S2	2	M2	Counter-clockwise
S3	1	M3	Clockwise
S3	Center	M3	Off
S3	2	M3	Counter-clockwise
S2 & S3	1	M2&M3	Clockwise
S2 & S3	2	M2&M3	Counter-clockwise

V. PERFORMANCE TEST AND COST ANALYSIS

A. Vehicle Maneuverability and Surveillance Unit Test

To check the ROV we have done dry test and wet test. Both the dry test and the wet test were done in day and night to check the visuallization efficiency. The wet test was the most significant test. The wet test has been done in two steps.

- Static
- Dynamic

As Dynamic wet test is the most accurate test where the actual performance of the ROV can be seen in TABLE II provided here.

TABLE II Dynamic Wet Test

Depth (in feet)	Motors	Direction	Light condition		Range of clear sight(in cm)	
			Night	Day	Night	Day
3	M1	Downwards	Poor	Good	25	29
3	M2&M3	Forward	Poor	Good	25	29
3	M1	Downwards	Poor	Good	24	29
7	M2&M3	Forward	Poor	Good	22	28
7	M1	Downwards	Poor	Good	22	28
14	M2&M3	Forward	Poor	Good	21	26.5

From TABLE II we can see the performance of the ROV in different light conditions. The flashlight becomes absolutely necessary for the deeper parts of a water body because of the low rate of light penetration. From the TABLE II we observe that the deeper the vehicle goes, the video output gets more and more underexposed and noisy. This happens severly during night time. The range of clear sight decreases significantly.Fig.7 is the view from 8 fts under water. A root of a water plant and bubbles are seen here in the video feed.



Fig. 7. Video feed from underwater

B. Cost Analysis

Our primary focus in the work was to reduce the cost to make a suitable ROV to be deployed into developing and under developed countries. Compared with the work in the developed world we have to say we succeeded in achieving that goal. World class ROVs is usually quite costly. In 2013 a group of researchers provided a model of a comparatively low cost underwater robot named GUPPIE which cost about \$1000 whereas our total cost to design and construct the vehicle has been around \$100 which makes it easier to deploy for developing and under developed countries with majority water bodies[25]. Table IV here projects the cost of our ROV.

TABLE IV Projection of the cost

Components and Process	Cost(\$)
Surveillance Unit	17
Control Box	32
Vehicle Frame	51
Waterproofing	6
Total cost	106

VI. CONCLUSIONS

We designed, assembled and finally construct a ROV and tested its surveillance system under the water. Using this we were able to recognize objects under water almost perfectly with the assistance of the flashlight.

Our ROV's main limitation is its dependence on wire length. To get rid of this constraint our ROV can be updated into AUVs (Automated Underwater Vehicle) incorporated with various types of underwater sensors. Turning the ROV into an AUV might increase the cost thus the challenge here will be to keep the cost minimum and develop further.

Using pattern or object recognition algorithm we can achieve a better rate of success in rescue missions and scientific expedition in the underwater realm. Instruments such as water sampler and instruments that measure water clarity, water temperature, water density, sound velocity, light penetration, and temperature can be attached to the ROV to help scientific explorations.

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A Cost-Effective Design and Development of A Surveillance Robot

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Abstract—This paper discusses the proposed design of a robot, which can move forward and detects obstacles in front of it. There is an ultrasonic sensor to detect the obstacle and motors connected to wheels of the body of the design. Other sensors and a microcontroller are used to make the designed robot operational. This design is programmed and run by an ArduinoUno. Power is supplied from a DC source, which is a lithium battery.

Keywords—*Arduino; ultrasonic sensor; motor; temperature sensor, digital display*

I. INTRODUCTION

A robot is a mechanical or virtual artificial agent, usually an electro-mechanical machine that is guided by a computer program or electronic circuitry. Robots have replaced human in the assistance of performing those repetitive and dangerous tasks which humans prefer not to do, or are unable to do due to size limitations, or even those such as in outer space or at the bottom of the sea where humans could not survive the extreme environments.

The type of robots that encounter most frequently are robots that do work that is too dangerous, boring, onerous, or just plain nasty. Most of the robots in the world are of this type. They can be found in auto, medical, manufacturing and space industries. In fact, there are over a million of these types of robots working today. The robot which we are planning to design is going to detect the obstacle and temperature. For further progress we can use different types of sensors like, water sensor, smoke sensor, humidity sensor etc. The existing robots like this, the range of obstacle detection is very low and added sensors are not available. The motor control has been very simple and problems have occurred when the robot has been driving through rough terrain. If a command or pathway is send to the motor, the robot should follow this pathway[1].

This has not been the case if, for instance, one or both wheels start to spin due to low friction between the wheels and the ground. When this happens, the computer in the robot thinks that the robot has moved from one place to another when the true fact is that the robot has just been spinning on the same spot. The project goal is to implement motor control with controllers have been unsuccessful on earlier versions of

the robot and the goal with this project is to create a better solution to the control problem than before by using Arduino.

II. DESCRIPTION OF THE PROBLEMS

Ultrasonic sensor will be placed in the front which will be monitoring the path ahead. If anything is detected, which in this case is an obstacle the sensor's output goes from logical low to high which is then given to the Adriano. The Adriano then controls the motors using the motor driver shield depending on the code written in it. Since we are detecting left right and front we will be using ultrasonic sensors. The obstacle is assumed to be in the front when the sensor goes high. The sensor being used by us, gives digital output and have a range of 100 centimeters so the obstacle is detected only when it is within this range. Using an ultrasonic sensor we can easily detect of presence of an object [2].

Our designed robot will follow the technique of bat to detect obstacle in its path. The ultrasonic sensor will always release ultrasonic wave. If the wave collides with an obstacle in front of the robot, the wave will bounce back to the sensor. If the receiver receives this reflected wave then we can be sure that there is an obstacle in front of the robot. The time difference between the transmission of the ultrasonic wave and the reception of the reflected wave is calculated. Since we know the speed of the sound wave we can calculate the distance of the obstacle.

$$\text{Distance} = \text{Speed} \times \text{Time} \quad (1)$$

The temperature sensor for hobbyists are cheap compared to the sensors on average, but they serve the same purpose – reading the temperature. Well, the sensor can eat the sun, but are ideal for homemade robotics and automation applications because they are easy to interface, accurate, and has a fast response time. After you place your finger on it, immediately the output of the sensor starts to rise. The lm35 is a cheap digital temperature sensor. The sensor is used in a wide variety of hobbyist applications for both beginner and others that are more experienced. This sensor has 1-wire interface, which means that require only one pin to communicate with the microcontroller. More than that, it is designed with a unique serial number that allows you to interface more sensors on the

same data bus. The accuracy of the measurements is high because the sensor does not depend on the accuracy of the microcontroller to measure the analog signal. And because this sensor has a digital output, you will not get any signal degradation even over long distances. The sensor is used in a large variety of applications including temperature sensing and monitoring robot, air temperature monitors, etc.

III. FEASIBILITY STUDY

To solve this problems, we have use ArduinoUno controller, the program has written such a way that it can produce a synchronization within the sensors and the direction of the robot. The temperature and obstacle sensors will be placed in such a way that they give the most accurate output from the system [3]. The Arduino program will be done very carefully to make sure the robot gets maximum dynamics as well as give the correct output. The feedback will be observed very carefully so that the robot can direct itself to the easiest path.

Another solution can be the robot will have more power than the existing robots. So that it can move on an unsmooth surface, this robot can be controlled by human rescue team which will roam around and give live feedback in the control room in case of an accident so that the rescue team can be immediately informed.

IV. SOLUTION ADOPTED

The solution has been adopted is the obstacle and temperature detection robot, because in this project we are giving emphasis on developing a system that can be controlled, will be low cost and embedded with the technologies such as obstacle detector and temperature detector. To complete this project we will be using Arduino and other micro controllers which have lesser power consumption and long life time. The robot will have long rage for detection and it can be used in places which are temperature sensitive. It also can be added with the other sensors and controllers [4] which can transmit and generate information from the places where the robot will be working.

V. EQUIPMENT

This part describes the equipment which are used for this robot and their working functions and details also. In which purpose they are used for also described here. The main components used in this design are Arduino Unomicrocontroller, HC-SR04 (Ultrasonic sensor), LM35 (temperature sensor), Digital Display, Motor(gear and servo), Chassis, L29 3D, MultiMeter, Battery, Bread board, Wires.

A. ARDUINO UNO R3 Microcontroller

The Arduino Uno is a microcontroller board based on the ATmega328. It has 14 digital input/output pins (of which 6 can be used as PWM outputs), 6 analog inputs, a 16 MHz ceramic resonator, a USB connection, a power jack, an ICSP header, and a reset button. It contains everything needed to support the microcontroller; simply connect it to a computer with a USB cable or power it with an AC-to-DC adapter or battery to get started Arduino Uno R3 Specification [1].



Fig. 1. An Arduino UNO R3

B. HC-SR04 (Ultrasonic Sensor)

The HC-SR04 ultrasonic sensor uses sonar to determine distance to an object like bats or dolphins do. It offers excellent non-contact range detection with high accuracy and stable readings in an easy-to-use package. From 2cm to 400 cm or 1" to 13 feet [5]. Its operation is not affected by sunlight or black material like Sharp rangefinders are (although acoustically soft materials like cloth can be difficult to detect). It comes complete with ultrasonic transmitter and receiver module [6].

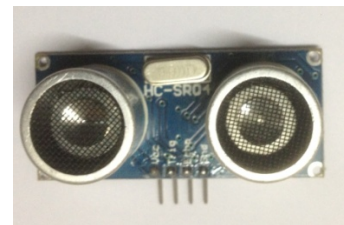


Fig. 2. An ultrasonic sensor

C. LM35 (Temperature Sensor)

The LM35 series are precision integrated-circuit temperature devices with an output voltage linearly-proportional to the Centigrade temperature [7]. The LM35 device has an advantage over linear temperature sensors calibrated in Kelvin, as the user is not required to subtract a large constant voltage from the output to obtain convenient Centigrade scaling.



Fig. 3. LM35 Sensor

D. Digital Display

A display device is an output device for presentation of information in visual or tactile form (the latter used for example in tactile electronic displays for blind people). When the input information is supplied as an electrical signal, the display is called an electronic display.



Fig. 4. Digital Display

E. DC Gear Motor

DC gear motors are used for this project with 12v voltage input and 400rpm rotational speed. Fig. 5 shows physical structure of this motor.

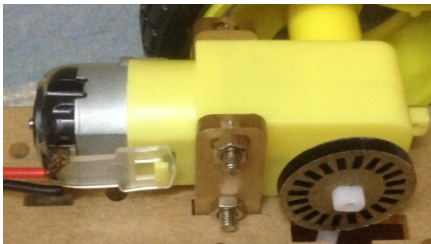


Fig. 5. DC Gear Motor

F. Chassis

The chassis is shown in Fig. 6. It is made in such a manner that a lot components can be attached to it. The chassis is not very heavy by itself.

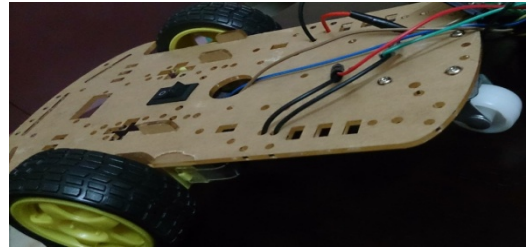


Fig. 6. Chassis

A. L293D Motor Driver

L293D is a dual H-bridge motor driver integrated circuit (IC). Motor drivers act as current amplifiers since they take a low-current control signal and provide a higher-current signal. This higher current signal is used to drive the motors.

1) Working Principle of L293D

The three 4 input pins for this l293d, pin 2,7 on the left and pin 15 ,10 on the right as shown on the pin diagram . Left input pins will regulate the rotation of motor connected across left side and right input for motor on the right hand side. The motors are rotated on the basis of the inputs provided across the input pins as LOGIC 0 or LOGIC 1. Just provide Logic 0 or 1 across the input pins for rotating the motor.

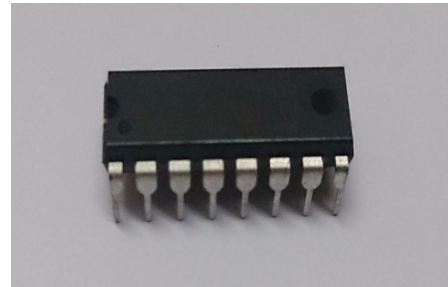


Fig. 7. L293D

2) L293D Logic Table

Let us consider a Motor connected on left side output pins (pin 3, 6). For rotating the motor in clockwise direction the input pins has to be provided with Logic 1 and Logic 0 [9]. Table I shows the output of L293D for different inputs to the controlling pins.

TABLE I. LOGIC TABLE

Pin 1	Pin7	Direction
1	0	Clockwise
0	1	Anticlockwise
0	0	Idle
1	1	Idle

B. Battery

ALithium battery (shown in Fig. 9) is used in this project.



Fig. 8. Battery

C. Design Layout

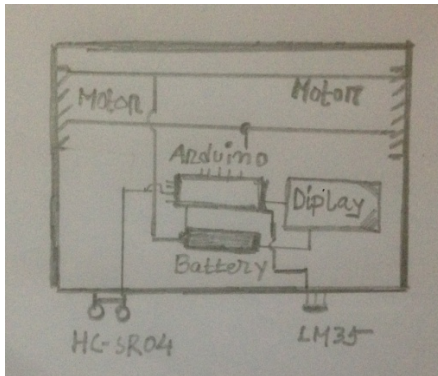


Fig. 9. Design Layout

The connectivity of the main components of the design are shown in Fig. 10.

VI. COST OF PRODUCTION

TABLE II. COST OF THE EQUIPMENT

Component	Cost (BDT)
Arduino Uno	750
Ultrasonic Sensor	350
LM35	45
Digital Display	170
Motors	2200
Chassis	1000
L29 3D	100
Multimeter	230
Battery	800
Bread board	150
Wire	100
Total	5895

The cost breakdown of making this robot is shown in Table II. The total cost is about BDT 5895, which is much less than comparable systems available in the market.

VII. CONCLUSION

The robot can successfully work as per the command. It can detect the obstacle in front of it and take the shortest path, also gives the reading of temperature in the display. The proposed design is very cost-effective in terms of performance and battery back compared to other designs. In future, some more features can be added to the project.

ACKNOWLEDGMENT

The authors thank everyone for sharing their knowledge and helping them to build this project. Finally, they thank their supervisor Dr. Hasan Uz Zaman who has always provided great help and encouragement to make this project a success.

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Development of a Bright Display Electrical Energy Meter using ATmega8 Architecture

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Abstract – At present times, the Electricity Company supplies LCD-type Electrical Energy Meters (LCDEM), and mount them on the outdoor panels. The users cannot time-to-time monitor the consumed energy units to assess that the consumption is exceeding their budget, and they should be aware to shut off the unnecessary loads. A ‘Bright Display Energy Meter (BDEM)’, when fitted indoor, will aware the consumers for conservative use of electricity as they can now view the readings of the meter. The users are psychologically motivated to control their consumptions; as a result, they save energy, pay less bill and help the nation like Bangladesh to mitigate electricity crisis. This paper has presented the design and development procedures of a BDEM using an ‘ATmega8 Microcontroller.’ The meter has been prototyped, tested and found to work as per specifications.

Index Terms – Bright Display Energy Meter, BDEM, LCD-type Energy Meter, LCDEM, ATmega8, and RMCKIT.

I. INTRODUCTION

In a developing country like Bangladesh where we cannot generate enough electricity to meet the growing needs of the public, we must strive hard to formulate some ways that will motivate people to use electricity with conservative minds. Introduction of a BDEM could be one of the ways to serve this purpose. In a BDEM (Fig. 1), the display unit that registers the units of the consumed electricity is made of bright-lighted 7-segment display devices. The meter is installed indoor and allows the house members to have all times sight into the readings of the meter. The meter reading develops a sensation in the mind of a user to check that if the consumption is exceeding his monthly budget for the electricity bill. A psychological urge prompts the user to switch off the unnecessary lights, fans etc. Thus, there is a saving of electricity, which could be useful for some other consumers.

II. DESIGN SPECIFICATIONS OF THE BDEM

We will not be required to design and manufacture all the constituent parts of the proposed BDEM; rather, we will transform an existing LCDEM into a BDEM with the placement of the following additional hardware components within the empty space of the LCDEM (Fig. 1):

- 6-digit cc-type 7-segment display unit in place of LCD-panel: It will register the units of the consumed electricity.
- Step-down transformer: It will supply unregulated DC supply for +5V and ‘line frequency for power failure detection circuitry.’
- A small printed circuit board (PCB): It will accommodate ATmega8 Microcontroller (MCU), + 5V electronics, and ‘power failure detection circuitry (Fig. 1).’
- A push button: The user will activate it every month to erase the kWhr reading of the meter.



Fig. 1 Various stages of the BDEM during its development

III. HARDWARE BLOCK DIAGRAM OF THE BRIGHT DISPLAY ELECTRICAL ENERGY METER (BDEM)

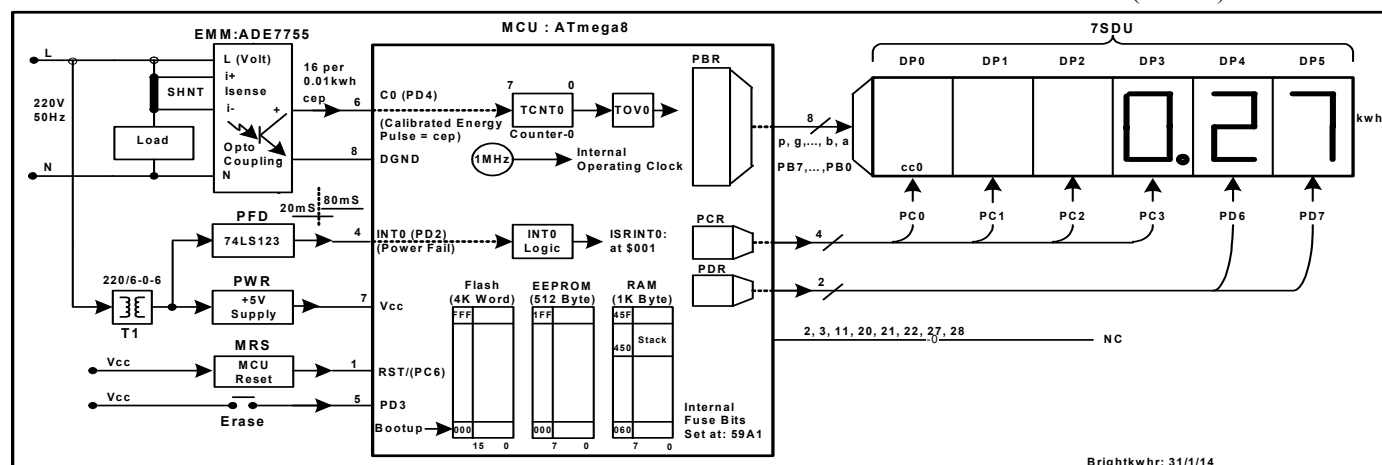


Fig. 2 Hardware block diagram of the Bright Display Electrical Energy Meter

A. Working Principle of EMM

The EMM (Energy Measurement Module) is an ADE7755 [1] based interface, which comes with the original LCDEM. It continuously measures the flow of real energy in the load and produces a *cep* (calibrated energy pulse) signal at the end of consumption of 0.000625 kwhr energy. The internal counter (TCNTC0) of the MCU counts every packet of 16 *ceps* and then updates the multiplexed-type display by 0.01. At the end of every month, the user presses the Erase button to delete the accumulated energy units from the internal memory, and the display unit shows 0.00. At power failure, the line frequency disappears; it immediately interrupts the MCU; the MCU goes to 'Interrupt Subroutine (ISRINT0)' and saves the last value of the display unit into its internal non-volatile EEPROM (Fig. 5). At power resumption, the MCU collects the saved kwhr from EEPROM and shows it on the display unit (Fig. 5).

C. Power Failure Detection Module (PFD)

The PFD circuit (Fig. 3) produces square wave signals (SQW) at every +ve zero-crossing points of the 50 Hz line frequency. It fires a re-triggerable oneshot (U1) of approx. 80ms duration. As a result, U1-13 sustains LH-state as long as the supply line remains active. At power failure, U1-13 goes to LL-state, and it immediately interrupts the MCU.

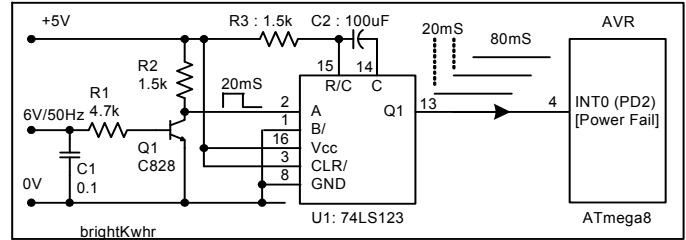


Fig. 3 Power Failure Detection Circuit

III. DATA STRUCTURES

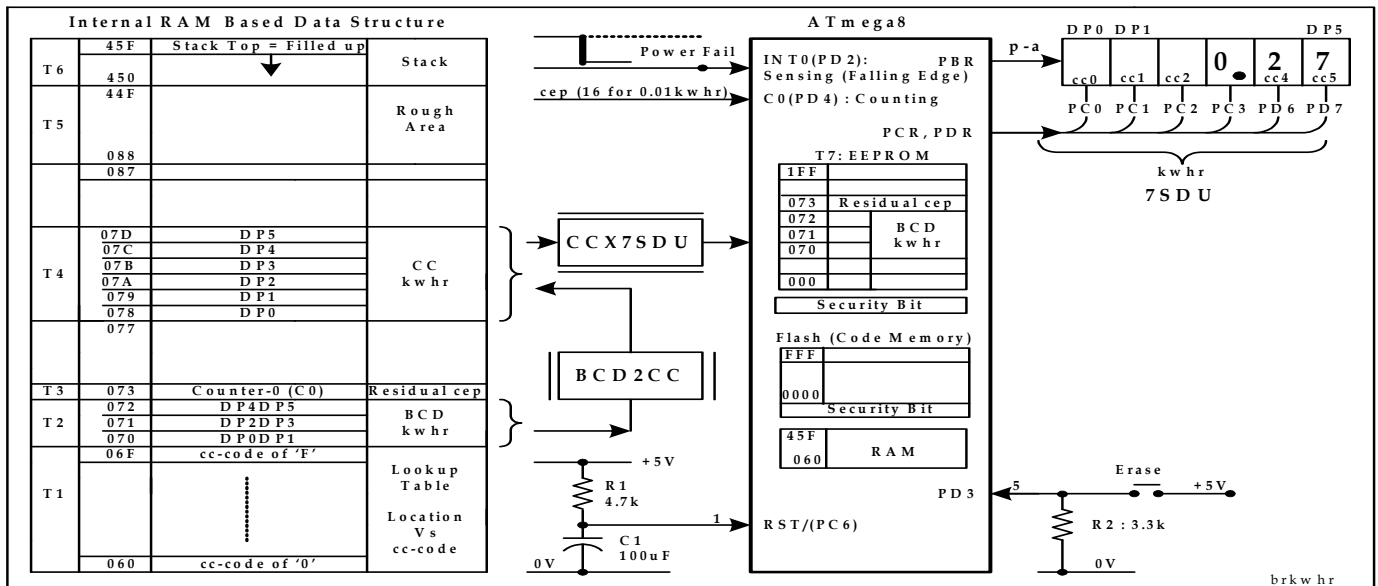


Fig. 4 Data Structure-I of BDEM for the Development of Control Program

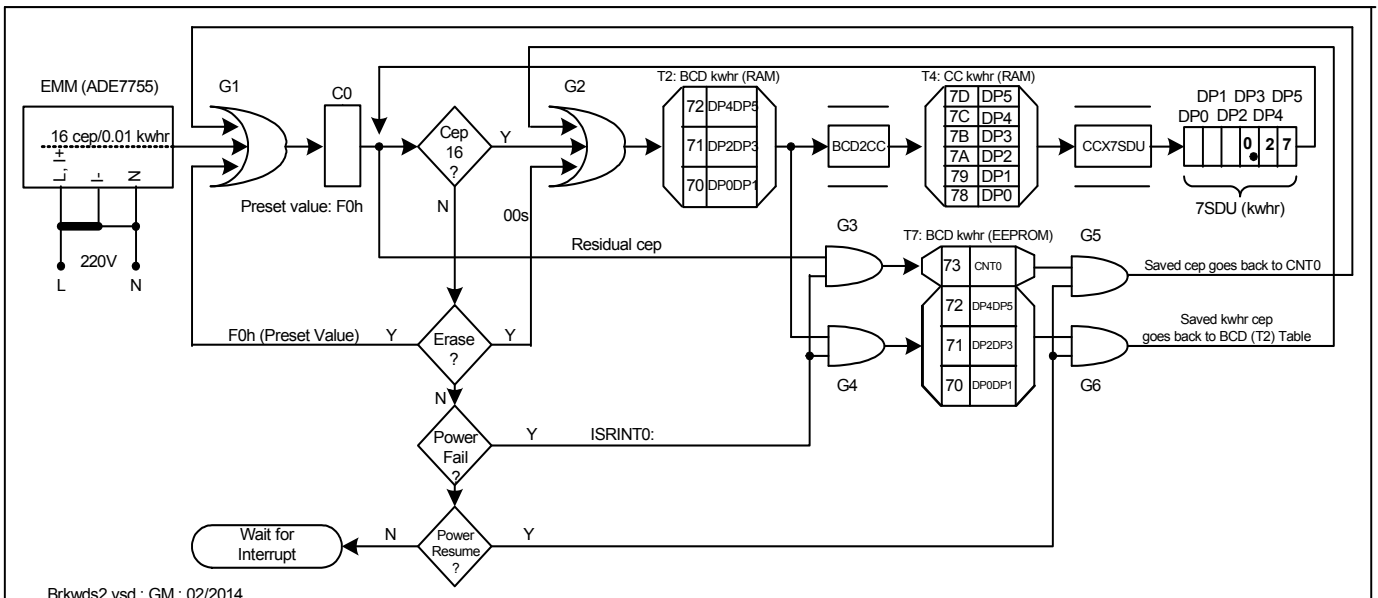


Fig. 5 Data Structure-II of BDEM for the Development of Control Program

A Data Structure is a memory map; it clearly indicates the memory locations (along with their numerical addresses) containing the input/output variables. Without a data structure, the MCU based designer would simply be unable to develop the Control Program (CP) for his system. As a CP (Fig. 6) is primarily a collection of subroutines (SUR), the data structure keeps the design engineer in the right tracks for modular development of the SURs. A correctly designed data structure ensures that the CP to be developed will work at the first run.

A. Data Structure-I

- The consumed (kwhr) electrical energy accumulated over the period from the last erase to present time is kept in T2 (Table 2) in BCD format with 2-digit precision. Counter-0 (C0) augments T2 by 0.01 kwhr when TOV0 (Timer-0 Overflow) occurs after counting 16 *ceps*. The algorithm is:
L1A: if ((TOV0) !=1)
Refresh display by calling CCX7SDU
Goto L1A
L1B: (T2) + 01h → r16 (contains incorrect BCD)
Adjust (r16) into correct BCD by BCDADJ SUR → T2
L1C: Convert (T2) into CC-code and save in T4
L1D: Transfer (T4) → 7SDU (7-segment display unit)
L1E: Goto L1A
- Content of T2 is converted into common cathode (CC) codes using BCD2CC subroutine and Lookup Table (T1); the converted codes enter into T4. CCX7SDU subroutine transfers the content of T4 onto multiplexed type 7-segment display unit (7SDU). The CP calls upon this SUR at regular intervals to refresh the multiplexed display unit.
- At power failure, the MCU saves the content of C0 (residual *ceps*) into T3; and then the contents of T2 and T3 are saved into the non-volatile EEPROM (T7) of MCU.
- The MCU samples the PD3-pin at regular intervals (Fig. 6) for active condition, and then erases the contents of T2, T3, and T4 by storing zeros (0s) in these Tables.

B. Data Structure-II

Data Structure-II is a vivid version of Data Structure-I. It expresses conditional information, which could not be accumulated in Data Structure-I. For example, the Gate-3 (G3) says that the residual content of C0 will enter into EEPROM (T7) only at the event of power failure. Data Structure-II is a great aid in the development of the complex Control Program for an apparently simple BDEM.

- G1 says that the C0 could be loaded in one of following three ways: (i) by EMM during normal operation, (ii) by the Erase command, and (iii) by T7 at power resumption.
- G2 says that the content of T2 could be loaded in one of the following three ways: (i) by C0 when overflow occurs after counting 16 *ceps*, (ii) by the Erase command, and (iii) by T7 at power resumption.
- Gate-3 (G3) says that the residual content of C0 will enter into EEPROM (T7) only at the event of power failure.
- G4 indicates that the content of T2 would be stored into EEPROM (T7) only at the event of power failure.
- G6 says that the BCDkwhr of EEPROM (T7) would be loaded back into T2 only at the resumption of power.
- At power failure, the MCU is interrupted (Fig. 2, 4) over INT0-pin; the MCU goes to ISRINT0 and saves residual *ceps* and BCDkwhr of T2 into EEPROM (T7) of the ATmega8 microcontroller

V. CONTROL PROGRAM OF THE BDEM

The Control Program (CP) of the BDEM resides in the Flash Memory of the MCU and regulates the operation of the meter. It is composed of the following items:

- A block of 'Program Codes' called the codes of 'Mainline Program (MLP)'. These codes continuously monitor energy consumption and then update the display.
- A block of 'Program Codes' called the codes for 'Interrupt Subroutine (ISRINT0)'. These codes save the critical data into the non-volatile EEPROM during power failure.
- A block of 'Data Bytes' called 'Look up Table (LUT)', which helps converting a BCD-digit into its 8-bit cc-code.

A. Flow Chart of the Control Program

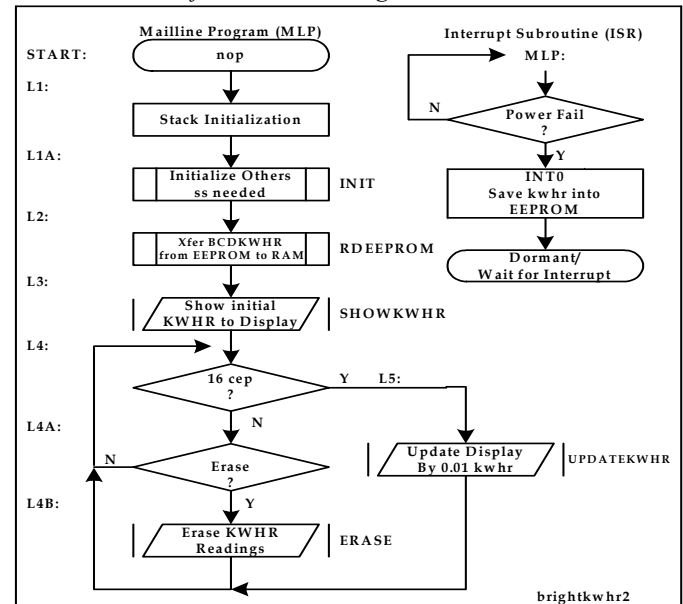


Fig. 6 Control program flow chart for BDEM

B. Complete Assembly Codes for the Flow Chart of Fig. 6

```

=====Main Line Program Coding=====
.include "m8def.inc"
.cseg
.org $000

RESET: rjmp START ;STKINIT

.org $001
rjmp SAVE2EEPROM ; save BCDkwhr and CEP to EEPROM

.org $020
START: nop
L1: ; stack init STKINIT:
ldi r16, 0x5F ; stack initialize
out SPL, r16
ldi r16, 0x04
out SPH, r16
L1A: ;--initialize others as needed-----
rcall INIT
L2: ;-- Update BCDKWHr and CNT0 of RAM from EEPROM
rcall RDEEPROM
L3: ;--show kwhr from RAM Table to display-----
rcall SHOWKWHr
L4: ;--chcek if 16 CEP pulses have occurred-----
in r16, TIFR ; reading TOV0 flag-bit
ror r16
brcs L5 ; 16 cef pulses are accumulated
rcall CCX7SD ; refresh display
L4A: sbis PIND, PIND3 ; checking active Erase button
rjmp L4
L4B: rcall ERASE ; Erase button is pressed
rjmp L4
L5: ;--16 CEP are accumulated; update display by 0.01 --
rcall UPDATEKWHr
rjmp L4 ; repeat the process-----
=====End of Main Line Program Codes=====
=====ISR Codings=====

```

```

SAVE2EEPROM:
    rcall    WR2EEPROM
HERE:    rjmp    HERE
;==End of ISR Codes=====
;==Subroutine Codes=====
UPDATEKWHR:
    ldi      r16, 0xF0
    out      TCNT0, r16    ; reload preset value
    ;--- clear TIFR Flag bit---
    in       r16, TIFR
    ori      r16, 0x01    ; clear TOV0 bit
    out      TIFR, r16
    ;--update Display by 0.01--
    rcall    UPDATE
    rcall    HEX2CC    ; BCD is a subset of HEX
    rcall    SUPPZER    ; suppress leading 00s on display
    ;--place point---
    lds      r16, $7B
    ori      r16, 0x80
    sts      $7B, r16
    ;-----
    rcall    CCX7SDU
    ;-----
    ret
;-----
;--exit
;==End of Assembly Program=====

```

V. DEVELOPMENT METHODOLOGY OF BDEM

We shall follow ‘Small Start Strategy (SSS)’ approach to develop the BDEM. In this approach, a basic (simple and elementary) hardware and the corresponding software routine are tested and then we add and test another basic part. The process keeps going on until the system appears.

A. Resource Requirements

- IBMPC with WINXP Operating System: It offers platform to develop assembly programs and down load it into the Flash of the ATmega8 via RMCKIT.
- RMCKIT: ATmega8 Writer/Trainer Kit (Fig. 7, 8) [3] The RMCKIT is a friendly and low-cost Microcontroller Learning/Trainer Kit. It allows developing and testing hardware circuits and software routines for ATmega8-based systems. It is equipped with ISP (In System Programming Interface) to fuse program codes and data into the Flash and EEPROM of the ATmega8.

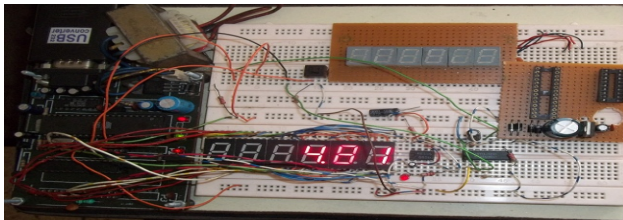


Fig. 7 Pictorial view of the RMCKIT Kit

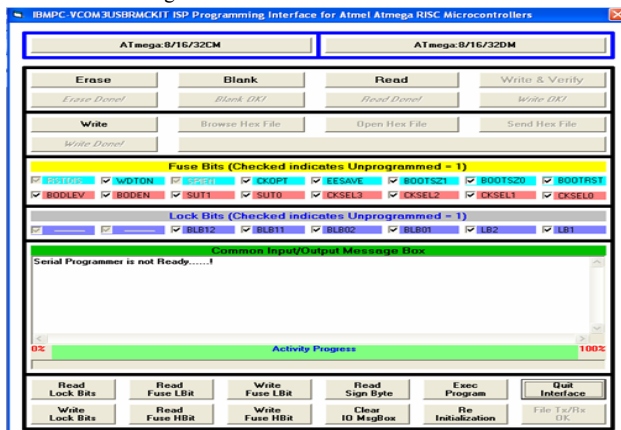


Fig. 8 Graphical user interface (VCOM3USBRMCKIT) for RMCKIT
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17-19 December 2015, Independent University, Bangladesh

- AVR Studio 4 [2]: It is an IDE platform; it allows developing and assembling application programs for Atmel’s Family of RISC Microcontrollers.
- Top2005 Universal ROM Programmer [4]: A commercial parallel ROM Programmer; it allows configuring the internal fuse bits and the security bits of the ATmega8.
- USB ↔ RS232 Conversion Cable and Driver [5]: This smart cable allows the operation of the RS232 serial port based RMCKIT using USB ports of the IBMPCs.
- Fully Assembled LCD-based kwhr Meter (Fig. 1) [6]

B. Development Procedures of Hardware and Software

- CC-type 6-Digit Display Unit and associated Software: Let us place the display unit on the breadboard (Fig. 7) and connect these with the ATmega8 of the RMCKIT as per diagram of Fig. 2. In this session, we shall develop and test the following subroutines: BCD2CC, and CCX7SDU. The Pseudo Codes of the program segment are:

```

L11: 123456 → BCDT of Fg. 6
L12: call BCD2CC    ; Section-VB for ASM codes
L13: call CCX7SDU   ; Section-VB for ASM codes
L14: goto L3

```

- Counting 16 *cep* Pulses and Advancing 7SDU by 0.01: Let us install a 555-based low-frequency oscillator circuit on the breadboard of the RMCKIT and connect its output into Counter-0 pin of the ATmega8. In this session, we shall develop the following subroutines: UPDATEKWHR, and BCDADJ. The Pseudo Codes are:

```

L21: if (CEP != 16)
    goto L1
L22: call UPDATEKWHR ; Section-VB for ASM codes

```

- Saving BCDKWHR at Power Failure: Let us build the PFD circuit (Fig. 3) on the breadboard and connect its output with the INT0-pin of the ATmega8. In this task, we shall develop the SAVE2EEPROM subroutine.

```

L31: if (Power Failure)
    goto ISRINT0    ; ISR for INT0 interrupt

```

```

ISRINT0:
SAVE2EEPROM:
    rcall WR2EEPROM ; for ASM codes Section-VB

```

- Developing Erase Function: Let us connect a Erase button (Fig. 4) at the PD3-pin of the MCU. The MCU keeps polling the active condition of the button and then erases the contents of the BCDkwhr (Table T2). In this session, we shall be developing the ERASE subroutine. The pseudo codes are:

```

L41: if (PD3 != LH)
    do other things
L42: rcall ERASE    ;

```

VI. CONCLUSIONS

In this paper, we have presented experimentally verified hardware setup and program codes, which the interested readers may exercise to acquire the ability of using ATmega8 technology for building their own projects. The BDEM could be a good starting point for developing ‘Prepaid Electrical Energy Meter’ by replacing the ATmega8 with ATmega32.

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Frequency Transient Analysis of Countrywide Blackout of Bangladesh Power System on 1st November, 2014

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Abstract— In this paper, real time frequency transient analysis with respect to frequency control of Bangladesh Power System (BPS) countrywide blackout on 1st November, 2014 is presented which occurred after HVDC station tripped. The inertia constant of a system can be a significant tool for investigating the frequency stability of a system and the frequency transient of power system is a vital index for power system operation and research. The dynamic frequency characteristic of power system is the basis of many important design works such as the design of under frequency load shedding scheme. Therefore, it is obvious and essential to analyze the dynamic frequency characteristic of BPS before countrywide blackout after disturbance. This paper first analysis real time frequency transient and determine the system inertia. Secondly it shows that maintaining nominal level of inertial reserve in real-time operations with unit commitment which can be crucial to ensure stable, secure and reliable operation of BPS. Beside this under frequency scheme should be revised and controlled. BPS network disintegration along with rate of change of frequency should be ensured for adaptive as well as self-healing of the power system.

Keywords—BPS (Bangladesh Power System); Inertial Reserve; df/dt ; HVDC (High Voltage Direct Current); BIPTC (Bangladesh India Power Transmission Center).

I. INTRODUCTION

As the last defense line of power shortage to keep the systems stability, frequency is considered the most reliable power shortage indicator [1]. In these circumstances, how to maintain and improve the safety and stability of systems, especially to maintain the systems stability after being subjected to high-power impact, as well as how to recovery frequency after the accident by load shedding under the premise of as little price as possible, have become issues of common concern by systems operation authority. For the frequency stability problem, obtaining accurate frequency response is the fundamental of studying frequency stability, and obtaining a reasonable analysis model is based on the comprehensive consideration of systems frequency dynamic characteristics [2]. Frequency response is defined as the automatic corrective response provided towards balancing demand and supply [3].

BPS network has almost no active spinning reserve because FGMOs (Free governor mode operation) are not active in

power stations of BPS. Consequently synchronous machines governor reserves (if any) are not automatically utilized during active power and frequency disturbance rather few generating units have reverse governor control mode (Say, Meghnaghat Power Plant) that exhibit positive response on system frequency (Generation declines with frequency fall & vice versa) [13] and worsen the frequency stability. Beside this it is experienced by NLDC (National Load Dispatch Center) engineers that when the system frequency becomes low the output power of CCPP reduces because the speed of motor of feed water pump and air compressor reduce [13]. So, during frequency decay for generation shortfall, BPS demands more active power input but active power injection to system reduces as described above and endangering the system security. In this circumstances, under frequency load shed scheme should be highly studied considering the system operating inertia during worst credible disturbed condition. Therefore, scheme settings (Frequency threshold, time delay, ROCOF, Islanding) optimized with critical inertial reserve of BPS should be adequate to keep the frequency stability of BPS.

Low levels of rotational inertia in a power system have implications on frequency dynamics and can influence the whole power system, in the worst case ending in fault cascades and blackouts. They are becoming faster in power systems with low rotational inertia [11]. This can lead to situations in which traditional frequency control schemes become too slow with respect to the disturbance dynamics for preventing large frequency deviations and the resulting consequences. The loss of rotational inertia and its increasing time-variance lead to new frequency instability phenomena in power systems [6].

As a Balancing Authority (BA), Power Grid Company of Bangladesh (PGCB) has a primary role is continuously balance demand and supply to maintain scheduled or nominal frequency. Bangladesh power system has an asynchronous Interconnection with its neighboring country India through a HVDC station at Bheramara, Kustia (Western part of BPS) and named as Bangladesh India Power Transmission Center (BIPTC). Typically, smaller system inertia is observed during

off-peak periods in BPS compared to peak periods because of less synchronously rotating mass online.

II. FREQUENCY DYNAMICS

Frequency at its nominal value reflects balance in supply and demand. Sudden loss of supply or demand will result in frequency deviation from the nominal frequency. The sudden disconnection of an active power source will initially produce rotor swings in the remaining generating units. The amplitude of the rotor oscillations of any given generator will depend on the amount of lost generation it picks up immediately after the disturbance occurs [7]. The dynamics the share of any given generator in meeting the lost load will depend on its electrical distance from the disturbance. The situation only last for a few seconds before the power imbalance causes all the generators in the system to slow down and the system frequency to drop. Then the active power share of any one generator in meeting the power imbalance depends solely on its inertia and not on its electrical distance from the disturbance [14]. The rate of change in frequency due to imbalance depends on the system inertia. System inertia is directly proportional to synchronously rotating mass in the system .The general equation for calculating rate of change of frequency using system inertia constant (H) is illustrated in below [9]:

$$\frac{df}{dt} = \frac{\Delta P}{2H} \times f' + \frac{D}{2H} \times \Delta f \tag{1}$$

Since we have no data on load characteristics of BPS hence for simplicity assumed ‘D’ is zero. Then the “(1)” becomes [9]

$$\frac{df}{dt} = \frac{\Delta P}{2H} \times f' \tag{2}$$

Where,

H: System Inertia Constant on system base (Seconds)

D: Power system load damping value in PU/Hz

f’: Frequency at the time of disturbance (Hz)

$\frac{df}{dt}$: Rate of change of frequency (Hz/Sec)

ΔP: Active Power Change in PU

III. REAL TIME FREQUENCY TRANSIENT ANALYSIS OF LAST BLACKOUT OF BPS

Bangladesh is now importing power around 500 MW from India through BIPTC. HVDC station in BIPTC does not add any inertia to Bangladesh power system. We observed the complete blackout of BPS on 1st November, 2014 for sudden outage of HVDC station causing the country wide blackout and restoration took place around 12 hours. The frequency transient after tripping of HVDC station is analyzed here. The frequency disturbance events are recorded by digital fault disturbances recorder (DFDR) which takes 1000 samples per second. From the recorded frequency transients on 1st November 2014 shown in Fig. 1 we observed that prior to HVDC station tripping, the network frequency was 50.327 Hz at 11:27:27.4098 hrs. Due to tripping of HVDC at 11:27:28:344 hours, import of 444 MW (9.49% of total generation) stopped and that generation shortfall which caused an imbalance between demand and supply in the network [13]. So, the system started to release kinetic energy consequently frequency declined rapidly and fallen to 48.636 Hz within 6.87 sec. Meanwhile under frequency scheme activated and

frequency remained between 48.631 Hz to 48.691 Hz. Then it started rising and reached at 49.066 Hz. Time span of this stage was 12.1 sec. The network frequency sustained over 49.0 Hz for 14.7 sec and finally it started to decrease.

From the existing under frequency scheme it is observed that 69 MW load at 33KV voltage level can be disconnected through the first stage (48.9 Hz). Similarly 356 MW and 576 MW load can be disconnected through second stage (48.8 Hz) and third stage (48.7 Hz) respectively. But since the black-out took place at off-peak hour at around 11:28 hours hence all the 33KV feeders were running well below their maximum load. Consequently, actual load shedding during frequency decline was far lower than expected load shed. The actual load sheds are tabulated in Table I.

From the frequency transient on 1st November, 2014 it is evident that a load cut of 23MW because of first stage operation of existing under frequency scheme could not affect the network frequency significantly. Action of second and third stages of under frequency scheme restricted the gradual fall of frequency. Prior to disturbance the system frequency was 50.327 Hz and down to 48.646 Hz within 6.8 sec. albeit under frequency scheme was activated but may be too late to remove the required amount of load to arrest the frequency decline and finally the scheme cannot prevent the black-out of BPS.

TABLE I
ACTUAL LOAD SHED

Sl. No.	Stage setting in Hz	Delay time (ms)	Expected Load Shed	Actual Load Shed
1	48.9	150	69 MW	23 MW
2	48.8	150	356 MW	152 MW
3	48.7	150	576 MW	240 MW
4.	48.6	150	523 MW	
5.	48.5	150	122 MW	

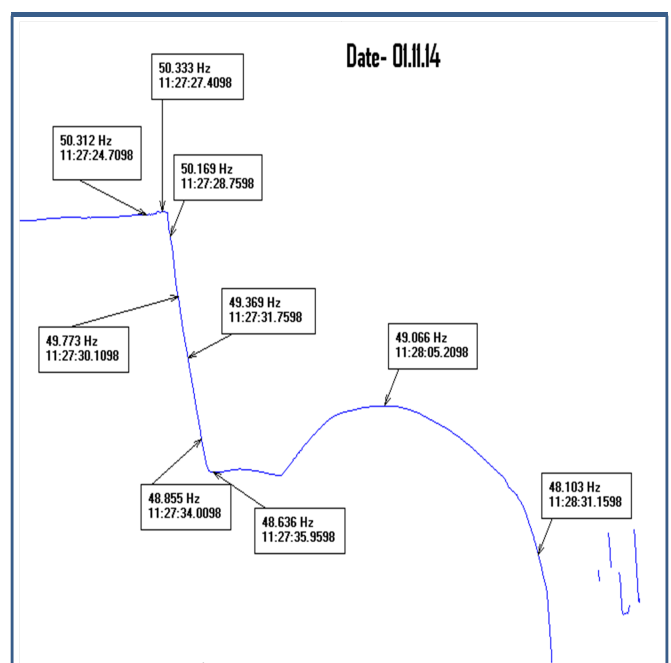


Fig. 1 Frequency Transient prior to blackout on 1st Nov. 2014

With the help of National Load dispatch Center (NLDC) recorded data we have found the generation from small engines (less than 5 MW) were 996 MW on 1st November, 2014 before BIPTC tripped and 996 MW was the 21.3% of total generation. It also indicates that the system was operating with lower system inertia. It is to be noted that on that day some big machines of western region went to maintenance and more small machines were running instead, in western region. When BIPTC tripped in western region, immediately after generation shortfall the machines in western region significantly share the impact according to their electrical proximity as expressed by the synchronizing power coefficients and finally according to their different inertias.

Since small engines have low rotating mass hence can store low kinetic energy and goes out of step after certain time because of post disturbance dynamic oscillation, although the frequency tries to rise before country-wide blackout. Therefore, lower system inertia and slower under frequency protection endangered the BPS network security. The BPS network time constant (H) for these two similar cases has been reckoned by using “(2)” and shown in Table II. Rate of change of frequency is determined for both cases from DFDR recorded frequency transient curve and with the help of DFDR software. The $\frac{df}{dt}$ values are used in calculation of network time constant (H) before activation of first stage of under frequency scheme. Successive Lower values of $\frac{df}{dt}$ in both scenarios after frequency goes below 48.9 Hz (First stage) & 48.8 Hz (Second Stage) show the sign of under frequency load shed activation.

From the frequency transient it is observed that during disturbance BPS network frequency travels down higher span of frequency and allows less time to network to reach into alarm/emergency state. From Fig. 3 before blackout we find that the frequency reaches to first stage frequency threshold to second stage frequency threshold within around 400ms. In BPS under frequency protection scheme, relay makes the trip contact after passing intentional delay time (here 150ms) and then master trip relay is energized through the relay trip contact and master trip relay makes its contact to actuate the tripping coil of circuit breaker and circuit breaker opens usually within three cycles. So at every step of load shedding scheme time adds up from relay pickup to circuit breaker opening and takes around 270ms. Consequently, there was chance that no load was shed before reaching second frequency threshold [13], because total time is (150+270) 420ms but frequency reaches to second threshold within 400ms as shown in Fig. 2.

TABLE II

Sl. No	Date	Time	Power through HVDC (MW)	Total system generation (MW)	HVDC percentage of Total gen. %	Inertia(H) Sec.	BPS Network
1.	01.11.14	11:27	444	4674	9.49	9.001663443	Collapsed

So, frequency load shed scheme have to be activated within minimum span of relay delay time to curtail optimized amount of load in the BPS power system until the available generation could supply the remained loads. Another important point regarding BPS under frequency scheme is that it does not mo-

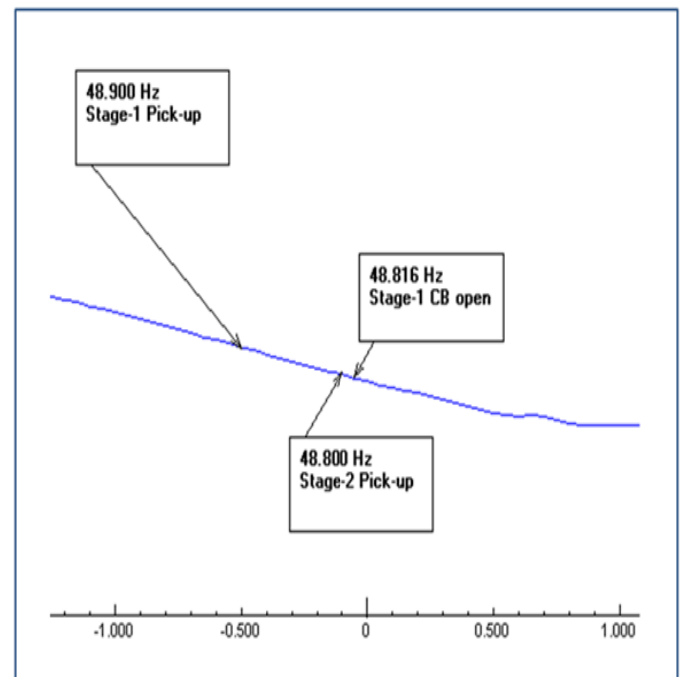


Fig. 2 Second stage picked up before 1st stage load shed after frequency disturbed

nitor how fast the frequency is declined. That is, the severity of the event causing the system frequency fall is not considered in this scheme. The traditional load shedding scheme monitors only the system frequency and the scheme is activated when the frequency reaches the threshold or set value and not take in count the rate of change of frequency (ROCOF) fall. The ROCOF is thus mainly determined by the magnitude of the power imbalance and the system inertia.

It is to be noted that in BPS under frequency scheme, there is no grid disintegration scheme as a last defense of countrywide blackout. There are two east-west interconnections (EWI) between eastern and western part of BPS network by 230KV double circuit line. According to NLDC data and from Fig. 3, it is to be noted that before blackout around 40 MW power was wheeling from eastern to western region and interestingly 40 MW power was returning to eastern part. So, power exchange between two parts was zero. In this circumstances, the disintegration of eastern part from western part as a last defense may prevent blackout. Analyzing the above frequency transient of our interest it is clearly evident that protection scheme is not designed considering the worst credible system inertia that can arrest the frequency decay during lower inertial reserve of BPS rather may create a state of blackout by iterative loss of gener-

ators. So, the critical duration of time delay of a stage and the number of stages of under frequency protection scheme may be revised considering the hardware limitations of relays and BPS network inertia.

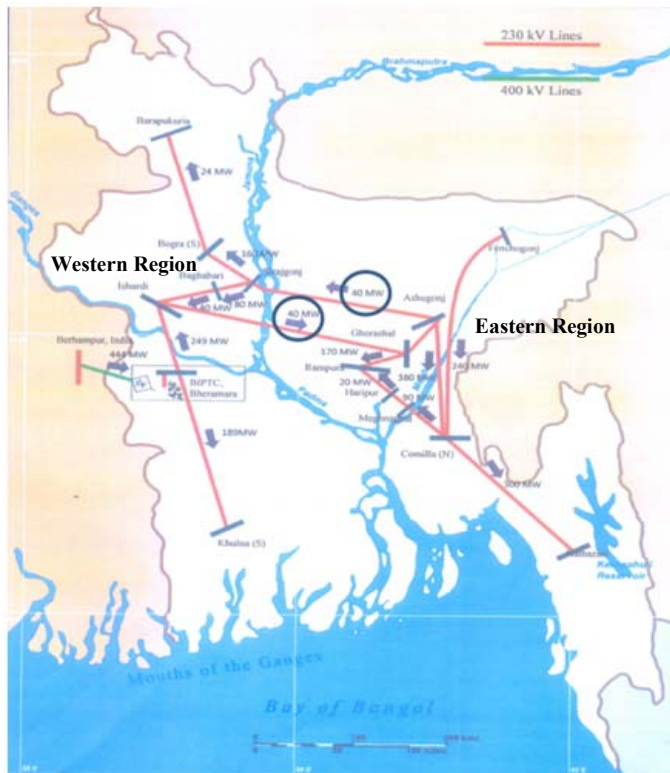


Fig. 3 Power flow between two EWI prior to disturbance

So, to prevent wide spread blackout of BPS either have to maintain the critical operating inertial reserve that must be greater than 9 sec, if PGCB don't want to revise under frequency scheme settings (frequency threshold and delay time) or scheme have to be revised taking in count, how minimum but credible operating inertial reserve may be found in BPS network. Along with the present under frequency scheme "under-frequency under-voltage load shedding (UFVLS)", ROCOF and controlled grid disintegration techniques should be adopted for adaptive and self-healing of Bangladesh power system. The critical system inertia and revised scheme settings can be determined by adequate event simulation studies of BPS network.

IV. CONCLUSION

The blackouts in recent year of Bangladesh illustrate the importance of having profound understanding of the characteristic of Bangladesh power system, including the stability of power angle, voltage and frequency. And the special defense lines are needed to control and protect the frequency stability of the BPS grid network. In this context, this paper has studied the dynamic frequency characteristic and suggested some additional features to add with present scheme and opens up new horizon to study the BPS network. This paper also shows that BPS network should be studied either to find the critical inertial reserve for existing present frequency scheme or to find the accurate and reliable faster under frequency scheme considering the minimum credible system inertial reserve during BPS network operation to prevent black-out caused by frequency instability.

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Implementation of Fuzzy Unified Power Flow Controller to Control Power Flow Dynamically in Transmission Line

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Abstract—This paper presents a fuzzy logic controller (FLC) based unified power flow controller (UPFC) for dynamic power flow control. FLC has been employed to provide dynamic control of series and shunt converters of UPFC. The proposed controller has been designed using PSCAD software and validated on IEEE 14 bus system network. The results show that the proposed method improves the dynamic power flow control, voltage profile and reduces the power loss in the transmission network. Moreover, the response of the proposed system is compared to proportional integral (PI) based UPFC.

Index Terms-- Flexible AC Transmission Systems, Unified Power Flow Controller, Dynamic power flow control, Fuzzy Logic.

I. INTRODUCTION

The demand of electric power over last few years has increased to such an extent that it has become essential to expand existing or build new power system network. But the expansion or building of transmission line is restricted by economic and environmental constraints. Therefore, system planners are forced to utilize the existing transmission resources up-to their maximum operating limit [1]. As a result, transmission lines become overstressed which has led the system designers to look for the new strategies to improve power system performance [2, 3].

FACTS devices have been efficiently utilized to control power flow through the transmission lines and enhance system controllability by mitigating power losses [4]. Though all the FACTS devices are quite effective for different power system applications but among them UPFC has got the epic popularity since it is capable of providing support for voltage regulation, series compensation, phase angle regulation, and control of the active or reactive power

separately transmitted over the line [5, 6].

Though UPFC implies many advantages, but its controller design still being a matter of challenge since it is a multi-variable controller. The challenge associated with control of UPFC is in UPFC two converters need to govern. Many control techniques had been implemented in literatures to control UPFC for various power system applications. In [7-9], decoupled control approach based UPFC has been designed to control the active and reactive powers flow in transmission line. However, the performance of a decoupling control system may vary significantly depending on the operating point of power system network. Cross-coupled controller in [10-12] has been implemented by considering the variation of the power system parameters. The drawback of both decoupled and cross-coupled controller based UPFC is the ignorance of the dynamics performance of the DC link capacitor. In addition, the interaction between the shunt and series converters of UPFC has not eliminated completely.

In addition, there is a common requirement of all controllers discussed above is the need of output feed proportional integral (PI) control system for UPFC to control power flow in power network [7, 13]. The disadvantage of PI based control algorithms is that to tune the optimum parameters i.e. K_p , K_i of PI controller, trial and error method must be performed. Also, any change in the power system dynamics cannot be resulted in changing in PI controller parameters [13]. To overcome the problems of the PI based UPFC controllers, fuzzy logic Controller (FLC) based UPFC has been proposed in this paper. The advantages of FLC over conventional PI controller are that it doesn't need an accurate mathematical model of the system and it is robustness to system parameters changes and dynamic operations [14, 15].

In this paper, a dynamic strategy to control power flow in transmission lines using dynamic UPFC model has been proposed. FLC is employed to develop the control algorithms for shunt and series converters of UPFC. Apart from this, the developed FLC based UPFC's performance

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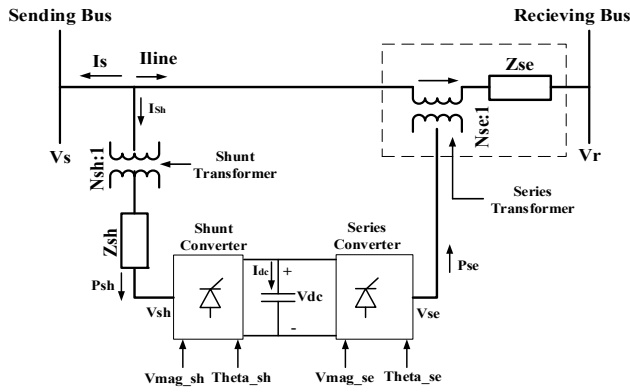


Fig. 1. Dynamic UPFC model

is compared with PI based UPFC's performance in terms of different system parameters. IEEE 14 bus system is considered as a case study.

The organization of the remaining paper is as follows: Section II introduces the dynamic model of UPFC developed in PSCAD. Section III presents description on shunt and series controllers of UPFC. Section IV executes the simulation results and discussions after implementing FLC based UPFC including a comparison with PI controllers. Finally, considering all the important aspects of the study, a conclusion has been reported.

II. DYNAMIC UPFC MODEL

A dynamic model of the UPFC is shown in Fig. 1. UPFC connects to the transmission line with shunt and series voltage source converters (VSC) which are coupled via a common DC link. Normally, the shunt VSC is considered as (STATCOM) and series one as a static compensator (SSSC). Low pass AC filters are connected in each phase to prevent the flow of harmonic generated due to switching. The transformer connected at the output of converters to provide the isolation, modify voltage levels.

III. CONTROLLERS OF UPFC

UPFC controller consists of shunt and series controllers for both shunt and series converters.

A. Shunt Controller

The control algorithm of shunt controller is shown in Fig. 2 which represents the FLCs of sending - end voltage and DC link voltage to generate (V_{d_sh}) and (V_{q_sh}). The shunt converter draws current from the sending end bus to keep the voltage of the DC link capacitor at its reference value and to maintain the sending end voltage at its reference value. To simplify the dynamic analysis of the shunt current (I_{sh}) controller, three phase shunt current (I_{a_sh} , I_{b_sh} and I_{c_sh}) have been converted to d-axis and q-axis components in a synchronously rotating $d-q$ frame ($I_{dsh_measured}$ and $I_{qsh_measured}$). The sending-end voltage ($V_{s_measured}$) is measured instantly, subtracted from its reference value (V_{s_ref}) which reveals V_{s_error} . This error signal and the rate of change of error ($V_{s_error_rate}$) have been given as inputs to another FLC

block to control the sending end bus voltage. The output of FLC gives reference quadrature component (I_{qsh_ref}) of the shunt current (I_{sh}).

The measured value of the quadrature shunt current ($I_{qsh_measured}$) has been compared with obtained reference quadrature shunt current (I_{qsh_ref}) and send to PI controller. The output of the PI controller will be multiplied with impedance (WL_{sh}) to produce the quadrature component of the injected shunt voltage (V_{q_sh}). Meanwhile, the difference between measured DC link capacitor voltage ($V_{dc_measured}$) and its reference value (V_{dc_ref}) is obtained to get V_{dc_error} and send to FLC to obtain the reference direct component (I_{dsh_ref}) of the shunt current. The measured value of the direct shunt current ($I_{dsh_measured}$) has been compared with obtained reference direct shunt current (I_{dsh_ref}) and send to PI controller. By multi-plying the output of the PI controller with impedance (WL_{sh}), a direct component of the injected shunt voltage (V_{d_sh}) can be found. The magnitude (V_{mag_sh}) and phase angle (α_{sh}) of the injected voltage by the shunt converter have been determined as follows:

$$V_{mag_sh} = \sqrt{V_{d_sh}^2 + V_{q_sh}^2} \quad (1)$$

$$\alpha_{sh} = \tan^{-1} \frac{V_{q_sh}}{V_{d_sh}} \quad (2)$$

To generate the switching firing signals, sinusoidal pulse width modulation (SPWM) has been used. For this purpose the phase angle of the reference sine wave has obtained from the difference ($angle_s - \alpha_{sh}$) of the two angles where the angle (α_{sh}) is determined from Eq. (2) and the angle ($angle_s$) is the sending-end voltage which has been extracted from phase locked loop (PLL). The magnitude (V_{mag_sh}) of the shunt injected voltage has been used as a magnitude for the reference sine wave. In the SPWM technique, the reference signals are compared with the carrier signal, which has a switching frequency of 4.5 KHz, to generate the firing signals for the shunt converter.

B. Series Controller

The controller of series converter is illustrated in Fig. 3. The series converter controls the active and reactive power flow through the transmission line by injecting a voltage in series with the line current having controllable magnitude and angle. The active and reactive power (P_{actual} and Q_{actual}) flow through the line are measured and subtracted from their references (P_{ref} and Q_{ref}). These revealed the error signals (P_{error} and Q_{error}). These two error signals and their rate of change (P_{error_rate} and Q_{error_rate}) are given as inputs to two fuzzy logic controllers (FLCs) as shown in Fig. 3. The outputs of the two FLCs provided the orthogonal components of the series injected voltage (V_{q_se} and V_{d_se}). The magnitude and phase angle of series injected voltage are obtained using equations (3) and (4).

$$V_{mag_se} = \sqrt{V_{d_se}^2 + V_{q_se}^2} \quad (3)$$

$$\alpha_{se} = \tan^{-1} \frac{V_{q_{se}}}{V_{d_{se}}} \quad (4)$$

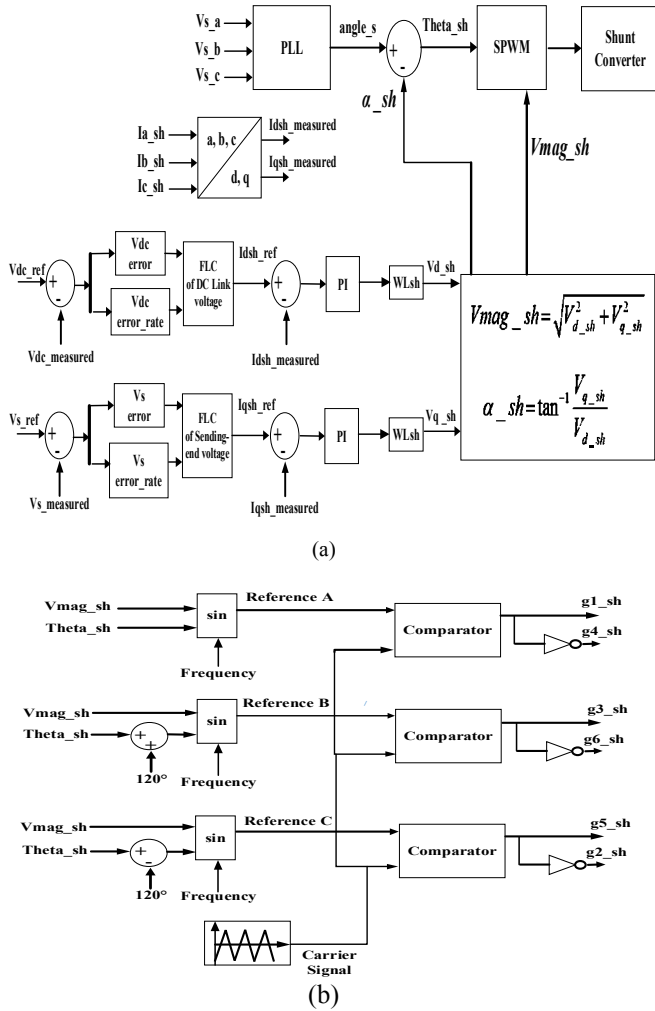


Fig. 2. (a) FLC of sending-end voltage and DC link voltage to generate V_{d_sh} and V_{q_sh} , (b) SPWM technique to generate firing signals for shunt converter of UPFC

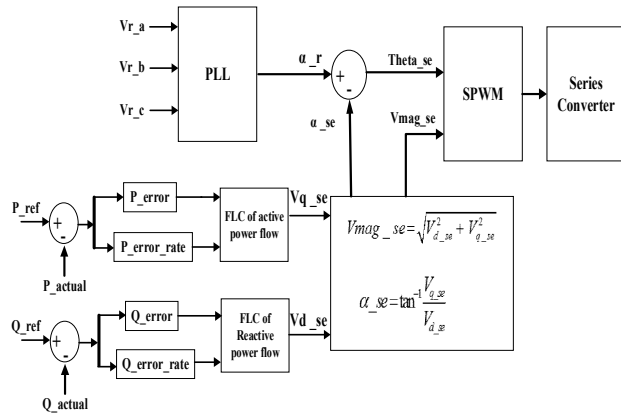


Fig. 3. Series controller of real and reactive powers to generate V_{mag_se} and α_{se}

SPWM technique has been used to generate the firing signals of the series converter as explained in the shunt controller of UPFC.

IV. RESULTS AND DISCUSSIONS

In this section, the simulation results for proving the robustness of the proposed controller are presented where IEEE 14 bus system is used as a case study. The parameters of the network is taken from [16]. The line selected for UPFC is based on the FVSI and VCPi stability indices. The detailed procedure to find the optimal location of UPFC using above stability indices can be found in [17]. The optimal location of UPFC for IEEE 14 bus system is 9-14.

The performance of the proposed FLC based UPFC is tested for the nominal operating state of the network. The reference values of power system parameters for line 9-14 are randomly chosen as: active power 0.072 p.u (7.2 MW), reactive power 0.09375 p.u (9.375 MVAR), and line voltage (1.01p.u).

The simulation is conducted till $t = 2$ sec without UPFC as shown in case "a" of Figs. 4 and 5. The measured values of active and reactive power flow through the line 9-14 without UPFC were 0.06251 p.u (6.21 MW) and 0.0998 (9.98 MVAR) respectively. UPFC based on PI and FLC are connected to the line 9-14 at $t = 2$ sec as shown in case "b" of Figs. 4 and 5. From Fig. 4, case "b", it has been found that with PI based UPFC the active power flow was 0.069 p.u (6.9 MW) which has been improved by 8.66 % of the nominal value, where with FLC based UPFC active power flow has been improved by 13.39% of the nominal value i.e. 0.07194 p.u (7.194 MW). With FLC based UPFC reactive power flow has reached to 0.09394 p.u (9.394 MVAR) which is closed to the reference value 0.09375 p.u (9.375 MVAR). In contrast, reactive power flow over the line with PI UPFC falls short by 0.00383 p.u from the reference value i.e. 0.09758 p.u (9.758 MVAR) as shown in Fig. 5, case "b".

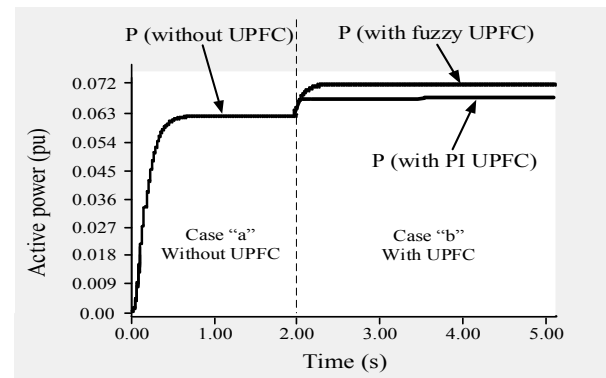


Fig. 4. Active power flowing through line 9-14 with and without UPFC installation

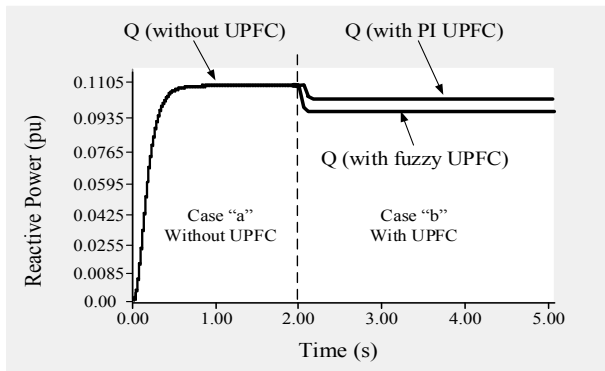


Fig. 5. Reactive power flowing through line 9-14 with and without UPFC installation

One of the advantages of UPFC is to maintain the line voltage of the transmission network. To address this case, the response of the proposed FLC based UPFC controller for line voltage is shown in Fig 6. Here also, the simulation is conducted till $t = 2$ sec without UPFC as shown in case "a" of Fig. 6. The measured value of the line voltage across without UPFC is 0.9886 p.u. According to Fig. 6, case "b", after Connecting UPFC at $t = 2$ sec, the line voltage has reached to 0.9931 p.u for PI based UPFC and this value has become 1.008 p.u for FLC based UPFC.

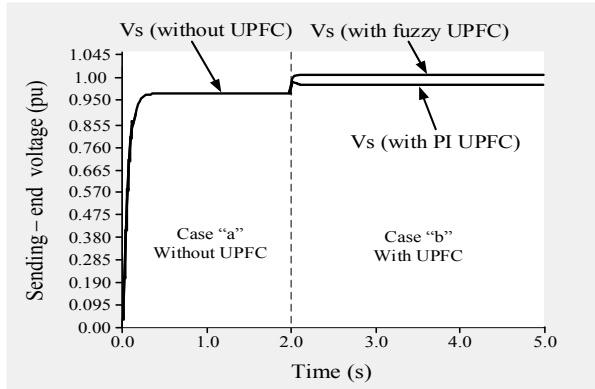


Fig. 6. Sending end voltage across line 9-14

The overall network voltage profile is demonstrated in Fig. 7 and it has been observed that with FLC based UPFC the voltages across all the buses of the network have been improved better than PI based UPFC.

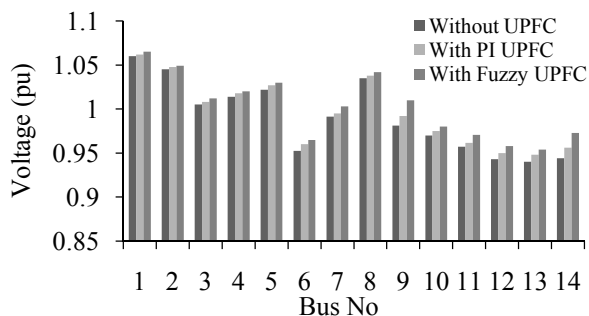


Fig. 17-19 Voltage profile across all the buses in IEEE-14 bus system

The response of the DC link capacitor voltage is shown in Fig. 8. As per Fig. 8, before $t = 2$ sec, the UPFC was not connected to the line, therefore DC link capacitor wasn't charging. At $t = 2$ sec, UPFC was connected to the line 9-14. The DC link capacitor was charged up to 9.45 kV based on PI UPFC which took 4.6 sec to reach to its steady state value. However, the capacitor charged to 10.435 kV with a time period of 2.75 sec when FLC based UPFC was in operation. For both the cases the reference charged value was set to 10.45 kV.

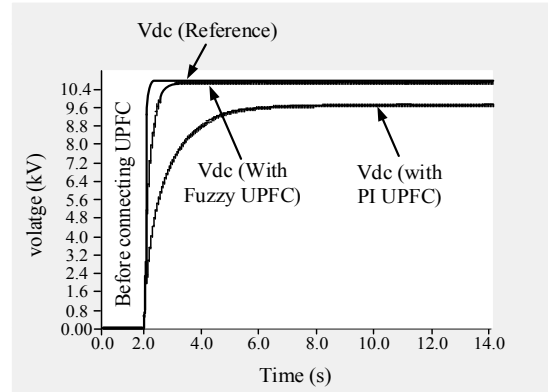


Fig. 8. DC voltage across DC link capacitor in IEEE-14 bus system

The waveforms of the series injected voltage for both PI and FLC based UPFC are presented in Fig. 9. From Fig. 9, it can be noticed that the UPFC was connected to the line 9-14 at $t = 2$ sec. The series injected voltage for both PI and FLC based UPFC were 4.5 kV and 5.4 kV respectively.

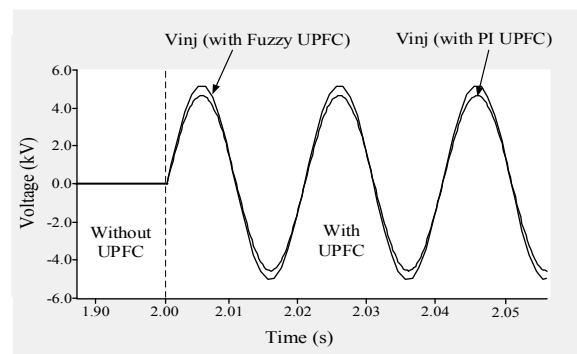


Fig. 9. Series injected voltage in line 9-14

V. CONCLUSION

This paper presented a real time implementation of FLC based UPFC to control the transmission line power flow on IEEE 14 bus network. The response of proposed FLC based UPFC is compared to PI based UPFC. The results showed that proposed FLC based UPFC has enhanced power flow ability, improved bus voltages and

reduced the power losses of the networks more than PI based UPFC.

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Different Perspectives of Energy Market in Bangladesh

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Abstract— Bangladesh, being a typical developing country, has undergone the energy crisis situation for a very long period of time. The energy system of the country has gradually been improving with the electricity accessibility ratio increasing year by year. The study has been based on different analytical information and data analysis regarding the Energy Market of Bangladesh. The study investigates on the current energy condition of Bangladesh which seemed to have significantly shown some improvement from the past 5 years regarding generation but it has a very uncompetitive energy market. Many researches have been made based on different aspects of the energy system in Bangladesh but mostly discussing problems with the system, policy making issues and efficiency. This study focuses on the drawbacks of the Bangladesh's energy system which is hindering the energy market of the country to flourish.

Keywords— energy market; monopoly; oligopoly; electricity

I. INTRODUCTION

Bangladesh is one of the densely populated countries in the world. Due to this over population, Bangladesh is having insufficient energy system. The energy supply stream is not properly managed which leads to the path of erroneous energy services and thus hindering the path of this sector's development. This creates an imbalanced condition between the demand and production of energy and leads to insufficient distribution. To fulfill the energy demand of the growing population, Bangladesh needs to set up a complete energy market that will conduct the key factors of energy services. In this paper, different perspectives of an energy market have been discussed and notable solutions to introduce efficient electricity market have been suggested.

II. ENERGY MARKET CONCEPT

The energy market is all about the mechanisms of trading energy and balancing the market and its individuals. Energy market is globally increasing and spreading over its individuals for several decades. Energy resources are an important factor in the market. Roughly 70% of natural gas (NG) flows across the globe are transported to market destinations within the country of production, while an additional 20% flows cross international borders through pipelines, and nearly 10% is moved to market destinations as liquefied natural gas (LNG). Approximately 30% of international NG trade is represented by LNG [1]. The use of coal for power generation and other purposes grew faster

globally than any other energy sources. Renewable energy is an enormous rise which leads the energy consumption to a new stage. Asia's emerging powers have accounted for about 65% of growth in the global oil demand over the past two decades [2]. Moreover, Europe and several emerging and developed countries of other regions have already been engaged themselves to develop the global energy market.

III. ENERGY MARKET OF BANGLADESH

A. Natural Gas

The current situation of NG sector is undoubtedly not able to fulfill the country's desire. It has become urgent to use mainland gas resources in a planned manner until it can explore and discover gas in offshore. It is reported from a survey [3] that if old boilers of the existing industrial units are transformed by new technology boilers, 150 MCF of gas can be saved per day. Failing to develop domestic coal or offshore gas, the govt. is now opting for an even costlier method of generating electricity using furnace oil and gas. It plans to generate 8,125 MW of electricity by installing seven power plants run on imported re-gasified LNG.

B. Power

The Electricity generation infrastructure of Bangladesh is still underdeveloped with the increasing demand and insufficient production. Approximately 38% of the population is in dearth of electrical support and these issues are improving very slowly. Though Bangladesh is aiming at mutual co-operation for power generation with different countries, this is not causing a great help now in this situation. As a result, Bangladesh might need to make efforts to explore off-shore energy.

C. Renewable Energy

Renewable technologies are massively used in most developing countries but it is still a very young system in Bangladesh. The technology requires a good amount of investment and feasible conditions to attain a sustainable solution. New moves are taken by government, private and commercial organizations to provide financial support and fiscal incentives. To encourage people to use renewable energy, the Infrastructure Development Company Limited (IDCOL) has initiated installation of 0.2 million solar home systems (SHS) since May 2013. As a result, about 0.8 million people in the rural areas now have access to solar power [4].

D. Economic Growth and Demand

The economic growth of Bangladesh will reach to another level of success if the wastages of power and gas can be reduced to a certain level. Maximum demand served so far is 7,817 MW where the consumer number is 29, 01,235 (FY-2014). At present, only 58% of population is served with grid connected electricity and per capita electricity consumption is only 285 kWh (FY-2013). According to the latest government figures, Bangladesh's gross domestic product (GDP) stands at Tk. 15.13 trillion in July to March period of the ongoing fiscal. GDP growth has a linear relationship with the increasing energy demand. As Bangladesh is progressing day by day with GDP growth rate, energy demand is also increasing with it.

E. Energy Budgetting

The government has proposed Tk. 9,400 crore as subsidy to the power and energy sector for this fiscal (2014-15) year. This amount was shown in the budget document as "net lending" to Power Development Board (PDB) and Bangladesh Petroleum Corporation (BPC), both state-owned entities. In this budget the government pointed out that power generation target was re-fixed at 24,000 MW instead of 20,000 MW by 2021 considering the growing demand and increasing generation capacity. Necessary steps have been claimed to be taken to raise power generation capacity to 18,162 MW by 2017 [5].

F. Private Investments

Private investment for electricity can be cumbersome due to bureaucracy issues. For maintaining transparency and accountability in the private pricing framework, Bangladesh Energy Regulatory Commission (BERC) arranges public hearings involving stakeholders, policy-makers, government officials etc. Compared to the 1970s, the cost of installation per MW has decreased, making small scale technologies more competitive. The new players are, therefore, investing in small scale fossil fuel and gas powered generation units which have higher marginal costs of production. Significant private investments are being made by renowned NGOs of the country for the advancement of the both the fossil fuel and renewable energy based power stations in Bangladesh [6].

IV. MONOPOLY ENERGY MARKET OF BANGLADESH

In Bangladesh, the electricity market is still in monopoly system which is hindering the system from significant development. Less competition is leading to lesser transparency of the system.

A. Electricity Utility Model

Bangladesh has a typically rigid electricity model but with very limited participants. Bangladesh Power Development Board (BPDB) is the authority for planning, construction and operation of power generation and transmission facilities throughout and for distribution in urban areas except metropolitan city of Dhaka and its adjoining areas. Few private owned companies are contributing in power generation as well. As clearly explained in Fig. 1, Power Grid Company

of Bangladesh Ltd. (PGCB) solely controls the electricity supply in Bangladesh with no significant competition. Operation & maintenance of grid sub-station and transmission line, transmission network planning and every other transmission details are controlled by PGCB. Whereas, the distributors like Dhaka Power Distribution Company Limited (DPDC), DESCO, WZPDC, REB/PBSs serves the customers with the demanded facilities [7].

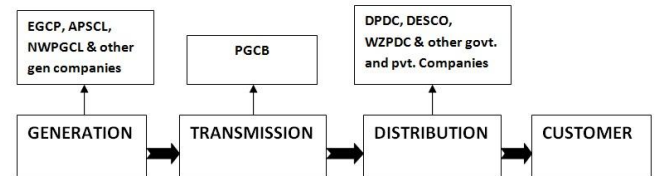


Fig. 1. Electricity Model of Bangladesh

B. Regulation

BERC makes provisions for the establishment of an independent and impartial regulatory commission for the energy sector. Fig. 2 comprises all the responsibilities regulated by BERC. The pricing policy for power and energy is formulated and implemented by BERC with support from related agencies. They observe the daily public demand of electricity and makes sure the generation companies are producing the sufficient amount on daily basis and also regulates the maximum amount that are being transmitted to the distributors.

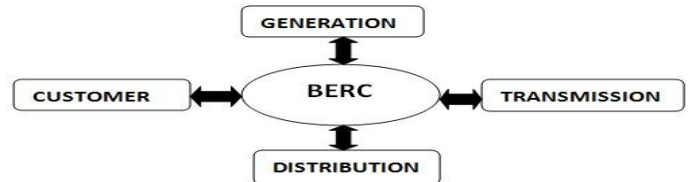


Fig. 2. Regulatory units of Electricity by BERC

C. Competition

The energy market of Bangladesh is quite rigid with a monotonous system. There are couple of generation companies which produce electricity for the assigned regions and therefore with no trace of competitiveness among the producers since they haven't have the culture of being taken over the projects by some new bidding companies. Each company generates electricity for a particular area of the country and the highly dense region like Dhaka city requires more than one generation plan to meet the demand. There is only single transmission company which is a public organization, PGCB, transmitting generated electricity throughout the country along with the distributors. On the other hand the distributing companies like DESCO, DPDC etc. are allocated with daily energy/power, by PGCB, based on their individual daily requirements [8].

D. Market Operator

Bangladesh doesn't have an independent electricity market and therefore no trading between generators, retailers and

other financial intermediaries before reaching the customers. The selling rates of the generation market are pre-determined so there is not much room for high profit by the investors in this sector. The only transmission authority, PGCB has the liberty to charge significant amount since they are independently on hold of the transmission market, so no biddings or bargains are eligible for this section of the market. It should be mentioned that BERC arranges public hearings involving stakeholders, policy-makers, govt. officials and eminent persons from civil society with a view to maintaining transparency and accountability in the pricing framework. Currently the case for providing both implicit (off-budget) and explicit subsidies in electricity by the govt. arises since in both cases the bulk of retail tariff rates are set below the supply cost of electricity. Therefore, BPDB as well as the distribution companies incur huge losses. The policy of reducing subsidies thus requires narrowing the gap between the selling prices and supply costs [9]. The govt.'s approach has involved increasing the bulk and the retail tariff rates in a gradual manner. Increasing the bulk tariff rate improves the financial situation of BPDB which, however, increases the supply cost of electricity at the retail level. This necessitates increasing the retail tariff rates for the power distribution companies otherwise they would face loss. Hence, increasing the bulk tariff rate should be based on a comprehensive assessment of the impact of a tariff hike on retail tariff rates paid by different categories of consumers. Fig. 3 shows the work cycle of the market operations.

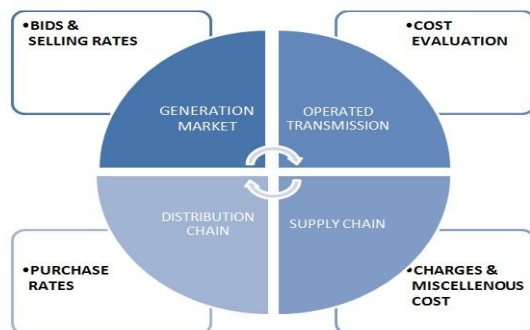


Fig. 3. Role of Market Operator

V. FACTORS OF ELECTRICITY MARKET

To create a perfect electricity market, there are some factors that participate to control balance in an electricity market. It's not possible to store electricity on a large scale. It should be generated and consumed at the same time. Therefore it is must to have a balanced market.

A. Bidding

Bidding Strategy is an efficient mechanism of a perfectly competitive electricity market. For a developing country, bidding should be an essential part of electricity markets to increase transparency and competition and it should focus on the theory of profit-maximization based on the residual demand curve. A perfectly competitive market discovers some benefits through the bidding process that largely influence the

future of the market. Cost effectiveness is one of the advantages of a bidding process which allows companies to control the costs. The costs are associated with the audit process. There is less chance of arising conflicts of interests through the bidding process because the contract is made on the basis of merit and open competition. Bangladesh needs to focus on the market participants or competitors who can be the key start of a bidding system. Initiating the bidding system will help achieving a competitive energy market.



Fig. 4. Factors enabled by strategic bidding mode [10]

B. Market Power

Market power is the ability which influences the pricing and supply decisions of a firm. At present, Bangladesh might be facing the condition of monopoly as there is one firm mainly playing the leading role and hence it is becoming dominant on the system without any ethical evaluation. Moreover, this price maker has the power to control the price it charges. Generating companies of Bangladesh are mainly playing the role of price maker as they are in charge of the selection process of generation resources. To get a perfectly competitive market with the proper involvement of the market players, many firms are needed to break this influencing power.

C. Options and Contracts

Electricity or Energy contract means “energy offers” or “energy plans”, the offers or plans which will help to accomplish virtual divestiture. If Bangladesh wants to focus on the standardized and exchange based trading system then future contract will be the right one to choose. In future contract, there has to be a new future price rewritten everyday which could be a little lengthy for the beginners. Again if the trading system is bilateral and based on the private agreement then it should focus on the forward contract which includes a specific expiration for the delivery and payment of the asset. So this contract follows a timeline which could be an advantage for the market. Again there is another contract which is called option contract that has a call and a put option which gives the holder the right to buy and sell. At present, Bangladesh is not able to follow any of the above mentioned methods because of the unorganized condition of the electricity market. By analyzing all the feasibilities and possibilities, Bangladesh needs to practice one of the contracts.

D. Trading Concept

Bangladesh needs a robust trading system which will be capable of risk hedging and other changes. A competitive electricity market holds the proper trading tools to restructure and reform the objectives of the market. If the country focuses on the trading system which include competition then it will be much more beneficial because of the preference of the customer's choice. In that case, the companies will be able to get an idea about the demands and supplies of electricity.

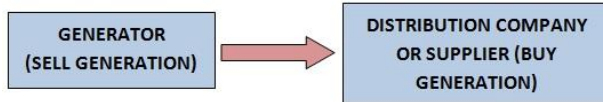


Fig. 5. Trading system between a generator and a supplier (Distribution Company)

In Fig. 5, generators are selling electricity generation to distribution companies. Distribution companies or suppliers buy this generation to meet the demand of their customers. This trading system follows the bilateral contract agreements and can be done by using forward and option contracts. Again, for a non-physical trader there is a slight difference in electricity trading. From Fig. 6, it is seen that nonphysical traders such as investment authorities do not sell any generation and they don't have to satisfy the customer either. Generators sell generation to nonphysical traders and non-physical traders will sell it to distribution companies or suppliers. This system is also following the bilateral trading system and hence forward and option contracts can be used for this system.

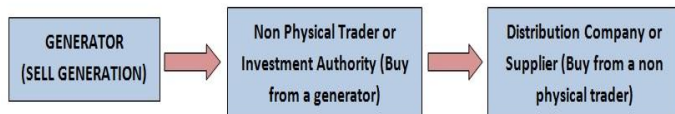


Fig. 6. Trading system of a nonphysical trader or Investment Authority

Sometimes electricity trading is done by power pools or power exchanges as well. Power pools and power exchanges are the two types of organized market. Power pools follow a public blueprint where the participation is obligatory. The main participants in power pools are the generating companies because generating companies hold the authority of proposing price-quantity pairs for the supply of electricity. Power exchanges are based on private scheme which follows willful participation. Here every participant can participate and specify desired price they want to pay or receive. Then the price will be set based on the supply and demand curves. Power exchanges ease the process of transactions between sellers and buyers [11].

VI. CHALLENGES TO OVERCOME

- Government has to initiate advanced laws and techniques to upgrade the system instead of technologically lagging behind. More rational investments are essential from both private and public sectors that will nourish the electricity market into a transparent and fair structure.

- The monopoly system needs to be broken and move towards oligopoly because enough of the days of "price makers". More companies, private or govt., have to come forward and increase the market competition.
- The bureaucracy of the energy sector needs to be simpler and more effective. The gap between the consumers and producers needs to be narrowed down so that efficient outcomes get the priority.
- Government should work effectively to utilize the abundant natural resources in Bangladesh and initiate tasks more punctually. The country has to be intuitive and have a secured energy system.

VII. CONCLUSION

The world is moving towards proper execution of Smart Grid technology. To facilitate the movement, significant steps have to be taken to change the conventional electricity sector, which includes introducing proper energy market. The unbalanced condition prevailing in Bangladesh has made the energy sector inconsistent [12]. The trading system and bidding strategies are recommended to follow along with more competitors or participants. An electricity market participant holds the responsibility of balancing service. Our observation says that the current monopolistic competition in Bangladesh's electricity market is lagging because of the absolute sovereignty to an individual (participant) rather than customer's choice.

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Comparison of Series and Combined Series-Shunt Compensation on East-West Interconnectors of Bangladesh Power System

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Abstract—Utilization of the existing transmission line capacity to its fullest is most important due to the continuous growth of the demand and rising cost of new transmission infrastructures. Due to the problems associated with voltage drop and stability limits, a transmission line can not be loaded up to its thermal limit. The potential benefits of installing Flexible AC Transmission System (FACTS) devices on the system for enhancing power transmission capacity and stability are well known. Two main members of FACTS family like thyristor controlled series capacitor (TCSC) and static var compensator (SVC) have been widely studied and reported in technical literature and have been shown as effective as well as time saving alternative to building new transmission lines for enhancing power transmission capacity and stability. This paper compares the application of series and series-shunt compensation by means of TCSC and SVC to show enhancement of power transmission capacity and stability of East-West Interconnected System (EWIS) of Bangladesh Power System (BPS). Two SVCs are connected in parallel with two buses in the western region of the interconnectors and two TCSCs are connected in series with the interconnectors. Bifurcation, steady state, and transient analyses are performed to compare positive effect of series and series-shunt compensation on transmission capacity and stability in BPS network.

Index Terms— Controller parameters, TCSC, SVC, oscillation, EWIS.

I. INTRODUCTION

It is important to optimize the utilization of the existing transmission systems in a competitive market environment due to the challenges associated with constructing new transmission lines. Evolution of different commercial schemes for the electric power industry, mainly deregulation of market, has made this requirement stronger [1]. A transmission line can be loaded up to its thermal limit; however, if the receiving zones do not have compensating reactive components, the voltage can drop below the limit constant by quality criteria [2]. In such situation, it is logical to limit the transmission of these lines to avert excessively low voltages in the receiving end as well as to avoid a possible voltage collapse [3]. Power transmission capacity and stability can be enhanced directly

by using series and shunt compensation [4]. The objective of FACTS technology is to bring a system under control and to transmit power as premeditated by control centre. Improvement of system stability is also observed with FACTS. The use of FACTS controlled series compensation and shunt compensation; particularly TCSC and SVC for stability enhancement have been reported in [5]. A Fixed level of series compensation with TCSC can improve transient stability of power system; but for enhancing more transient stability, a damping controller can be added with TCSC for modulating the reactance of TCSC [6, 7].

To suppress inter-area mode oscillations and to improve power system damping in [7], J. X. Ling et. al. proposed adaptive TCSC damping controller based on oscillation energy. In the proposed method no detailed information about system data and configuration of the studied power system is required. Implementation of this controller involves only active power flow along the tie line as feedback where the TCSC is installed. Following a disturbance in the system kinetic energy and potential energy of the oscillations as well as their sum increases from zero until the disturbance is cleared. Conversion rate of potential to kinetic and kinetic to potential energy determines the magnitude of oscillations.

Power transmission capacity and system stability improvement by Static Var Compensator (SVC) have been presented in [8]. To control bus voltage, SVC involves voltage control loop which is composed of voltage regulator, TCR and Thyristor Switched Capacitor (TSC) firing circuit, and voltage magnitude transducer. H. K. Tyll [9] presented detailed information about how SVC can play an important role in real power system by appropriate reactive power compensation. The author demonstrated that at all load conditions like heavy load, light load, outage of any line during heavy load condition, load rejection at the end of the line, SVC can minimize voltage deviations and can maintain good voltage profile. In the stressed system the first peak cannot be influenced because the SVC control must first observe the increase and can only react afterwards. Having sensed first peak of oscillation by the SVC damping controller, system stability margin is maintained within the prescribed limits by

damping the oscillations of the. In [10], R. Griinbaum et. al. presented several examples of benefits of FACTS like TCSC and SVC in power transmission systems. For the system with increasing transmission line loading over long distances, TCSC and SVC are shown to be very robust and effective means to improve damping. The damping effect of TCSC and SVC has different benefits like the effectiveness of these FACTS members for controlling of power swings increases with high levels of power transfer, improvement of voltage regulation, reactive balance and transient stability improvements. FACTS devices have been studied for voltage stability and damping oscillation improvement of power system in [11-12].

Most of the thermal capacity of east-west interconnectors (EWIs) of Bangladesh Power System (BPS) is unexploited [6]. Besides, inter-area oscillation in the system is a stability problem. Application of controlled series compensation has been studied and it has been found that even with highest level of series compensation 15% capacity of first EWI and 45% capacity of second EWI are left unused [13]. This paper studies combined application of series and shunt compensation to enhance transmission capacity and stability and compares it with series compensation.

Section II provides a brief description of analytical techniques used in this work. TCSC and SVC are discussed in section III. System modeling is presented in Section IV. Simulation results and comparisons are provided in Section V. Finally, section VI presents the conclusion of this work.

II. ANALYTICAL TECHNIQUES

To capture static and dynamic behavior of the EWIS steady state, bifurcation and transient analysis have been performed. Steady state analysis is performed to determine transmission angle between sending and receiving end buses. Transient stability analysis is conducted to determine system stability upon occurrence of a large disturbance. Static voltage collapse phenomena may be explained as a dynamic effect of a generic saddle-node bifurcation. Continuation power flow analysis is used to find out such saddle-node bifurcation [5, 15].

III. TCSC AND SVC

A. Thyristor Controlled Series Capacitor (TCSC)

The TCSC is the member of FACTS family that is used for the purpose of series compensation. The basic TCSC structure consists of a thyristor controlled reactor (TCR) module and a fixed capacitor connected in parallel as shown in Fig. 1. This fixed capacitor and TCR branch act together to produce preferred reactance.

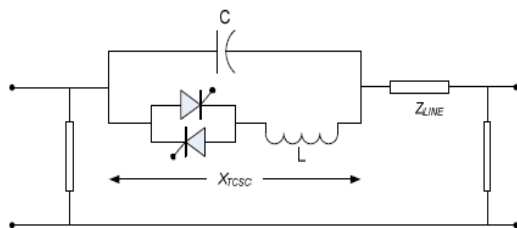


Fig.1 Schematic Diagram of TCSC and SVC

The equivalent reactance of TCSC represented as a function of the firing angle of thyristor (α) is given by the following equation [6].

$$x_e = -x_c \left[1 - \frac{K_x^2 (\sigma + \sin \sigma)}{(K_x^2 - 1)\pi} + \frac{4K_x^2 \cos^2(\sigma/2)}{\pi(K_x^2 - 1)^2} \left(k_x \tan \frac{K_x \sigma}{2} - \tan \frac{\sigma}{2} \right) \right] \quad (1)$$

Where,

$$\sigma = 2(\pi - \alpha)$$

$$K_x = \sqrt{\frac{x_c}{x_L}}$$

x_c and x_L represent capacitive and inductive reactance respectively at fundamental frequency.

B. Static Var Compensator (SVC)

SVC is the shunt compensation device whose main role is to regulate voltage at a selected bus by adjusting the reactive power injection. SVC increases steady state power transmission limit and hence enhances the transient stability of the system [8]. Common structure of SVC and TCSC is shown in Fig. 1. The main differences between these two devices are: SVC is shunt connected to a bus whereas TCSC is series connected to the transmission line, and SVC connected to the bus through a step-down transformer whereas TCSC is directly connected to the transmission line.

The controllable reactance of TCR of SVC, x_v is given by following equation.

$$x_v = x_L \frac{\pi}{2(\pi - \alpha) + \sin 2\alpha} \quad (2)$$

Where, x_L represents fundamental frequency inductive reactance of TCR.

IV. SYSTEM MODELING

Bulk power is generated in the eastern grid of BPS network. This power is transmitted to the western grid by means of two tie line that are Ghorasal to Ishurdi (1st EWI) and Ashuganj to Sirajganj (2nd EWI). Fig. 2 shows EWIS system of BPS network. This topology has been derived from the actual network condition of BPS.

In this work series and combined series-shunt compensation is applied to the two existing interconnectors between eastern and western grid. In order to be able to apply series compensation in the EWIs by means of TCSC two additional buses are connected adjacent to the Ishurdi and Sirajganj buses. To apply shunt compensation in the EWIs, two SVCs are connected to the Ishurdi and Sirajganj buses.

V. SIMULATION RESULTS

Saddle Node Bifurcation (SNB) and Hopf Bifurcation (HB) analyses are performed to find out maximum loading point. Steady state analysis is carried out to demonstrate improvement in transmission angle and bus voltage by different compensations. Finally, stability improvement by series and combined series-shunt analyses is observed by performing transient simulation. UWPFLOW software tool has been used for bifurcation analysis [16]. Steady state and transient analyses have been performed using Power System Toolbox [17].

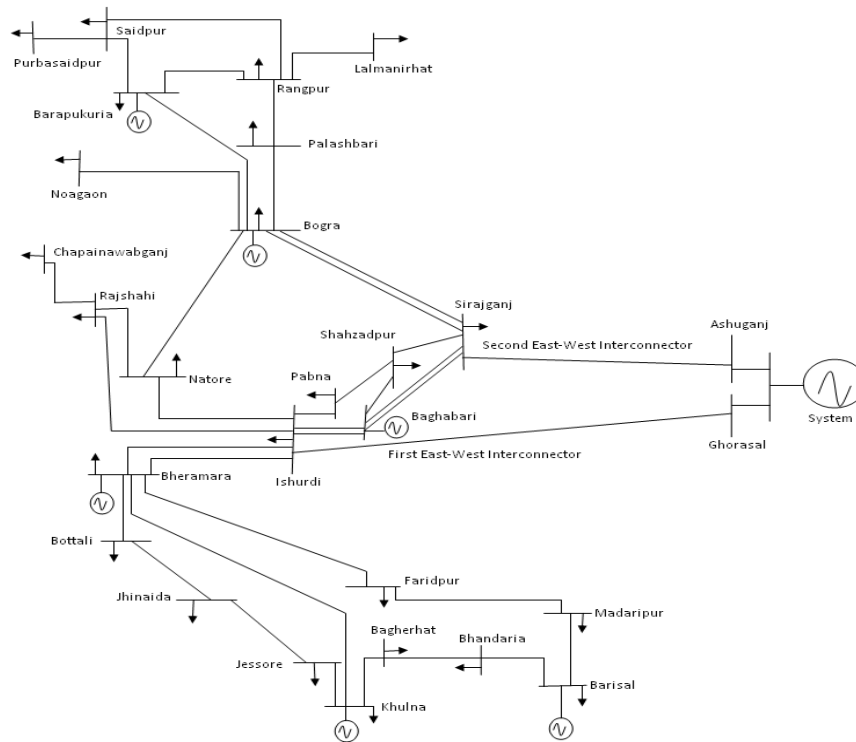


Fig. 2 East-West interconnected System

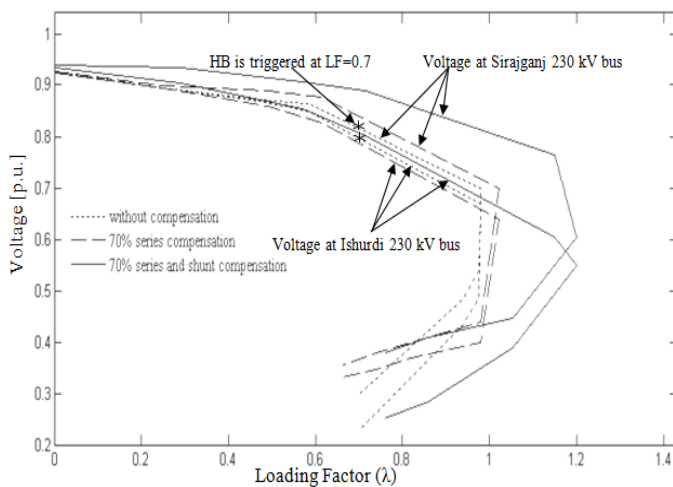


Fig. 3 Comparison of bifurcation diagrams of EWIS with different compensation levels

Bifurcation diagram shown in Fig. 3 clearly demonstrates that system without compensation experiences voltage collapse at a loading factor of around 0.98. However, before this loading level HB is triggered at a loading factor of 0.7. So, without compensation maximum loading factor is 0.7. After this the system experiences oscillatory instability before voltage collapse corresponding to loading factor of 0.98. Application of 70% series compensation on EWIS by means of TCSC increases system loading level at around 1.022. No HB is triggered up to this loading level. With combined 70% series and shunt compensation system loadability is further increased up to a loading factor of 1.2.

Steady state analysis has been performed on different loading levels without, with 70% series, and with 70% series

and shunt compensation. Table I summarizes steady state analysis of the system. Without compensation maximum 768 MW power can be transmitted from eastern to western grid. The figure is 1064 MW with 70% series compensation and 1228 MW with combined series and shunt compensation as shown in Table I. Ishurdi and Sirajganj bus voltages and transmission angles are also shown in Table I.

TABLE I
STEADY STATE CONDITION OF EWIS WITH DIFFERENT COMPENSATION LEVELS

Compensation level	Maximum loading factor	Import to western grid [MW]	Bus voltage [P.U.]		Transmission angle [Deg.]	
			I	S	G-I#	A-S*
Without	0.7	768	0.83	0.85	17.82	15.30
With 70% series	1.022	1064	0.80	0.82	9.05	6.050
With 70% series and shunt	1.2	1228	0.82	0.84	11.58	8.062

Ghorasal-Ishurdi *Ashugang-Sirajganj

Table II presents the percentage enhancements in power transmission capacity achieved with different compensations. It is seen that a 38.59% increase in power transmission capacity has been achieved with series compensation. The figure is 59.89% with combined series-shunt compensation.

TABLE II
ENHANCEMENT OF POWER TRANSMISSION CAPACITY WITH DIFFERENT COMPENSATIONS

Compensation level	Maximum power transmission [MW]	Cause of Limitation	Enhancement [%]
Uncompensated	768	HB	----
70% Series compensation	1064	SNB	38.57%
70% series and shunt compensation	1228	SNB	59.89%

Improvement in transient stability with series-shunt compensation compared to the system without and with series compensation has also been observed. Transient simulations have been performed at $LF = 0.7$, which is the maximum loading factor of the EWIS without compensation considering dynamic stability limit. A three phase fault is applied at the Ishurdi bus in the western grid during time domain simulation. The fault is cleared in three cycles. Time domain simulation profiles of faulted bus voltage and machine speed deviations of Baghabari unit are presented in Figs. 4 and 5, respectively. The figures show the improvements in system transient performance with series-shunt compensation compared to the system without and with series compensation.

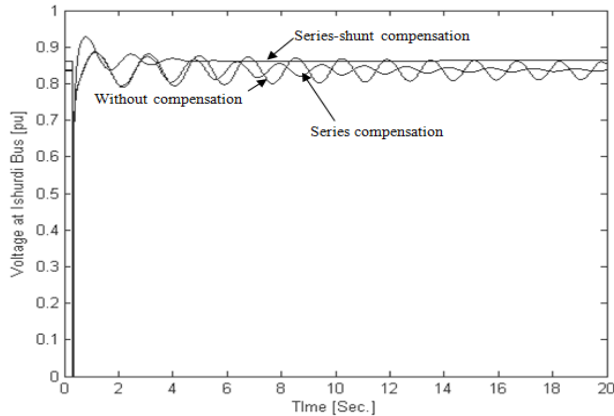


Fig. 4 Voltage profile at Ishurdi bus

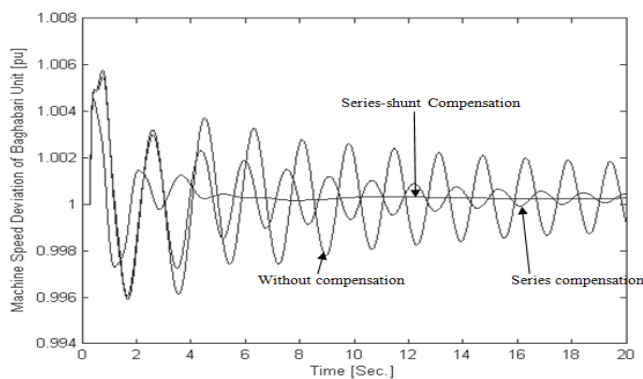


Fig. 5 Machine speed deviation

VI. CONCLUSIONS

The combined application of series-shunt compensation has been studied in this work to enhance power transmission capability and stability of EWIS of BPS network and has been compared with series compensation. Improvement of system stability as well as transmission capacity of the EWIS have been observed. Transmission capacity enhancement of 38.57% has been achieved with series compensation and 59.89% with combined series-shunt compensation. Transient simulations show improvement of damping of oscillations as well as voltage profile with series and combined series-shunt compensation.

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Torque Sensor Based Electrically Assisted Hybrid Rickshaw-Van with PV Assistance and Solar Battery Charging Station

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Abstract— A significant percentage of population in Bangladesh is directly or indirectly dependent upon the traditional tri-wheeler rickshaw. In order to advance the lifestyle and modernize the traditional tri-wheeler, battery powered rickshaws were introduced and thus it gained enormous popularity. However, after a few months, the Bangladesh government decided to ban import and assembly of electric vehicles, as it consumed energy from the already overloaded grid. Aiming to modernize the traditional tri-wheeler and make efficient use of renewable energy, Control & Applications Research Centre (CARC), BRAC University has proposed a field tested solution with the use of PV panel, torque sensor pedal and solar battery charging station to make the whole system independent of national grid. The paper describes the design and implementation of the PV array and torque sensor into the system and findings from the field tests along with the implementation of the solar battery charging station. It was found that 73% energy is saved when both PV array and torque sensor were integrated into the system. The distance travelled by the rickshaw-van was around 2.8 times greater than normal battery operated rickshaws.

Keywords— Electrically assisted; PV array; rickshaw-van; torque sensor; solar battery charging station

I. INTRODUCTION

Rickshaw, being the most popular and comparatively faster mode of transportation in Bangladesh, the modernization in this sector and electrification of rickshaws caused more energy consumption from the national grid, resulting in more load shedding and thus the government decided to ban the battery-run electric rickshaw [1]. Work on reducing this energy consumption from the national grid by installing PV array on rickshaw[2] and replacing the throttle input with a torque sensor pedal and a solar battery charging station has already been done [3] individually. To modernize the rickshaw-van, Control and Application Research Centre (CARC), BRAC university has designed motorized rickshaw-van with PV assistance and torque sensor pedal with solar battery charging station to charge the batteries [4,5] This paper represents the design and performance analysis of the

rickshaw-van with PV array and torque sensor pedal installed as shown in Fig.1 to determine the percentage of energy saved. A solar battery charging station is also implemented to charge the batteries and thus making the whole system independent of national grid [4].

II. PROJECT OVERVIEW

Initially, the tri-wheeler was a full throttle controlled electric vehicle, manufactured by Beevatech Limited. The motorized rickshaw-van is made with light weighed modern steel body and a different architecture compared to the traditional rickshaw or vans. Usually a traditional rickshaw-van as shown in Fig.1 [6] is capable of carrying 4 or 5 passengers, whereas the motorized rickshaw-van is capable of carrying 6 to 8 passengers. The modernized tri-wheeler with PV array and torque sensor pedal installed is shown in Fig. 1. The motorized rickshaw-van has a 750W, 48V brushless DC gear motor, 48V (4 X 12V, 25Ah connected in series) lead acid battery, a controller unit, a throttle, main power key, emergency motor stopper, traditional front wheel brake and an extra rear wheel break, charge controller, charge indicator, headlights and direction indicators. A 400Wp (4 X 100Wp connected in series) panel is installed on the roof of the rickshaw-van that provides a share of the required electrical load alongside the battery bank to reduce the energy consumption from the battery bank and thus increase the battery cycle duration. Another purpose of this array is to charge the batteries when the tri-wheeler is at rest, thus reducing the need to charge the batteries in the middle of the day when the grid is already supplying power at peak condition. A torque sensor pedal is introduced in the system to reduce the over-use of the battery bank. The rickshaw-puller will get assistance from the motor while paddling. A torque-



Fig. 1. Traditional rickshaw-van and rickshaw-van with PV array and torque sensor pedal installed

transducer pedal will output the corresponding voltage of torque applied upon pedal by the puller, which will be compared to a predetermined value (the minimum torque). If the applied torque is more than the predetermined value, the motor will be allowed to assist the puller. After crossing this threshold, the more the torque (voltage) is excess to the threshold, the more assistive power will be provided to the puller, proportionately. Certainly, there will be a cut-off; which means that there will also be a maximum value of torque. After that torque, the motor will either not be allowed to assist the puller, or be assisting with its maximum level preset by the designer. These will be done for the safety of the motor and also the driver [7]. The operation of the vehicle with both PV array and torque sensor is shown in Fig. 2. A solar battery charging station is implemented on the rooftop of BRAC University Campus to charge the batteries shown in fig 11. This avoids consuming power from the already overloaded grid during peak hours and allows efficient use of renewable energy. The whole project is independent of national grid.

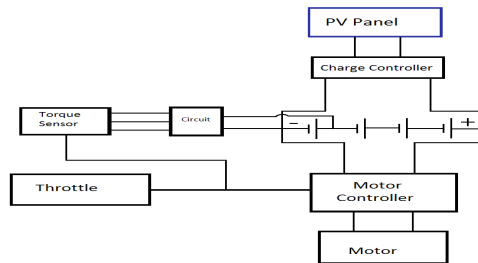


Fig. 2. Block diagram of the whole system

III. TORQUE SENSOR DESIGN

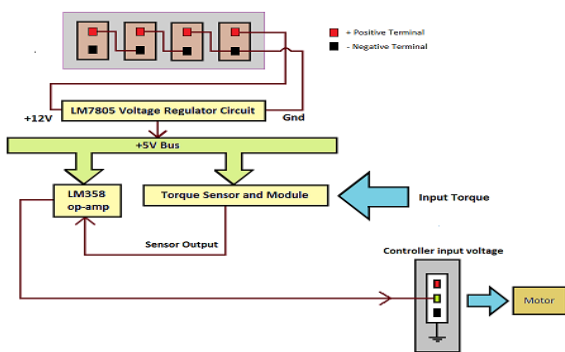


Fig. 3. Block diagram of the torque adjuster

The specification of the torque sensor is available at [8]. The torque adjuster diagram of the system is illustrated in Fig. 3. The key devices used here are LM7805 and LM358. LM7805 is a voltage regulator which gives an output of +5 V when the voltage more than 7.5 V is provided as input. The output is always fixed and it is +5 V. The amplifier's maximum output is 3.6V when a voltage of +5V is supplied. This characteristic will help the system to maintain a maximum tolerable voltage input to the CU without interrupting the whole system by motor switching or sudden-stop-runs. For instance, if gain is set to 2, an input of 1.5V will give 3V as output, but an input

of 2.5V will give a fixed 3.6V instead of giving 5V as output. This advantage is utilized in the circuit diagram. While practically running the vehicle, it was seen that when puller uses torque sensor pedal, the voltage generated is much higher around 3-4 V. So to minimize this voltage, a voltage divider is used in such a way that output voltage from torque sensor pedal to the circuit is 0.6 times.

IV. CONTROL ALGORITHM

The torque sensor or throttle signal controls the speed of the BLDC motor. When there is no signal from the torque sensor or throttle, the rickshaw is at stationary. The PV array is generating power but rickshaw is at stationary, the total generating power will be used to charge the battery only. Fig 4 shows the flowchart of the whole system. The conditions that may occur depending upon the power required by the motor by using throttle or torque sensor pedal are listed below.

1. The PV array cannot generate any power due to solar irradiation, in this case battery supply the load motor solely.
2. The PV array generates more power than required by the load, in such situation, excess power is used to charge the battery.
3. The PV array is generating power but not adequate enough to run the load, in this situation load power is shared by the PV array and the battery bank.

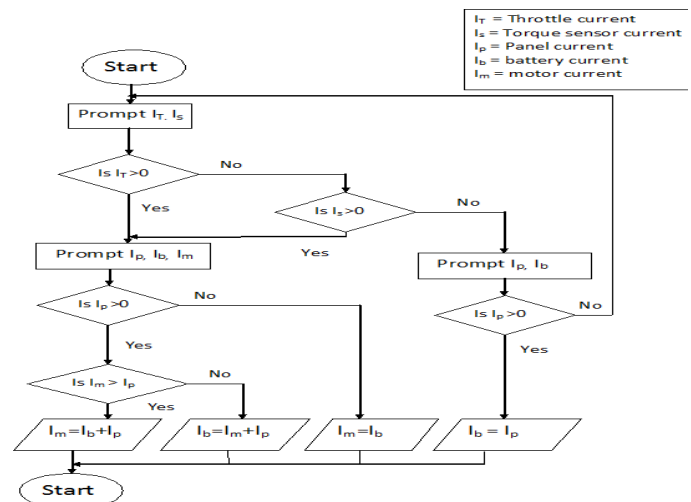


Fig. 4. Flowchart of the whole system

V. FIELD TEST

The objective of the field test was to find out the percentage of energy saved by using both PV array and torque sensor pedal. The data was obtained during the peak sun hours from 12pm to 3pm. The test was conducted in an area which had free road for running the rickshaw-van and sufficient amount of solar irradiation to determine the impact of the PV array on the performance of the rickshaw-van. The voltage and current readings for the PV array, the battery and the load were recorded. A GPS tracking system was used to estimate the average speed of the rickshaw-van and the distance travelled. The test was carried out until the battery was

discharged down to 50% SOC and all the tests were carried out at full load condition. The field test was carried out in four steps:

- Without any PV array assistance and torque sensor pedal
- With PV array assistance only
- With torque sensor pedal and no PV array assistance
- With torque sensor pedal and PV array assistance.

VI. ANALYSIS OF FIELD TEST RESULTS

A. Without PV array assistance and Torque Sensor Pedal.

Initially field test was carried out to determine the energy consumed by the battery without PV support and Torque Sensor pedal. The battery voltage and current drawn by the load from the battery were recorded. Fig 5 shows the power supplied by the battery to the load. It was seen that the rickshaw-van travelled 12.455km in 1.24 hrs with the battery discharged down to 50% SOC. At maximum torque, maximum current and power drawn by the load was 33A and 1700W respectively. The energy consumed by the load is calculated from the area under the curve of battery power vs. time graph shown in Fig 5. It can be seen that the average power supplied by the battery is 483W and the energy supplied by the battery to load is 2.184MJ.

B. With PV array assistance only

The test was carried out during the peak sun hours from 12pm to 2pm. Panel voltage, battery voltage, current drawn by the load from battery and current supplied by the panel were record. Fig 6 shows the panel, battery and load power curves. It was found that the rickshaw can travel 17.2 km in 1.69hrs with the battery discharged down to 50% SOC. From the area

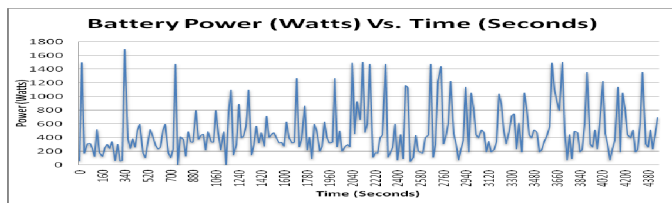


Fig. 5. Power supplied by the battery

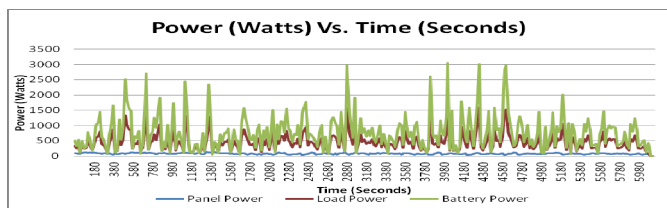


Fig. 6. Panel power, motor power, battery power curves

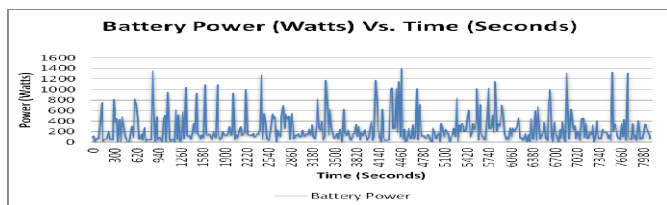


Fig. 7. Power supplied by the battery

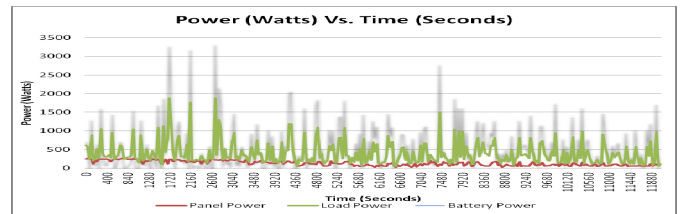


Fig. 8. Panel power, motor power, battery power curves

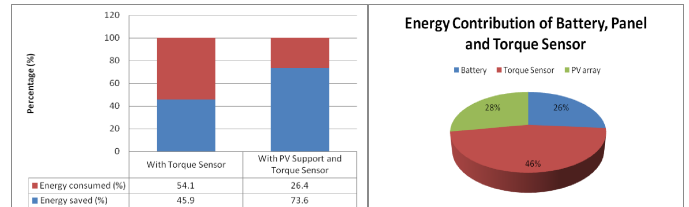


Fig. 9. Comparison between energy saved with Torque sensor and with both Torque Sensor and PV array and energy contribution

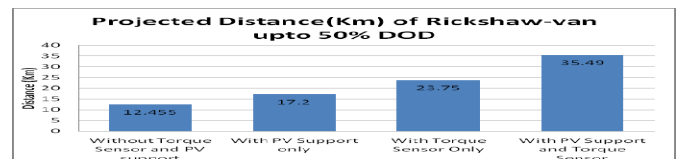


Fig. 10. Projected distance of the rickshaw-van up to 50% DOD of the battery



Fig. 11. Solar battery charging station on the rooftop of BRAC University Campus

under the power curves in Fig 6, it can be seen that the energy supplied by the panel is 0.49MJ at an average power supply of 80W, energy supplied by the battery is 1.94MJ at an average power supply of 310W and energy consumed by the load is 2.42MJ at an average load power of 390W.

C. With Torque Sensor Pedal and no PV array assistance

The field test was carried out by eliminating the use of throttle from the system and by using torque sensor pedal. Battery voltage and current drawn from the battery were recorded. The torque sensor input voltage to the controller unit and output voltage from the torque sensor pedal to the torque adjuster were recorded as well. Fig 7 shows the power supplied by the battery at this condition. It was found that the rickshaw-van travelled 23.75km in 2.25hr with the battery discharged down to 50% SOC. The energy consumed by the load is calculated from the area under the battery power curve in Fig 7. It is seen that the energy consumed by the load during test without torque sensor and PV support is 2.183MJ in 1.24hrs whereas the energy consumed by the rickshaw –van with torque sensor is only 1.18MJ in 1.24hrs. The system consumed 54.05% of the throttle controlled design and it saved 45.95% energy.

D. With PV array assistance and Torque Sensor Pedal.

Field test was carried out during the peak sun hours from 12pm to 4pm. Panel voltage, battery voltage, current supplied by the PV panel and current drawn by the load were recorded.

The torque sensor input voltage to the controller unit and output voltage from the torque sensor pedal to the torque adjuster were recorded. Fig 8 shows the battery, panel and load power curves. It was found that the distance travelled by the rickshaw-van was 35.49km in 3.36hrs with the battery discharged down to 50% SOC. The energy consumed by the load is calculated from the area under the load power curve. Energy supplied by the PV array is calculated from the area under the panel power curve and battery energy consumption is calculated from battery power curve. It is calculated that the PV array supplied 1.563MJ energy, battery supplied 1.963MJ energy and the load consumed 3.51MJ energy. The energy consumed by the load during the test without torque sensor and PV support is 2.183MJ in 1.24hrs whereas the energy consumed by the rickshaw –van up to that time interval is only 575.67KJ. From this we can say that with torque sensor and PV support, the system consumes 26.3% energy of the throttle controlled design and saves 73.6%.

VII. IMPACT OF PV ARRAY AND TORQUE SENSOR-COMPARITIVE STUDY

A. Energy Saved

From the field test data, it was determined that the rickshaw-van can travel 23.03km extra with the both PV support and torque sensor pedal. With torque sensor and PV support, the system consumes 26.3% energy of the throttle controlled design and saves 73.6% energy. With torque sensor individually, 45.9% energy is saved. Thus, we can say that the PV panel provides 27.7% energy. Fig 9 shows the energy contribution of battery, panel and torque sensor.

B. Projection of distance coverage for the rickshaw van

For the batteries used in the rickshaw-van, 50% DOD has been considered for the longevity of the battery-bank. Fully discharging the batteries causes the battery life to decrease and thus adding up extra cost for replacing batteries. It was found that the rickshaw-can travel 12.455km without PV support and torque sensor pedal, 16.18km with PV support only, 23.75km with torque sensor pedal only and 35.49km with both PV support and torque sensor pedal. The projected distance coverage at each field tests are represented in Fig 10.

VIII. FUTURE WORK

The basic goal of the whole project was to transform the existing traditional rickshaw-van to the motorized modern tri-wheeler and analyze its performance. There are also scopes of future research to develop more power assisted tri-wheeler in the form of ambulance and cargo hauler. To make it commercially feasible it is important to add extra features. The system is sometimes unstable on uneven road surfaces. To avoid this, suspension can be added. Special features such as special compartment for keeping spare wheels, toolbox, windshield for driver's safety, lock system, casing for charge controllers can be added to make it much more user friendly. An automatic reverse gear system should be added in future.

Easy battery swapping techniques should be found for the ease of the driver for swapping batteries from solar battery charging station to the van and vice versa.

IX. CONCLUSION

A large number of people will be benefited with the commercialization of such vehicle nationwide. This will improve the lifestyle and income of the rickshaw pullers. The slow speed non-motorized traditional rickshaw-van which is often blamed to be the reason for physical exhaustion and stress on the rickshaw-van pullers can be replaced by this modernized tri-wheeler [9]. The cost of the overall system is comparatively higher but faster payback overcomes this problem. Through the implementation of these technologies, lifestyles of the millions of people those who are directly or indirectly dependent upon rickshaws or rickshaw-vans can be improved tremendously. The designed tri-wheeler with PV support, torque sensor pedal causes zero pollution to the environment and thus it is a green project. It does not consume power from national grid for charging batteries. Instead, it uses renewable energy that is inexhaustible and thus the whole system is a complete off grid solution that can be sustainable.

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Design and Performance Analysis of a Single Phase Rectification System by using Cuk converter and EMI filter.

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Abstract – The demand of DC equipments are increasing day by day due to rapid development of power electronics devices and green energy. But most of the power generating systems engenders power in the form of AC. Consequently an efficient rectification system is strongly recommended to convert AC into DC. A close study of usual AC to DC conversion system arises a number of problems such as high THD, low power factor, low voltage regulation etc. This paper offers an analysis of rectification systems. In every system converters are fed through the rectifier and output has taken through a filter. Their subsequent effect on the voltage, power and THD waveforms have observed. In full-bridge rectifier, the amount of THD is so high. Here a conventional boost PFC converter is analyzed with a regular filter (LC) in which the power factor is not better and amount of THD is so high. So the boost converter is replaced by cuk converter and responses are analyzed for various regular filters in which amounts of THD were not in permissible limit. Finally in favor of a rectification system this paper presents an efficient rectification system amalgamated of cuk converter and EMI filter. In case of cuk converter with EMI filter, it is seen that the amount of THD is greatly low and power factor is better than the previous traditional boost converter. All the circuits are simulated in MATLAB Simulink environment and the desired outputs have obtained.

Keywords—Rectification system; EMI Filter; Power Factor; THD; Boost Converter; Cuk Converter.

I. Introduction

The rectification system has covered a great place in power electronics. Some green energy systems produce power in ac form with unenvied frequency. This frequency can be matched with loads or grid by rectification system. To vary the voltage after rectifier various converter are used. The rectifier draws a current from the line which is perverted resulting in a high Total Harmonic Distortion (THD) and low Power Factor (PF). It also reduces the efficiency of the converter added with the rectifier. Low power factor prolong the losses in circuit doubly and high THD abates the life time of the components of the circuit. So various filters are used to reduce THD and improve power factor [1].

The conventional boost PFC converter can play an important role in alleviating THD and improving power factor. But, it is observed that this PFC converter cannot reduce THD significantly.

For this reason the boost converter is replaced by a cuk converter and responses are observed for various regular filters. But desired responses are not found in every filter topology. Finally, proposed cuk PFC converter is designed.

This configuration consists of a rectifier; a cuk converter and an EMI filter which are added in parallel with each other. The THD of the converter become low and power factor become nearer to unity. The problems which were arising in the boost converter are eliminated by this cuk converter [2].

II. CONVENTIONAL BOOST PFC CONVERTER

At present the rectification systems are designed by cascading a rectifier, a dc-dc converter and regular filters (RC, LC, π and T filter etc.). In regular filters, it is seemed that the LC filter is better than the other filter topology. So, a conventional boost PFC converter circuit with LC filter is discussed here.

The output voltage equation is [3]: $V_o = V_s / (1 - D) \dots \dots \dots (1)$

A. Simulation of Conventional Boost PFC Converter Circuit

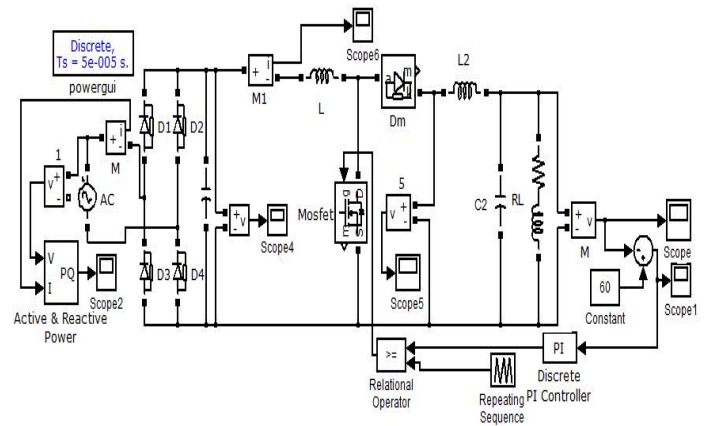


Fig. 1: Boost converter circuit with LC filter.

B. Output Curves of Conventional Boost Converter

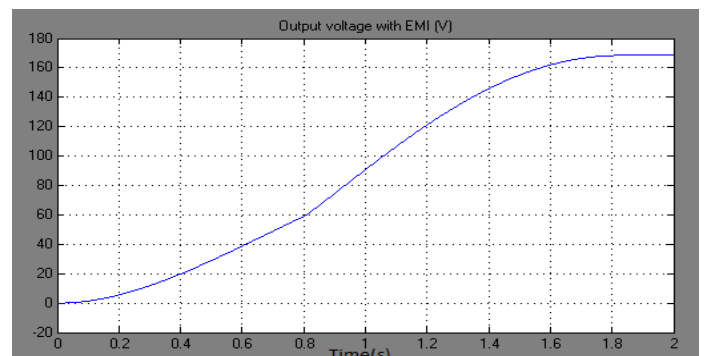


Fig. 2: Output voltage curve for LC load.

From fig.2 it is observed that the output voltage become stable after few seconds that is transient period is high.

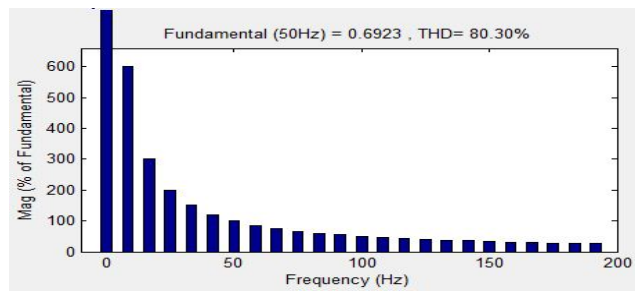


Fig. 3: THD analysis curve Boost PFC Converter.

The fig.3 represents the amount of THD is 80.30 % which is very high.

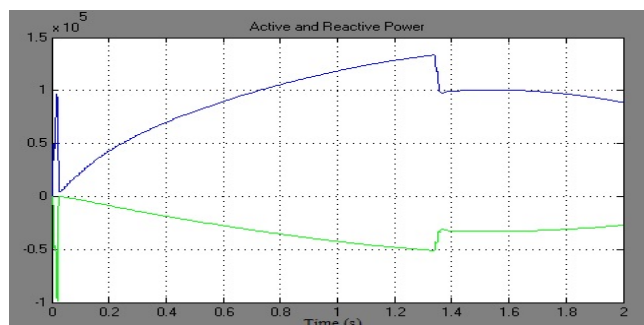


Fig.4: Active power (W) and Reactive power (Var) curve.

Fig.4 represents the active and reactive power curve of conventional boost PFC converter. From the curve it is observed that the power which is drawn by the load is not stable. At transient period the power rises to extra high level which is not desired for any stable system.

From the above analytical study of the conventional boost PFC converter, it can be said that the instability of power and high THD cannot bear good result for better power factor.

As well as the input current through the inductor is smooth but the output current that flows through the capacitor load combination is pulsating. So it requires a very large capacitor filter to make it smooth. In the case of Cuk converter, the input and output current through the inductor is smooth, so there is no need of a large capacitor filter unlike boost converter. It can also step up and down the output voltage. So the boost converter can be replaced by a cuk converter [4].

III. PROPOSED PFC CONVERTER WITH RESPONSE

The output voltage equation of the cuk converter is:

$$V_o = -V_i D / (1 - D) \dots \dots \dots (2)$$

A. Simulation of Cuk PFC Converter Circuit

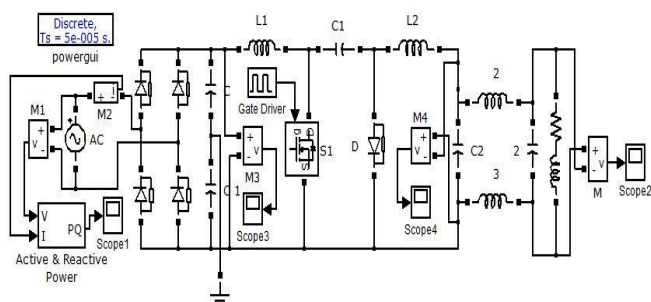


Fig. 5: Cuk PFC converter with filters.

Here the cuk converter is cascaded with different types of filters such as RC, RL, Pi and Tee.

Output voltage responses of cuk PFC converter with RC, LC, Pi and Tee filters are shown:

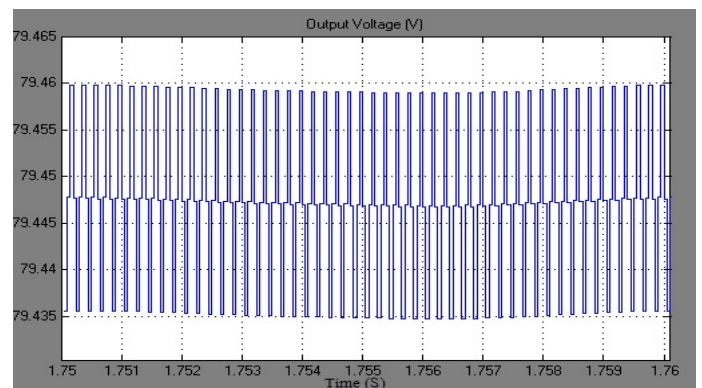


Fig. 6: Output voltage curve for RC filter.

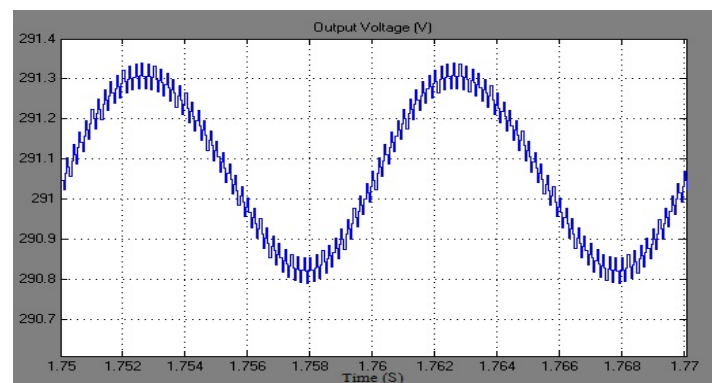


Fig. 7: Output voltage curve for LC filter.

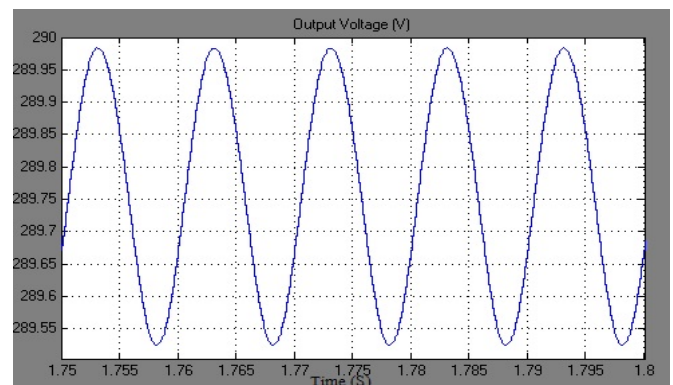


Fig. 8: Output voltage curve for π filter.

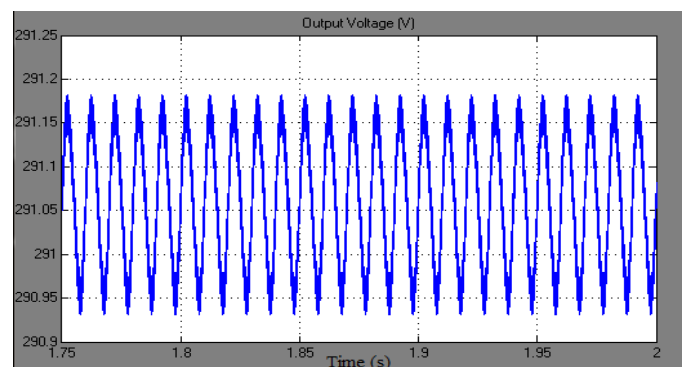


Fig. 9: Output voltage curve for T filter.

From the above curves (fig.6-fig.9) of the output voltages it is seen that every curves is highly distorted and contains more ripples.

The THD analysis curves for various regular filters are shown below:

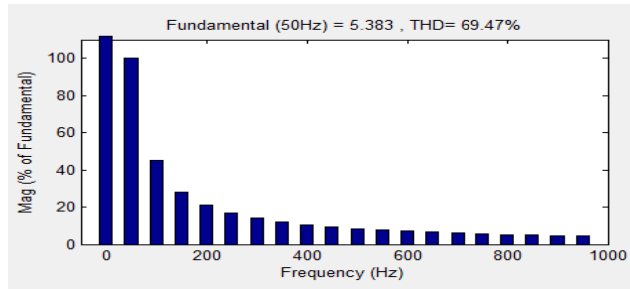


Fig. 10: THD analysis curve for RC filter.

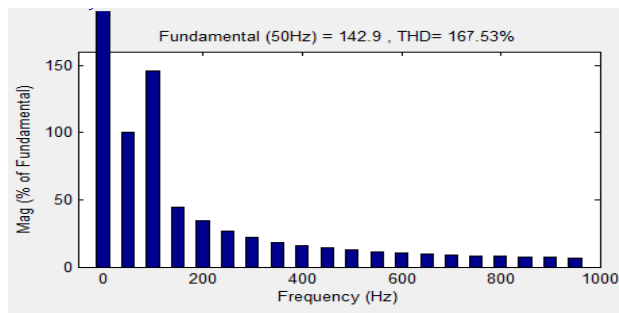


Fig. 11: THD analysis curve for LC filter.

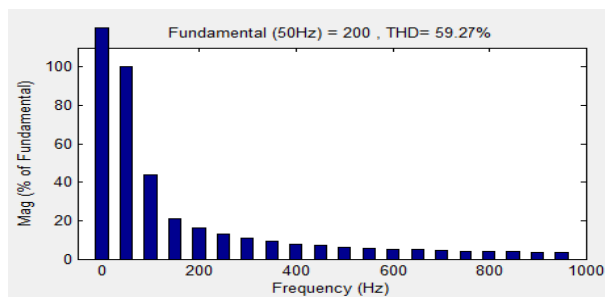


Fig. 12: THD analysis curve for π filter.

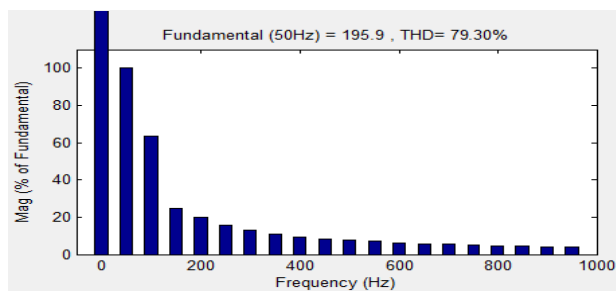


Fig. 13: THD analysis curve for T filter.

It is seen (fig.10-fig.13) that THD of the cuk converter with various filter topologies are so high. As well as the output voltages for various topologies are highly distorted. For these reasons a cuk converter with EMI filter topology is designed.

B. EMI Filter

EMI means Electromagnetic interference. An EMI filter is a passive electronic device used to suppress conducted interference present on any power or signal line [5].

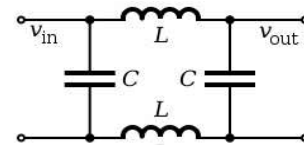


Fig. 14: A typical EMI filter configuration.

EMI Filter Design Procedure: 1) Measure the base line DM, (V_{DM} , measured) dB and CM (V_{CM} measured)dB EMI noise. 2) Determine CM, (V_{req} , CM)dB, and DM, (V_{req} , DM)dB, attenuation requirement [6]. 3) Determine the minimum corner frequency of second order L-C filter for CM and DM according to the attenuation requirement for all frequency spectrums [7]. 4) Calculate the inductor and capacitor component values of the conducted EMI filter from the corner frequencies found on the previous step [8].

$$f_{R,CM} = \frac{1}{2\pi\sqrt{L_{CM}C_{CM}}} \dots \dots \dots (3)$$

For $f_{R,DM}$

$$f_{R,DM} = \frac{1}{2\pi\sqrt{L_{DM}C_{DM}}} \dots \dots \dots (4)$$

C. Simulation of Proposed Cuk PFC Converter with EMI filter:

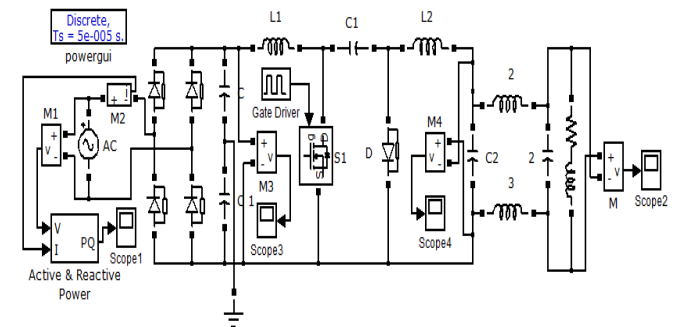


Fig. 15: Cuk PFC converter with EMI filter topology.

The output voltage and THD responses of Proposed Cuk PFC Converter with EMI filter are shown below:

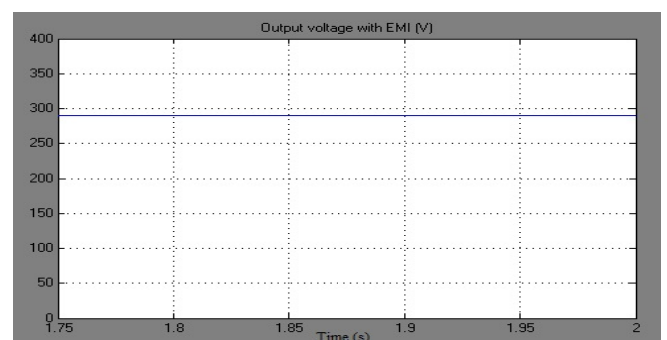


Fig. 16: Output voltage curve with EMI filter.

From fig.16 it is seen that the output voltage with EMI filter does not contain any ripple and it get approximately a smooth DC output voltage.

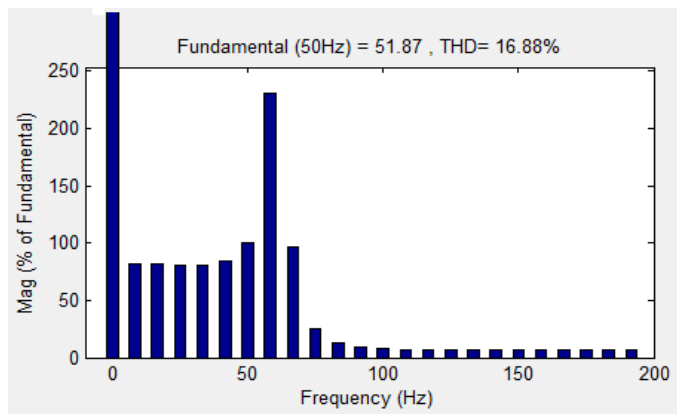


Fig. 17:THD curve analysis with EMI filter.

From the THD curve (fig.17) analysis it is seen that the EMI filter reduces the amount of THD significantly.

TABLE I: THD COMPARISON FOR VARIOUS FILTER TOPOLOGIES WITH EMI TOPOLOGY

No.	Cuk converter with various filter topology	Amount of THD (%)	THD difference with EMI filter (%)
1	Cuk converter with EMI filter	16.88	
2	Cuk converter with RC filter	69.47	49.17
3	Cuk converter with LC filter	167.53	147.23
4	Cuk converter with π filter	59.27	38.97
5	Cuk converter with T filter	79.30	59.00

From the table-I it is clear that the THD of Cuk converter with EMI filter is better than the other topology and amount of THD is 16.88%. So, the cuk converter with EMI topology is suitable for any stable system.

Simulation Curve of Power for Every Load:

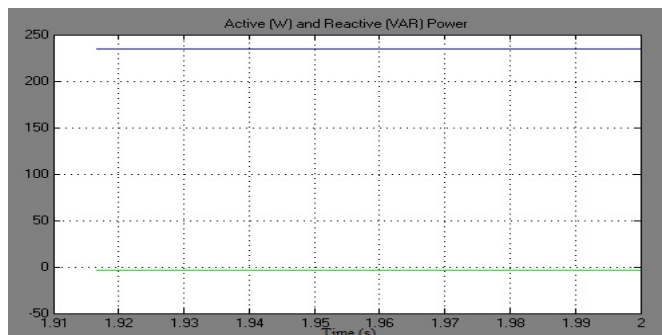


Fig. 18: Ac power (W) and Reactive power (Var) curve.

The cuk converter gives a higher constant power (fig. 18). So, the proposed cuk converter is more efficient than the designed boost converter.

IV. RESULTS AND COMPARISON

From the power triangle [9], we get that

$$VA^2 = W^2 + VAR^2 \text{ or } VA = \sqrt{W^2 + VAR^2} \dots\dots\dots(5)$$

Now, Power factor = Real Power/Apparent power

$$\cos = W/VA \dots\dots\dots(6)$$

From the above equation, the value of power factor can be determined.

TABLE II: POWER, POWER FACTOR AND THD COMPARISON BETWEEN CONVENTIONAL BOOST AND PROPOSED CUK CONVERTER

	Boost Converter	Cuk Converter
Active Power	88.99 kW	234.3 W
Reactive power	-2.722 kVAR	-3.97VAR
Apparent power	89.03 kVA	234.334 VA
Power Factor	99.95%	99.98%
THD	80.30 %	16.88%

V. CONCLUSION

The main objective throughout the paper is to improve the input Power Factor with simultaneous reduction of input current harmonics of converter and thus reducing THD. The conventional boost PFC converter cannot offer good power factor and low THD. So, the cuk PFC converter with EMI filter is designed which offers better PF than that of the previous one along with impressively low THD.

FUTURE WORK

Presently the converter has designed for low voltage. In future the converter will be designed for high voltage. Here, the THD of proposed cuk converter is 16.88%. The research will be continued to find the THD approximately 1.5% which will be more efficient.

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Transformer less, Lower THD and Highly Efficient Inverter System

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Abstract - Total Harmonic Distortion (THD) of a voltage or current signal causes power system losses, apparatus overheating and increases the power bills. This paper proposes the design and implementation of a transformer less single-phase inverter system that produces an output voltage of desired magnitude and frequency at a very lower THD. In 'general inverter system', a lower DC voltage is inverted into lower AC voltage which is further boost up to distribution level through a step up transformer. This system contains higher THD. But, the 'proposed inverter system' efficiently converts a 12V DC into 220V AC at a lower THD. By analysis on a real system, both topologies are simulated. Here PSIM simulation software is used. From the comparison of the simulation output, lower THD is demonstrated for 'proposed inverter system' than 'general inverter system'.

Keywords- Total Harmonic Distortion (THD), Inverter, Booster etc.

I. INTRODUCTION

Inverters are widely used today mainly in uninterruptible power supply (UPS) systems or in solar power transmission system. The use of inverter is increasing day by day. A new challenge has established on efficiently conversion of 12V dc to 220V AC at a lower THD with higher efficiency. [1]

The total harmonic distortion (THD) of a signal is a measurement of the harmonic distortion present and is defined as the ratio of the sum of the powers of all harmonic components to the power of the fundamental. It provides an indication of the degree to which a voltage or current signal is distorted. [2]

In 'general inverter system' a lower dc input voltage is inverted into lower ac voltage which is fed to a step up transformer.[3] Now, if this lower ac input voltage contains even a very lower harmonics, the step up transformer boost up this harmonics along with its stepped up output voltage which depends upon the turn ratio 'k'. This is why THD is higher in 'general inverter system', which in turn reduces the efficiency. [4]

The 'proposed inverter system' is able to reduce these problems, where a low voltage DC power is firstly boost up to high voltage DC using DC to DC boost converter. Inverter converts high DC voltage into high AC Voltage using Pulse width modulation. The output of inverter is passed through a low pass filter which gives 220V 50Hz AC output voltage at

Lower Total Harmonic Distortion (THD). Now, this output voltage of lower THD is given to the load.

II. GENERAL TOPOLOGY (WITH TRANSFORMER)

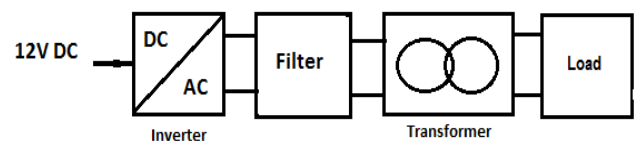


Fig 1: Block diagram of 'general inverter system'.

A. Inverter

An inverter is a circuit which converts a DC power into an AC power at desired output voltage and frequency. The AC output voltage could be fixed or variable voltage and frequency. This conversion can be achieved either by controlled turn on and turn off devices. [5]

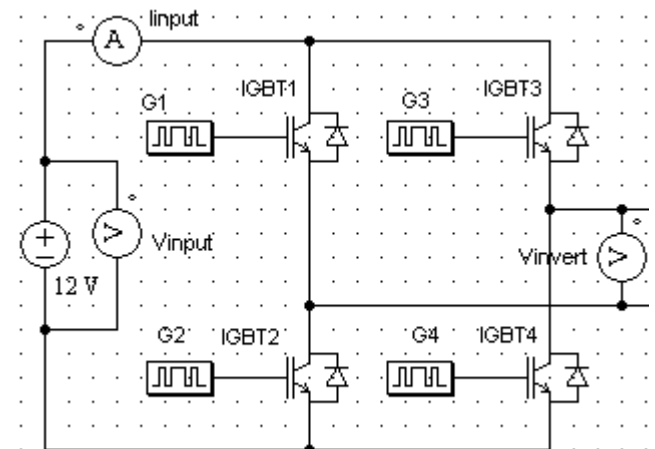


Fig 2: Inverter (DC to AC converter).

The circuit diagram consists of four distinct IGBTs such that they are connected as bridge circuit. The input to the circuit is 12V DC. The IGBTs are triggered accordingly such that the AC output voltage is obtained at the output. The IGBT1 and IGBT4 are turned on by triggering the gates of them. During this time the current starts from the supply positive, IGBT1, IGBT4, load and to the negative of the supply.

During the next cycle the IGBT2 and IGBT3 are turned on by giving trigger pulse to the gate of those IGBTs. During this period the input voltage is applied at the output but in the

negative direction. The current conduction starts from the supply, IGBT2, IGBT3, load and to the negative of the supply. [6]

B. Transformer

A transformer is an electrical device that transforms electrical energy from one circuit to other circuit at same frequency. There are no direct electrical connections, Energy transfers magnetically.

$$\text{In case of transformer, } \frac{E_2}{E_1} = \frac{N_2}{N_1} = k$$

Where E_1 , E_2 are the primary & secondary voltage; N_1 , N_2 are the number of turns of primary & secondary windings, And 'k' is called the turns ratio. [7]

Now varying the value of 'k', the transformer can be either step up or step down. Here, the transformer is used to boost the Output voltage of inverter. [8]

Here, Turns ratio, $k = 12$

Transformer input RMS Voltage, $E_1 = 18.24$ V

Transformer output RMS Voltage, $E_2 = 218.96$ V

C. Filter

A Filter is used to remove the undesired noise from desired signal. The high frequency inverter output signal contains fluctuations, distortion and harmonics. [9]

Here, a low pass filter is used. A low pass filter passes low frequency signals but attenuates signals with frequency higher than the cut off frequency. Hence, the low pass filter eliminates the harmonics of the inverter output voltage. The output frequency of filter should be in line with application ratings. [10]

Here, $L = 0.01$ H $C = 1000$ uF

Filter cutoff frequency,

$$f = \frac{1}{2\pi\sqrt{LC}} \\ = 50 \text{ HZ}$$

III. PROPOSED TOPOLOGY (TRANSFORMER LESS INVERTER)

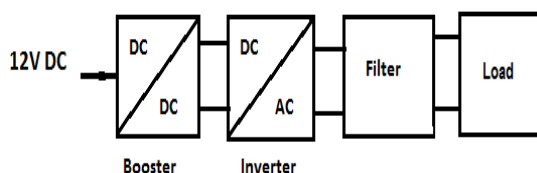


Fig 3: Block diagram of 'proposed inverter system'.

Figure-3 shows a block diagram of proposed inverter where a boost converter has introduced as a replacement of transformer to perform step up the lower voltage level.

A. Booster

A Booster or boost converter can step up a lower voltage level to a higher voltage level with a very high efficiency. The

average output voltage of a boost converter can be found from the equation,

$$V_o = \frac{V_s}{1-d}$$

Where V_o & V_s are average output voltage and supply voltage respectively, and 'd' is duty ratio of IGBT. [11]

DC component of the output V_o is controlled by varying the duty ratio d,

$$d = \frac{t}{T}$$

Where, t =on period, T = total period (On period + off period). [12]

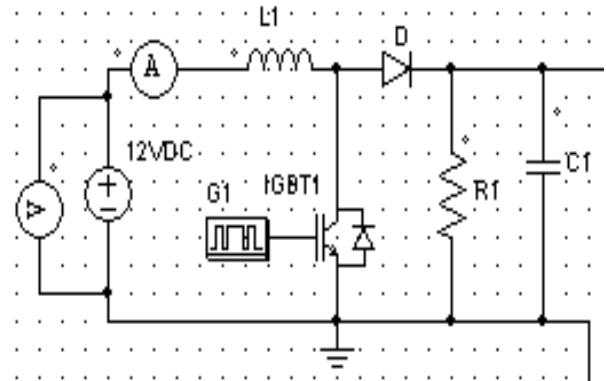


Fig 4: DC to DC boost converter.

When the switch (IGBT) is closed for on period, the inductor current rises and energy is stored in the inductor. If switch is open for off period the energy stored in the inductor is transferred to the load through the load. [13]

The value of inductor and capacitor can be found from,

$$L = \frac{t \cdot V_{in}}{\Delta I} \quad C = \frac{t \cdot i_o}{\Delta V_o}$$

Where, L = Inductance

ΔI = inductor ripple current,

t = on time of IGBT1

C = capacitance,

ΔV_o = output ripple voltage of booster,

And t = on time of IGBT1 [14]

Here,

duty ratio $d = 0.85$, $L = 10$ mH,

$R = 100$ ohm and $C = 1000$ uF.

Input voltage level of the booster = 12V

Output voltage level of the booster = 234V.

The booster output is supplied to the inverter as shown in figure-6.a. This inverter switch is fired with pulses from a Pulse width generator enables to alternatively switch two of its diodes at every single instant, thereby converting the direct current to an alternating current. Universal bridge consists of IGBTs which perform the inverter functions. The amplitudes

of both the modulating signal and the carrier signal is kept the same to maintain the modulation index $m_a=1$.

The inverter model consists of four IGBT's named individually as G1, G2, G3 and G4. This four-diode model bridge is fired pulses from a Pulse width generator. The pulses cause the alternate diodes G2 and G3 to turn ON while the diodes G1 and G4 remain OFF. In the next sequence, diodes G1 and G4 are turned ON while the diodes G2 and G3 remain OFF. This alternating switching of diodes converts the input direct current to an alternating current. The output voltage of an Inverter is a square waveform. This output can contain higher order harmonics which can create harmonic problem. So this inverter output is fed to LC filter of 50Hz to produce a pure sinusoidal output waveform.

IV. SIMULATION ANALYSIS

A. Simulation of 'general inverter system'

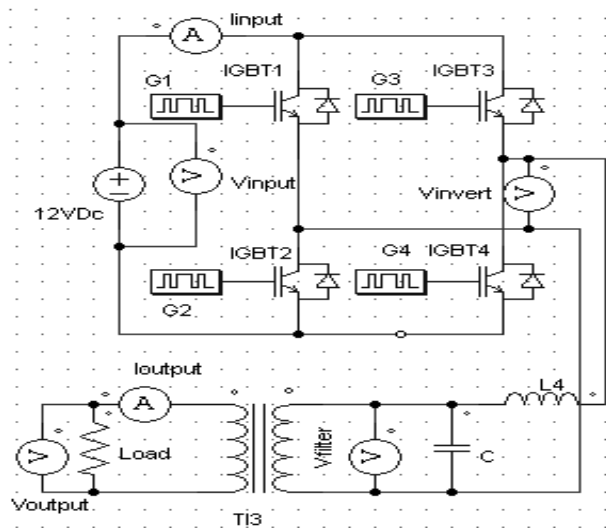


Fig 5.a: Circuit Diagram of 'general inverter system' output.

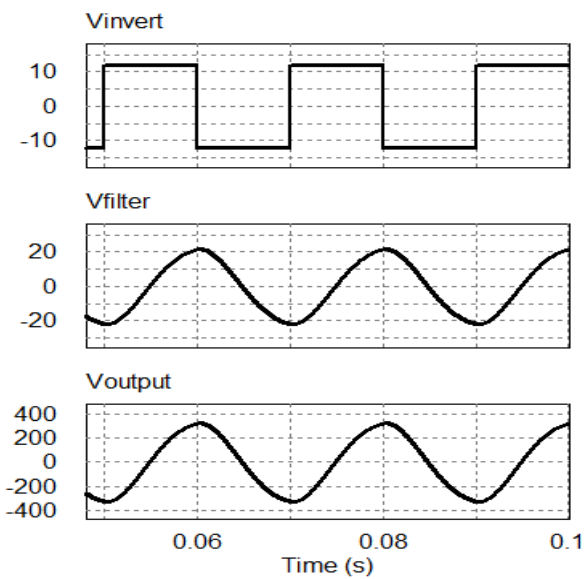


Fig 5.b: Inverter output voltage; Filter output voltage and Transformer output voltage of general Invert system.

B. Simulation of 'proposed inverter system'

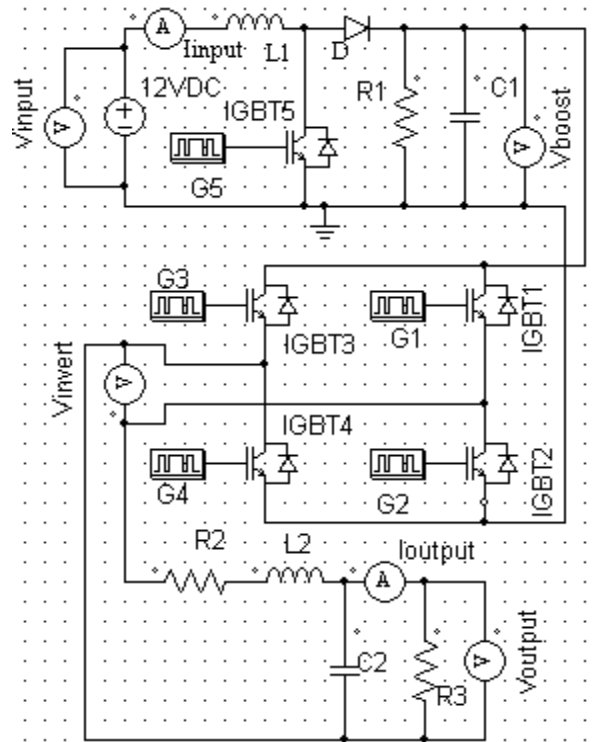


Fig 6.a: Circuit Diagram of 'proposed inverter system'.

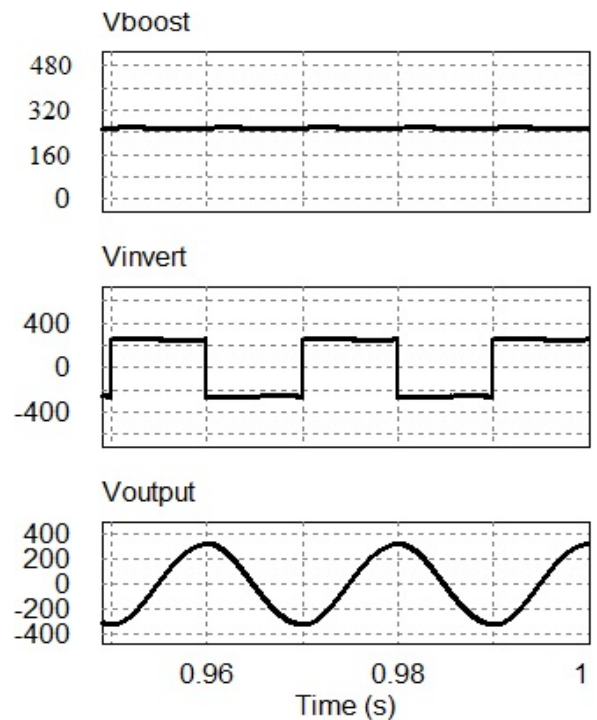


Fig 6.b: Booster output voltage, Inverter output voltage and Filter output Voltage of proposed inverter system.

Figure-5.a and 5.b shows the simulated circuit diagram of general inverter system and output of different stages respectively. Figure-6.a, 6.b shows the simulated circuit diagram of proposed inverter system and its output waveforms.

V. RESULT ANALYSIS

Table -01: Observation of both topologies

Topics	General inverter system	Proposed inverter system
Input DC Voltage	12V	12V
Output RMS Voltage	221V	221V
Output RMS current	0.00221 A	0.002219 A
Form Factor	18.41	18.41
Input Power	0.488 W	0.49282 W
Output Power	0.5105 W	0.5105 W
Overall Efficiency	95.59%	96.53%
THD	0.0295526182	0.01088456

VI. CONCLUSION

This paper represents the simulation of an efficient inverter system and also a comparison of 'proposed inverter system' to the 'general inverter system'. The proposed system eliminates the use of transformer, a booster is used to boost the DC input voltage, and then boosted DC is inverted to AC distribution level. From simulation output and comparisons, the lower THD and Higher efficiency was obtained. From PSIM simulation, the obtained THD for 'general inverter system' is 0.0295526182 while the obtained THD for 'proposed inverter system' is 0.01088456. The difference is 0.018668058. So, it is found that the 'proposed inverter system' reduces **63.17%** of THD of that of 'general inverter system'. For the simulation of both topologies, the parameters of circuit diagrams were kept same to understand the comparisons easily.

As a part of future work, feasibility will be checked and cost analysis of both systems will be done. Finally the system will be implemented practically with photovoltaic arrangement.

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The Next Generation of Irrigation

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Abstract— In the mid-20th century, the advent of diesel and electric motors led to systems that could pump groundwater out of major aquifers faster than drainage basins could refill them hence revolutionizing the whole irrigation system and paving a path for modern technology to play its part in it. In this paper we propose an irrigation system that comes in par with the development in technology. Incorporating a microcontroller, a soil hygrometer and a weather sensor, we plan to develop a relatively low cost irrigation system that would be completely self-sustaining and require almost no human intervention.

Keywords—automated system; weather prediction; water wastage prevention; watering crops; smart irrigation

I. INTRODUCTION

Huge amount of fresh water is being used, every day, throughout the world, for irrigation purposes; the agricultural sector consumes about 70% of the planet's accessible freshwater [1]. Consequently, a large amount of that water is wasted over-watering the crops and in transportation. Currently, according to the UN Food and Agriculture Organization (FAO), an astonishing 60 percent of the water diverted or pumped for irrigation is wasted due to irregular irrigation methods. 2,500 trillion litres of fresh water is used throughout the world each year, 1,500 trillion litres of that water is wasted [1]. We are squandering away more than half of what is necessary. According to the United Nations, two-third of the world's population is estimated to face water scarcity by 2025. In a large scale, irrigation can become a tedious task requiring a lot of time or manpower, in most cases both. The primary objective of this project is to minimize the overhead cost, reduce time consumption and money and most importantly, prevent the inefficient use of water in irrigation and minimize waste.

II. EXISTING SYSTEMS

A few automated irrigation systems already exist at present times, all of which function in more or less the same way. The water delivery system is connected to a central processor and a switch to control the system. The problem with this system is it is still dependent on a person's cognitive ability to judge when and if to turn on the system. It may be less tedious and require fewer people to monitor, but it still lacks one essential element. It depends on the reasoning and judgment of one person to determine if the plants or crops have enough water or not. A way to work around this problem is to set the system to go off on certain times of the day using a timing circuit thereby eliminating the need for human intervention, one system in particular, developed by Amitabh Saran is quite

intuitive but this system is also flawed since it does not take into account the future prospects of rainfall [2]. There is no point in wasting water and electricity to irrigate a huge mass of land when nature itself is going to take care of it.

III. GENERAL IDEA BEHIND THE PROPOSED WORK

For the system to meet our criteria, we need to integrate a soil hygrometer and a weather sensor (namely barometric sensor). The sensors will collect data periodically and feed it to the central computer, where the data is processed and the decision is made “when” and “if” to turn the watering system on. Given below is a flow chart of the system's working principle.

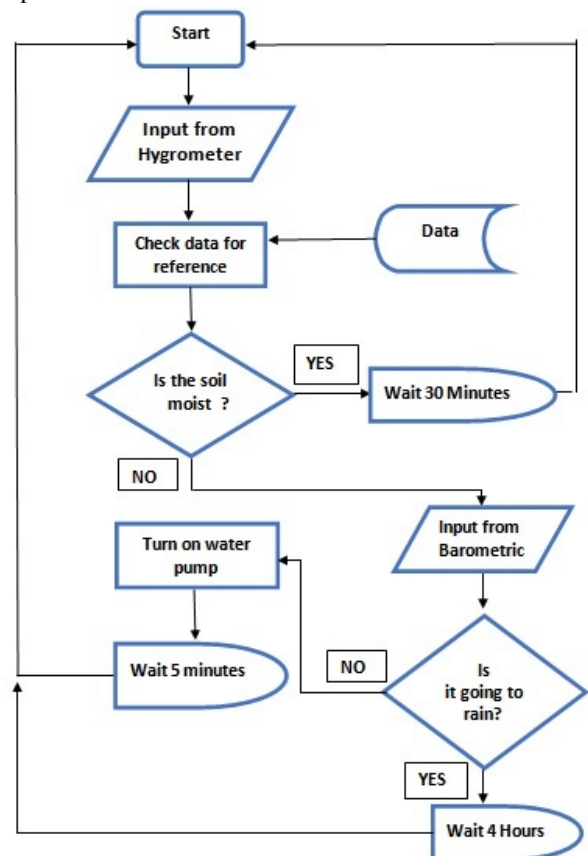


Fig 1: Overview of the System's working principle

The soil moisture sensor would monitor the present state of the soil and the weather sensor to determine the prospects of future rainfall. Taking data feed from the sensors, the central computer will be programmed to turn the switch on if the levels fall below a certain threshold. For instance; depending on the texture of the soil, if the soil moisture level fall below

1ml/cubic inch, the central computer would request an immediate feed from the barometric sensor, checking to see whether it is going to rain in the next four hours or so, if the system has adequate evidence that points towards rainfall in the recent future, the system will not trigger the switching mechanism. However, after checking the determinant factor, if the computer gets no evidence from the sensor about any rainfall, the system would be turned on and automatic Irrigation would begin.

IV. SETUP

For our system we choose the "YL-69" soil hygrometer and "BMP 180" Barometric pressure and temperature sensor as our sensory input devices and an Arduino Uno R3 board for our control purposes. We also require a few other components, for our test purposes we used a TIP 122 NPN Epitaxial Darlington transistor for the "automatic switch". A 1N4007 Diode which prevents excess voltage from the water pump to flow back into the delicate circuitry and damage it [3]. A detailed description of the essential components and their working methodology is given below.

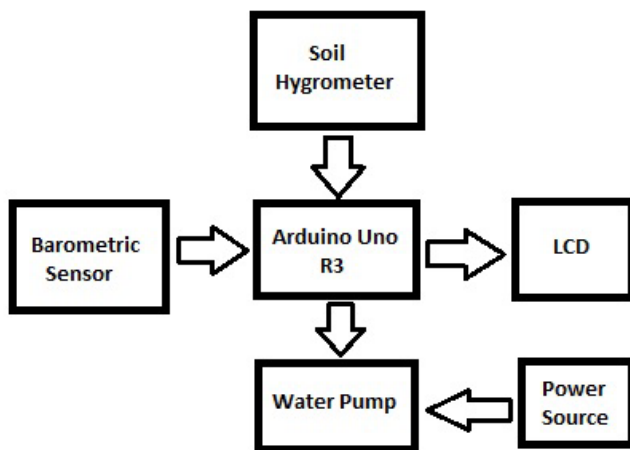


Fig. 2. A simple Block diagram of all the components

A. Soil moisture sensors

It measures the water content in soil. The soil moisture probe is made up of multiple soil moisture sensors. These sensors measure some other property such as electrical resistance; dielectric constant etc. the conversion circuit provides the measurement as a voltage output. The use of a relatively thin dielectric coating on each of the electrodes increases the capacitance and reduces the soil resistance as a factor in the soil moisture measurement [4]. Based on the shape of the sensor electrodes and dielectric coating, a linear relationship is developed between capacitance and the electrode length in contact with soil particles containing moisture, which increases the accuracy of soil moisture measurement at soil depths of interest [5]. Measuring soil moisture is important for agricultural applications to help farmers manage their irrigation systems more efficiently. Knowing the exact soil moisture conditions on their fields, not only are farmers able to generally use less water to grow a crop, they are also able to increase yields and the quality of the

crop by improved management of soil moisture during critical plant growth stages.

B. Darlington Transistor

TIP 122NPN Epitaxial Darlington Transistor is used as a medium power linear switching application. It is ideally suited for our experiment purpose since it can handle up to 100V. Here, we are using the transistor as in electronic switch for the water pump. The water pump has been hooked up with a 12V DC external power source so as not to draw any power from the Arduino and cause any interference with the other sensors.

C. Barometric Sensor

It is used to measure atmospheric pressure and determine the prospects of future change in weather by measuring subtle changes in the atmospheric pressure. This is the most essential component for our system. Most electronic barometers use some sort of strain gauge transducer to directly convert pressure to voltage and true weather bureau performance can be difficult to achieve with these pressure sensors due to their significant sensitivity to temperature changes. Using barometric pressure and the pressure tendency (the change of pressure over time) has been used in weather forecasting since the late 19th century [6]. Reasonably accurate short-term forecasts can be made from the rise or fall in the localized atmospheric pressure. If the pressure drop is rapid, a low pressure system is approaching, and there is a greater chance of rain. Rapid pressure rises are associated with improving weather conditions. Since we will only be monitoring the weather pattern, the effects of altitude has to be removed. The barometric sensor will provide results under the assumption that it is working at sea-level.

V. RESULTS AND DISCUSSION

After Setting up the soil hygrometer we did a test run. Given below is a picture of the setup:



Fig 3: Setup of the Soil moisture Sensor

For visual aid, we have added 3 LED of distinct coloration. Green LED indicates that the soil moisture content is low and

this would trigger the pump motor to turn on. The yellow LED shows that the water content of the soil is at an optimum level. This is where the system turns off the water pump. Red LED indicates that the moisture content is too high, this could only happen due to excess rainfall and under those circumstances the system will stay turned off.

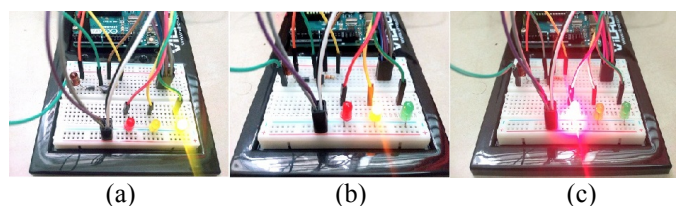


Fig 4: (a) Dry Soil (b) Moist Soil (c) Over Hydrated

Incorporating the barometric sensor proved to be a challenge. Since the BMP 180 is an IIC circuit, failing to initialize this sensor effectively shuts the whole system down [7]. This is the most delicate piece of equipment implemented in the whole system and thus special care has been taken to ensure its safety. The whole system has been encased in sealed box but special measures had to be taken to make sure the BMP 180 had room to “breathe” i.e. it had access to the atmospheric air. The sensor is extremely sensitive to temperature as well hence special vents had to be made on the protective casing to ensure proper ventilation is taking place. The enclosing also serves as a protective unit against strong air currents, which could easily muck up the readings and provide false data. Given below are the results obtained from the Barometric sensor:

```
provided altitude: 15 meters, 49 feet
temperature: 30.47 deg C, 86.85 deg F
absolute pressure: 1011.75 mb, 29.88 inHg
relative (sea-level) pressure: 1013.55 mb, 29.93 inHg
THE WEATHER IS GOOD NOW
```

```
1003
Sensor is not in the Soil or DISCONNECTED
```

```
provided altitude: 15 meters, 49 feet
temperature: 30.26 deg C, 86.47 deg F
absolute pressure: 1011.68 mb, 29.88 inHg
relative (sea-level) pressure: 1013.48 mb, 29.93 inHg
THE WEATHER IS GOOD NOW
```

```
1002
Sensor is not in the Soil or DISCONNECTED
```

```
provided altitude: 15 meters, 49 feet
temperature: 30.11 deg C, 86.20 deg F
absolute pressure: 1011.67 mb, 29.88 inHg
relative (sea-level) pressure: 1013.47 mb, 29.93 inHg
THE WEATHER IS GOOD NOW
```

Fig 5: Temperature and Pressure readings



Fig 6: LCD Readings

We could also incorporate a timing mechanism into the system which would allow for the water lines to run for even shorter amounts of time consequently saving more water.

The amount of water the system can deliver is completely independent of the structural components; thereby eliminating any need for further tweaking of the framework's internal mechanism. The soil texture and type varies from place to place so just calibrating the moisture sensor to the soil type should suffice. As a result of its modular nature, and depending on the water delivery system; the amount of landmass that can be incorporated by this irrigation system is virtually limitless. Acres and acres of land could be under the jurisdiction of only one of these system although it would be impertinent to incorporate such a large land-mass under one system since the moisture content changes from region to region. The water delivery system could be either rotary heads, sprinkler nozzles, drip irrigation or a combination of all three, this of course is left entirely up to the user's discretion.

VI. BILLS OF EQUIPMENT AND APPROXIMATE COST

One of the objective of this project was to develop a cost effective solution, to that that end we tried to keep the cost of implemented hardware to a bare minimum. Given below is a list of components and their corresponding price:

TABLE I. APPROXIMATE COST OF COMPONENTS

Name of Material	Cost in BDT
Arduino Uno R3	750
YL-69 Soil Hygrometer	500
BMP180 Barometric Pressure sensor	800

Custom made Enclosing	700
LCD	150
TIP 122 NPN Transistor	30
Colored LEDs	30
Wires, Resistors, Diode and other components	200
Total	3,160

VII. PRACTICAL IMPLEMENTATION

All the hardware used in the system has been chosen for the sole purpose of a small scale demonstration. For practical implementation however, we need to make some alterations in the system's internal mechanism.

The TIP 122 transistor cannot handle the 220V AC from the mains hence we need to use a 5V relay module instead. This is the only significant change we need to make the system sustainable for practical purpose. With some other minor tweaks, the remaining equipment are perfectly suitable for practical utilization.

The relatively low tech of the system makes it easier to maintain the system and keep it online. Unless there are any component failure, the only regular maintenance the system would require would be changing the soil hygrometer. This is because the dielectric coating on the probe corrodes over time and this interferes with the moisture content reading. A better quality hygrometer would last longer but that would increase the overall cost of the system so the user has to decide on the tradeoff.

VIII. CONCLUSIONS

We have been successfully in achieving all our goals that we had set for ourselves when we first started working on this project. Every technological development has certain rooms for development and our system of course is no exception. This system has been developed keeping the weather conditions of Bangladesh and neighboring countries in mind. To implement the system in harsher climates would require a few minor changes and some additional components.

Given the modular nature of the framework, we could make it function completely off the grid using solar power. If, for whatever reason, the automation fails to trigger the system on, we could imbue the system in the "internet of things" and trigger it manually, wirelessly, over the internet.

This automated system could vastly improve irrigation and provide a much needed boost to the agricultural sector. Not only can it save huge amounts of fresh water every day, it is also cost effective and a much more efficient means of irrigations.

ACKNOWLEDGEMENT

We wish to acknowledge Amitabh Saran for his brilliant work on the integration of nature and technology [2]. He has been the primary source of inspiration regarding our project work. We would also like to thank our instructor and

supervisor Dr. Hasan Uz Zaman for his patience and kind words of encouragement in dealing with us. Without him, our work would not have had any physical manifestation. Finally, we would also like to thank Tahsin Momin for his involvement in our project and providing us with his time and knowledge on various issues.

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Transformerless Dual Inverter Based Series Injected Shunt Active Power Filter

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Abstract— The following paper proposes a topology for reactive power compensation and harmonic reduction for medium voltage three phase three wire systems during unbalanced load conditions. Main three phase inverter produces harmonic compensation. Three single phase inverters connected in series with the three phase inverter for separate compensation of reactive power in three different phases are present. The three phase inverter is a hybrid active power filter which is an HFI (High Frequency Inverter) and uses Hysteresis control. The three single phase inverters are LFIs (Low Frequency Inverters) which also operate using sinusoidal triangular PWM control. The DC bus voltages of the three single phase inverters are common. The topology specifically caters to the need of real-time load which is largely unbalanced and requires both a reduction in harmonic levels as well as reactive power and negative sequence compensation simultaneously. Simulation results are also provided to validate the claims made in this paper regarding the hybrid topology and associated control strategy.

Keywords—Shunt active power filter; var compensator; unbalanced load; DC link voltage control; transformerless;

I. INTRODUCTION

The modern age trend to improve energy efficiency is to use ASDs (Adjustable Speed Drives). These are far economical in terms of energy consumption than usual industrial and residential load comprising of motors. However the cost utilities pay for promotion of ASDs is diminishing power quality as these ASDs employ rectifiers at the front end. These power electronic devices inject voltage and current harmonics. However high frequency harmonics are themselves mitigated by the transmission line as it is a low pass filter but extra initiative is required to remove lower order harmonics.[1].

Active filters have more robust performance as compared to passive filters as passive filter are easily detuned when supply frequency changes and easily fall prey to harmonic resonance with source and load. The reduction in harmonics in output voltages can be accomplished by using multi-level converter topologies [2]-[3] some of which are : three level converters with neutral point clamped, flying capacitor based technology. Five-level or seven-level converters find usage in

case of higher voltage rating applications. However, these configurations inherently suffer from imbalance in DC voltage and require truly involved PWM control measures to maintain DC bus voltage.

The use of Active power filters (series or shunt) provides better compensation for much wider frequency ranges albeit requiring higher rating devices for low and medium voltage applications. The advent of hybrid active power filters has facilitated the use of lower rated switches in series with the passive filter as reported by Akagi *et al.* [4] and other researchers[5]-[9]. This arrangement helps in operating the active part of the filter at conveniently lower voltages and currents. The passive filter which comprises of series capacitor and inductor is tuned to eliminate lower order harmonics while the low voltage series square wave inverter supplies reactive power at fundamental frequency. The reduced switch configuration consists of only two legs instead of three with the third phase connected to DC bus midpoint and hence is more economical.

One important aspect of these shunt active power filters is the controller design. Different types of controllers have been reported in literature including time based, frequency based and evolutionary methods based controllers. Linear control schemes are implemented using outer voltage loop and an inner current loop. Discontinuous PWM switching strategy was given by Asiminoaei *et al.*[10] to reduce switching losses. Bacterial foraging based compensation was reported by Mishra *et al.*[11]. Bhattacharya *et al.* have reported Adaline based prediction and compensation techniques[12]-[14]. DFT based techniques are also prevalent.

The inherent lack of control in passive components when used for harmonic elimination in shunt active power filters has been addressed by Sahu *et al.*[15]. They have proposed a new topology where a main high voltage inverter has been used in conjunction with several low voltage cells.

The paper has been presented in six identifiable parts. Part I serves as the introduction. Part II introduces the reader to the novel APF topology and its circuit implementation. Controller design is discussed in Part III. The section serves to present

simulation results. Part V leaves scope for experimental results. The work has been concluded in Part VI

II. SYSTEM CONFIGURATION

The three phase high frequency inverter is a voltage source inverter. Since this inverter is used for harmonic compensation, hence the DC link voltage is comparatively low. This allows the usage of MOSFETs as switching devices. This brings down system cost as MOSFETs are far cheaper than IGBTs. It provides harmonic compensation to the system. This is achieved by Synchronous Reference Frame Controller which extracts the real and reactive part of the load current and transforms them into vectors in a synchronously rotating frame. The harmonic reference currents are extracted by passing these through a high pass filter.

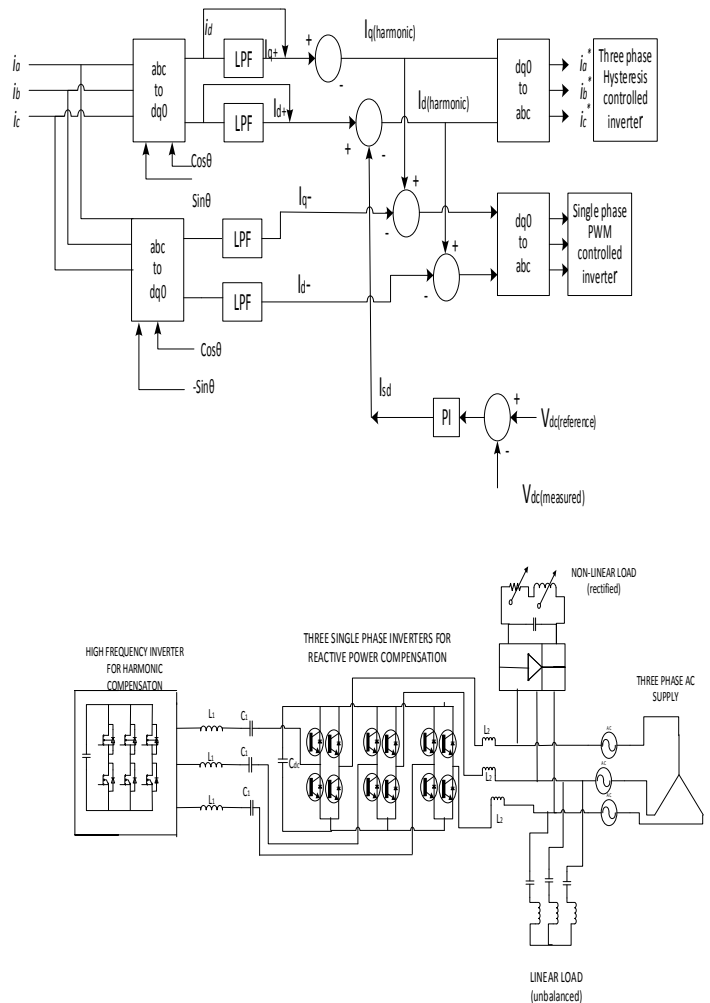
Further to sink harmonic currents effectively a passive filter comprising of an inductor and capacitor is placed in series in each of the three phases. They are tuned to effectively eliminate 5th and 7th order harmonics.

The reactive power compensation is performed for three different phases separately. Since most of the times, the load is unbalanced and its effect is more prominent on reactive power compensation than on harmonic compensation. Hence, our proposed topology uses three different single phase inverters to compensate reactive power for unbalanced load. Both negative and positive sequence load currents are separately heterodyned to achieve control signals for generating negative and positive sequence compensating currents. Since these inverters are series connected, the voltage stresses to which they are individually subjected are considerably reduced. Also, The single phase inverters use a reduced bridge topology further bringing down the cost of implementation.

III. CONTROLLER

The basic controller controls all the four inverters used in the circuit all the three single phase inverters are maintained by a single DC link capacitor the three filter voltages along with the DC link voltage of the capacitor are used to maintain the inverters. We use sinusoidal PWM controller for the three single phase inverters. The aim of the three phase inverter is to provide harmonic reduction by the use of hysteresis control. This helps in reducing the complexity of the controller. For obtaining the control signals, the load current is converted from abc to dq frame using park's transformation. The cut off frequency of the LPF connected to the reactive power compensating single phase inverters is kept in such a way that it just allows the fundamental component of ac reactive current into the system. The constant band hysteresis controller ensure that drifting of switching frequency is content within 1% of mean switching frequency. Use of single controller for maintaining the DC link voltages not only reduces controller complexity but also brings down cost of implementation. As shown in [14], the controller has been

implemented on similar lines while the compensating currents are generated using the scheme shown in Fig 2.



IV. SIMULATION RESULTS

The simulations have been performed on Matlab SIMULINK. The Simscape –Sim power electronics block has been used to implement various circuit parts like the Low and High Frequency Inverters, The unbalanced load and the three phase AC supply. The three phase inverter that is compensating harmonics has been implemented using MOSFETs whereas the three single phase inverters which have higher voltage ratings have been implemented using IGBTs. The simulations have been carried out using discrete time step solution in Power GUI block. The mean switching frequency of LFI is around 2.5 KHz whereas that of HFI is around 20 KHz. To consider the effect of non-linearity as well as load unbalance, a three phase diode rectified RL load as well as simple unbalanced linear load have been considered as shown in Fig 1. Thus the following results have been obtained.

- A. Figure 3 shows the steady state operation of the Low Frequency Inverter. It can be observed that both source voltage and load current are co-phasor. Also the waveform (d) shows the nature of compensating current.
- B. Figure 4 shows the performance of same LFI for unbalanced load conditions. The source voltage, source current, load current and compensating current are shown in respective order.
- C. Figure 5 illustrates the operation on both the HFI and the single phase LFI in tandem. Almost perfectly sinusoidal load current in phase with source voltage can be observed which proves the utility of the proposed topology for prescribed purpose
- D. The Figure 6 illustrates the starting of the APFs. The design of the inverters is further validated by the “glitch free” starting of the two inverters which ensures that the load is not subjected to unduly high transients during start-up and smooth start is ensured.
- E. Finally the last figure shows the effect of step load change. Inverters show resilience to glitches and smoothly transit from one load condition to other ensuring that load is not subjected to switching hazards like high transient currents and voltages.

F Figures and Tables

Fig 4: (Waveforms for) Performance of LFI for unbalanced load.
 (a): Source voltage (Scale: 50V/div)
 (b): Source current (Scale: 10A/div)
 (c): Load current (Scale: 10A/div)
 (d): Compensating current of low frequency inverter 10A/div
 Time Scale: 20ms/div

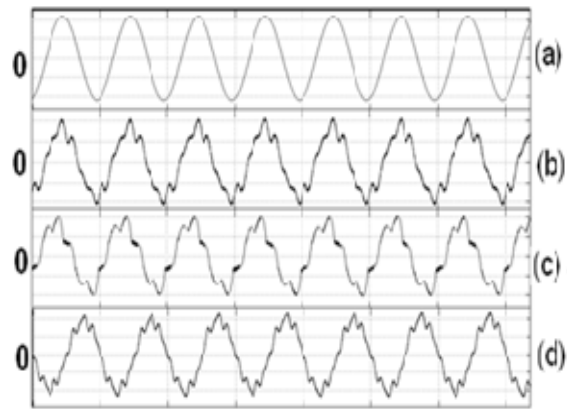


Fig. 3:

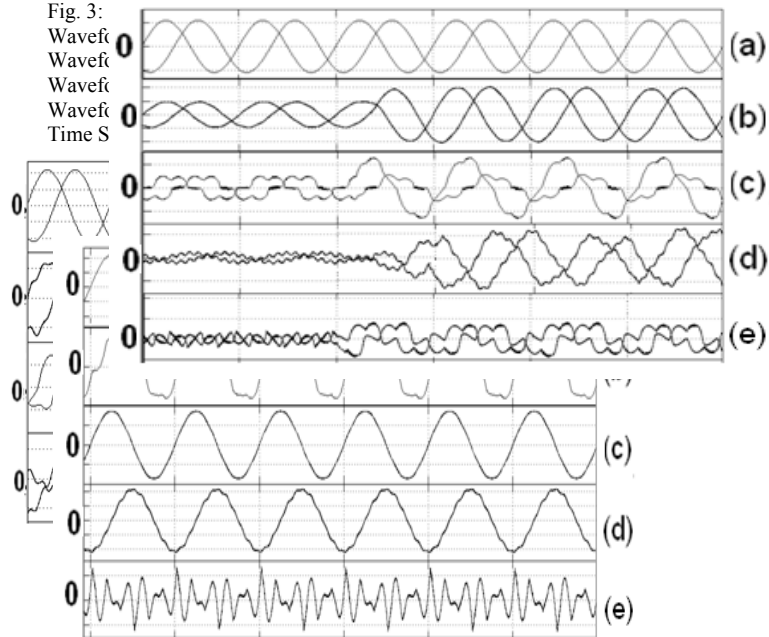
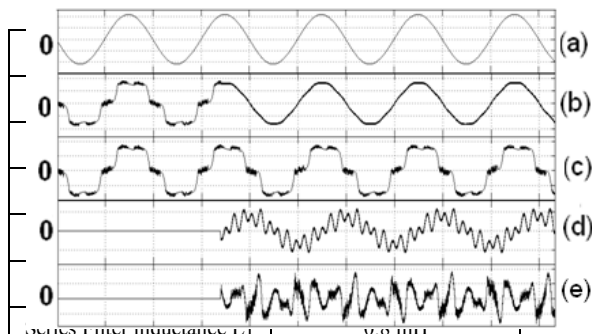


Fig 5: Steady State operation (waveforms) of APF when both the APF are operating.

(a): Source voltage (Scale: 50V/div)
 (b): Load current (Scale: 10A/div)
 (c): Source current (Scale: 10A/div)
 (d): Compensating current of low frequency inverter (Scale: 10A/div)
 (e): Compensating current of high frequency inverter (Scale: 5A/div)
 Time Scale: 20ms/div

Fig 7: Effect of step change in load on APF operation.

(a): Source voltage of phase A&B (Scale: 50V/div)
 (b): Source current of phase A & B (Scale: 10A/div)
 (c): Load current of phase A & B (Scale: 10A/div)
 (d): Compensating current of low frequency inverter for phase A & B (Scale: 10A/div)
 (e): Compensating current of high frequency inverter for phase A & B (Scale: 5A/div)
 Time Scale: 20ms/div.



Series filter inductance L1	0.6 mH
Quality factor of L1 and C1	25
DC link voltage	40V
DC link capacitor	3.6 mF
Switching frequency of 3-phase inverter (mean)	2.5 kHz
Switching frequency of line inverters	20 kHz

Fig 6: Waveforms showing starting of the APF

(a): Source voltage (Scale: 50V/div)

(b): Source current (Scale: 10A/div)

(c): Load current (Scale: 10A/div)

(d): Compensating current of low frequency inverter (Scale: 10A/div)

(e): Compensating current of high frequency inverter (Scale: 5A/div)

Time Scale: 20ms/div

V. CONCLUSION

The paper has proposed topology, based on four shunt active power filters, that gives enhanced control over individual phase reactive power and harmonic performance in reduced cost. The HFI connected through the shunt path has the capability to handle the 5th and 7th harmonics of the load waveform. The three LFI's that are connected to each individual phase of the supply are responsible for compensation of the reactive power.

Through the simulation results we have seen that the load power factor has been really improved and the harmonic content in load current is also considerably reduced.

The control over each phase helps us in practically mitigating the problems of unbalanced load (both real and reactive). The proposed topology has been extensively simulated on SIMULINK. Researchers have planned to implement the topology on an experimental set up using dSPACE-1104 and the results of simulation and hardware implementation will be shown to match during final submission of the paper.

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Experimental Assessment of PV Arrays Connected to Buck-Boost Converter using MPPT and Non-MPPT Technique by Implementing in Real Time Hardware

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Abstract— Solar photovoltaic (PV) arrays are a great source of generation of renewable energy. Harvesting clean energy from photovoltaic cell is reliable, pollution and noise free, low maintenance and long lasting energy source. However, typical Voltage-Current (V-I) and Power-Voltage (P-V) characteristics curves of PV cells depend on the weather conditions such as temperature and irradiation. It is always difficult to fix Maximum Power Point (MPP) on V-I and P-V curve. In order to achieve maximum efficiency from PV cells, Maximum Power Point Tracking (MPPT) is used. This paper presents a non-inverting Buck-Boost converter based on MPPT and Non-MPPT technique by implementing modified P&O algorithm through MPPT control circuit. The modified P&O algorithm is a simple algorithm that does not require previous knowledge of the PV characteristics. A PV system incorporating a solar panel, Buck-Boost converter and load as battery is modeled and implemented through hardware. Entire range of the duty cycle needs to be used for using this modified algorithm. Without adding much additional cost a desired efficiency is obtained. A microcontroller based PV system is used in this work.

Key words — Renewable Energy; Photovoltaic (PV) Array; P&O Algorithm; Buck-Boost Converter; MPPT, Duty Cycle; Microcontroller.

I. INTRODUCTION

Solar photovoltaic (PV) array is becoming an increasingly attractive way to generate power around the world in the last few years. The fact is that solar energy can be directly converted into electrical energy without any intermediate conversion process. Since the present world faces both energy crisis and global warming. Photovoltaic energy can mitigate these global problems. It does not emit CO₂ emission resulting in environmentally friendly and pollution free. However, extracting energy from photovoltaic arrays is still expensive and inefficient in terms of technological view point compared to other sources of energy. So it is important to optimize the efficiency of PV panel. A lot of research has been conducted to address this matter [1] [2]. A typical solar panel characteristic basically depends on the weather conditions changing i.e. temperature, illumination. The technology which is widely used is a device that continually finds the maximum power point of a solar array termed as maximum power point tracker (MPPT) [4]. Many sophisticated algorithms have come

up with control this device [3] [5].

Many researchers have developed and implemented many methods in this research fields. Such as, in [6], proposed the output sensing method implementing P&O just slightly modified it using MATLAB simulator and measured converter efficiency. In [9], proposed an adaptive P&O MPPT for a photovoltaic module connected to the buck-boost converter using MATLAB for current, voltage and output power for different solar irradiation. In [10], proposed the comparison between two algorithms P&O and incremental conductance method through buck, boost and buck converter using MATLAB. In [11], authors presented in details implementation of P&O MPPT using buck and buck-boost converter through MATLAB.

In this paper, a modified P&O algorithm is implemented through buck-boost converter controlled by duty cycle using automatic controlled with the help of microcontroller. This research article implemented modified algorithm with real time hardware.

The remaining paper is organized as follows. In Section II, theoretical background is described. In Section III, system design and description is given. In Section IV modified MPPT algorithm is explained, system performance and analysis are given in Section V. The paper is concluded in Section VI.

II. THEORETICAL BACKGROUND

A. PV Module Modeling

A solar PV array is considered as a P-N junction semiconductor. When light is incident on a semiconductor material DC current is generated. The produced current is linearly dependant with solar irradiance. One single diode equivalent circuit of a solar cell is considered as one of the most popular mathematical model for solar arrays, almost all practical solar cells work under this mathematical principle [12][7]. Equivalent Circuit of PV Solar Cell is as shown in Fig. 1.

The current through the diode is given by:

$$I_d = I_o \left[e^{\frac{(V+IR_s)}{AVT}} - 1 \right] \quad (1)$$

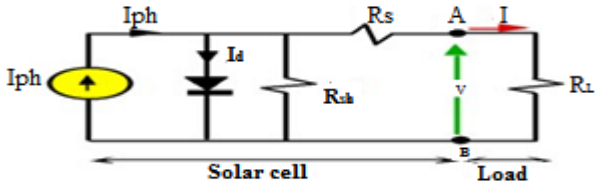


Fig. 1. Equivalent circuit of PV solar cell.

While, the solar cell output current:

$$I = I_{ph} - I_d - I_{sh} \quad (2)$$

$$I = I_{ph} - I_0 \left(e^{\frac{V+IR_s}{AV_t}} - 1 \right) - \frac{V+IR_s}{R_{sh}} \quad (3)$$

Where, I is the cell current (A), I_L is the light generated current (A), I_0 the diode saturation current, A is diode quality factor, V_t is thermal voltage, R_s and R_{sh} are cell series and shunt resistance (ohms) and V is the cell output voltage (V). Making a better mathematical model, the calculation of I_{ph} , I_d , and R_s parameter including temperature effects should be taken into account.

B. MPPT in PV Solar System

MPPT [4][8] is for Maximum Power Point Tracking for the case of solar cells to draw maximum power possible from them, for the specified environmental and structural condition. Main causes to search for MPP are ambient temperature and irradiance related as shown in Fig. 2.

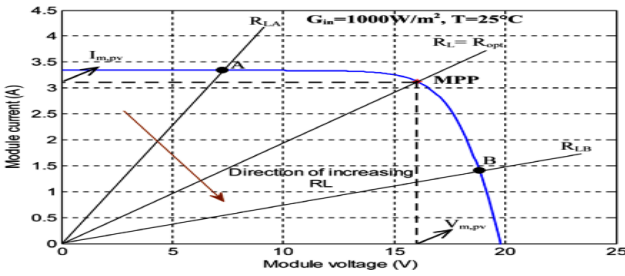


Fig. 2. Solar panel IV curve with MPPT.

III. SYSTEM DESIGN AND DESCRIPTION

The modified MPPT algorithm is used to control the maximum power transfer from a PV panel. This algorithm is executed by a PIC16F73 microcontroller using the PV voltage and current data to compute the maximum power. This computed maximum power is compared with instantaneous actual power from PV panel. The difference between the maximum power (reference power) and actual power drives the converter which is controlled by PWM from the microcontroller. The system block diagram is as shown in Fig. 3.

A. Microcontroller

The MPPT control circuit is implemented in a microcontroller PIC16F73 that has 8-bits analog-to-digital (A/D) converter and two PWM mode signals. The buck-boost converter is controlled by the microcontroller. It reads the voltage and current of the solar panel through the A/D port of the controller and calculates the output power. The PIC16F73 is a perfect combination of performance, features, and low power consumption for this application [7].

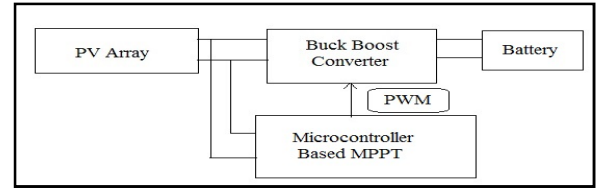


Fig. 3. System block diagram with MPPT.

B. Buck-Boost Converter

A non-inverting buck-boost converter is essentially a cascaded combination of a buck converter followed by a boost converter, where a single inductor-capacitor is used for both. This converter requires the use of two active switches and is designed by combining a buck converter and boost converter design in the same topology. Due to this design, this converter can work as buck-only, boost-only or buck-boost converter. It provides an output voltage that may be less than or greater than the input voltage, hence the name buck-boost. Basic buck-boost converter circuit is as shown in Fig. 4.

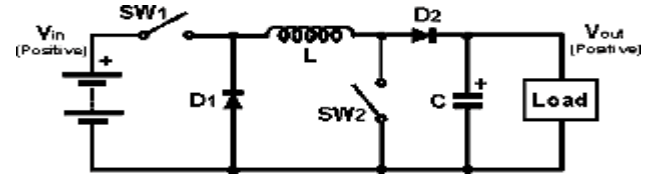


Fig. 4. Basic buck-boost converter.

IV. MODIFIED MPPT ALGORITHM

A generic I-V characteristic of a PV module shows the maximum power point (MPP),

$$P_{max} = \frac{V_{max}}{I_{max}} \quad (4)$$

Where, V_{max} and I_{max} are the PV max voltage and current respectively. Since these quantities vary in accordance with climate changes, particularly temperature and irradiance of sunlight, it is always critical to set a maximum power point. To get a maximum power point, a technique is required which can locate the exact position of MPP called the MPPT algorithm.

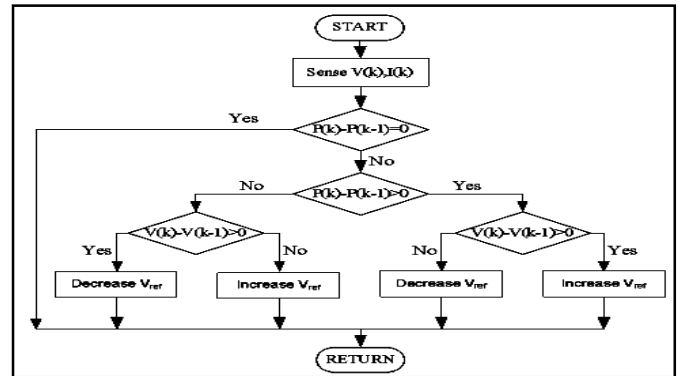


Fig. 5. The flowchart of the P&O MPPT method.

The P&O MPPT algorithm is as shown in Fig. 5. The idea behind the traditional P&O algorithm is that it changes the PV-voltage or current until reaching the optimum point termed as MPP. The modified algorithm works the same as the traditional one except some modification is introduced. The flow chart of

this modified algorithm is presented in Fig. 6. This algorithm starts with setting the PWM duty cycle value at half of duty cycle or any other values.

The flow chart of the modified method containing the additional block set of the duty cycle at its initial value. Reading the voltage and current from PV panel, this algorithm calculated the present power. Comparing the difference of present power and past power which is put into algorithm as zero initially and then algorithm decide the next step.

When the power difference is greater than zero, it increment the duty cycle until the position of correct MPP and vice versa. At this point, the present power is equal to past power. Here the duty cycle value is incremented or decremented with a fixed step size.

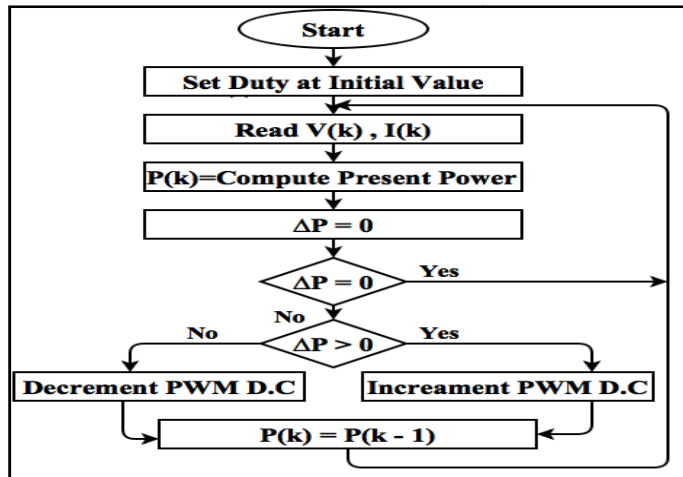


Fig. 6. Flowchart of the modified MPPT algorithm.

The algorithm implemented in this work is called power reading algorithm since it works sensing the power of PV array. The Significant difference between the modified and conventional P&O algorithm is that the modified algorithm is used to govern the MPPT directly drawn the power from PV array, whereas conventional P&O algorithm is used to control the MPPT by periodically increasing or decreasing the solar array voltage.

The MPPT tracking process is as shown in Fig. 7. This curve is obtained under standard testing condition at irradiation level 1kw/m^2 and at 25°C .

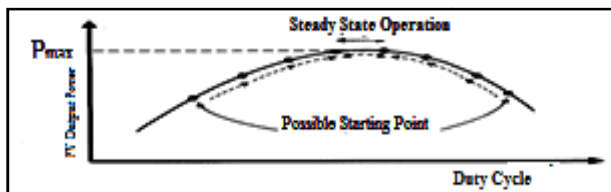


Fig. 7. Ideal MPPT tracking process.



Fig. 8. Measuring data with solar panel.

V. SYSTEM PERFORMANCE AND ANALYSIS

The specification of PV module is as shown in Table I.

TABLE I. SPECIFICATION OF THE PV MODULE

Maximum power	50 W
Open circuit voltage	21.60 V
Short circuit current	3.23 A
Voltage at max power	17.20 V
Current at max power	2.91 A
Cell Temperature	25°C
AM	1.5
Irradiation	1000 W/m ²
Weight	5 Kg

In this section we present in details implementation of modified P&O MPPT and Non-MPPT technique using buck-boost converter.

The data shown in Table II was measured on 17th July, 2015, Chittagong in Bangladesh using buck-boost converter with Non- MPPT showing circuit response.

TABLE II. MEASURED VALUE FROM CIRCUIT RESPONSE

Time	Input Voltage	Input Current	Duty Cycle (%)	Output Voltage	Output Current
11:40	17.8	0.41	0.420	12.9	0.27
11:50	17.2	0.56	0.429	12.9	0.50
12:00	17.1	0.56	0.429	12.9	0.49
12:10	17.08	0.54	0.430	12.9	0.49
12:20	17.05	0.53	0.431	12.9	0.46

The Table III and Table IV data were measured on 17st July, 2015, Chittagong in Bangladesh using modified MPPT algorithm through buck boost converter at different times.

TABLE III. MEASURED VALUE BY MPPT ALGORITHM FROM CIRCUIT RESPONSE

17 st July, 2015 at 10.00 AM, Chittagong in Bangladesh					
Time	Input Voltage	Input Current	Duty Cycle	Output Voltage	Output Current
10.00	17.5	2.46	173	12.9	3.17
10.10	18.1	2.33	155	12.9	3.19
10.20	18.6	2.53	160	12.9	3.48
10.30	17.9	2.67	180	12.9	3.17
10.40	17.7	2.52	178	12.9	3.30
10.50	18.3	2.25	150	12.9	3.03
11.00	17.6	2.50	176	12.9	3.25
11.10	18.4	2.06	138	12.9	3.79
11.20	17.8	2.35	158	12.9	3.10
11.30	18.7	2.32	170	12.9	3.21

TABLE IV. MEASURED VALUE BY MPPT ALGORITHM FROM CIRCUIT RESPONSE

17 st July, 2015 at 2.20 PM, Chittagong in Bangladesh					
Time	Input Voltage	Input Current	Duty Cycle	Output Voltage	Output Current
2.20	17.5	2.36	152	13.15	2.99
2.30	17.9	2.45	157	13.2	3.23
2.40	18.5	2.10	138	13.2	2.84
2.50	18.9	2.22	140	13.2	3.03
3.00	16.6	2.76	165	13.2	3.31
3.10	17.8	2.08	95	13.26	2.69
3.20	18.3	2.10	105	13.29	2.73
3.30	18.6	2.23	170	13.35	2.96
3.40	17.8	2.37	155	13.4	3.02
3.50	19.00	2.23	160	13.42	3.02

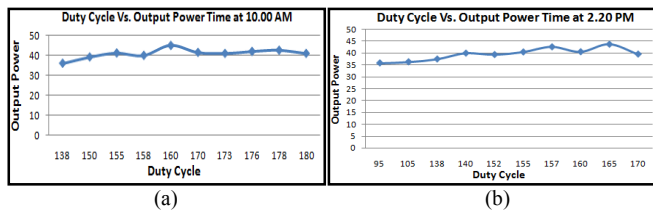


Fig. 9. Duty cycle Vs. Output Power characteristics: a) at time 10.00 AM
b) at time 2.20 PM.

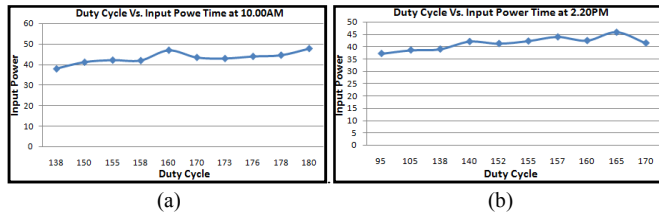


Fig. 10. Duty cycle Vs. Input Power characteristics: a) at time 10.00 AM
b) at time 2.20 PM.

Input and output power at different instant of tracking process also changes between different periods of day time as shown in Fig. 9(a), 9(b), 10(a) and 10(b). Initially, the output power varies significantly with duty cycle indicates by the slope of the curve in figures.

Average efficiency of the buck-boost converter at different day time is as shown in Table V.

TABLE V. AVERAGE EFFICIENCY AT DIFFERENT DAY TIME

Input current	Input voltage	Input power	Output current	Output voltage	Output power	Efficiency
Time at 10.00 AM						
18.06	2.40	43.34	12.9	3.26	42.05	97.02%
Time at 2.20 PM						
18.09	2.29	41.43	13.26	2.98	39.51	95.37%

Average input and output power from MPPT and Non-MPPT is as shown in Table VI.

TABLE VI. AVERAGE INPUT AND OUTPUT POWER

Input Power		Output Power	
MPPT	Non-MPPT	MPPT	Non-MPPT
42.39	9.33	40.78	5.7

The equation to calculate the efficiency of MPPT than Non-MPPT is shown in Eq. (5).

$$\% \text{Efficiency} = \frac{\text{Total power (with MPPT)} - \text{Total power (Non - MPPT)}}{\text{Total power (with MPPT)}} \times 100 \quad (5)$$

Results of Comparative efficiency of MPPT than Non-MPPT Systems are discussed in Table VII using the Eq. (5).

Table VII. COMPARATIVE RESULT OF MPPT AND NON-MPPT

MPPT Efficiency than Non-MPPT	
According to average Input	According to average Output
77.99%	86.02%

From Table II we see that the input voltage and current is greater than output voltage and current showing the characteristic of buck-boost converter. From Table III we see that the output powers are fluctuate very little. In this solar panel the indicated power is 50W. Average output power is measured from solar panel for MPPT and Non-MPPT is 40.78W and 5.7W respectively which are around 81.56% for MPPT and 11.4% for Non-MPPT.

From above analysis we observed that applied modified algorithm works efficiently regardless of weather conditions.

VI. CONCLUSION

This paper presented experimental evaluation of PV cells designed on MPPT and Non-MPPT scheme. Results reveal that significant amount of power is wasted when Non-MPPT scheme is introduced. According to measured data 81.56% average power is achieved compared to Non-MPPT technique which is 11.4%. Moreover, the result shows that MPPT is more efficient than Non-MPPT which is about 86.02% according to average output power.

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A Novel DMC based SST for Voltage and Frequency Regulation of PMSG based WECS

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Abstract—This paper presents a novel topology for voltage and frequency regulation of wind energy conversion systems (WECS) equipped with Permanent Magnet Synchronous Generators (PMSG) using Direct Matrix Converter (DMC) based Solid State Transformer (SST). The venturini based algorithm together with the voltage oriented vector control (VOC) scheme is used to regulate the generated power from WECS also called as Power Regulation Technique, which can effectively regulate the generated voltage and frequency at output side of SST. The variation in generated power with varying voltage & frequency due to the change in wind speed are processed by proportional & integral (PI) controllers to regulate the voltage gain of Matrix Converter (MC) through which the desired regulation of powers, voltage and frequency is carried out. Matlab/Simulink based model is tested for evaluating the performances of proposed scheme. From the simulation results it is confirm that PMSG based WECS under varying wind speed using SST with adequate input filters and proposed control scheme is able to directly fed power to the existing grid or local load at desired output frequency and voltage at nearly unitary input power factor.

Keywords— DMC, WECS, VOC, PMSG

I. INTRODUCTION

Wind energy conversion systems (WECS) equipped with Permanent Magnet Synchronous Generators (PMSG) plays an important role in contribution of electrical energy from renewable energy sources. But it possesses some undesirable drawbacks such as voltage and frequency variation under varying wind speed. Under such intermittent wind conditions power electronic converters are required to synchronize the PMSG based WECS with the grid or for supply power to local loads. [1-2]

Generally two stage AC-DC-AC power converter (rectifier-inverter) is used for power conversion and regulation but due to an intermediate DC-link with capacitor banks, an increase in losses, total weight, size and costs of the equipment, as well as lifetime decreases. Due to these demerits, direct AC-AC conversion is developing more interest in research field. Recently Solid State Transformers (SSTs) are gaining more attention for such works.[3-4]

SST is mainly composed of power electronics based components. It can effectively regulate the magnitude of voltage, current or frequency according to the application & requirements. Direct Matrix Converters (DMC) based SST are single stage AC/AC bidirectional power converters capable of establishing a desired output frequency and voltage at nearly unitary input power factor.[5-7] It is one of the most appropriate converter for harnessing wind power because of

its high merit over traditional back-to-back voltage source converter like free from commutation problems, improved voltage gain with simplified control, compact in size, light weight, high reliability due to absence of dc capacitor and extremely fast transient response.[3-4][8-11]

Various control strategies have been proposed to regulate the power exchange between converter with source and grid, such as current hysteresis control (CHC), voltage-oriented control (VOC) and proportional-resonant (PR)-based control. [12-14] But as a close relationship exists between reactive power and voltage magnitude, reactive power support is the mean used to maintain the desired voltage profile, both during normal and contingency conditions. [15]

In VOC based scheme the errors between the active and reactive components of the line current and the reference values are fed to proportional-integral (PI) controllers in the synchronous reference frame, which generate the reference voltage for the converter. This voltage is then applied to the converter using a voltage modulator.[12-14]

The aim of this work is to interface variable speed PMSG with grid using DMC based SST. The power from WECS is processed using venturini based algorithm together with the voltage oriented vector control (VOC) scheme with conventional proportional and integral (PI) controller to regulate active and reactive power at converter side which effectively regulate the voltage gain of DMC.[2][10-16]

II. SYSTEM MODEL & CONTROL STRATEGY

The block diagram of the proposed DMC based SST with PMSG based WECS is shown in fig. 1.

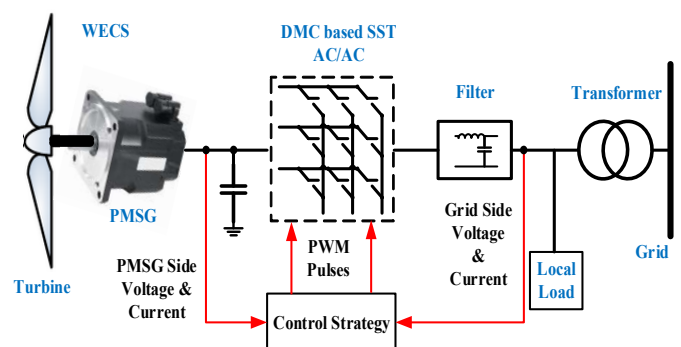


Fig.1 Proposed System with control strategy model

A. Solid State Transformer

W. McMurray et.al in 1970 first introduced the concept of a high frequency link called as Power Electronic Transformer (PET). PETs are also known as solid-state transformer (SST) or intelligent universal transformer (IUT). These SSTs are capable of performing a large number of functions which can bolster the power delivery quality and stability of the electrical grid simultaneously. Not only do these transformers provide higher flexibility and efficiency, but are also capable of providing reactive power support to the grid according to demand. As a result these SST provide power at a better power factor and quality.[17] There are various topologies of SST, out of which Direct Matrix Converter (DMC) based SST is main focus for this work.

Direct Matrix Converter

A direct matrix converter (DMC) is a single stage AC-AC forced commutated converter that can directly convert an AC power supply voltage into an AC voltage of variable amplitude and frequency without any energy storage elements or it does not have any dc-link circuit. [2] It uses a fully controlled bi-directional switch, which allows high frequency operation in all four quadrants.

Modelling of Matrix Converter using Venturini algorithm

In venturini based algorithm: [8]

The Output voltage vector, $V_o = T(S_{ij}) * V_i$

Input Current Vector, $I_i = T(S_{ij})^T * I_o$

Where, $T(S_{ij})$ is instantaneous transfer matrix for DMC and represented as:

$$T(S_{ij}) = \begin{bmatrix} S_{Aa} & S_{Ba} & S_{Ca} \\ S_{Ab} & S_{Bb} & S_{Cb} \\ S_{Ac} & S_{Bc} & S_{Cc} \end{bmatrix}$$

Where,

$V_o = [V_a \ V_b \ V_c]^T$, $V_i = [V_A \ V_B \ V_C]^T$,

$I_o = [i_a \ i_b \ i_c]^T$, $I_i = [i_A \ i_B \ i_C]^T$ and

$S_{ij} = \frac{t_{ij}}{T_s}$, $\{i = A, B, C \text{ and } j = a, b, c\}$ is the duty cycle and t_{ij} is the 'on' time of the switch S_{ij} . T_s is the sampling period.

$$\begin{aligned} S_{Aa} + S_{Ba} + S_{Ca} &= 1; \\ S_{Ab} + S_{Bb} + S_{Cb} &= 1; \\ S_{Ac} + S_{Bc} + S_{Cc} &= 1 \end{aligned}$$

Duty cycles must satisfy the above conditions to avoid short circuit on the input side and open circuit at output side [6]

The generated duty ratios are given below:

$$\begin{aligned} S_{Aa} &= S_{Bb} = S_{Cc} = \frac{1}{3} [1 + 2q_v \cos(\omega_m t + \theta)] \\ S_{Ba} &= S_{Cb} = S_{Ac} = \frac{1}{3} \left[1 + 2q_v \cos \left(\omega_m t + \theta - \frac{2\pi}{3} \right) \right] \\ S_{Ca} &= S_{Ab} = S_{Bc} = \frac{1}{3} \left[1 + 2q_v \cos \left(\omega_m t + \theta - \frac{4\pi}{3} \right) \right] \end{aligned}$$

B. Control Strategy

VOC scheme to regulate active and reactive power

The control scheme, shown in Fig. 2 is used to regulate power flow in the system. The grid side voltages V_R , V_Y & V_B are transformed into $d^s - q^s$ in stationary reference frame and further into $d_e - q_e$ in synchronously rotating reference frame transformation. The actual active power (P_s) and reactive Power (Q_s) flowing through the grid are compared with the reference active power (P_s^*) and reactive power (Q_s^*) to generate the reference voltages ($V_d I^{*e}$ & $V_q I^{*e}$). Again this reference voltages are transformed into V_{abcI} which will become reference modulating voltages for matrix converter. Here q_v is the voltage gain of reference modulating signal. [13-15]

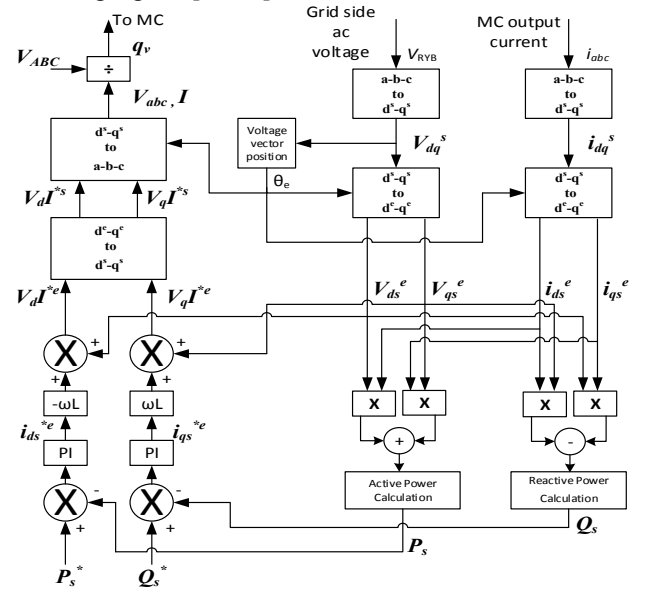


Fig.2 VOC based control scheme

III. SIMULINK MODEL

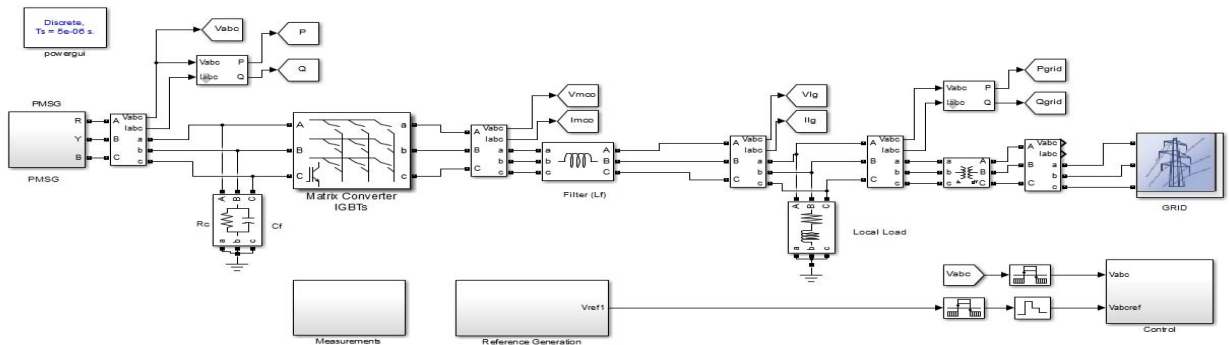


Fig. 3 Simulation model of the proposed system

IV. RESULTS & DISCUSSIONS

The simulation model of the WECS using PMSG with DMC based SST connected to grid through a transformer and local load is developed using MATLAB/simulink is shown in fig 3. Fig.4 shows that performance parameters of PMSG at varying wind Speed. In fig 4(a) variation of wind

speed is shown. Corresponding to this wind speed the PMSG generated voltage and current is shown in fig 4 (b). From the results it's seen that with the decrease in wind speed from time interval 3 to 4 and 5 to 6, the PMSG generated voltage, current and frequency is decreased accordingly.

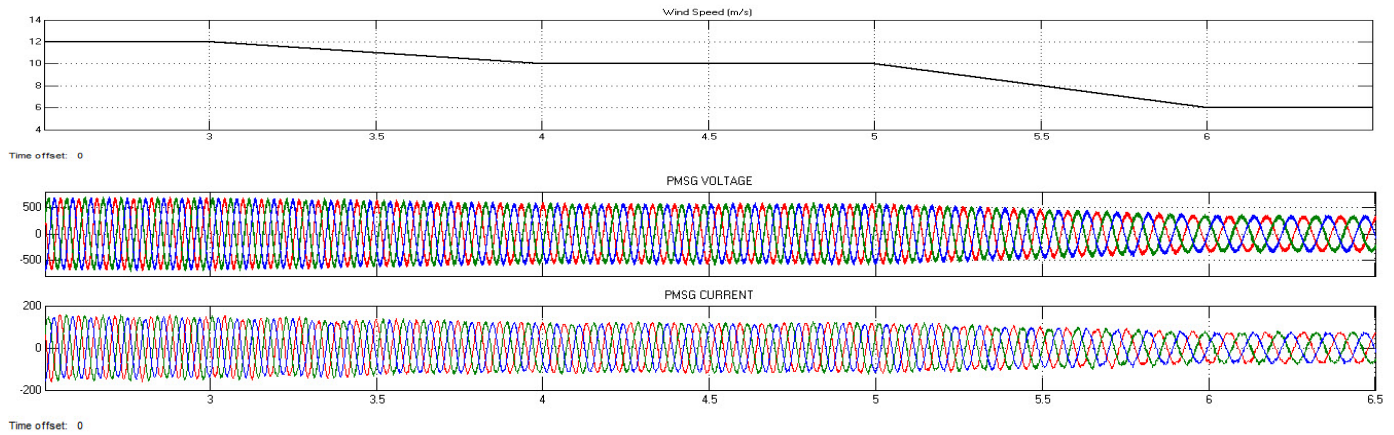


Fig.4 Performance Parameters of PMSG at varying wind Speed a) Wind Speed b) Voltage & Current

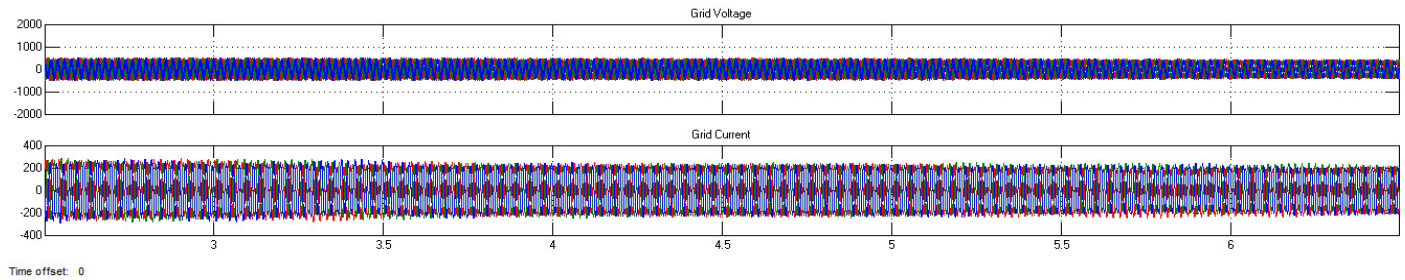


Fig.5 Voltage & Current at grid Side at varying wind Speed

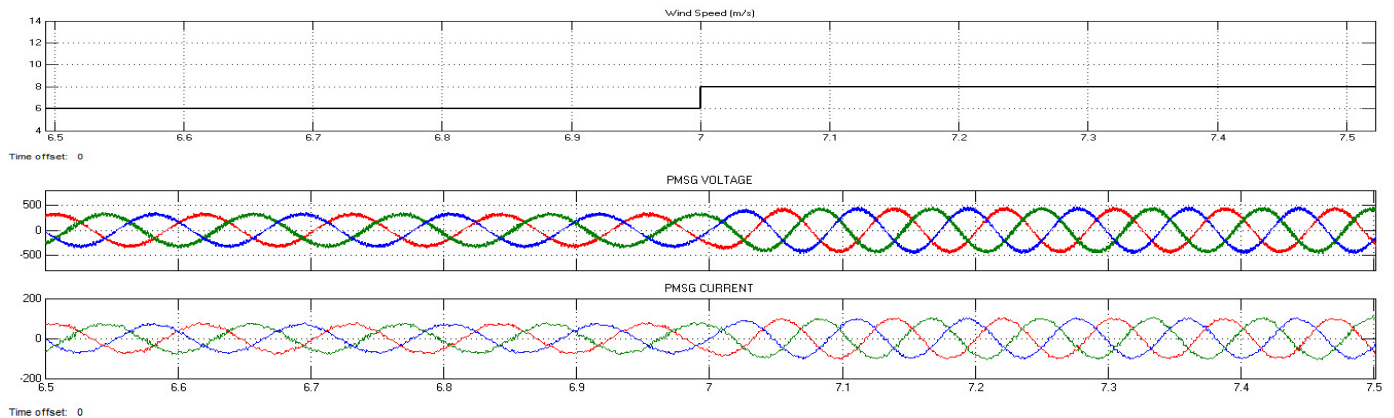


Fig.6 Performance Parameters of PMSG at step change a) Wind Speed b) Voltage, Current & Frequency

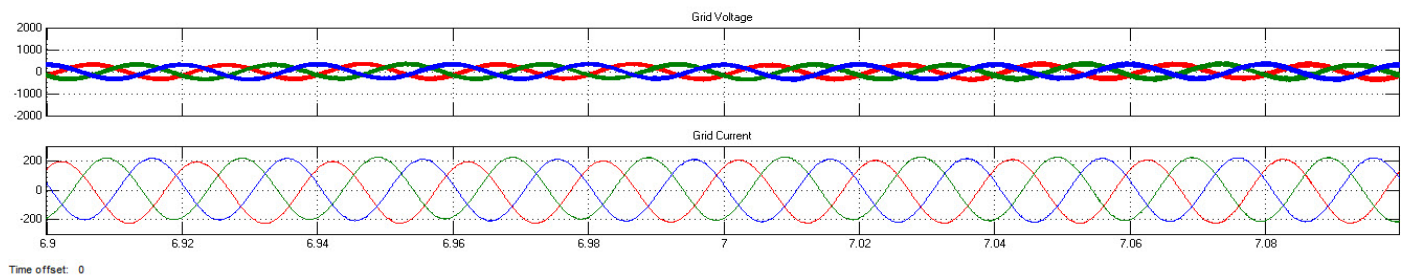


Fig.7 Voltage, Current & Frequency at step change in wind Speed at grid side

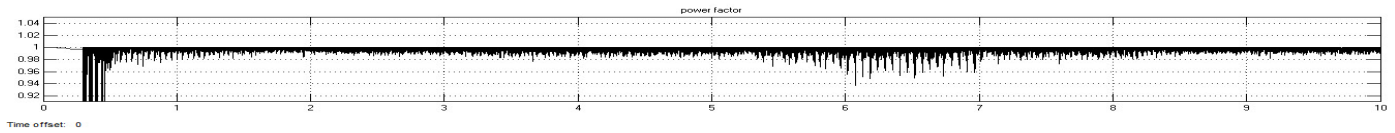


Fig.8 Power Factor at input side of matrix converter

Fig 5 shows the performance parameters at grid side under varying wind Speed. In fig 5 it is seen that the voltage Waveform of the system at grid side remains constant but the current is decreased between time interval 3 to 4 and 5 to 6 as the wind speed changes correspondingly generated power changes. While the frequency of the system remains constant

In fig 6(a) there is a step change at time instant 7 in wind speed. Due to which there is change in generated voltage and current of PMSG as well as the frequency of the generated voltage and current changes from 9 Hz to 12Hz approximately as shown in fig 6(b).

Fig 7 shows the voltage and current waveforms at the grid side or output of SST after filter. It is observe that the voltage and frequency of the system remains constant 415V (rms) and 50 Hz, while the current is increased.

Fig.8 shows that the input power factor of the DMC based SST remains nearly unity under all speed variation.

From the obtained system performance, it is found that the proposed algorithm works satisfactorily under all wind speed variation.

V. CONCLUSION

A PMSG based WECS with DMC based SST connected to grid and local load has been modeled using MATLAB/Simulink based platform. The aim of the proposed work was to regulate voltage and frequency of variable speed PMSG based WECS by using DMC based SST so that power can be directly transferred to the existing grid and local load has been achieved. The proposed venturini based algorithm together with the voltage oriented vector control (VOC) scheme with conventional proportional and integral (PI) controller to regulate active and reactive power at converter side has been simulated and results obtained from it are satisfactory.

On the basis of results it's projected that the proposed scheme has fast dynamic response, robust and supple for voltage and frequency as well as power regulation for VSWECS.

VI. APPENDIX

TABLE I: PARAMETERS OF PMSG

Rated Power	2[MW]	Stator Resistance	0.00344[p.u]
Rated Voltage	690[V]	d-axis inductance	0.4026[p.u]
Frequency	11.25[Hz]	q-axis inductance	0.7685[p.u]
No. of Poles	60	Field Flux	0.8332[p.u]

Filter: Inductor, $L_f = 5\text{mH}$, Capacitor, $C_f = 25\mu\text{F}$, Resistance, $R_c = 10\Omega$

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Development of a Single Phase Prepaid Electrical Energy Meter using 89S8252 Microcontroller Architecture

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Abstract—The Electricity Company implements the ‘Prepaid Electrical Energy Meter (PEM)’ scheme among the electricity consumers to meet the objectives of: (i) collecting advance money and (ii) motivating the customers to make conservative use of electricity. The 2nd objective can be effectively achieved with an indoor type PEM fitted with 7-segment bright display unit. The meter designer aims, using appropriate hardware modules and software routines, to guide the PEM for cycling through the following steps: (i) deduct cost of every 0.01kwhr energy consumed, (ii) display balance, (iii) check balance, (iv) disconnect load at zero balance, (v) issue alarm signal to engage correct prepaid card, (vi) read the prepaid card, (vii) zeroing the prepaid card, (viii) update balance, (ix) save balance in nonvolatile EEPROM at power failure, and (x) retrieve balance from EEPROM at power resumption. This paper has described, based on extensive experiments on real hardware using AT89S8252-ADE7755-AT90S2313 chip set, the implementation methodology of a PEM with bright display unit. A prototype meter has been built, tested and found to work as per standard specifications.

Keywords—PEM; Prepaid Energy Meter; Calibrated Energy Frequency; Prepaid Carr; CMCKIT

I. INTRODUCTION

In a PEM (Fig. 1), a consumer pays advance money for certain amount of electrical energy. For example, at a rate of Tk 4.30/kwhr, a consumer named Mr. X pays Tk 1000.00 against 232.558140 kwhr energy. Considering an average load of 1kW and running for 12Hrs a day, Mr. X would consume electric energy for about 19 days. Mr. X inscribes the money as initial balance into his PEM via a Prepaid Card (PRPC, (Fig. 1c) and the balance appears in a 6-digit bright display unit. The PEM is expected to be installed indoor having attachment with the cover plate of the entry switchboard (Fig. 1d) so that the balance remains visible to the house members at all times. The PEM updates balance by deducting the cost of every quanta of 0.01kwhr energy consumed. For 1kW load, there will be 36-sec to go to consume 0.01kwhr energy. So, at the end of 36-sec time, the PEM will show a balance of Tk $1000.000 - 0000.0430 = 999.95(70)$. The 2-digits within () of the balance are not shown in the display; but, they reside in memory (Fig. 5a, 5b) and are taken care of during subtraction. A PEM of this kind develops psychological awareness in the minds of the house members to switch off the unnecessary loads in order to prevent the balance from going down though it is only Ps 4.3 in a step. There is savings of electricity, which could be consumed by some other user.

The design of the PEM is a challenging job as we need to develop hardware (Fig. 2) and software (Fig. 4) to guide the meter for cycling through the steps: (i) deduct cost of every 0.01kwhr energy, (ii) display balance, (iii) check balance, (iv) disconnect load at zero balance, (v) issue alarm to engage PRPC, (vi) read PRPC, (vii) zeroing the PRPC, (viii) update balance, (ix) save balance into EEPROM at power failure, and (x) retrieve balance from EEPROM at power resumption.

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A. Literature Review

There are many companies [1, 2] who manufacture PEMs; but, unfortunately they do not post any information as to the technical design of various subsystems (Fig. 2) of the meter.

There are traces of conference papers [3-5] on PEMs; but, they seem to remain concentrated on the analytic/simulation aspects of the meter rather than dealing with the practical activities like ‘designing communication link between the meter and prepaid card’, ‘preventing the EEPROM from being corrupted during power failure’, and etc.



Fig. 1(a)

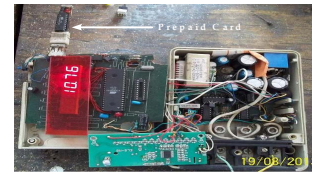


Fig. 1(b)



Fig. 1(c)



Fig. 1(d)

B. Motivation for the Project

In Bangladesh, we have shortage of electricity; it causes people and industries to suffer from load shedding. In this situation, a PEM can motivate people to make conservative use of electricity. Use of ‘Locally Developed PEM’ could contribute a lot in creating such conservative attitudes within the consumers. Imported meters would not help much in this regard owing to (i) very high cost of meters, (ii) payment in foreign currency, (iii) proprietary technology, (iv) lack of local expertise to repair imported meters, (v) prompt replacement of the faulty/burnt-out meters, (vi) resolving balance disputes for burnt-out meters, (vii) legal laws to deal with hackers, and (viii) security management of the PRPC.

C. Design Strategy

To realize the PEM, we will adopt ‘SSS (Small Start Strategy) Design Philosophy’, where an incremental approach is followed to design, develop and test the hardware components and the software routines of the system on trial-and-error basis. In SSS methodology, a minimum hardware is tested along with the activating software routine. And then an additional hardware/software is added with the previous one and is tested. The process keeps going on until the last component of the system is added and tested. The SSS requires that the development team has documented the initial rough design of the hardware block diagram (Fig. 2), schematics of the hardware subsystems (Fig. 3), flow chart of the control program (Fig. 4), and data structures (Fig. 5).

II. HARDWARE BLOCK DIAGRAM AND DESCRIPTION

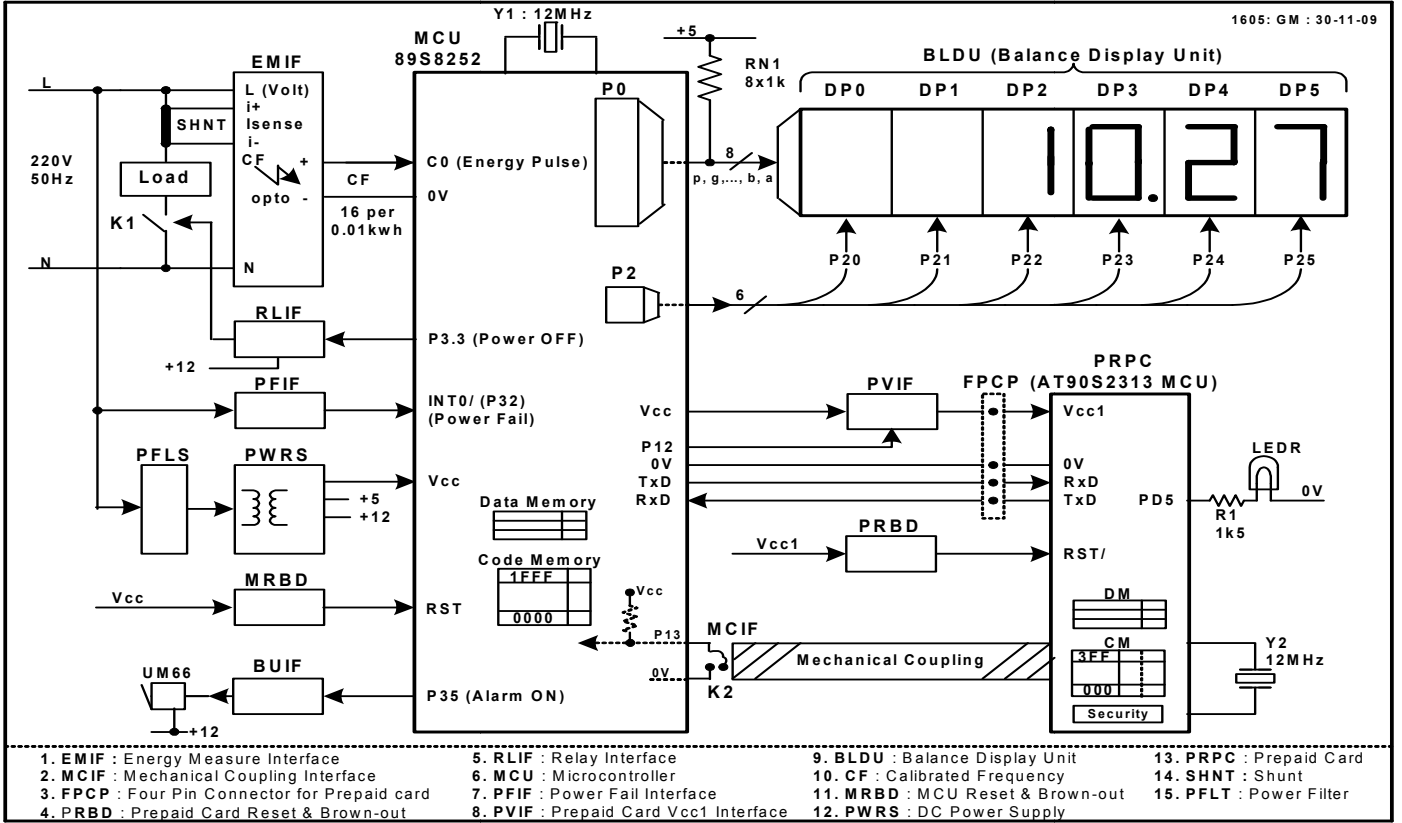


Fig. 2 Hardware Block Diagram of Prepaid Electrical Energy Meter

A. Brief Working Principle

EMIF is an ADE7755 based pre-fabricated calibrated commercial module (Fig. 3a) [6]. The module continuously monitors real power flow into the load. After the consumption of 0.000625kwhr energy, it sends a CF-pulse (called Calibrated Frequency, *cef*) to the Counter-0 (C0) of the MCU via an opto-coupler. The C0 accumulates 16 CFs to amount for 0.01kwhr and then deducts Ps 4.3 (at the tariff rate of Tk 4.30/kwhr) from the current balance. The balance appears on 7-segment bright display unit, BLDU.

When the balance arrives at a preset value (say, Tk 10.00), the 89S8252 MCU [7] initiates a musical alarm via BUIF (Fig. 3e), which alerts the user to recharge the meter by engaging the Prepaid Card (PRPC, Fig. 3i). After physical engagement of a correct PRPC, the MCU reads the recharge amount, inscribes it into the internal data structure (Fig. 5, 6) of the meter and then zeros the recharge field of the PRPC (Fig. 4).

At zero balance, the MCU cuts off the load via RLIF (Fig. 3f) and generate audible alarm asking for the recharge of the meter. At power failure, the MCU saves the last balance into its nonvolatile EEPROM (Fig. 5). At power resumption, the MCU retrieves the last balance from EEPROM, updates the internal data structure, shows it on the display unit and then resumes the operation by connecting supply line to the load.

B. Various Hardware Modules

1) EMIF: Energy Measure Interface

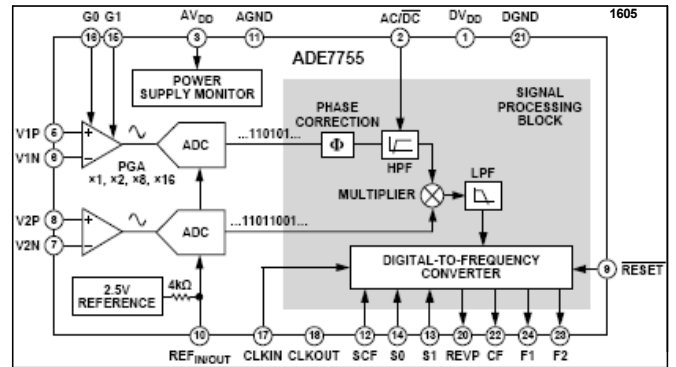


Fig. 3a EMIF Module and Internal Architecture of ADE7755 [8]

The ADE7755 is an active energy measurement IC with high accuracy. It complies with IEC 1306 standard. It directly couples with the power lines using simple attenuators and shunt. The chip outputs calibrated energy frequency, which could be counted and processed by a host microcontroller.

2) MRBD: MCU Power-on Reset and Brown-out Detector

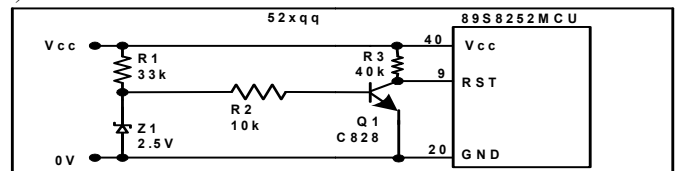


Fig. 3b Power-on Reset and Brown-out Detector for 89S8252

The circuit serves two purposes: (i) generates Power-on Reset (POR) signal to ensure that the 89S8252 MCU will always start up at power up time and (ii) protects the EEPROM from being erratically written during power failure. When power fails, the Vcc supply keeps falling down; when it arrives at +2.5V or less, the MCU stays in the reset condition.

3) PFLS: Power Supply Filter and Suppressor

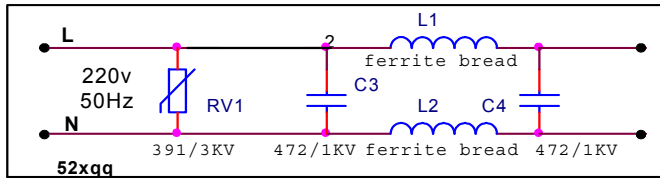


Fig. 3c Filter and Suppressor for the AC Supplies of the PEM

4) PWRS: DC Power Supplies

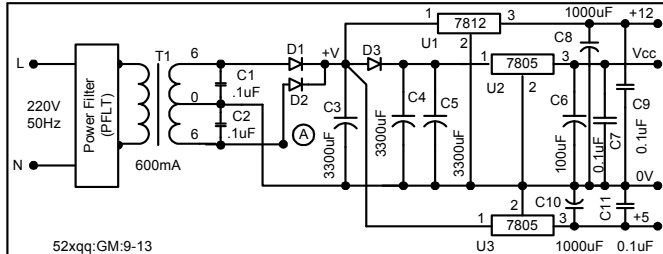


Fig. 3d DC Power Supplies for MCU, Relay and Electronics

Vcc is reserved for MCU. Two large capacitors C4 and C5 hold enough energy, which is sufficient to finish 'write up' task into EEPROM during power failure. +12V supply is for the Relay. Unregulated +V is for the Buzzer and +5V supply for all other electronics including the 7-segment display unit.

5) BUIF: Buzzer Interface

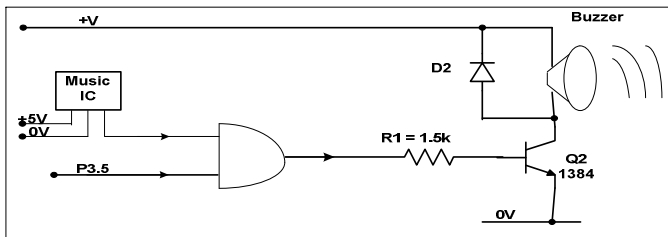


Fig. 3e Buzzer Circuit to Initiate Musical Alarm

6) RLIF: Relay Interface

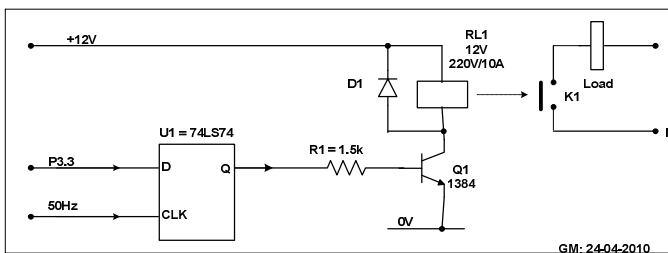


Fig. 3f Relay Interface Circuit

The Relay (RL1) becomes ON and OFF depending on the current balance of the PEM. The ON/OFF conditions are synchronized with +ve zero crossing point of the line voltage to ensure minimum sparks in the relay contacts. The control signal (P3.3) comes from MCU and enters into D-pin of 74LS74 latch. The clocking signal of the latch is the positive zero crossing point of the AC line voltage.

7) PRBD: PRPC Power-on Reset & Brown-out Detector

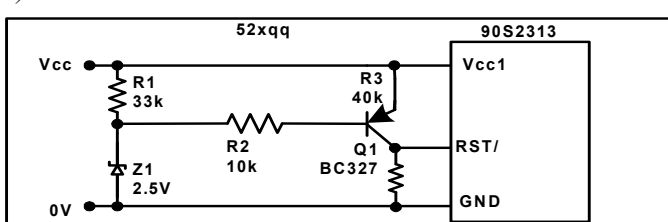


Fig. 3g Power-on Reset and Brown-out Detector for PRPC
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8) PFIF: Power Failure Interface

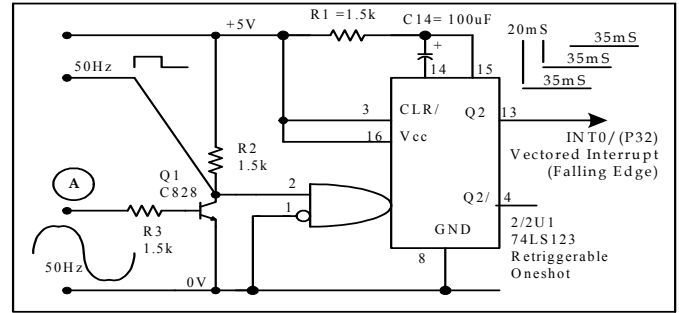


Fig. 3h Zero Crossing and Power Failure Detection Circuit

A 50Hz square wave is available at the collector point of Q1 at every positive zero crossing point of the line frequency. U1 is re-triggered by the 50Hz signal and its output at Q0-pin remains high for 35mS. Since, the triggering signal arrives at every 20mS interval; the output of U1 always remains at Logic-H. When line voltage fails, line frequency disappears and there is no triggering signal. As a result, the output of U1 goes to Logic-L state, which interrupts the MCU. The MCU goes to ISR and saves the balance into EEPROM.

10) RPC: Prepaid Card

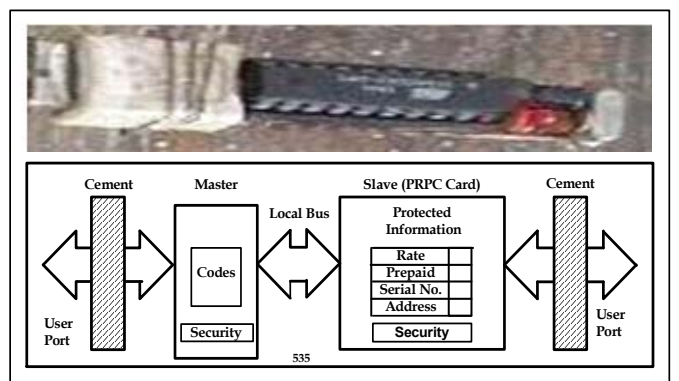


Fig. 3i Pictorial View and Block Diagram Prepaid Card (PRPC)

Prepaid Card is a one kind of 'Secured System.' The PRPC for our PEM has been designed using 16-pin AT90S2313 RISC Microcontroller. The MCU has security bit, which prevents others to gain access into the code/data of the chip.

The MCU contains information like: 'recharge amount', 'tariff', 'Meter Serial Number', and the address of the meter owner. Once the meter is loaded with the 'recharge amount', the host MCU erases the balance field of the PRPC. The PRPC should be recharged by a proprietary vending machine.

9) PVIF: Prepaid Card V_{CC1} Interface

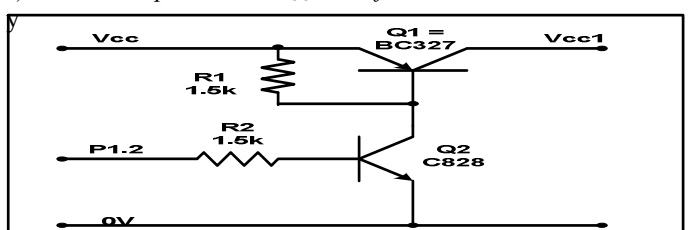


Fig. 3j Prepaid Card V_{CC1} Interface Circuit

As we can see in Fig. 5b, the host MCU activates the +5V into the PRPC once the PRPC has been properly engaged with the meter by affecting a mechanical contact (MCIF, Fig. 2). The circuit of Fig. 3i depicts the electronic control for the +5V (V_{cc1}) supply of the PRPC card.

III. CONTROL PROGRAM FLOW CHART OF THE PREPAID ELECTRICAL ENERGY METER

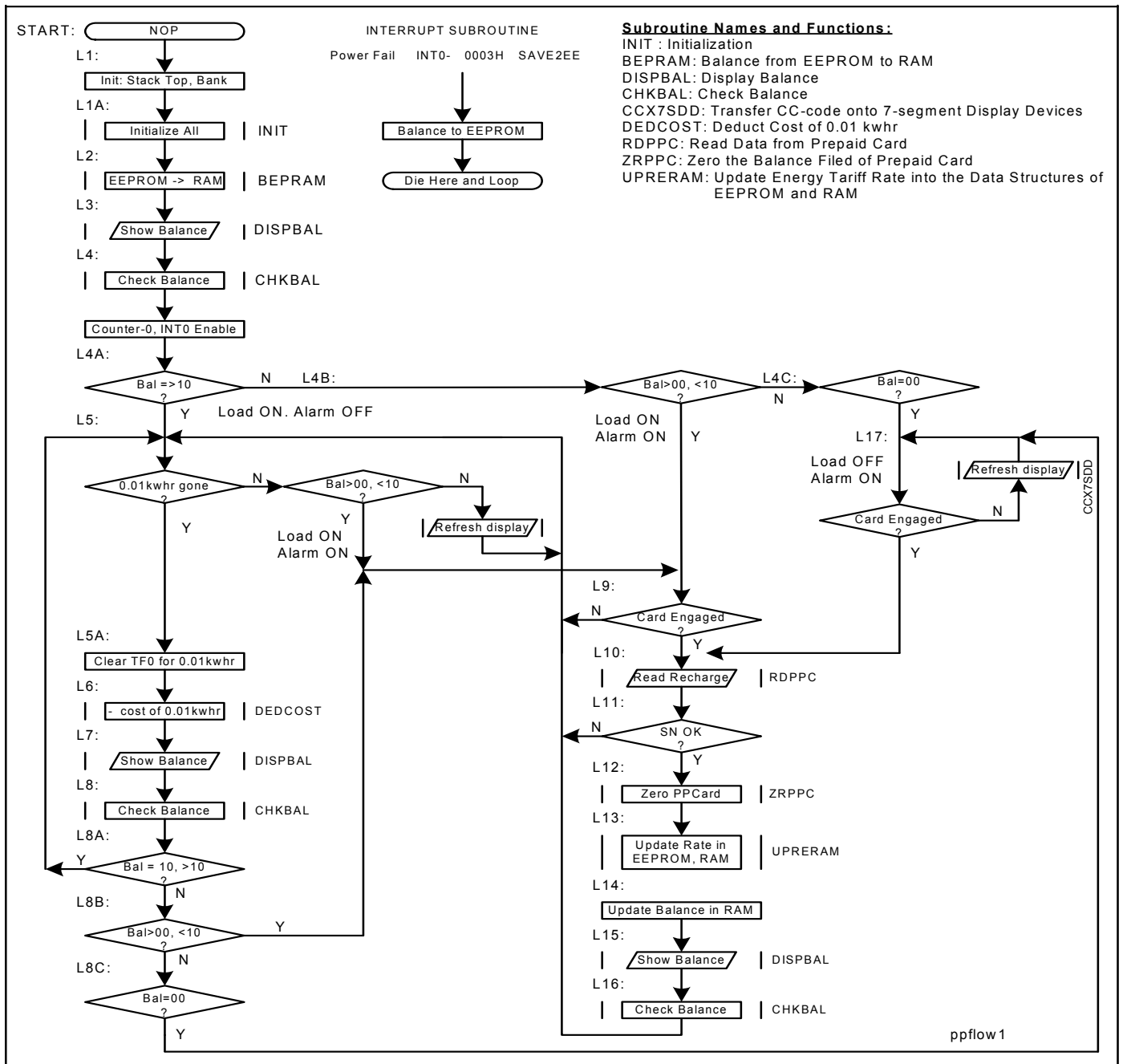


Fig. 4 Flow Chart for the Control Program of Prepaid Electrical Energy Meter

IV. DATA STRUCTURES FOR THE PREPAID ELECTRICAL ENERGY METER

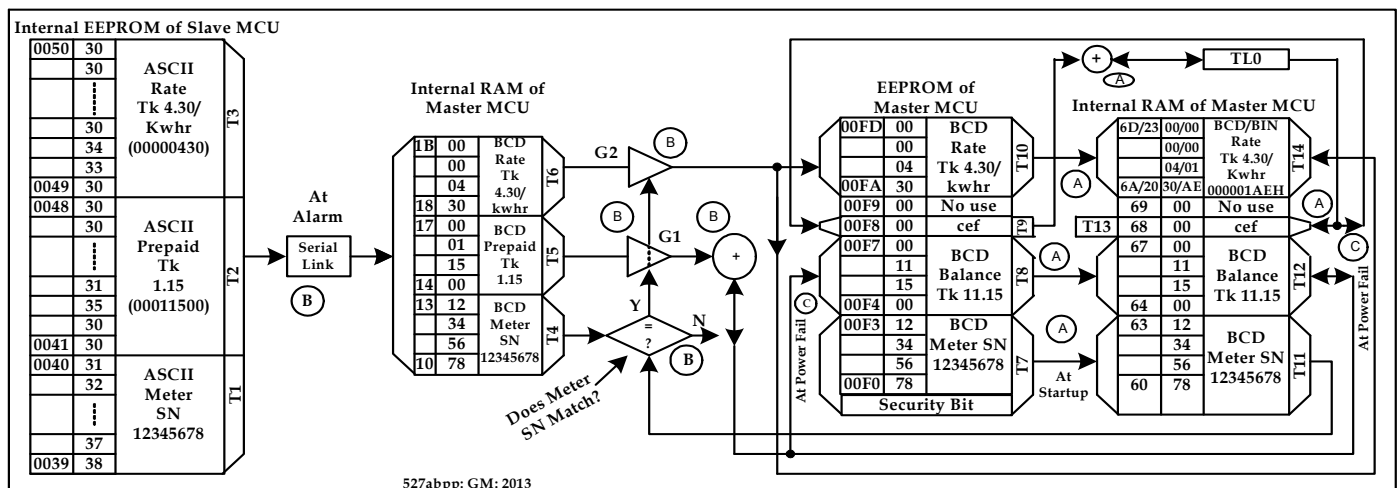


Fig. 5a Data Structure-I for Control Program of Prepaid Electrical Energy Meter

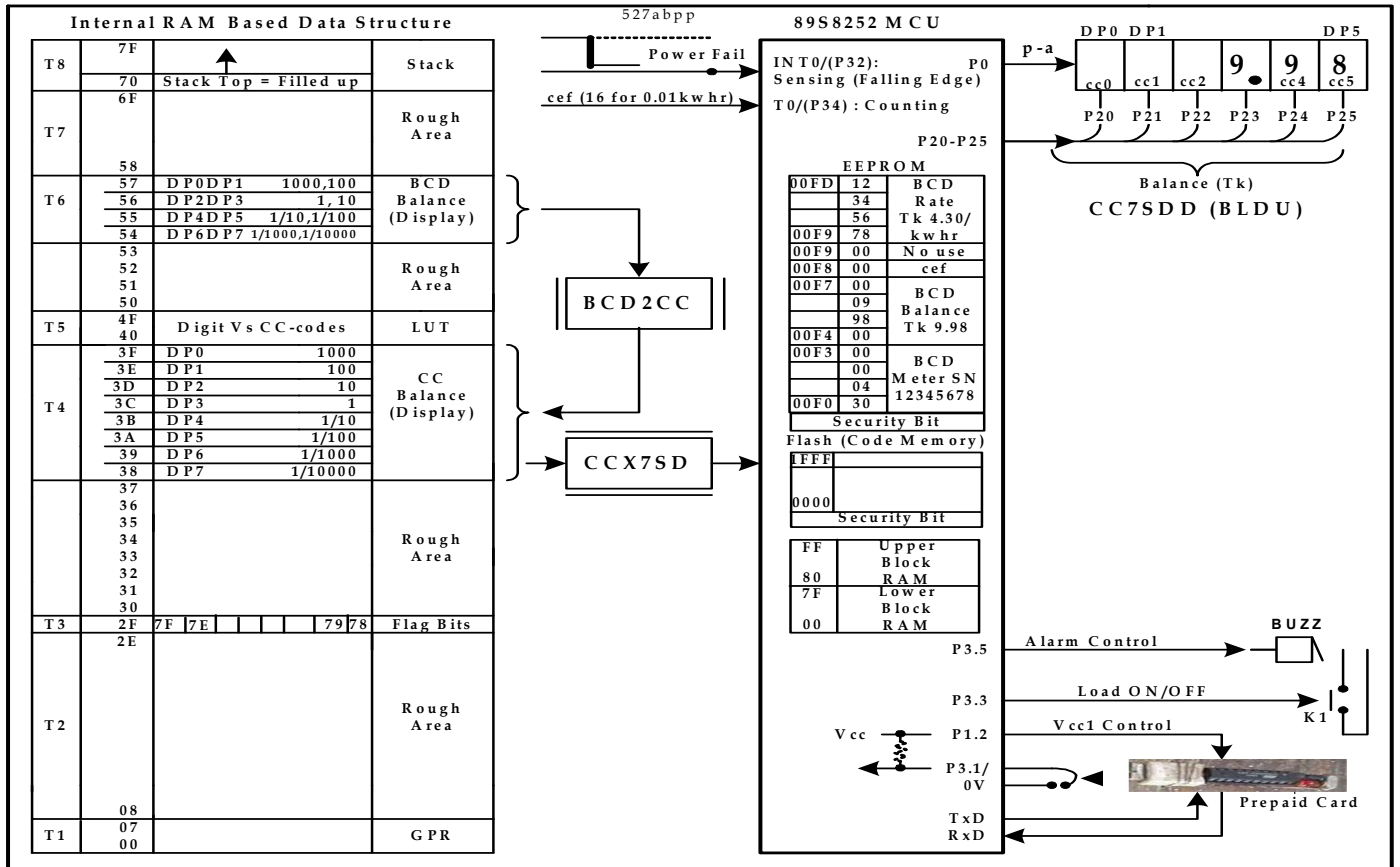


Fig. 5b Data Structure-II for Control Program of the Prepaid Electrical Energy Meter

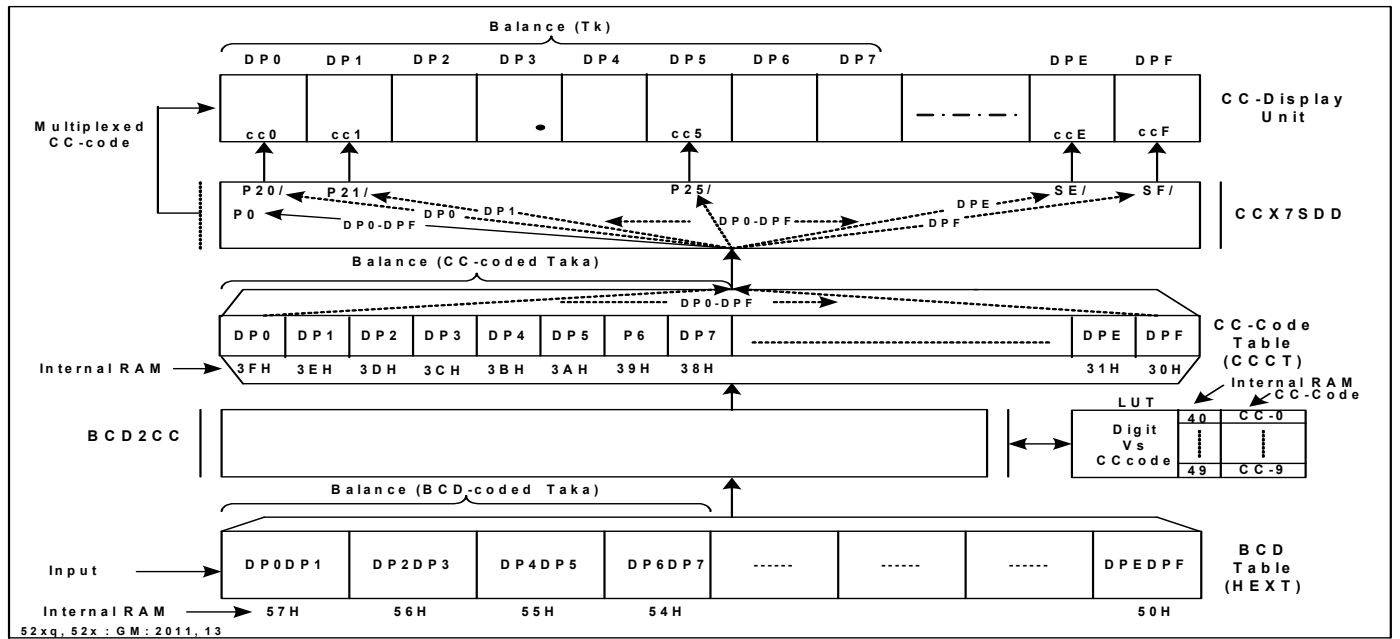


Fig. 5c Data Structure-III for the Control Program of Prepaid Electrical Energy Meter

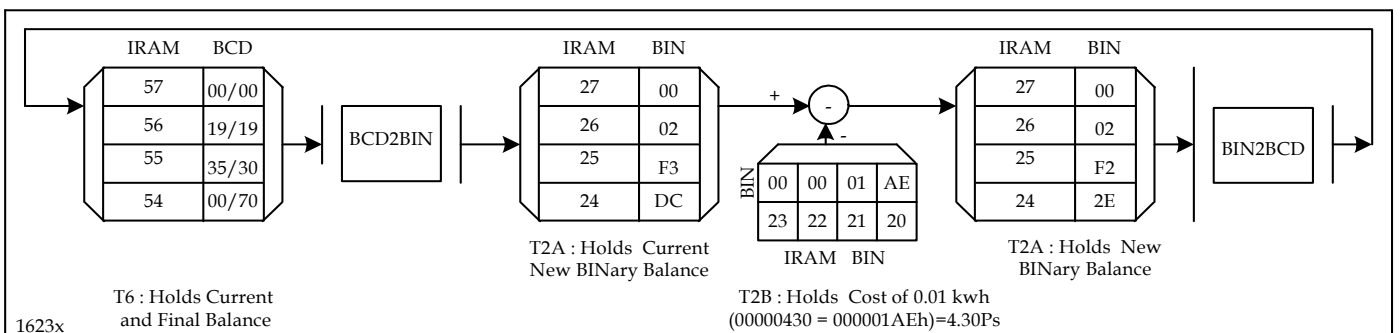


Fig. 5d Data Structure-IV for the Control Program of Prepaid Electrical Energy Meter (for Decimal Adjustment after Subtraction, DAS)

A Data Structure is a map of the input/output variables; it is usually created in the RAM space of the microcontroller. A well designed data structure is a mandatory item for the coding of the Control Program Flow Chart of any MCU based system like the PEM. Since, a Control Program (CP) is primarily a collection of subroutines (Fig. 4); the data structure accurately guides the design engineer to the modular development of the subroutines. A correctly designed data structure always ensures the possibility of a workable CP.

A. Data Structure-I

- The mapping of the internal EEPROM of the PRPC (Fig. 3i) clearly shows the ASCII based 'Rate Field', Prepaid Field', and the 'Meter Serial Number (MSN) Field'. At alarm time, the user engages the PRPC into the host meter; data enters into the RAM area of the master MCU.
- MSN is checked for validity; G1 and G2 are enabled; recharge amount and new tariff rate (if any) enter into EEPROM and RAM of the master MCU.

B. Data Structure-II

- After the consumption of 0.01 kwhr energy, the CP immediately deducts the cost (Ps 4.30) from T6 (Table T6); the new BCD balance is converted to CC-code using the BCD2CC subroutine; the cc-balance of T4 goes to the multiplexed display with the help of CCX7SDD subroutine. The BCD digits are converted to corresponding cc-codes using LUT (Lookup Table) of T5.
- T6 maintains balance up to 4-digit precision. For example, Tk 19.35 is maintained as 00193500. This is to facilitate the subtraction of Ps 4.3 (=Tk .0430), cost of 0.01 kwhr, from T6 by calling the algorithm of Fig. 5d. The display shows the balance with 2-digit precision.
- The master MCU executes the following algorithm to understand that the PRPC has been properly engaged:

```

L1A:  If ((P3.1) !=0)
        Deduct cost of 0.01 kwhr if there is balance
        Goto L1A
L1B:  Activate P1.2 to apply +5V to the PRPC.
L1C:  Activate serial communication link with PRPC

```

C. Data Structure-III

This data structure required to develop the instruction codes for the BCD2CC and CCX7SDD (Fig. 5b) subroutines. This is, in fact, an enhanced version of the Data Structure-II (Fig. 5b) with the inclusion of the scenario of driving the multiplexed display unit using the port lines of the MCU.

D. Data Structure-IV

This is actually an algorithm to perform subtraction on multi-byte BCD numbers. As the 89S52 MCU does not support the DAS (Decimal Adjustment after Subtraction) instruction, we need to apply this algorithm to obtain correct BCD balance after the subtraction of the cost of 0.01 kwhr (Tk 0000.0430) from the current balance (say, Tk 0019.3500). The algorithm works this way: (i) binary value for the cost of 0.01 kwhr (000001AEh) is kept in T2B; (ii) the current BCD balance of T6 is converted to binary and is kept in T2A, (iii) binary subtraction is made between T2A and T2B; binary result goes to T2A, and (iv) Binary to BCD conversion is done on T2A; the correct BCD balance enters back to T6.

IV. DEVELOPMENT OF THE PREPAID ENERGY METER

We have already documented the hardware circuits and the software routines for the PEM. Now, we may organize facilities for developing and testing these objects.

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A. Resource Requirements

- CMCKIT (Fig. 6, [9]): It is an 89S8252 based friendly Trainer Kit fitted with breadboard and edge connector for circuit prototyping. It contains RS232/USB and ISP interfaces to receive program code/data from IBMPC and get them fused inside the Flash and EEPROM of the MCU.

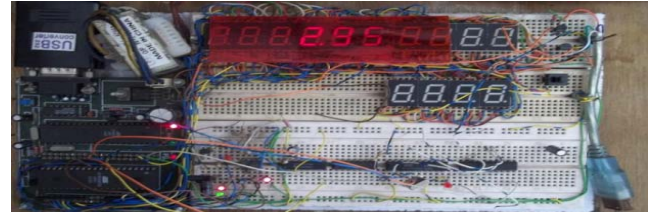


Fig. 6 Pictorial View of CMCKIT

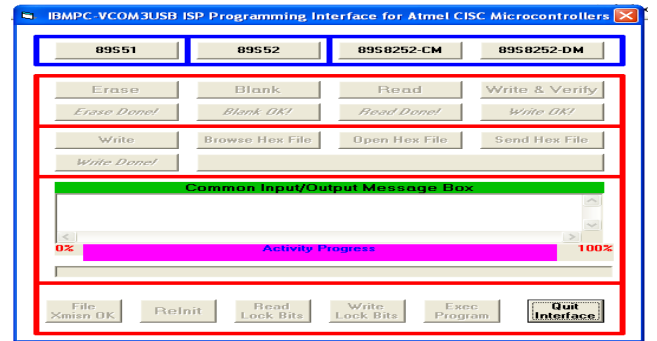


Fig. 7 GUI Interface to Communicate with CMCKIT

- MIDE-51 [10]: Assembler for 89S52 MCU.

B. Development Methodology

- Place 8 CC-type display devices on bread board, and test them by executing the codes of Data Structure-III.
- Build 555-based low frequency oscillator circuit on the bread board to simulate the CF pulses.
- Write ASM codes to test DAS algorithm of Fig. 5d.
- Follow SSS strategy to develop and test the routines and subroutines of Fig. 4 and Fig. 5b.
- The details of PRPC could be obtained from [9].

V. CONCLUSIONS

The PEM has been successfully developed and tested based on methodology that solely belongs to Bangalee Intelligence. It is now exposed to public who may build their own versions with advanced feature like adding interface for charging the meter via internet. The author emphasizes that the following issues are to be addressed for a reliable PEM: (i) at power up reset, the MCU must start, (ii) any nearby electromagnetic noise such as electric welding, thundering, and radio communication must not interfere the meter operation, (iii) EEPROM must not be corrupted during power failure.

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Continuous Wavelet Transform Based Analysis of Low Frequency Oscillation in Power System

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Abstract— Low frequency oscillation is a threat to the stability of interconnected power system. Proper identification of the low frequency oscillation is crucial. In this work continuous wavelet transform (CWT) with complex morlet wavelet is used to extract modal information of the power system oscillations. The modal information of the measured signal such as frequency and damping is determined from the magnitude and phase plot of the complex valued wavelet coefficients. The proposed technique is applied to the measured signal of a two area power system to detect inter area mode of oscillation and results are compared with Eigenvalue analysis method.

Keywords—Wavelet transform; low frequency oscillation; complex morlet wavelet

I. INTRODUCTION

Electricity is the driving force for the development of modern civilization. In the present trend of deregulated electricity market environment the power system are more likely to operate closer to their rated capacity due to the increased competition among the power producers as well as demand for the large block of power from the consumer side [1]. These factors increase the vulnerability of power system to the disturbances. The stability of power system is crucial for supplying electrical power to the consumer without loss of reliability. The power system stability is defined as the capability of the system to come back to normal operating state after having gone through some sort of disturbances [2, 3]. The poorly damped low frequency electromechanical oscillation is one of the major disturbances faced by the synchronous machines of the power system. There are two types of low frequency oscillations in the power system namely local area mode and inter area mode oscillation. Local area mode corresponds to swinging of machines of one particular area with respect to each other whereas machines of one area swings with respect to machines of others area in inter area mode of oscillations [4]. Typical, this oscillatory phenomenon lies in the range of 0.1 to 1 Hz which can cause the system outage due to the instability if effective preventive measures are not taken timely [5]. So, it is of great importance to detect and extract the properties of these oscillations in order to provide adequate damping through control [6].

As found in the literatures, lots of techniques have been employed to determine the properties of low frequency oscillations in the power system. The conventional method of determining oscillation parameters is the eigenvalue analysis based on linearized model of the power system. The linearization of the system is done around certain operating points. Since the operating point of a real power system is varying, this method is not effective outside the predefined operating points [2]. Moreover, it is an offline method for observing the disturbances in the system.

The online measurement based techniques are of interest as it does not require modeling of system and gives information in the near real time operations. These methods identify the system model and operating conditions from the measured signal [7].

Signal processing techniques such as Prony analysis [8] and Fourier spectrum analysis [9] have been used for estimation of low frequency oscillations in the power system. Prony analysis suitable for analysis of stationary (time invariant) signal but the power systems dynamic is time varying and non stationary in nature. Fourier spectrum analysis gives the information only in the frequency domain and corresponding time domain characteristics of the signal are not present.

Kalman filtering technique has been used to detect the temporal variation of mode parameters of oscillations [10]. But, the implementation of Kalman estimator requires prior knowledge of the system transfer function which may not be available in real time.

The wavelet transform is a popular signal processing tool for analysis of non-stationary signals. It can extract signal features both in time and frequency domain. The wavelet transform has been applied in the analysis of various power system phenomena like fault analysis [11], detection and identification of power quality problems [12] and detection and classification of power system transients [13].

This work proposes the application of complex morlet wavelet based continuous wavelet transform (CWT) to detect and extract the frequency and damping of low frequency electromechanical oscillations in a two area four machine power system.

II. CONTINUOUS WAVELET TRANSFORM

Continuous wavelet transform of any signal $x(t)$ can be expressed mathematically by the following equation [7]

$$W(a,b) = \frac{1}{\sqrt{a}} \int_{-\infty}^{\infty} x(t) \psi^*\left(\frac{t-b}{a}\right) dt \quad (1)$$

Here, $W(a,b)$ and $\psi(t)$ are the wavelet coefficients and the mother wavelet function respectively. The parameters a and b accounts for the scaling factor and translation factor of the mother wavelet respectively. Scaling and shifting of the mother wavelet produces son wavelets. The scaling factor controls the frequency of the son wavelets; higher the scales lower is the frequency and vice versa. Wavelet coefficients are calculated from the convolution of son wavelets and the signal. In order to preserve energy at each scale the convolution sum is multiplied by the factor $\frac{1}{\sqrt{a}}$. The wavelet transform (WT) represents the signal in both time and frequency domain. The capability of localization of time and frequency with proper resolution makes the wavelet transform suitable for the analysis of signal having non-stationary characteristics.

There are different types of wavelet function such as Gabor, Morlet, Mexican hat and Haar. It is found that the complex morlet wavelet provide good results for the analysis of ring down signal as it possesses the good time and frequency localization property to uncover the oscillatory modes of the signal [14]. Mathematically, the complex morlet wavelet is given by the following expression

$$\psi(t) = \frac{1}{\sqrt{\pi f_b}} e^{j2\pi f_c t} e^{-\frac{t^2}{f_b}} \quad (2)$$

Where, f_b is the bandwidth parameter of the mother wavelet, f_c is the center frequency, t is the time, and j is the imaginary unit. The bandwidth parameter controls the shape of the wavelet.

The fourier transform of the mother wavelet function is given by following equation

$$\psi(af) = \frac{1}{\sqrt{\pi f_b}} e^{-\pi^2 f_b (af - f_c)^2} \quad (3)$$

The value of the $\psi(af)$ will reach to the peak when $af=f_c$. The value of localized frequency component of the signal can be retrieved as follows

$$f = \frac{f_c}{a\Delta t} \quad (4)$$

Where, Δt is the sampling period of the signal.

III. PROPOSED METHOD

The electromechanical oscillations of power system can be expressed as sum of damped sinusoids as given below.

$$x(t) = \sum_{k=1}^n x_k(t) = \sum_{k=1}^n A_k e^{-2\pi f_k \zeta_k t} \cos(2\pi f_{dk} t + \theta_k) \quad (5)$$

Where, $x(t)$ is the signal of n damped sinusoids, A_k , ζ_k , f_{dk} and f_k are the amplitude, damping ratio, damped natural frequency and undamped natural frequency of k th mode, respectively. The continuous wavelet transform of the signal $x(t)$ using complex morlet wavelet can be implemented by employing equations (1)-(3). The coefficients of wavelet transform obtained as follows

$$W(a,b) = \frac{\sqrt{a}}{2} \sum_{k=1}^n A_k e^{-2\pi f_k \zeta_k b} e^{-\pi^2 f_b (af - f_c)^2} e^{j(2\pi f_{dk} b + \theta_k)} \quad (6)$$

The wavelet coefficients are complex number and can represent the multimode signal as a single mode signal [15]. The value of wavelet coefficient will be maximum when $a_k f_k = f_c$ occurs for a certain value of scaling parameter a_k and this scale a_k of wavelet coefficients contains the modal information of k th mode of signal. So, k th mode of the signal with replacing b with t can be written in the amplitude and phase angle form as

$$W(a_k, t) = \frac{\sqrt{a_k}}{2} \sum_{k=1}^n A_k e^{-2\pi f_k \zeta_k t} e^{j(2\pi f_{dk} t + \theta_k)} = |W(a_k, t)| e^{j\phi_k(t)} \quad (7)$$

$$|W(a_k, t)| = \frac{\sqrt{a_k}}{2} \sum_{k=1}^n A_k e^{-2\pi f_k \zeta_k t} \quad (8)$$

$$\phi_k(t) = \arg[W(a_k, t)] = 2\pi f_{dk} t + \theta_k = 2\pi f_k \sqrt{1 - \zeta_k^2} t + \theta_k \quad (9)$$

Taking logarithm of the modulus of instantaneous amplitude of wavelet coefficient presented above in equation (7), we get,

$$\ln|W(a_k, t)| = -2\pi f_k \zeta_k t + \ln\left(\frac{\sqrt{a_k}}{2} A_k\right) \quad (10)$$

Therefore, natural frequency of oscillation f_k and damping ratio ζ_k can be obtained by calculating the slope of $\ln|W(a_k, t)|$ and $\arg[W(a_k, t)]$ of equations (9)-(10). The slope can be determined using least square regression analysis. The small time range in which amplitude and phase plot becomes linear contains the modal information of the signal.

IV. POWER SYSTEM MODEL

The well known four machines two area power system shown in fig. 1 has been considered to simulate the low frequency oscillation phenomena. The power system model and associated data can be found in [2]. All the generators have been modeled using sub-transients model, static exciter and simple turbine governor system. In order to simulate the inter area mode of oscillation 5% step increase in the exciter reference voltage of generator 2 (area I) and generator 3 (area II) has been applied. The Power System Toolbox (PST) functions have been used for time domain simulation of 20 seconds. The PST is an open source matlab based tool box for load flow and stability analysis of the power system. For this work Power System Toolbox Version 3.0 has been used.

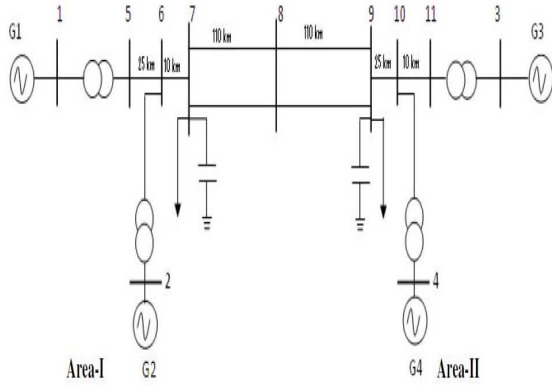


Fig. 1. Four machine two area power system

V. SIMULATION RESULTS

A. Eigen Value Analysis

The frequency and damping information of oscillatory Eigen mode have been calculated and plotted in the fig. 2, the green marker in the figure corresponds to minimum damping.

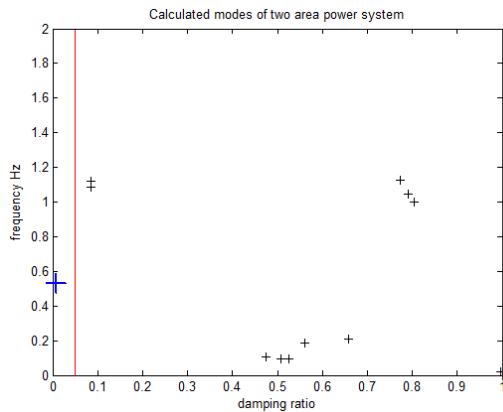


Fig. 2. Modes of the power system

The Eigen mode of 21 exhibits least value of damping ratio; therefore, it is the dominant mode of oscillation as shown in Table I.

TABLE I

For the Eigen Mode 21		
Eigen value	Frequency (Hz)	Damping ratio (%)
-0.0232-3.3212i	0.5286	0.70

B. CWT Based Analysis

The active power flow between the two area measured at the line between bus 6 and 7 is shown in the fig.3. The proposed scheme of continuous wavelet transform has been applied to extract the information of the measured signal. The center frequency (f_c) and bandwidth parameter (f_b) are taken as

0.55 Hz and 15Hz respectively with sampling time 0.01 second. The contour plot of the measured power flow signal is shown in fig. 4. From the figure it is seen that most of energy of the wavelet coefficients lies at the starting and end of signal around scale 107. So, there is only one dominant mode at scale 107 and signal is said to be stationary in this particular case. The linear amplitude and phase plot of wavelet coefficients are shown in fig. 5, and fig. 6, respectively.

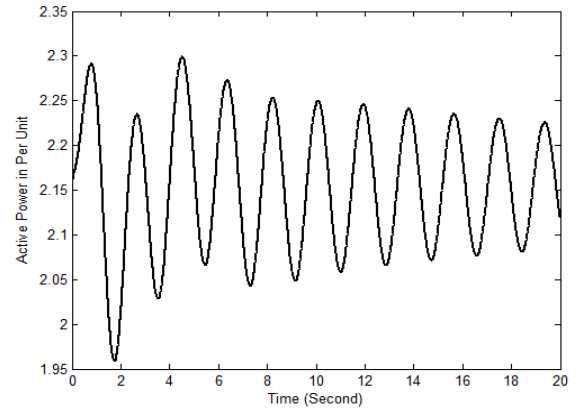


Fig. 3. Active power flow

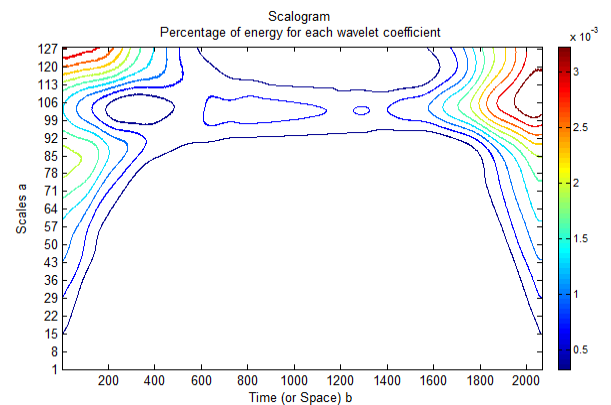


Fig. 4. Contour plot of the energy of wavelet coefficients of the measured signal

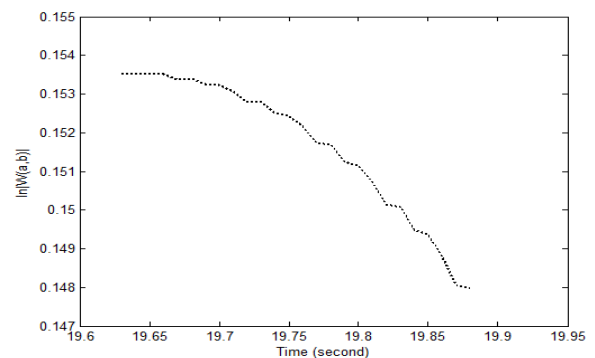


Fig. 5. Amplitude plot of the wavelet transform of the measured signal

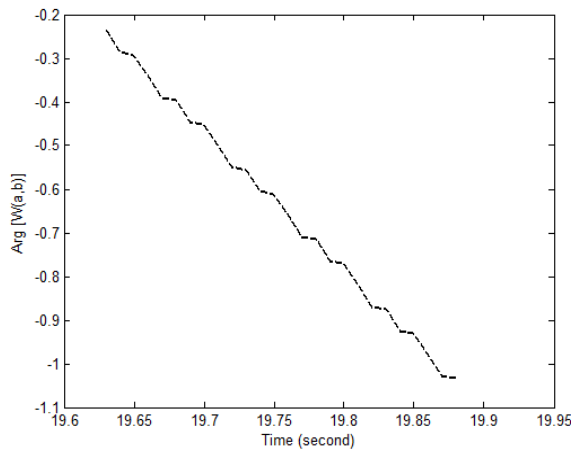


Fig. 6. Phase plot of the wavelet transform of the measured signal

The frequency and damping information obtained from the proposed technique are presented in Table II along with the result obtained from Eigen analysis.

TABLE II

For the Single Mode Signal			
Method	Frequency (Hz)	Real part	Damping ratio (%)
Eigen Analysis	0.5286	-0.0232	0.70
Wavelet Transform	0.5088	-0.0226	0.7082

The frequency and damping information obtained using continuous wavelet transform based technique are very close to the results of Eigen value analysis method. This proves the efficacy of proposed method in extracting the modal information of low frequency oscillations in power system.

VI. CONCLUSION

In this work, continuous wavelet transform has been applied to find the modal information of the power system low frequency oscillations. The proposed method has been presented mathematically. The results obtained from the proposed method closely matches with the values obtained from the Eigen value analysis method. This demonstrates the effectiveness of the continuous wavelet transform based technique in identifying the low frequency oscillations in power system.

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Problem Analysis and Probable Solution of Existing Interconnected Grid System of Chittagong Region

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Abstract—Power crisis is one of the foremost concerning problem in Bangladesh. The gap between demand and production is increasing day by day. System loss, transmission loss and established connection to grid without upcoming load plan are main issues regarding this crisis. It is possible to control this crisis by taking some necessary steps such as system loss reduction, establishing efficient interconnection, assembling necessary substation etc. Priority should be given to transmission and distribution of electrical energy through the proper way. This paper represents a case study and analysis to find out the key problems of existing power system including generating station, interconnected grid system and substations of Chittagong. Existing generating station with their generating capacity, transmission lines with their length as well as conductor grade and substation with its load are studied to figure out the major problems and finally it offers some probable solutions of these problems.

Keywords— *Interconnection, power scenario, Power Grid, Chittagong.*

I. INTRODUCTION

During the early years small local generating Stations supplied power to respective local loads. Each generating station needed enough installed capacity to feed the local peak loads. Problem arises when load becomes greater than the generating capacity. Interconnection system was considered as a solution for this problem. The interconnected system is very efficient and very reliable in transmission and distribution of electric power if the modeling and designing is properly done. [1]

Energy is the fundamental resource for economic development of a country. In this modern world, the dependence on electricity is greater so that it has become a part of our life. The over-increasing uses of electric power for domestic, commercial and Industrial purposes necessities to provide bulk electrical power economically. This achieves with the help of suitable power producing units known as power plants or electric power generating station. [2]

As an underdeveloped country, the socio-economic structure of Bangladesh is gradually increasing. So the demand of power is extending day by day hence the importance of Transmission and Distribution are becoming more complicated.

Each and every year power from new power plants is added to the national grid. Many power plants are still under construction. But the total power is transmitted and distributed through existing infrastructure with some old rating transformer, conductors and other equipments. The overloading of lines and transformers causes system loss, transmission and conduction loss take place which in turn causes load shedding or power shortage. [3][4]

Though this paper has analyzed the key problems of existing power system in Chittagong region, we have also observed that almost same conditions are found for the every important city of this country.

II. POWER SECTORS SCENARIO OF CHITTAGONG

In 1960, the first power plant of Chittagong (as well as Bangladesh) was built at Siddhirganj with a capacity of only 10 MW. At that time Kaptai dam was under construction. The unit size of Kaptai was 40 MW, which for that time was considered to be a large power plant. Necessary steps were taken to construct the side by side 132 KV Dhaka-Chittagong transmission line. Construction of Kaptai dam and commissioning of Dhaka-Chittagong 132 KV transmission took place in 1962 which was a very important milestone of power development of Bangladesh.[5]

Chittagong has become very important city for its port and a large number of industries from small size to heavy. Around 40% of the heavy industrial activities of the country is located in this city and neighboring areas, which include dry dock, Dock Yards, Oil Refinery, Steel Mills, Power Plant, Cement clinker factory, automobile industries, pharmaceutical plants, chemical plants, cable manufacturing, textiles mills, jute mills, urea fertilizer factory along with other private sector medium size industrial developments and activities. So electricity demand is increasing day by day in Chittagong.

Presently, there are ten private and public power stations in Chittagong. Among them five power plants run by gas, another four power plants run by diesel and one is hydro-electric power plant. Karnafuli hydro-electric power plant is the only hydro-electric power plant of the country which produces 230 MW power.

Table-01 shows the list of power plant and total power production in Chittagong.

Table -01: List of Power plant in Chittagong

Name of Power Plant	Generation Capacity (MW)
Karnafuli hydro-electric power plant	230
Raojan gas power plant	210
Shikalbaha steam power plant	60
Shikalbaha diesel power plant	150
Shikalbaha peaking power station	55
Regent gas power plant	22
Malancho diesel power plant	44
Dohazari diesel power plant	100
Hathazari gas power plant	100
Julda diesel power plant	100
Total generation Capacity	1071

As a royal city, more and more industries are being established in Chittagong. So electricity demand is increasing at a decreasing rate. **Table-02** shows the demand statistics of power in Chittagong. The total generation of Chittagong including all public and private power plants is about 1071 MW in 2013 and the total load is 1150 MW.

Table -02: Demand Statistics of Chittagong.

Year	Load(MW)
2005	460
2006	512
2007	563
2008	625
2009	687
2010	755
2011	945
2012	1035
2013	1150

III. EXISTING POWER SYSTEM IN CHITTAGONG

In Chittagong grid system, mainly two types of transmission line are used according to voltage level. They are 230KV and 132KV AC transmission line. The total length of

230 KV transmission line is 175 Km (Double circuit) and the total length of 132 KV transmission line is 630 Km ACSR (Aluminum conductor steel reinforced) conductor having Capacity of 600A for both case.

From the analysis of existing interconnection system of Chittagong, it is found that two 230 KV transmission lines are used to connect 230/132 KV grid substation with Raojan and Comilla. Other substations are interconnected through 132 KV transmission line. Almost all sub-stations use double circuit line.

According to transmission voltage level there are two types of substation in Chittagong. They are 230/132 KV and 132/33 KV sub-station. A substation of 230/132 KV is placed in Hathazari having power capacity of 600 MVA. There are sixteen 132/33 KV grid sub-station with a power capacity of 1502 MVA which are situated at Kaptai, Chandraghona, Hathazari, Madunaghat, Sikalbaha, Dohazari, Cox's bazaar, Haliashahar, Khulshi, Baraulia, Julda, Bakalia, Shamirpur, T.K Chemical Complex, Modern Steel A.K Steel in Chittagong .

IV. PROBLEMS AND PROBABLE SOLUTIONS FOR EXISTING GRID INTERCONNECTED SYSTEM IN CHITTAGONG

A. Incompetent Grid Connection

Dhaka-Chittagong national grid system is interconnected by Hathazari grid substation with Hathazari-Comilla by 230KV transmission line and Hathazari-Feni by 132KV transmission line (Both line are double circuited). In case of any major fault (such as Busbar or line protection failure) at Hathazari grid substation, the total interconnected grid system of Chittagong will be disconnected from the national grid. [6]

Solution:

This problem will be effectively solved if a 230 KV transmission line is established from Raojan to Madunaghat and Madunaghat- to Sikalbaha substation. Because the Raojan 420MW power station is linked only with Hathazari Sub-station.

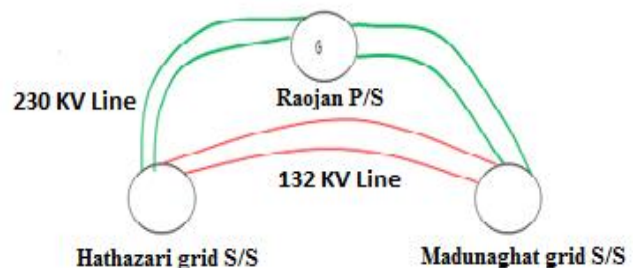


Figure 1: Proposed Dhaka-Chittagong Grid interconnection.

B. Conductor overload

The double circuit transmission lines from Hathazari to Madunaghat and from Hathazari to Baraulia use ACSR (Aluminum conductor steel reinforced) conductor having Capacity 600A. Both lines are so old and they can't carry the full load. Hence, overload is occurred in these lines which in turn cause load shedding. If the Generation of Chittagong fails due to shortage of gas and water, then all loads will come from Dhaka through Hathazari grid Sub-Station and both lines will be overloaded. [7]

Solution:

To solve this problem first we have to increase the power generation of Chittagong. And the line conductor which is used for the power transmission should be changed by AAAC (All Aluminium Alloy Conductor) or Finch conductor. If the weight of AAAC or Finch conductor do not support by existing tower then a new 132 KV parallel transmission line have to be installed.

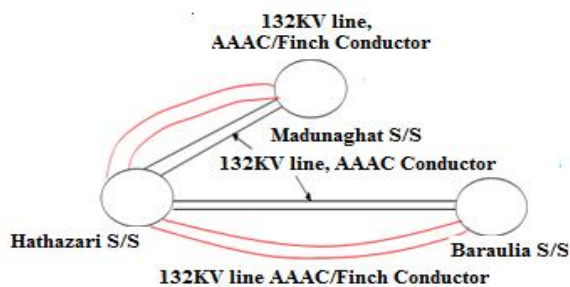


Figure 2: solution for Conductor overload

C. Transformer Overload

A 40 MVA transformer is used in Madunaghat substation to feed the load, where the total load of this substation is 33 MW to 40 MW. So this old existing transformer cannot full fill the demand properly and the substation is not reliable with this single transformer. If the overload or any major fault occurs in this transformer the people under Madunaghat substation will suffer by load shedding. [8]

Solution:

This problem can be solved by installing another transformer of 40 MVA parallel with the existing one. It is very essential to enhance the reliability of Madunaghat grid substation.

D. Substation Overload

The present capacity of Khulshi grid substation is 126MVA and the maximum load capacity is 110MW where the total load on this substation is more than 155 MW. If one

transformer trips by any cause then another one will trip due to over current. On the other hand Khulshi grid substation has not enough space to install another transformer and expansion the 132KV Bus bar. So regularly Khulshi substation faces overload problem.

Solution:

A 132KV substation can be installed at Agrabad area. As Khulshi-Halishahar 132KV transmission lines pass through Agrabad area, it eliminates the installation of more 132KV transmission line. Hence the solution is reasonable.

E. Transmission loss

The load of hill area (i.e Rangamati, Khagrachhari, Bandarban) of greater Chittagong is supplied from the Hathazari and Chandroghona grid substation by 33KV line. Due to the low voltage level and long line length this line has high line loss and voltage drop. So the consumer of this area cannot operate electric load properly due to low voltage problem. [9]

Solution

132/33KV grid substation should be installed at Rangamati, Khagrachhari and Bandarban area. This substation will be able to draw power from Chandraghona grid substation. 132KV supply from Kaptai substation will be economic. But Kaptai substation is operated by BPDB and there is no facility to expand the existing 132KV bus bar.

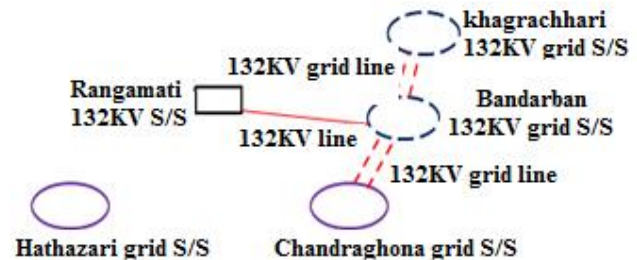


Figure 3: Transmission loss overcome.

V. PROPOSED INTERCONNECTION SYSTEM FOR CHITTAGONG

The best solution for Chittagong is to increase power generation. The power generation of Chittagong should be increased up to 1600MW in 2020 and new power station should be installed except Gas. Because the existing 420MW thermal power station and Sikalbaha (60+150=210) MW thermal power station cannot continue due to shortage of gas [10]. In this situation it will great be a mistake to install gas dependent power station. If we install heavy farness oil power station then per unit generation cost will be increased.

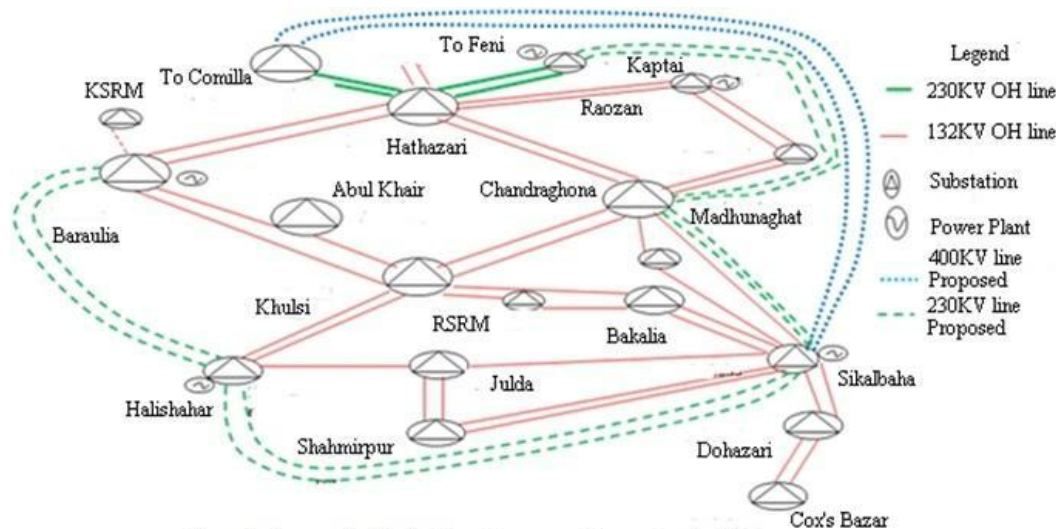


Figure 4: Proposed with Existing interconnection system in Chittagong

Therefore, Chittagong region should think about nuclear or coal power station. If we are able to install that type of power station in Chittagong, the power generation & transmission problem will be solved and then load shedding in Chittagong will be negligible.

For better performance and reliability, 400 KV double circuit grid interconnection lines from Sikalbaha substation to Comilla (north) substation and another 230KV grid ring should be established in Chittagong. Also, it is the best time to give our attention on Renewable energy.

Figure-4 shows the proposed interconnection system for Chittagong which solves the major problems of existing interconnection system.

VI. CONCLUSION

This paper shows an analysis of problems and probable solutions for existing interconnection system of Chittagong. It requires infrastructural changes for higher efficiency, reliability and performance. Finally a complete power system interconnection is proposed to meet the demands and can be considered as a milestone for this region.

As a part of future work, the problems and all probable solutions will be analyzed for the total interconnected system of Bangladesh. Finally, a unique power system design will be proposed for the whole country with details cost analysis.

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Loss Minimization in Rural Distribution Systems from Double Digit to Single Digit

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Abstract—Due to the expansion of rural distribution systems (RDSs), distribution loss is increasing day by day. This paper presents a novel approach to minimize the distribution loss of RDSs from double digit to single digit. To minimize the system loss, a simplex algorithm for linear programming is used. The proposed method is tested on a RDS and the test results indicate that it is able to determine the appropriate conductor size for reducing system loss of distribution systems to the intended value.

Keywords—minimization; power loss; rural distribution system (RDS); simplex method.

I. INTRODUCTION

In Bangladesh, several electricity distribution utilities including Bangladesh Rural Electrification Board (BREB) distribute power to consumers. BREB started its journey before 38 years ago with the mandate to electrify rural Bangladesh. At present, after a long walk, BREB, being the largest of all utilities, rendering its service through 2,83,079 Km of distribution line [1]. This rural system has 678 different capacity 33/11 kV substations with the distribution capacity of 6900 MVA to 1.22 crores of different category consumers. After electrifying 54007 villages, it is now consuming 1880 GWhr energy per month and delivering 1610 GWhr to the consumers. Therefore, the distribution loss is about 270 GWhr per month, which is 12.92% of total consumed energy. This distribution loss includes technical loss as well as commercial loss [1], [2].

The electric distribution power system is a complex interconnection of over-head lines, transformers, voltage regulators, and capacitors for the distribution of electric power over long distances from the step down sub-stations to the consumers. It is the medium through which electric power is distributed from the sub-station to the various consumers [3]. It plays very significant role in the power system as it works as a link to consumers for supplying power. If the system consists of improper size conductor, it will contribute more energy loss to the system, and so more financial loss [4], which leads to poor quality of electric power to the end users [3],[4].

From the inception, BREB includes the standard to construct distribution lines with #3ACSR (swallow) (aluminium conductor steel reinforced), #1/0 ACSR (raven) considering lower load of rural area and lower cost for rural

people. #3, #1/0 ACSR conductor has lower ampacity than DOG, WASP conductor, which are used in the other electricity distribution utilities in Bangladesh. If we start with a few span of distribution line of lower ampacity, due to ACRE (Area Coverage Rural Electrification) [1], until renovated to higher size, it will have to be expanded and therefore, it may become a long line of lower size conductor which ultimately creates a system of higher distribution loss.

As days are passing, consumers are connected to the existing distribution line, which is increasing line current, thereby, increasing system loss. In order to reduce distribution loss, there are lots of factors to be considered [5]-[8]. However, how to reduce system loss to a single digit effectively is a major question to solve. In this paper, the size of conducting medium through which power flows and distributes over the whole system is taken into consideration to reduce the loss. For this purpose, the Microsoft (MS) Excel solver is used. The proposed approach finds out an effective solution for reconductoring the distribution system.

II. DISTRIBUTION LOSSES

Since all the energy supplied to a distribution feeder does not reach the end users, a large amount of energy is lost in the distribution system which should be reduced by eliminating or minimizing the causes of losses [9], [10].

The distribution loss refers to the amount of energy injected through a feeder which is not paid for by the users. It is the difference between the energy injected or input and the units billed to the ultimate consumers or energy delivered to the users, which is generally expressed as percentage of units injected. If P_I is the power injected and P_D is the power delivered to the connected users; then power loss P_{loss} will be

$$P_{loss} = P_I - P_D$$

It is calculated for each month and a period of one financial year, which is expressed as year to date value.

Hence, Distribution losses (%) = [(Energy injected – Energy delivered) X 100]/ Energy injected {For a month or a financial year}[3].

For ideal condition, distribution losses should be 3 to 6%. In advanced countries, the range of distribution losses is 6 to

10% [3], but in developing countries, it is more than that and ranges even up to 25-30%. For utilities working in Bangladesh, in recent years, total distribution losses are given in Table I [11] from which it can be observed that BREB has the highest loss compared to other utilities.

TABLE I. DISTRIBUTION LOSSES OF DIFFERENT UTILITIES

Organization	Distribution loss (%) in 2014
Bangladesh Rural Electrification Board (BREB)	13.72
Bangladesh Power Development Board (BPDB)	11.89
Dhaka Electric Supply Company (DESCO)	8.43
Dhaka Power Distribution Company (DPDC)	8.99
West Zone Power Distribution Company (WZPDCO)	10.97

A. Components of Losses:

Distribution losses generally have two components: (1) Technical losses and (2) Commercial or Non-technical losses.

1) *Technical losses*: It is inherent to a distribution system and occurs naturally in the equipment installed in the system. In RDS, this loss is due to high tension (HT) (11/6.35 kV) lines, low tension (LT) (0.415/0.23 kV) lines, distribution transformers, service drops, etc. [3].

2) *Commercial losses*: Commercial losses are caused by non-technical or commercial factors such as theft, defective meters, errors in meter reading, right of way (ROW) clearing and errors in accounting and record-keeping [3].

B. Formulation of Technical Losses:

The technical power loss, P_{loss} (KWh) is equal to the total line resistance (R_T) of the conductor multiplied by the square of the current (I_L) flows through that portion of line in Amps and loss factor with total hour of a specified month [2].

So, power loss is

$$P_{\text{loss}}/\text{Month} = (I_L^2 \times R \times \text{LsF} \times \text{Total Hour})/1000 \text{ kWh}$$

where, R = No. of phase $\times r \times$ line length in Km

r = Resistance per phase per Km of line.

LsF = Loss factor

$$= 0.85 * (\text{Load factor})^2 + 0.15 * (\text{Load factor}).$$

Load Factor is the ratio of average power consumed to maximum power delivered at a certain period. The value of load factor for RDS is 45% to 55%.

III. WHY EXCEL SOLVER AS LOSS MINIMIZER

At present, monthly turnover of BREB is about 900 crores. This indicates 1% loss costs 9 crores of Taka per month. Therefore, it is the great time to think about the national interest through money saving by reducing the distribution losses. Therefore, determining accurate cost of losses, in kWh, is essential in the economic evaluation of all loss reduction projects. In this paper, Microsoft (MS) Excel solver, which incorporates simplex algorithm for linear programming, is used to minimize loss which, in turn, reduces cost.

The MS Excel Solver has become the most widely used general-purpose optimization modeling system, since its introduction in February 1991 [12], [13]. It can easily solve the real life problems like profit maximization, cost optimization of industry, location optimization in warehouse, etc. The aforementioned reasons are the motivation to the writers of this paper to use the Excel solver to determine the proper size of conductor for loss minimization. In Excel solver, for linear problems, John Watson and Dan Fylstra, of Frontline Systems, Inc. [12] implemented the simplex algorithm, which is developed in 1947 by an American mathematical scientist named George Bernard Dantzig. The flowchart of the simplex method is shown in Fig.1. After initialization, it performs optimality test through iteration.

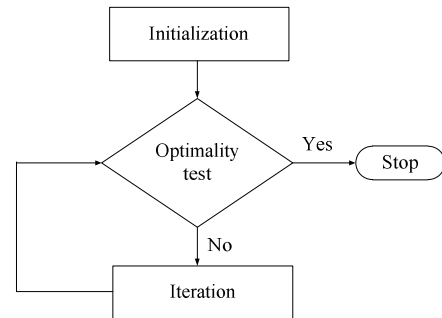


Fig. 1. Flowchart of the simplex algorithm.

The simplex method is an algebraic method, whose underlying concepts are from geometry and works with the same logic that of graphical solution. In the simplex algorithm technique, a series of solutions in tabular form, called tableaux is created. It is clear by inspecting the bottom row of each tableau, whether it represents the optimal solution or not. Standard form of linear problem with several variables that can be solved by simplex method is [14],

Maximize (or minimize) $z = c_1x_1 + \dots + c_nx_n$ subject to $a_{i1}x_1 + \dots + a_{in}x_n \leq (\text{or } \geq \text{or } =) b_i, i = 1, \dots, m, x_1 \geq 0, \dots, x_n \geq 0$

The function, z is called the objective function, constraints $x_1 \geq 0, \dots, x_n \geq 0$ are called non-negativity constraints, a, b and c are known constants.

So, from the standard form, it is obvious that an optimization model has three parts: the target cell (objective function), the changing cells (variables), and the constraints. We want to either minimize, maximize or equalize the amount in the target cell considering the condition of constraints.

In the Excel solver, the optimization model is defined by algebraic formulas which appear in spreadsheet cells. Excel's formula language can express a wide range of mathematical relationships which provides both interactive and user-programmable ways to specify spreadsheet cells to serve the desired roles [13].

The solver perform linearity test after it had obtained a solution using the simplex method. It uses these solution values x^* and the initial values x^0 for the variables, to check

Therefore, $P_{\text{loss}}/\text{Month} = (I_{\text{TL}}^2 \times R_T \times LsF \times \text{Total Hour})/1000 \text{ kWh}$

Here, $R_T = 1 \times 1.3946 \times 2 \text{ Ohm} = 2.7892 \text{ Ohm}$

$P_{\text{loss}}/\text{Month} = [(30)^2 \times 2.7892 \times 0.2875 \times 720]/1000 \text{ kWh}$
 $= 519.63 \text{ kWh}$

Calculation for other sections and taps are summarized in Fig.3.

	A	B	C	D	E
1	Microsoft Excel 12.0 Answer Report				
2	Worksheet: [calculation of loss.xlsx]single phase (original) (2)				
3	Report Created: 28-08-2015 6:31:09 PM				
4					
5					
6	Target Cell (Value Of)				
7	Cell	Name	Original Value	Final Value	
8	\$E\$1	Loss, % Loss per month, kwh	12.63	9.50	
9					
10					
11	Adjustable Cells				
12	Cell	Name	Original Value	Final Value	
13	\$E\$7	Resistance	1.3946	1.053934752	
14	\$E\$13	Resistance	1.3946	0.3	
15	\$E\$15	Resistance	1.3946	0.3	
16	\$E\$17	Resistance	1.3946	0.364087666	
17	\$E\$19	Resistance	1.3946	1.053934752	
18	\$E\$27	Resistance	1.3946	0.3	
19	\$E\$29	Resistance	1.3946	0.3	
20	\$E\$34	Resistance	1.3946	0.713269504	
21	\$E\$40	Resistance	1.3946	0.41348414	
22	\$E\$42	Resistance	1.3946	1.030845221	
23					

Fig. 4. Snapshot of the answer report generated from Excel solver.

From the Fig. 3, it is seen that the total distribution loss is 12.63%, which includes HT line loss (8.12%), other line loss (2.5%) (LT+service drop), equipment loss (1.25%) (distribution transformer + line voltage regulator, etc.), and commercial loss (0.75%). Considering other losses are almost same, this paper shows the proposed conductor size for reconductoring of the HT line for which total distribution loss will reduced from double digit (12.63%) to single digit i.e. 9.5%. Here, target cell (E1) is set as '9.50' in the MS Excel solver and this value may be achieved by changing the conductor size i.e. resistance value i.e. cell no. E7, E13, E15, E17, E19, E27, E29, E34, E40, E42 assuming the constraints of this cell value not below 0.3. Maximum value is set as 1.4 considering available size used in BREB system. These cells are taken into consideration because corresponding sections of these cells bears more current than the others, as shown in Fig. 2. These sections have more effect on the loss than others.

The answer report for the selection of the proposed size of conductors is shown in Fig. 4. The answer report shows the final value of the resistance of the conductors to minimize loss to target value.

If the calculated value of resistance is not coincide with the resistance of conductor available in BREB system, then next higher size or nearest size can be chosen. It is obvious that rehabilitation of conductor, i.e., reconductoring involves monetary consideration, which includes the overall cost of annual energy losses, and increase of depreciation for the

capital investment due to changed conductors. For this case of reconductoring, benefit to cost ratio (B/C) should be maintained as ≥ 1.5 [15].

V. CONCLUSION

In this paper, a novel approach to reduce distribution loss to single digit is presented. It involves identifying the high loss sections and proposing the conductor size of the selected sections by simplex algorithm used in the MS Excel solver. The proposed approach reduces the computational effort for engineers. As it is extremely important to reduce the distribution loss, strategically measures should be taken and set the implementation priority to minimize the loss to a single digit. The proposed method can be effectively used to plan and design rural distribution systems to reduce loss.

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A Tabu-based Heuristic Optimization Algorithm for Load Shedding Minimization

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Abstract—Efficient supply and distribution of energies in the power sector is a very important issue for development. The smart grid technologies allow the distribution and production of the commodity flow so that the proper communication is maintained and optimized depending on the demand. All the demand sites will be connected to the grid and the load over each sources are to be distributed in such a way that the load is minimized per source so that they can keep some of the power as residue. The power flow is required to be maintained continuously by choosing the best path which has minimum cost and transmission loss. For this different optimization technique could be used to be finding an optimal or maximal flow path for transmission. In this paper, we propose a deterministic graph model to represent the flow in the power grid and apply tabu-based heuristic local search to find the optimal flow such that the load shedding is being minimized. We applied a number of local search algorithms on different data sets and the results shows the potential for applying such meta-heuristic algorithms to solve the problem.

Keywords—power grid, local search, tabu-list, optimization.

I. INTRODUCTION

Power transmission is a very important issue for the developing industrial sector of a country. In the context of Bangladesh, even if the power sources produces excess of energy than what is required by the demand sites, due to poor transmission lines and distribution techniques the power is not utilized optimally. This had been the case is recent time. Moreover due to increased demands, load shedding minimization by utilizing the existing communication links is a great challenge. Power flow optimization is a rising topic now a days where different researchers are using dissimilar techniques to optimize power and use renewable sources to reduce load shedding.

Power optimization is done either using optimizing source side management (SSM) or Demand side management (DSM). DSM is a an option where demand is being optimized by the modification of consumer demand for energy through various methods like power shifting [1], peak load reduction [2] and consumer demand shifting for using appliances [3]. In case of SSM, least number of sources can be used to supply more demand so that some of the power can be saved for future purpose. The demand is selected according to most desired one, means the source will be supplying those demands which have highest need for power [4]. The source will keep few residues within themselves, the sources will be not being kept null. While transmission, there is always loss and these

increases with distance and overload, an optimized path need to be chosen that all the demand get the required power and less loss. Even after transmission, there is enough power so that it can supply if more power is needed during any necessity. Various algorithms are being used to solve power flow optimization problem [5], [4].

In this paper, we have proposed a deterministic optimization model for the load shedding minimization problem. We have also devised a tabu-based heuristic optimization algorithm to solve the problem. The experiments are being performed on artificially generated datasets and the effectiveness is tested by comparing it with other optimization methods. Our algorithm significantly outperforms other methods. The main contribution of this paper are as follows:

- A deterministic model for load shedding minimization in power grids.
- Application of stochastic local search techniques to solve this hard combinatorial optimization problem.
- A novel tabu-based meta-heuristic to solve the problem efficiently.

The rest of the paper is designed as follows: Section II presents our deterministic model for load shedding minimization problem and the necessary background; Section III describes some related work in the literature; Section IV describes the algorithm being used; Section V discusses and analyzes the experimental results; and finally, Section VI presents our conclusions and outlines our future work.

II. PRELIMINARIES

A key limitation of electric power is that, with minor exceptions, electrical energy cannot be stored, and therefore must be generated as needed. A sophisticated control system is required to ensure electric generation very closely matches the demand. If the demand for power exceeds the supply, generation plant and transmission equipment can shut down, which in the worst case may lead to a major regional blackout. Transmitting electricity at high voltage reduces the fraction of energy lost to resistance, which varies depending on the specific conductors, the current flowing, and the length of the transmission line. Electricity is transmitted at high voltages (120 kV or above) to reduce the energy losses in long-distance transmission. So for generation of high electricity, we need powerful cables so it increases the transmission cost. The transmission flow is heterogeneous as it needs to consider the

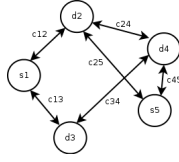


Fig. 1. Connections between different source and demand sites.

cost, power loss, need to supply the nearby demand as well the demand which has more consumption. Provided, all the demand should be connected to the source so that there is no black out for any consumer. In Bangladesh, two season changes are being focused like summer and winter. In summer, consumption is high where else in winter its less. So the remaining power which is not being used during winter, will be used when the consumption level is high during summer.

If we find the distance from all the sources, we will choose those demands for the source which is nearby, as distance reduces transmission loss provided the maximum consuming demand will be preferable. All the sources and demands will form an array, the demand will be connected to those sources who will fulfill all those constraint shown in 1. During all this, none of the demand should be left out. A typical scenario is depicted in Figure 1. Here we could notice a small grid with heterogeneous links between demand (d_2, d_3, d_4) and supply (s_1, s_5) sites with varying capacities. In the next section we describe the deterministic model we adopt in this paper.

A. A Deterministic Model

In this model, we there are two types of nodes: demand and supply. The supply nodes may represent the power generating sites and the demand nodes are residential locality or industrial area. These nodes are connected to each other by power lines with varying capacity depending on the connection types. Supply of a node i is indicated as s_i and demand is indicated as d_i . However these quantities are in the same units and only varying in the sign. The grid is given a weighted graph, $G \equiv \langle V, E \rangle$. Each edge, $(i, j) \in E$ indicates links between two nodes, $i \in V, j \in V$. The set of vertices $V = S \cup D$, where S is the set of supply nodes and D is the set of demand nodes. The flow capacity of each of the edges are given as a matrix C . Each element, $c_{ij} \in C$ denotes the flow capacity of the link between node i and j .

1) *Variables*: A set of flow variables in matrix X where each individual flow from node i to j is indicated as x_{ij} . This variable is used to identify the power flow of the model. The flow is unidirectional. A flow in a particular direction means a negative flow in the reverse direction.

2) *Constraints*: Flow in each link must not exceed the associated capacity of the links.

$$|x_{ij}| \leq c_{ij}$$

The power flow within the transmission lines is less than or equal to the capacity of transmission lines so that there is no eddy current loss within the lines.

$$\sum x_{ij} \leq s_i, \forall i \in S$$

where s_i is the supply at site i . However, for the demand sites the total demand might not be fulfilled. However, the total amount must not violate the flow balance, which could be denoted by the following constraint:

$$\sum x_{ij} \leq 0, \forall i \in D$$

Now, there is a load shedding l_i associated with each demand site negative in sign defined as following:

$$l_i = d_i + \sum x_{ij} \leq 0, \forall i \in D$$

Our main objective is to meet the demand, due to shortage of source capacity or leakage in transmission lines the demand is greater the transmitted power.

3) *Objective*: The objective function is to minimize the total load shedding in the grid defined by the following equation:

$$L = \sum l_i, \forall i \in D \quad (1)$$

Heuristic optimization is being discussed which means a way of solving a problem more quickly when classic methods are too slow, or for finding an approximate solution when classic methods fail to find any exact solution. Using this, optimal flow is found where minimal of load shedding occurs. Among them, Tabu search optimization is being considered which enhances the performance of local search. Even worsening flow is accepted if no improving flow is available. In addition, tabu tenure is introduced to avoid repetition. It has got memory structure to avoid local minima and loadshedding is not minimized then the tabu list is not being updated, its marked as "tabu" (forbidden).

III. RELATED WORK

Different optimization techniques have been proposed to solve the smart grid related problem [6]. During load cuts, communication [7] and self heal is necessary so that there is no black out. Demand side management and optimal load shedding with minimum load cuts are proposed using self healing and preventive methods [5]. Self healing is done through demand side where optimal power flow determined through load shedding scheme by Bacterial particle swarm optimization [8]. Demand side management along with flow control and cost management is being analyzed previously with combination of Fuzzy Logic and Genetic Algorithm [9].

To balance the power flow from supplier to consumer, the mechanism of power shiftable and time shiftable appliances is being used [3]. Load management is done using Integer Linear Programming where hourly peak load is being reduced [3]. In this case, the customers can take decisions regarding energy consumption, reduce peak load demand and reshape profile. Heuristic based evolutionary algorithm is used for solving minimization problem dividing into three sectors: residential, commercial and industrial area [4]. Evolutionary algorithm allows implementation of features that model load demand pattern based on customer needs [4]. Some devices have high priorities that are taken consideration by using evolutionary algorithm where load shifting according to time steps are used.

Integer linear programming (ILP) models designed to identify optimal combinations of supply sources, demand sites for them to serve, and the pathways along which the reallocated power should flow [10]. Stochastic search algorithms are applied to solve a related problem finding optimal power flow, minimizing load shedding and cost management [4].

For data collection in real time from remote locations, SCADA (supervisory control and data acquisition) is a category of software application program for process control in order to control equipment and conditions [11].

IV. OUR APPROACH

In this section, we present our tabu-base heuristic algorithm to solve the load shedding minimization problem.

A. Search Procedure

Our search algorithm is depicted in Algorithm 1. It starts with a constructive initialization procedure that initializes the flow matrix. In each iteration of the algorithm, it generates a candidate list using a very simple local search operator called *singleflip*. The best candidate is selected from this generated candidate list and selected it for the next iteration. Ties were broken depending on a uniform random distribution. However each of the links are maintained in a tabu list. After a certain number of iterations the best solution found so far is returned.

Algorithm 1: TabuSearch(*numberOfIterations*)

```

1 stagnation  $\leftarrow$  numberOfIterations *.25
2 nonImproving  $\leftarrow$  0
3 iteration  $\leftarrow$  0
4 flows  $\leftarrow$  generateInitialSolution()
5 initializeTabuList()
6 bestLoadShedding  $\leftarrow$  calculateLoadShedding(flows)
7 bestSolution  $\leftarrow$  flows
8 while iteration + +  $\leq$  numberOfIterations do
9   candidateList  $\leftarrow$  generateNeighbors(flows)
10  bestCandidate  $\leftarrow$  selectBest(candidateList)
11  if notTabu(bestCandidate) or aspiration criteria
    then
12    flows  $\leftarrow$  bestCandidate
13    updateTabuList()
14    loadShedding =
    calculateLoadShedding(flows)
15    if loadShedding > bestLoadShedding then
16      bestLoadShedding  $\leftarrow$  loadShedding
17      bestSolution  $\leftarrow$  flows
18      nonImproving  $\leftarrow$  0
19    else
20      nonImproving + +
21  if nonImproving  $\geq$  stagnation then
22    flows  $\leftarrow$  restartSolution()
23    clearTabuList()
24    nonImproving  $\leftarrow$  0
25 return bestSolution

```

Rest of this section describes various components of our search algorithm.

1) *Initialization*: We used a constructive initialization algorithm to generate initial flows in the grid. From each of the supply nodes, flows were generated towards the demand nodes by selecting the edges randomly and the residual power was also carried using the same procedure.

2) *Operator*: The *singleFlip* local search operator generated the candidate list by selecting each list and randomly flipping the flow value associated with it. This operator is simple in manner and cost effective. Further intensification is possible on the values to provide more direction towards search.

3) *Tabu List*: Memory based methods are applied to escape local minima and to provide alternative paths to search [12]. In our algorithm, we maintain a tabu list that prohibits the selection of a particular link to be altered for a specific number of iterations denoted by the parameter, *tabuTenure*. However, an aspiration criteria was used whenever an improving value was found. This made our search more flexible.

4) *Handling Stagnation*: We used the random restart technique [13] to handle stagnation in our search. In each iteration, our algorithm keeps track of the improvement done on the load shedding value through a counter called *nonImproving* and whenever the algorithm fails to improve for consecutive number of iterations denoted by *stagnation* a random restart is applied to improve further.

V. EXPERIMENTAL RESULTS

We have implemented the algorithms in Matlab 2012b and have run our experiments on an Intel Quad core processor having 2.66Ghz machine 4 GB ram running 64 bit Windows seven operating system.

A. Datasets

The benchmark datasets are being generated a program written in Java. We varied the size of the graph by changing the number of nodes. The demand and supply data was approximated using the data from the national grid data of Bangladesh. We collected information information provided in the PDB website. For any given size of the graphs, the mean and the variance of the supply were estimated. Sparse graphs were generated assuming the average line voltages as capacities. In Table I, we present results for 20 such graphs. Each instance data is denoted by 'graph x - n ' where n is the instance identifier and x is the number of nodes.

B. Results

We compared our algorithm with three other algorithms: Monte Carlo, Gibbs Sampling and a Greedy selection algorithm. We performed experiments by running different algorithms on each of the benchmark datasets. We executed each of the algorithms for 5 times and 1000 iterations each time reported best and average load shedding values achieved in Table I. The best values in Table I are shown in bold faced fonts. An infinity (∞) indicates the initial load shedding value or no improvement.

TABLE I
COMPARISON OF LOAD SHEDDING VALUES ACHIEVED BY DIFFERENT ALGORITHMS FOR VARIOUS DATASETS WITH DIFFERENT.

graph instance	Tabu Search		Monte Carlo		Greedy Algorithm		Gibbs Sampling	
	best	avg	best	avg	best	avg	best	avg
graph5-0	0	63.64844	555.5462	556.7652	555.0016	555.00798	555.927	558.34502
graph5-1	0	0	0	0	0	0	0	0
graph5-2	0	5.94774	116.4614	128.90372	64.0485	72.66468	126.9701	135.8478
graph5-3	0	13.77974	74.9999	83.65462	50.0267	50.04296	65.8928	85.93336
graph5-4	0	40.99412	113.4751	121.94246	79.0398	79.0919	126.8523	145.74364
graph10-0	0	0	18.4088	33.43964	0	2.845	31.5535	47.99244
graph10-1	0	0	0	0	0	0	0	15.1487
graph10-2	0	4.462	149.1593	205.06794	62.0142	86.44544	178.1823	188.9234
graph10-3	23.9821	34.967	180.2594	191.5367	86.409	105.65126	177.4741	214.6886
graph10-4	0	0	74.0882	112.47566	0	0	91.2324	98.5033
graph15-0	0	0	0	21.08372	0	0	10.351	17.1249
graph15-1	10.0105	21.14136	65.4529	71.68938	22.0117	22.0192	102.0269	127.61176
graph15-2	0	0.05146	112.3944	137.7436	0	0	132.3236	160.5073
graph15-3	0	0	49.7488	86.32676	0	0	∞	∞
graph15-4	0	0	47.1956	64.27512	0	0	41.5419	66.29152

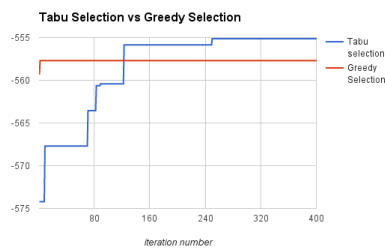


Fig. 2. Search progress for a typical run of Tabu Selection and Greedy Selection algorithm.

C. Analysis

From the results reported in Table I, it is evident that tabu selection works better than any other algorithm. Though it worked similarly with greedy selection in some cases. Initially, Monte carlo and Gibbs sampling was giving results with graphs of smaller sizes but as the size of node increases, there behavior declines as there is no intelligent decision has been taken in the search.

In tabu selection, intelligence is being applied where both source and demand is being optimized. By introducing its flow depending on the size of source so there is no overflow from the source the demand is not oversupplied of the flow and in case demand does not get enough supply load shedding is measured. These intelligence being checked for each source and demand. It is depicted that with increase in size both greedy Selection and tabu selection behave spontaneously better by giving optimal flow and nearly zero load-shedding. Whereas, tabu selection works better with both less sample size and larger sample size. Tabu selection shows continuous improvements of best accuracy achieved. The situation is depicted in Figure 2.

VI. CONCLUSION

In this paper, we presented a tabu selection algorithm with a constructive initialization for load shedding minimization problem. Our algorithm performs restarts to handle after stagnation. Our tabu selection algorithm produces significantly better results with increase of number of nodes compared to other algorithms. In future, we would like to solve the problem

with larger graph sizes and deploy the algorithm to solve for real data sets. We would also like to extend our model to incorporate the stochastic behavior of the power grid and load.

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Temperature Effect on Electronic Properties of Armchair Graphene Nanoribbon

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Abstract—Graphene Nanoribbon (GNR) is considered as a forthcoming material for the next generation of nanoelectronic devices. The more exciting thing in GNR is that it shows considerable bandgap. Armchair GNR (A-GNR) is one type of GNRs which demonstrate semiconducting electronic properties. In this research, it has been observed that due to edge effect and thermal effect on A-GNR, the electronic properties vary. In this paper, the electronic properties like bandgap energy, carrier concentration, capacitance and fermi probability for a definite energy level have been analyzed by considering the above mentioned effect.

Keywords—Graphene nanoribbon, bandgap, carrier concentration, quantum capacitance.

I. INTRODUCTION

Graphene is the latest discovery in the field of nanotechnology based research, because each layer of graphene sheet is close to one atomic thickness. For this negligible amount of thickness it is assumed as two dimensional material with respect three dimensional based world where each layer of carbon atoms and tightly C-C bond form honeycomb structure [1]. Because of exceptional and unique electronic properties of graphene, it has attracted attention of most of the nanoscientists [2]-[4] in the world. It has been considered as important element in future carbon-based nanoelectronics where electronic properties can be tuned by its width or edge structures [5]. Graphene is also a basic building block of carbon nanotubes (CNTs), buckyballs, graphite as well as graphene nanostrips and some of them show metallic or some of them show semiconducting properties or some of them show both.

Graphene nanoribbon is the nanostrips of monolayer graphene, again a potential candidate for its metallic or semiconducting properties. One of the GNR types whose edges are like “arm of chair” is called Armchair GNR which shows semiconducting electronic properties mostly. In addition to this properties, thermal properties, thermal effect on GNR’s electronic properties (here carrier density, capacitance) have also fundamental and theoretical importance [6]. Along with thermal effect, another is the edge effect in the time of manufacturing of A-GNR has a significant impact on electronic properties.

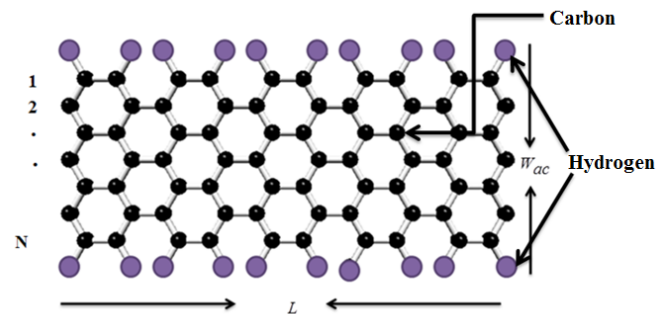


Fig. 1. Schematic of a A-GNR. chirality(5,5) The black circle denotes carbon and blue circle denotes hydrogen. The dimline, width and length of A-GNR are denoted by N , W_{ac} , L .

In this paper we will first observe the energy band structure of A-GNR at room temperature. Then we will observe the energy for considering the thermal effect of A-GNR as a channel. Then we will observe the carrier concentration and quantum capacitance of A-GNR.

II. THEORY

The useful property of A-GNR is that it shows a considerable bandgap because it has been seen that there is an inverse relation of bandgap and width of A-GNR. It also shows direct bandgap which means valence band and conduction band meets at a specific point which has been clearly observed through energy dispersion relation by considering the surrounding edge effect of A-GNR. The energy dispersion relation of it is expressed as [7]

$$E = \sqrt{\left(\frac{E_g}{2}\right)^2 + (\hbar v_s k)^2} \quad [1]$$

where $\hbar$ is the reduced plank’s constant and bandgap energy is [8]

$$E_g = -\frac{2\pi(\gamma_1 - 2\gamma_3)}{\sqrt{3}(N+1)} + \frac{6(\gamma_3 + A_{\gamma_1})}{N+1}$$

and

$$(\hbar\gamma_s)^2 = (3d)^2 \left\{ -\frac{1}{2}\gamma_1 s \cos \frac{p\pi}{N+1} \times [\gamma_1 + \gamma_3 (2 \cos \frac{2p\pi}{N+1} + 1)] \right. \\ \left. + 4 \frac{(\gamma_3 + A_{\gamma_1})}{N+1} \sin^2 \frac{p\pi}{N+1} \right\} - \\ \gamma_3 (\gamma_1 + 2\gamma_3 \cos \frac{p\pi}{N+1} + 4 \frac{(\gamma_3 + A_{\gamma_1})}{N+1} \sin^2 \frac{p\pi}{N+1}) \}$$

where $\gamma_1 = -3.2$ eV is the first nearest neighbor hopping parameter, $\gamma_3 = -0.3$ eV is the third nearest neighbor hopping parameter, $A_{\gamma_1} = -0.2$ eV is the correction of γ_1 .

A-GNR is used as a channel in a contact with source and drain. As source and drain regions are coupled to the channel and then electrons will flow in and out of the device bringing them all in equilibrium with a common electrochemical potential μ at temperature T , according to the Fermi probability function. The Fermi probability function means the finding probability of an electron in an energy level with energy E . But due to thermal changes of surrounding environment of A-GNR, the Fermi probability changes for a definite $E - \mu$ [9]

$$f_o(E - \mu) = \frac{1}{1 + e^{\left(\frac{E - \mu}{k_B T}\right)}} \quad [2]$$

As said, due to changes of temperature it also changes the carrier (electron) movement per unit volume in a transistor. That means the changes on intrinsic carrier concentration (density) for a definite width of A-GNR [10].

$$n_i = \left(\frac{4}{W_{ac}}\right) \left(\sqrt{\frac{\pi k_B T}{E_g}}\right) e^{-\frac{E_g}{k_B T}} \quad [3]$$

where $k_B = 8.61 \times 10^{-5} \text{ eV K}^{-1}$ is the Boltzmann constant, W_{ac} is the width of A-GNR which is calculated as [11]

$$W_{ac} = (N-1) \frac{\sqrt{3}}{2} a$$

Here $a = 0.142 \text{ nm}$ is the C-C distance and N is the number of dimerline.

Due to quantum effect of A-GNR, a condition arises in carrier concentration as well as capacitance. This is called non-degenerate regime where Fermi level is located in between A-GNR bandgap. The carrier concentration becomes

$$n_{ND} = n_i e^{\eta} \quad [4]$$

$$\text{where } \eta = \frac{E_g}{2k_B T}$$

Capacitance is one of the important electronic properties in semiconducting device. For the applied gate voltage V_g to a metallic gate, extra carriers (electron) are induced with the density n to the A-GNR with same amount of opposite carrier resides to the gate. The applied gate voltage shifts the chemical potential μ of the A-GNR from the charge neutrality point μ_0 . For this, as usual capacitance splits into two different capacitances where classical capacitance is responsible for changing the charge neutrality point. And another capacitance is responsible to excite the carrier from changed neutrality point to quantum mechanical band. Electron fills up these bands according to the Pauli Exclusion Principle rules. Again quantum capacitance splits into two regimes. These are 1) Degenerate regime (DR) and 2) Non-degenerate regime (NDR) [12].

The degenerate regime denotes when Fermi level remains in any state, from CB to positive infinite energy band, which is written as [13]

$$C_{QD} = \frac{e^2}{3\pi} \left(\frac{x + \frac{E_g}{2k_B T}}{\sqrt{x + x \frac{E_g}{k_B T}}} \right) \quad [5]$$

Again quantum capacitance in non-degenerate regime denotes when Fermi level remains in between the closest but not in touch with our imagined Valence band (VB) and Conduction band (CB) diagram, which is written as

$$C_{QND} = C_{QD} e^{\eta} \quad [6]$$

Actually DR works for both positive and negative gate voltage, but NDR works only in negative gate voltage.

III. RESULTS AND DISCUSSION

For simulating these electronic properties, MATLAB has been used. At first, we will observe the band structure of A-GNR as given in Eq.(1) from where we will get a clear idea about considerable bandgap for a definite width of A-GNR (here 4nm).

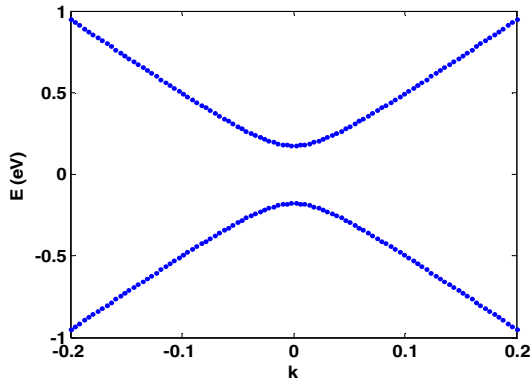


Fig.2. Energy (eV) vs. wavevector for 4 nm A-GNR.

From Fig.2 we have seen that there is a clear BG between VB and CB energy. More the width, lower will be the BG. Also this value will be used in later analysis. Here we have used for 4nm, as nowadays if we consider sub-nm ranged device, it actually denotes device width less than 10nm. There is an inverse relation between BG and width of A-GNR. For a considerable amount of BG, we have used 4nm and from using E_g , the BG is 0.1832eV but if the edge effect is not considered, it will be 0.3414eV. It should be reminded that the carrier can easily be excited choosing lower amount of BG which can be controlled through sub-nm ranged device structures.

Now we will observe the energy level and Fermi probability from Eq.(2) within a few $k_B T$ of chemical potential and they are partially empty or partially filled.

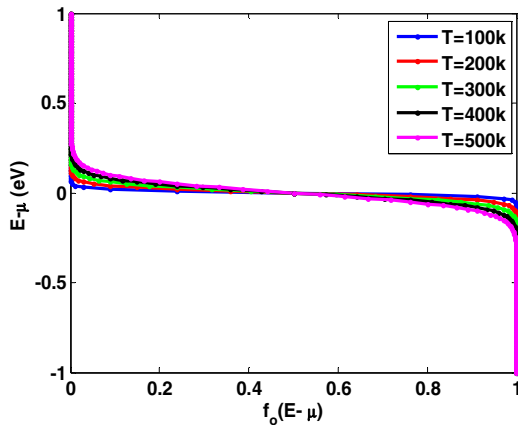


Fig.3. $E-\mu$ vs. Fermi probability of carrier for different temperature.

From the Fig.3 it has been seen that for increasing temperature the curve more deviates than 0 value of $E-\mu$. Because of more temperature more carriers move for which it fills the level more as well as more levels remain empty.

Now we will observe the carrier concentration characteristics as given in Eq. (3) and (4) and their responses to varying temperature.

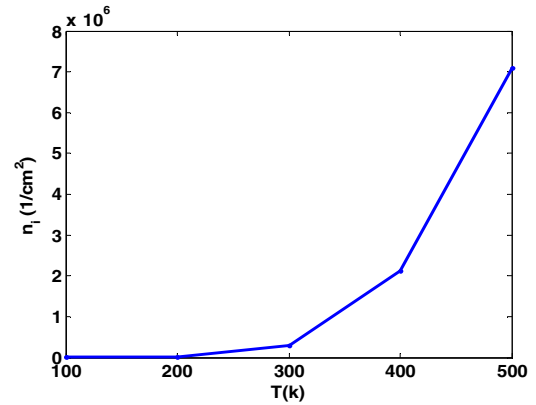


Fig. 4. : Intrinsic carrier concentration (cm^{-2}) vs. Varying Temperature (K) for $W_{ac} = 4\text{nm}$.

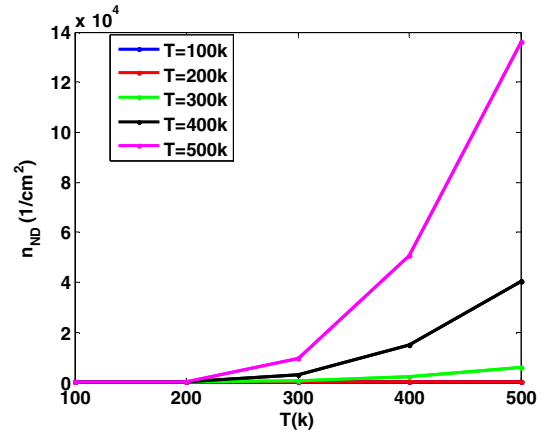


Fig.5. Carrier concentration (cm^{-2}) vs. Varying Temperature (K) for $W_{ac} = 4\text{nm}$.

From Fig.4 and Fig.5, it has been seen that carrier movement and density per unit volume or area is increasing for increasing the temperature. Because of more temperature, more electrons excite from its VB to CB and create denser media in conduction region. That means carrier concentration in non-degenerate regime increases for increasing temperature but lower than intrinsic carrier concentration, as in this regime carrier moves in forbidden energy gap, which actually decreases the carrier density. We know that DR denotes when Fermi level can be situated from CB to $+\infty$ eV. When E_F locates from $+0.0916\text{eV}$ (actually from CB valley to 1eV), then the device will work as degenerate semiconductor.

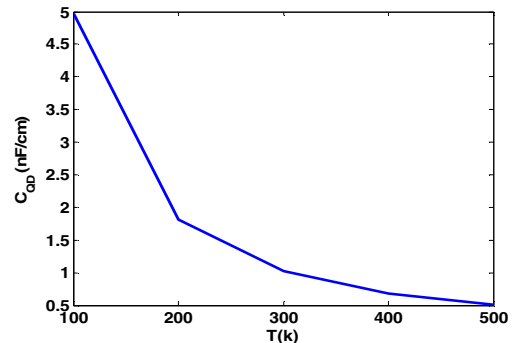


Fig.6. Quantum capacitance in degenerate regime (nF/cm) vs. Varying Temperature (K) for $W_{ac} = 4\text{nm}$.

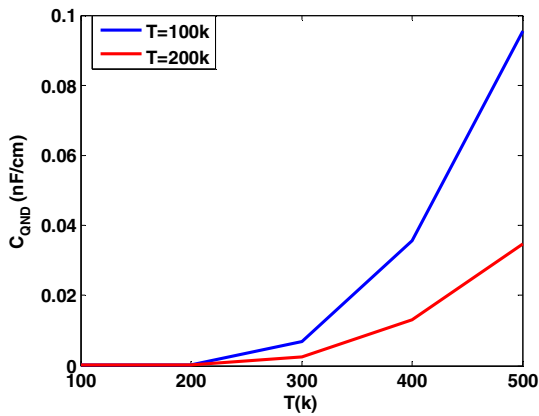


Fig. 7 (a). Quantum capacitance in non-degenerate regime (nF/cm) vs. Varying Temperature (K) for $W_{ac} = 4nm$.

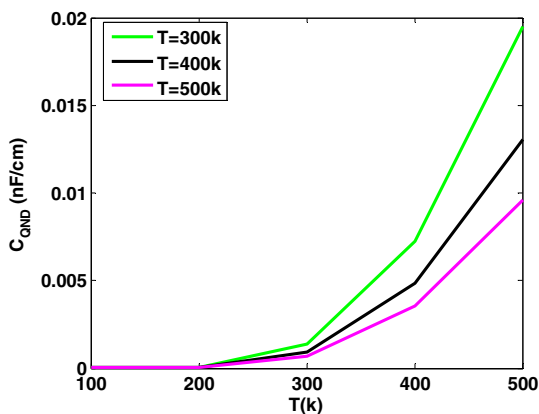


Fig.7 (b). Quantum capacitance in non-degenerate regime (nF/cm) vs. Varying Temperature (K) for $W_{ac} = 4nm$.

From Fig.6, Fig.7 (a) and Fig.7 (b), we have seen two different natures of carrier in quantum capacitance. Actually when the Fermi level is above the conduction band, then the device conducts, or below the valence band region it is in off state. As we have used the method of conduction region, the capacitance normally decreases along with the increase of temperature due to denser carrier through A-GNR. So normal behavior is like the capacitance in degenerate regime. But in non-degenerate regime, the Fermi level resides in between VB and CB for which an exponential relation arises which increases the capacitance. Due to large difference of capacitance from 100k, 200k than 300k, 400k, 500k the curve has given in separate diagram.

IV. CONCLUSION

In this paper, we have first observed the band structure of A-GNR as a channel. Then we have observed the energy and Fermi probability responses for varying the temperature. From this, we have decided that more the temperature, more energy levels have a probability within range $0 \leq f_o \leq 1$. Then we have observed the intrinsic carrier concentration and non-degenerate carrier concentration which increase for according

to increase of temperature. That means more will be the temperature, more will carrier density in a definite area. But capacitance in degenerate regime decreases with respect to temperature where in capacitance in non-degenerate regime it reverses the nature. Actually as the carrier movement and carrier density are increasing, carrier does face the barrier in movement or delay due to capacitance formed in A-GNR. Because we know that capacitance impedes the movement of carrier. As non-degenerate regime works only in between VB and CB but saturates in conduction region, so it reverses normal nature of capacitance with respect to varying temperature. Temperature in an experiment has a great significant into its electronic properties. Capacitance actually causes the decrease of the switching speed as seen from scientific work, where temperature sometimes decreases the capacitance, like for 500k the non-degenerate regime capacitance is decreased upto 0.01nF/cm. Hence, a compromising decision between temperature and device width must be chosen so that transistor made by A-GNR will provide a higher switching speed or lower amount of delay.

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Strain-dependent Optical Properties of [113]-oriented InGaAs/GaAs Quantum Well

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Abstract—A numerical approach is presented to study the optical properties of compressively strained [113]-oriented InGaAs/GaAs quantum well (QW) architecture by solving an eight-band $k \cdot p$ Hamiltonian using finite difference method including spin-orbit coupling. Euler's rotation technique is used to modify the wave vector and Hamiltonian matrix in conventional [100] crystal orientation. It is found that there is a substantial correlation between magnitude of strain and optical gain spectra. From the MATLAB simulation results, it can be settled that the energy band dispersion profile, momentum matrix and optical gain deviates exclusively with the increase of strain magnitude. The regular optical gains are inspected as 2700, 2810, 3080 and 3300 cm^{-1} when the well is compressively strained by 0.50, 0.90, 1.15 and 1.60% respectively at the carrier injection density of $2.5 \times 10^{18} \text{ cm}^{-3}$ which shows that highest optical gain and lowest effective mass are attained for 1.60% compressive strain.

Keywords—Compressive strain; InGaAs; $k \cdot p$ method; Transition momentum matrix; Optical gain

I. INTRODUCTION

The zinc-blend (ZB) III-V semiconductors have received considerable attention in the last decades due to its unique characteristics such as high electron mobility and superior carrier saturation velocity. Consequently, its application scope broadens up to the operation as the active region of different optoelectronic devices. The earliest and perhaps the most established hetero-structure laser technique is AlGaAs/GaAs structure. The range of available wavelength (λ) for lattice matched AlGaAs/GaAs QW hetero-structure is from 0.80-0.65 μm . But the range of wavelength (λ) 0.88-1.1 μm is not readily available in any III-V lattice matched laser material system. There are, however, a number of important applications that require emission in this range, especially pump lasers for rare earth doped silica fiber amplifiers [1-4]. A simple demonstration on direct gap III-V compound semiconductor alloys suggests that, perchance, the best material system for attaining the emission wavelength in range of 0.88-1.1 μm is $\text{In}_x\text{Ga}_{1-x}\text{As}$. There is, however, no suitable binary substrate that allows lattice matched composition of $\text{In}_x\text{Ga}_{1-x}\text{As}$ in the wavelength range of interest. Thus, only a hetero-structure material technique in which a large lattice mismatch must be accommodated (as great as 3%) may be considered [5].

Recently, it has become visible that optical gain and threshold current of lasers can be improved by introducing compressively strained quantum well (QW) into the active region. The major reasons are that the compressively strained QW has a reduced effective mass for the top heavy hole band which reduces density of states and hence, Auger recombination rate so, lower threshold current. The optical gain and threshold current strongly depend on the conduction and valence band structures. Both the conduction and valence band energy states and wave functions can be modified according to the magnitude and state of strain in the QW.

Therefore, strain is a key parameter for controlling the device performance [6-7].

From theoretical aspect, several authors have investigated the emission profile and threshold current density for ZB [100]-oriented structure for different strains and temperatures[8-9]. In our previous work, we focused our attention on studying the effects of crystal orientation and piezoelectric field on the optoelectronic properties of 0.9% compressively strained InGaAs/GaAs QW and found [113]-oriented QW showed nice performance with orientation-dependent tunable characteristics[10-11]. M.M.Hasan et al. recently reported compressively strained [113]-oriented InGaSb mid-infrared quantum well laser structure with excellent optical properties [12]. In this study, the strain-dependent optical properties of [113]-oriented InGaAs/GaAs QW are presented in details. To best of our knowledge, the optoelectronic properties of strained non-[100] oriented ZB QW is unavailable. In this paper, the optical properties of 0.5, 0.9, 1.15 and 1.60% compressively strained ZB [113]-oriented $\text{In}_{0.4}\text{Ga}_{0.6}\text{As}/\text{GaAs}$ QW's have been investigated by using the 8×8 $k \cdot p$ matrix. Here, 8 nm quantum well is sandwiched between 7 nm barrier layers. To inspect the optical properties in [113] orientation, unitary transformations are used to modify the wave vector and Hamiltonian matrix in conventional [100] crystal orientation. The conduction and valence band wave functions and Eigen energies of the proposed QW structure are evaluated using the finite difference method. The momentum matrix element, emission profile and carrier density-dependent optical gain are reckoned using the energy dispersion profile and wave functions.

The layout of this proposal is outlined as follows: the theoretical details to calculate optical properties of the InGaAs/GaAs QW are illustrated in section 2. Results and discussions on the energy band structure and optical gain performances are delineated in section 3. Conclusively, a summary is depicted in section 4.

II. THEORY

A. The $k \cdot p$ Method of Band-Structure Calculation near Γ -point:

The electronic band structure is solved in the envelope approximation using an eight-band $\mathbf{k} \cdot \mathbf{p}$ Hamiltonian. The modified Kane 8×8 Hamiltonian matrix can be asserted by:

$$H\psi_h(z) = E\psi_h \quad (1)$$

While, H is resultant Hamiltonian matrix that can be expressed by:

$$H = H_0 + H_\epsilon \quad (2)$$

The components of resultant Hamiltonian matrix are[12]:

$$H_0 = \begin{bmatrix} A & 0 & V^* & 0 & \sqrt{3}V & -\sqrt{2}U & -U & \sqrt{2}V^* \\ 0 & A & -\sqrt{2}U & -\sqrt{3}V^* & 0 & -V & \sqrt{2}V & U \\ V & -\sqrt{2}U & -P+Q & -S^* & R & 0 & \sqrt{\frac{3}{2}}S & -\sqrt{2}Q \\ 0 & -\sqrt{3}V & -S & -P-Q & 0 & R & -\sqrt{2}R & \frac{1}{\sqrt{2}}S \\ \sqrt{3}V^* & 0 & R^* & S & -P+Q & S^* & \frac{1}{\sqrt{2}}S^* & \sqrt{2}R^* \\ -\sqrt{2}U & -V^* & 0 & R^* & S & -P+Q & \sqrt{2}Q & \sqrt{\frac{3}{2}}S^* \\ -U & \sqrt{2}V^* & \sqrt{\frac{3}{2}}S^* & -\sqrt{2}R^* & \frac{1}{\sqrt{2}}S & \sqrt{2}Q & -P-\Delta & 0 \\ \sqrt{2}V & U & -\sqrt{2}Q & \frac{1}{\sqrt{2}}S^* & \sqrt{2}R & \sqrt{\frac{3}{2}}S & 0 & -P-\Delta \end{bmatrix} \quad (3)$$

$$\Psi_h(z) = \begin{bmatrix} \Phi_{1/2,1/2}(z) \\ \Phi_{1/2,-1/2}(z) \\ \Phi_{3/2,1/2}(z) \\ \Phi_{3/2,-1/2}(z) \\ \Phi_{3/2,3/2}(z) \\ \Phi_{3/2,-3/2}(z) \\ \Phi_{1/2,1/2}(z) \\ \Phi_{1/2,-1/2}(z) \end{bmatrix} = \begin{bmatrix} g^{(1)}(z) \\ g^{(2)}(z) \\ g^{(3)}(z) \\ g^{(4)}(z) \\ g^{(5)}(z) \\ g^{(6)}(z) \\ g^{(7)}(z) \\ g^{(8)}(z) \end{bmatrix} \quad (4)$$

Where, ϕ_{j,m_j} or $g^{(j)}$ is the conduction and valence band wave function component and E is the energy of conduction band and three valence bands.

The matrix elements are:

$$A = E_c - \frac{\hbar^2}{2m_0} (k_x^2 + k_y^2 + k_z^2) \quad (5)$$

$$P = -E_v - \gamma_1 \frac{\hbar^2}{2m_0} (k_x^2 + k_y^2 + k_z^2) \quad (6)$$

$$Q = -\gamma_2 \frac{\hbar^2}{2m_0} (k_x^2 + k_y^2 + k_z^2) \quad (7)$$

$$R = \sqrt{3} \frac{\hbar^2}{2m_0} [\gamma_2 (k_x^2 - k_y^2) - 2i\gamma_3 k_x k_y] \quad (8)$$

$$S = -\sqrt{3} \gamma_3 \frac{\hbar^2}{2m_0} k_z (k_x - ik_y) \quad (9)$$

$$U = \frac{-i}{\sqrt{3}} P_0 k_z \quad (10)$$

$$V = \frac{-i}{\sqrt{6}} P_0 (k_x - ik_y) \quad (11)$$

P_0 is the coupling between the conduction and valence bands, E_c and E_v are the (unstrained) conduction and valence band energies respectively, Δ is the spin orbit splitting and γ_i 's are modified Luttinger parameters. The other component of the resultant Hamiltonian matrix, H_ϵ depicted in equation (2) is the strain Hamiltonian matrix that can be revealed by [12]:

$$H_\epsilon = \begin{bmatrix} a_c e & 0 & -v^* & 0 & -\sqrt{3}v & \sqrt{2}u & -u & -\sqrt{2}v^* \\ 0 & a_c e & \sqrt{2}u & \sqrt{3}v^* & 0 & v & -\sqrt{2}v & -u \\ -v & \sqrt{2}u & -p+q & -s^* & r & 0 & \sqrt{\frac{3}{2}}s & -\sqrt{2}q \\ 0 & \sqrt{3}v & -s & -p-q & 0 & r & -\sqrt{2}r & \frac{1}{\sqrt{2}}s \\ \sqrt{3}v^* & 0 & r^* & s & -p-q & s^* & \frac{1}{\sqrt{2}}s^* & \sqrt{2}r^* \\ \sqrt{2}u & v^* & 0 & r^* & s & -p+q & \sqrt{2}q & \sqrt{\frac{3}{2}}s^* \\ u & -\sqrt{2}v^* & \sqrt{\frac{3}{2}}s^* & -\sqrt{2}r^* & \frac{1}{\sqrt{2}}s & \sqrt{2}q & -a_v e & 0 \\ -\sqrt{2}v & -u & -\sqrt{2}q & \frac{1}{\sqrt{2}}s^* & \sqrt{2}r & \sqrt{\frac{3}{2}}s & 0 & -a_v e \end{bmatrix}$$

The elements of H_ϵ matrix are:

$$p = a_v (e_{xx} + e_{yy} + e_{zz}) \quad (12)$$

$$q = b [e_{zz} - \frac{1}{2}(e_{xx} + e_{yy})] \quad (13)$$

$$r = \frac{\sqrt{3}}{2} b (e_{xx} - e_{yy}) - i d e_{xy} \quad (14)$$

$$s = -d (e_{xz} - i e_{yz}) \quad (15)$$

$$u = \frac{-i}{\sqrt{3}} P_0 \sum_j e_{zj} k_j \quad (16)$$

$$v = \frac{-i}{\sqrt{6}} P_0 \sum_j (e_{xj} - i e_{yj}) k_j \quad (17)$$

Here, e_{ij} is the strain tensor; a_c is the conduction band deformation potential; a_v is the valence band deformation potential; b and d are the shear deformation potentials.

The energy dispersion profile along [100] orientation is calculated as a function of k for the 8-band k.p theory by diagonalizing the Hamiltonian using finite difference scheme.

B. Crystal orientation dependent wave vector and strain

If the active region is grown on arbitrary crystal orientation, the wave vectors on [100] crystal can be attained by the subsequent expression [13]:

$$\begin{pmatrix} k_1 \\ k_2 \\ k_3 \end{pmatrix} = O_R \begin{pmatrix} k_x \\ k_y \\ k_z \end{pmatrix} \quad (18)$$

The strain on (100) crystal orientation can be reckoned by:

$$\begin{pmatrix} \epsilon_{11} & \epsilon_{12} & \epsilon_{13} \\ \epsilon_{21} & \epsilon_{22} & \epsilon_{23} \\ \epsilon_{31} & \epsilon_{32} & \epsilon_{33} \end{pmatrix} = O_R \begin{pmatrix} \epsilon_{xx} & \epsilon_{xy} & \epsilon_{xz} \\ \epsilon_{yx} & \epsilon_{yy} & \epsilon_{yz} \\ \epsilon_{zx} & \epsilon_{zy} & \epsilon_{zz} \end{pmatrix} O_R^T \quad (19)$$

Here, ϵ_{ij} ($i, j = x, y, z$) are the strain tensors in (hkl)

direction. The values of diagonal elements are:

$$\epsilon_{xx} = \epsilon_{yy} = \epsilon_{zz} = (a_{well} - a_{barr}) / a_{barr}$$

$$\epsilon_{zz} = -\frac{C_{11}}{2C_{12}} \epsilon_{xx}$$

Here, C_{11} and C_{12} are the elastic constants of the well material and the values of off-diagonal matrix element are set to zero in case of in-plane strain condition.

The expression of the rotation matrix is as follows [14]:

$$O_R = \begin{pmatrix} \cos\theta \cos\phi & -\sin\phi & \sin\theta \cos\phi \\ \cos\theta \sin\phi & \cos\phi & \sin\theta \sin\phi \\ -\sin\theta & 0 & \cos\theta \end{pmatrix} \quad (20)$$

Where, $\theta = \tan^{-1}(\frac{\sqrt{(h^2 + k^2)}}{l})$ and $\phi = \tan^{-1}(k/h)$

Practicing the wave vector in the (100) crystal orientation, the Hamiltonian matrix for [100] crystal orientation can be calculated according to section 2.2. Then the Hamiltonian matrix in (hkl) orientation can be counted by the consecutive equation [15]:

$$H^{hkl} = U H^{(100)} U^* \quad (21)$$

Where, $U = R(\theta)R(\phi)$

In this equalization, $R(\theta)$ and $R(\phi)$ are symbolized as rotations which transform the [100] orientated valence band Hamiltonian to arbitrary crystal orientated Hamiltonian matrix. The argument of $R(\theta)$ and $R(\phi)$ can be defined by:

$$R(\theta) = \begin{pmatrix} \alpha^2 & -\sqrt{3}\alpha^2\zeta & \sqrt{3}\alpha\zeta^2 & \zeta^3 \\ \sqrt{3}\alpha^2\zeta & \alpha^3 - 2\alpha\zeta^2 & -2\alpha^2\zeta + \zeta^3 & \sqrt{3}\alpha\zeta^2 \\ \sqrt{3}\alpha\zeta^2 & 2\alpha^2\zeta - \zeta^3 & \alpha^3 - 2\alpha^2\zeta & -\sqrt{3}\alpha^2\zeta \\ \zeta^3 & \sqrt{3}\alpha\zeta^2 & \sqrt{3}\alpha^2\zeta & \alpha^3 \end{pmatrix} \quad (22)$$

$$\alpha = \cos \frac{\theta}{2} \text{ and } \zeta = -\sin \frac{\theta}{2}$$

$$R(\phi) = \begin{pmatrix} e^{i(3/2)\phi} & 0 & 0 & 0 \\ 0 & e^{i(1/2)\phi} & 0 & 0 \\ 0 & 0 & e^{-i(1/2)\phi} & 0 \\ 0 & 0 & 0 & e^{-i(3/2)\phi} \end{pmatrix} \quad (23)$$

C. Optical Gain

The optical gain as a function of energy for quantum well architecture can be addressed by [15]:

$$g(E) = \frac{2q^2\hbar}{n\epsilon_0 m_0^2 L E} \times \sum_{n,m} \frac{k_i M_{nm}(k_i) \Gamma / (2\pi)}{(E_{cn}(k_i) - E_{kpm}(k_i) - E)^2 + (\Gamma / 2\pi)^2} (f_c^n - f_v^m) dk_i \quad (24)$$

Here, q is denoted as free electron charge, $\hbar$ as reduced Planck constant, n as the index of refraction, ϵ_0 as free space dielectric constant, C as speed of light, E as photon energy, E_{cn} as the n^{th} conduction sub-bands, E_{kpm} as the m^{th} valence sub-bands, and M_{nm} as the momentum matrix element in strained quantum well architecture. $\Gamma = \frac{\hbar}{\tau}$; τ is the photon relaxation time. F_c , F_v are the Fermi levels of conduction and valence band. The momentum matrix element for transition can be written as [15]:

$$M_{nm}(k_i) = \frac{3}{4} \left[\langle \psi_{n(k_i=0)} | g^{(1)}_m \rangle^2 + \frac{1}{3} \langle \psi_{n(k_i=0)} | g^{(2)}_m \rangle^2 \right] M_b^2$$

Where, $\langle \psi_{n(k_i=0)} | g^{(1)}_m \rangle = \int_{-\infty}^{\infty} \psi_{n(k_i=0)}(z) g^{(1)}_m(z) dz$, M_b is the bulk average crystal momentum matrix that can be noted by:

$$M_b = \left(\frac{m_0}{m_e} - 1 \right) \frac{m_0 E_g (E_g + \Delta)}{6 (E_g + \frac{2}{3} \Delta)} \quad (25)$$

While, E_g and Δ are the energy band gap and split-off valence band separation respectively.

III. RESULTS AND DISCUSSIONS

Based on the mathematical model presented in Section 2, the strain-dependent optical properties of 8 nm ZB [113]-oriented InGaAs/GaAs QW architecture has been calculated. The barrier material $\text{In}_{0.40}\text{Ga}_{0.60}\text{As}$, $\text{In}_{0.46}\text{Ga}_{0.64}\text{As}$, $\text{In}_{0.52}\text{Ga}_{0.48}\text{As}$, $\text{In}_{0.58}\text{Ga}_{0.42}\text{As}$ are so selected that the QW is compressively strained by 0.5, 0.9, 1.15 and 1.60%. The simulation is carried out in MATLAB environment at room temperature. The conduction band dispersion profiles (C1) for different strains are shown in fig.1. It is found that at the band edge the conduction band energy shifts towards higher energy with increasing the value of strain. The conduction band edge energies found to be shifted from 0.53eV to 1.19eV for changing the compressive strain from 0.50% to 1.60%. The wave functions of the first heavy hole (HH1) and light hole (LH1) at the Γ -point for different strained QWs are shown in fig 2(a) and (b) respectively. In both cases the wave function of holes are more confined in the well with increasing strain. Because, the valence band potential energy for hole decreases in the well with increasing strain. The valence band dispersion profile obtained for same QW for different strain conditions are shown in fig. 3 where HH, LH and SH represents heavy hole, light hole and spin orbit split-off hole. The energy separations between valences sub-bands strongly depend on the state of strain. From fig.1 and 3, it is found that energy separations between C1 to HH and LH at the Γ -point increase due to increasing strain from 0.50% to 1.60%. The separation between HH and LH bands also increases as seen in fig.3. This indicates that the band-mixing effect between two bands is reduced with increasing strain. As the QW is compressively strained, the maximum transition occurs between C1 to HH. The momentum matrix element for the transitions between C1 to HH for different values of strain is calculated and shown in fig. 4. It is found in Fig. 4 that the values of wave vector-dependent momentum matrix elements increase with increasing strain. It is due to the fact that the overlapping of wave function of electrons and holes increases with increasing the magnitude of strains. The carrier density-dependent optical peak gains are calculated and shown in fig. 5. It is found that the optical gain starts approximately at the carrier density $1.90 \times 10^{18} \text{ cm}^{-3}$ for

all values of strain and then gradually increases with increasing carrier density and magnitude of strain. However, the change in optical gain is not so significant with strain when the injection current density is lower than $2.0 \times 10^{18} \text{ cm}^{-3}$. From fig. 4, it is found that the maximum optical peak gain is evaluated when the well is 1.60% compressively strained. The optical gain changes significantly with carrier density when the well is compressively strained by 0.5 to 0.9%. In contrast, the change in optical gain is not visible when the well is strained by 0.9 to 1.60% with injection carrier density.

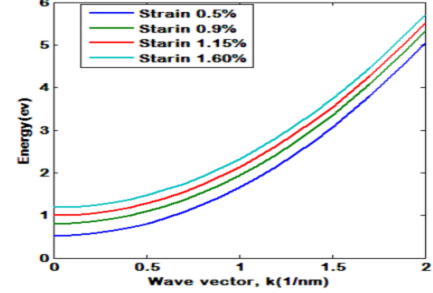


Fig 1: Compressively strained conduction band (C1) dispersion profile of [113]-oriented InGaAs/GaAs QW

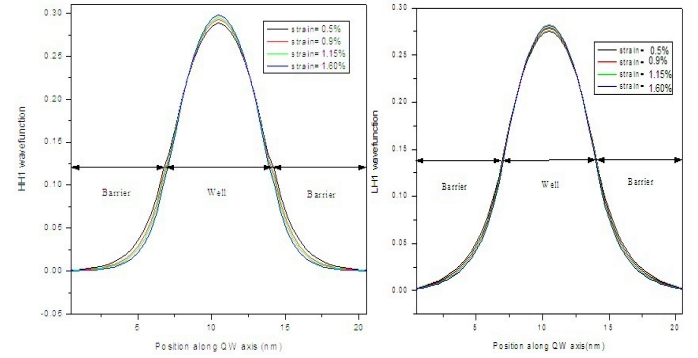


Fig 2: First Heavy Hole (HH1) and Light Hole (LH1) wave function of [113]-oriented InGaAs QW for different strains

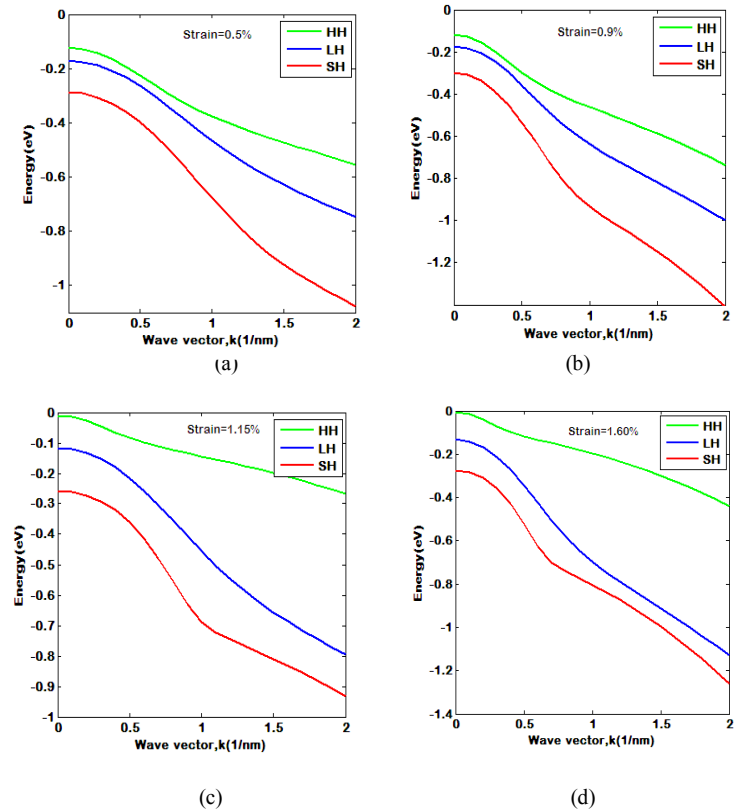


Fig 3: Compressively strained valence band dispersion profile of [113]-oriented InGaAs/GaAs QW

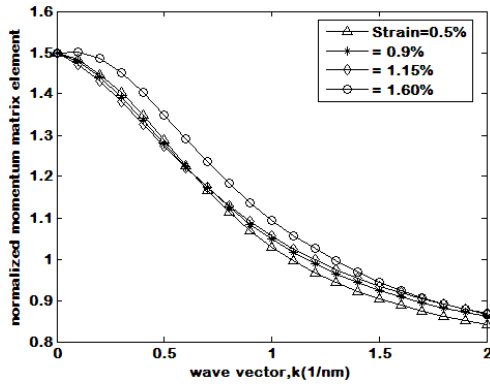


Fig. 4: Strain-dependent momentum matrix element for C1-HH transition

The optical gain spectra are calculated for different values of strain. The results are summarized in fig. 6. For the injection carrier density $2.5 \times 10^{18} \text{ cm}^{-3}$ the peak gains are evaluated to be 2700 cm^{-1} , 2810 cm^{-1} , 3080 cm^{-1} , 3300 cm^{-1} when the well is compressively strained by 0.50%, 0.90%, 1.15%, and 1.60%

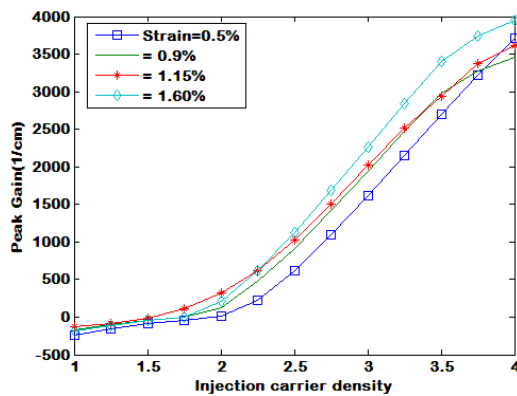


Fig. 5: Carrier density dependent optical gain for compressively strained ZB [113]-oriented InGaAs/GaAs QW

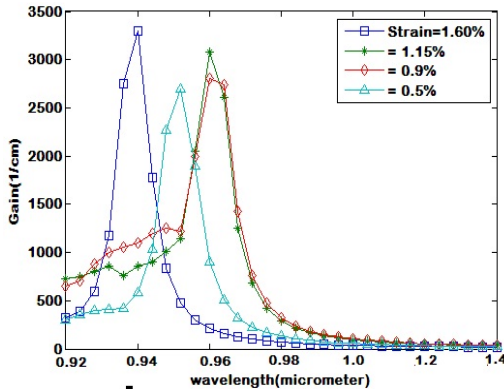


Fig. 6: Optical gain evaluated in different strain conditions for [113]-oriented ZB InGaAs/GaAs QW.

respectively. The value of peak gain increases with increasing strain due to increase of magnitude of momentum matrix element for C1-HH transition. The increased energy separations between HH and LH also contribute in achieving higher gain. The emission wavelength found to change from $0.93 \mu\text{m}$ to $0.97 \mu\text{m}$ as the energy separation increases for C1-HH transition with strain values.

IV. CONCLUSION

To recapitulate, the impact of compressive strain on optical properties of [113]-oriented zinc-blend InGaAs/GaAs QW are theoretically investigated by practicing the eight band $k \cdot p$ method including spin orbit coupling. The theoretical results depict the exclusive variation of conduction and valence band energy band dispersion profile with magnitude of strain. Due to minimum interaction from distant band, larger splitting between heavy hole and light hole band is observed when the QW is 1.60% compressively strained. As the maximum value of momentum matrix element is evaluated when the well is compressively strained by 1.60%, the maximum optical gain is found to be 3300 cm^{-1} . The emission wavelength for the QW shifts towards the lower wavelength with increasing the compressive strain. This result regards that compressively strained [113]-oriented ZB InGaAs/GaAs QW is a highly potential material for the fabrication of high speed optoelectronic devices for exclusive range of $\lambda \sim 0.88\text{-}1.1 \mu\text{m}$.

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Effect of Lightly Doped Drain on the Electrical Characteristics of CMOS Compatible Vertical MOSFETs

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Abstract—Effects of lightly doped drain (LDD) on the 100 nm CMOS compatible vertical MOSFETs are investigated for different body doping. It is found that with the increase of LDD doping drive current of vertical MOSFET increases whereas sub-threshold performance is degraded. The degradation of sub-threshold performance is found to be more prominent at low body doping values. In addition, the threshold voltage of vertical MOSFET is found to decrease with the increase of LDD doping. These effects are explained by the reduction of the effective channel lengths and decrease in the source/ drain series resistances with the increase of LDD doping. This analysis reveals that the body doping of 100 nm vertical MOSFETs should be around $6 \times 10^{17} / \text{cm}^3$ to $1 \times 10^{18} / \text{cm}^3$ with an appropriate anneal to deliver a LDD doping around $5 \times 10^{19} / \text{cm}^3$ for a good sub-threshold characteristics which is very significant for choosing appropriate body doping and LDD doping values for fabricating 100nm CMOS compatible vertical MOSFETs.

Index Terms—Fillet Local Oxidation (FILOX), CMOS, Silicidation, vertical MOSFETs.

I. INTRODUCTION

Thin pillar, fully depleted, surround gate vertical MOSFETs are being researched as candidates for end-of-roadmap CMOS technology because they have advantages such as improved short channel effects and improved drive current [1-2]. However, these devices require advanced lithographic techniques and/or complicated processing to define the ultra thin channel. Thick pillar, surround gate, vertical MOSFETs are also of interest because they offer lithography-independent channel length scaling, decoupling of the gate length from the packing density and an improved current drive per unit silicon area compared with conventional lateral CMOS [3-4]. This approach is particularly attractive for RF applications, since it enables co-integration of short channel vertical RF transistor and conventional planar MOSFETs due to the bulk silicon starting material in the matured CMOS technology.

For several years the challenges of vertical MOSFETs for this application are thought to be high overlap capacitance, susceptibility to dry etch damage and the lack of an appropriate silicidation technology. We have previously reported a CMOS compatible Fillet Local Oxidation (FILOX) Process [4] to reduce the overlap capacitance at the top and bottom of the pillar. Damage occurring during pillar dry etch can be eliminated by sacrificial oxidation/ H_2 anneal [5] and we have also shown how the device architecture and the FILOX process can be optimized to eliminate dry etch damage during silicon gate etch and deliver excellent values of sub-threshold slope [6]. Most recently we have reported a CMOS compatible silicidation technology for surround gate vertical MOSFETs. The technology uses a double spacer comprising a polysilicon spacer for the surround gate and a nitride spacer for the silicidation and is successfully integrated with a Fillet Local Oxidation (FILOX) process, which thereby delivers low overlap capacitance and high drive-current vertical devices [7]. While the aforementioned reports demonstrates an excellent

promise of CMOS compatible ion implanted vertical MOSFETs, these devices have specific source/drain architecture which requires to be properly taken care off for RF applications. Unlike planar MOSFETs, CMOS compatible vertical MOSFETs source/drain region is not symmetrical and no specific extension implantation is provided for lightly doped drain (LDD). Source/drain region is formed by single implantation and the lightly doped drain region is provided by controlling the anneal time. For this reason depending on anneal time the doping of the LDD region may vary which could have unavoidable effects on vertical MOSFET's electrical characteristics. So far there is no report in the literature addressing the effect of LDD doping on the vertical MOSFETs characteristics.

In this article, the effects of lightly doped drain (LDD) doping values on the 100 nm CMOS compatible ion implanted vertical MOSFETs are investigated for different body dopings to see the effect of LDD on different device parameters like, drive current, sub-threshold slope, DIBL and threshold voltage. It is found that with the increase of LDD doping drive current of vertical MOSFET increases whereas sub-threshold performance is degraded. The analysis provides insight into appropriate body and LDD doping for vertical MOSFETs fabrication. It is found that the body doping of 100 nm ion implanted vertical MOSFETs should be around $6 \times 10^{17} / \text{cm}^3$ to $1 \times 10^{18} / \text{cm}^3$ with an appropriate anneal to deliver a LDD doping around $5 \times 10^{19} / \text{cm}^3$ for a good sub-threshold characteristics.

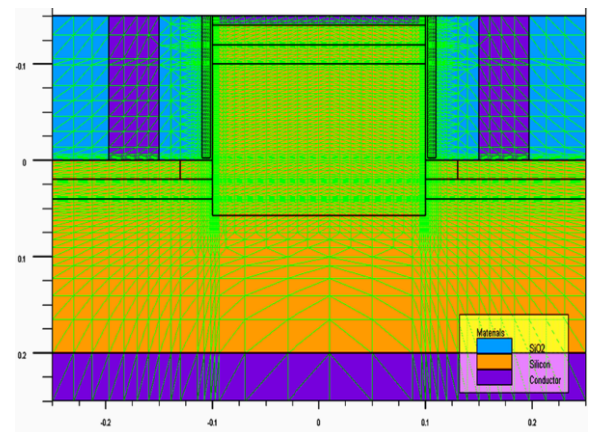


Figure 1: Cross-sectional view of 100 nm CMOS compatible vertical MOSFET.

II. SIMULATION METHODOLOGY

Analysis of the vertical MOSFET was performed with the aid of numerical simulations using the Silvaco Atlas device simulator [8]. A 100 nm channel length vertical ion-implanted double gate nMOSFET as shown in fig. 1, was simulated for different LDD doping values. The gate oxide thickness was 2 nm and the gate electrode chosen was a metal with a work function of 4.5 V. A 2D coupled Poisson's drift-diffusion

solver was used to investigate the device operation. To invoke the dependence of carrier mobility on the parallel and transverse fields in non-planar devices like vertical MOSFETs, the Lombardi CVT model was used [8]. In this model, the dependences of the transverse field, doping and temperature are incorporated that are combined using Mathiessen's rule. The model for carrier emission and absorption processes proposed by Shockley-Read-Hall (SRH) was used to reflect the recombination phenomenon within the device. To account bandgap narrowing (BGN) effects in heavily doped source/drain regions, Slotboom and de Graaf's analytical BGN model was used. The simulation was performed for a heavily doped drain region with a uniform doping concentration of $1 \times 10^{21} / \text{cm}^3$ and for LDD doping values varying from $5 \times 10^{18} / \text{cm}^3$ to $1 \times 10^{20} / \text{cm}^3$. Several body doping concentrations were investigated with values varying from $1 \times 10^{16} / \text{cm}^3$ to $1 \times 10^{18} / \text{cm}^3$.

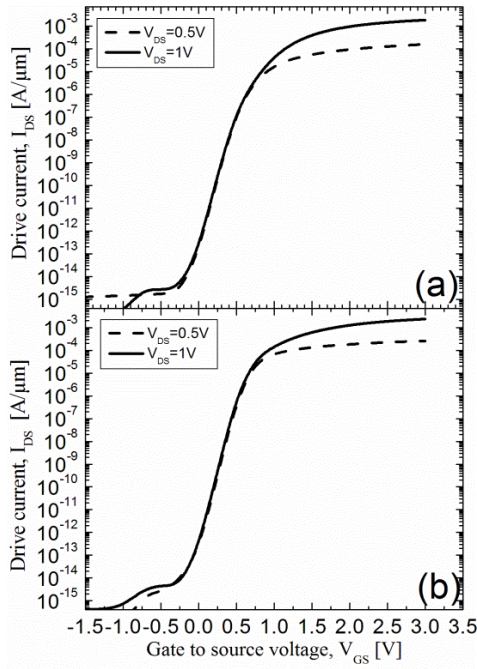


Figure 2: Transfer characteristics of 100nm vertical MOSFETs for a body doping of $1 \times 10^{18} / \text{cm}^3$ and for LDD doping values of (a) $5 \times 10^{18} / \text{cm}^3$ and (b) $5 \times 10^{19} / \text{cm}^3$.

III. ELECTRICAL RESULTS

Figure 2 shows the transfer characteristics of vertical MOSFETs for a body doping of $1 \times 10^{18} / \text{cm}^3$ at two different LDD doping values. For a LDD doping of $5 \times 10^{18} / \text{cm}^3$, in fig. 2(a), the sub-threshold slope of the device is 71 mV/dec and DIBL is 6 mV/V. The threshold voltage is estimated from the transfer characteristics using constant current method and is found to be 1.15 V. For a LDD doping of $5 \times 10^{19} / \text{cm}^3$ in fig. 2(b) the sub-threshold slope, DIBL and threshold voltages of the device is found to be 74 mV/dec, 15 mV/V and 0.85V respectively. This result imply that with the increase of LDD doping value, the sub-threshold performance is slightly degraded at the body doping of $1 \times 10^{18} / \text{cm}^3$. In addition, a slight reduction in the threshold voltage can also be observed with the increase of the LDD doping at this body doping concentration.

Figure 3 shows the output characteristics of vertical MOSFETs for a body doping of $1 \times 10^{18} / \text{cm}^3$ at two different LDD doping values. For a LDD doping of $5 \times 10^{18} / \text{cm}^3$, in fig.

3(a), the drive current for 1 V of gate voltage overdrive at $V_{DS}=1.5\text{V}$ is found to $1241 \mu\text{A}/\mu\text{m}$. For the LDD doping of $5 \times 10^{19} / \text{cm}^3$, the drive current of the device is found to be $1270 \mu\text{A}/\mu\text{m}$. This result implies that with the increase of the LDD doping the drive current of the device increases.

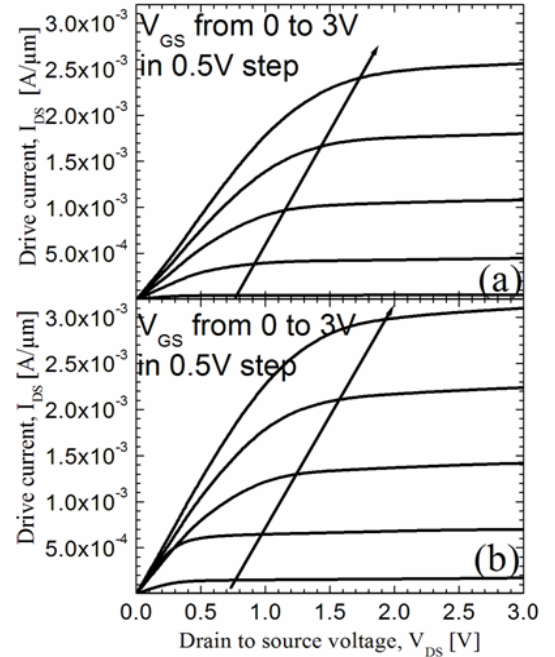


Figure 3: Output characteristics of 100nm vertical MOSFETs for a body doping of $1 \times 10^{18} / \text{cm}^3$ and for LDD doping values of (a) $5 \times 10^{18} / \text{cm}^3$ and (b) $5 \times 10^{19} / \text{cm}^3$.

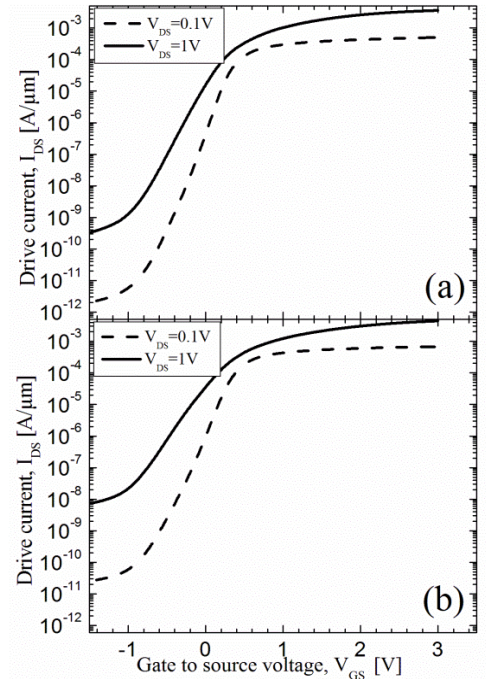


Figure 4: Transfer characteristics of 100nm vertical MOSFETs for a body doping of $1 \times 10^{19} / \text{cm}^3$ and for LDD doping values of (a) $5 \times 10^{18} / \text{cm}^3$ and (b) $5 \times 10^{19} / \text{cm}^3$.

Figure 4 shows the transfer characteristics of 100 nm vertical MOSFETs for a very low body doping of $1 \times 10^{16} / \text{cm}^3$ at two different LDD doping values. For a LDD doping of $5 \times 10^{18} / \text{cm}^3$, in fig. 4(a), the sub-threshold slope, DIBL and the threshold voltage of the device is found to be 232 mV/dec, 406 mV/V and 0.240 V. For a LDD doping of $5 \times 10^{19} / \text{cm}^3$ in fig. 4(b) the sub-threshold slope, DIBL and

threshold voltages of the device is found to be 294 mV/dec, 575 mV/V and 0.16 V respectively. These results again indicate that with the increase of the LDD doping the sub-threshold performance of the device is degraded. A reduction of the thresholds voltages (V_T) of the device is also observed at low body doping values. However, it is worth noting that sub-threshold slope (S) and DIBL values are significantly worse at the body doping of $1 \times 10^{16} / \text{cm}^3$ in comparison to the body doping of $1 \times 10^{18} / \text{cm}^3$.

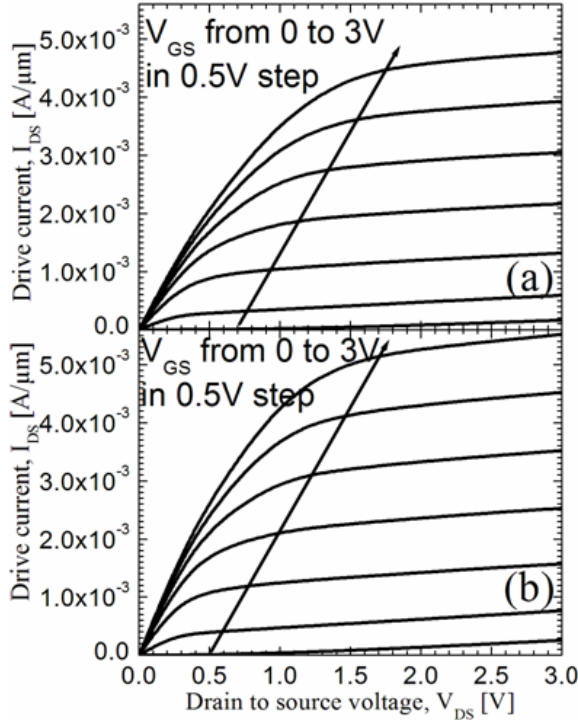


Figure 5: Output characteristics of 100nm vertical MOSFETs for body doping of $1 \times 10^{16} / \text{cm}^3$ and for LDD doping values of (a) $5 \times 10^{18} / \text{cm}^3$ and (b) $5 \times 10^{19} / \text{cm}^3$.

Figure 5 shows the output characteristics of vertical MOSFETs for a body doping of $1 \times 10^{16} / \text{cm}^3$ at two different LDD doping values. For a LDD doping of $5 \times 10^{18} / \text{cm}^3$, in fig. 5(a), the drive current of the device for 1 V of gate voltage overdrive at $V_{DS}=1.5\text{V}$ is found to be $1728 \mu\text{A}/\mu\text{m}$. For the LDD doping of $5 \times 10^{19} / \text{cm}^3$, the drive current of the device is found to be $1834 \mu\text{A}/\mu\text{m}$. These results again imply that with the increase of the LDD doping the drive current of the device increases.

To investigate the device characteristics in detail, we have analyzed vertical MOSFET characteristics for several LDD doping with values varying from $5 \times 10^{18} / \text{cm}^3$ to $1 \times 10^{20} / \text{cm}^3$ and for several body doping concentrations with values varying from $1 \times 10^{16} / \text{cm}^3$ to $1 \times 10^{18} / \text{cm}^3$. Figure 6 shows sub-threshold slope of vertical MOSFETs as a function of lightly doped drain (LDD) for different body doping values. For body doping of $1 \times 10^{18} / \text{cm}^3$ sub-threshold slopes are found to be 71, 73, 74 and 75 mV/dec for LDD doping of $5 \times 10^{18} / \text{cm}^3$, $1 \times 10^{19} / \text{cm}^3$, $5 \times 10^{19} / \text{cm}^3$ and $1 \times 10^{20} / \text{cm}^3$ respectively. This result imply that there is mild degradation of sub-threshold characteristics with increasing LDD doping even in a relatively a heavy body doping of $1 \times 10^{18} / \text{cm}^3$. This trend is found for each body doping values. However, for the lowest body doping $1 \times 10^{16} / \text{cm}^3$ sub-threshold slopes are found to be 232, 294 and 305 mV/dec respectively for the LDD doping of $5 \times 10^{18} / \text{cm}^3$, $1 \times 10^{19} / \text{cm}^3$, $5 \times 10^{19} / \text{cm}^3$ and $1 \times 10^{20} / \text{cm}^3$.

cm^3 showing that the sub-threshold slope degradation with increasing LDD doping is more prominent at $1 \times 10^{16} / \text{cm}^3$ body doping in comparison to $1 \times 10^{18} / \text{cm}^3$ body doping.

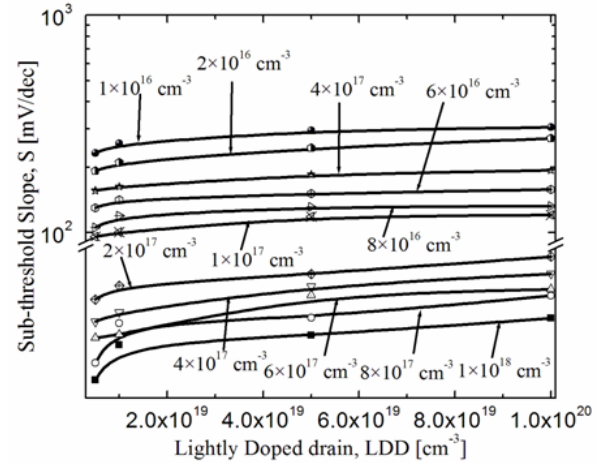


Figure 6: Sub-threshold slope of 100 nm vertical MOSFETs as a function of lightly doped drain (LDD) for different body doping values.

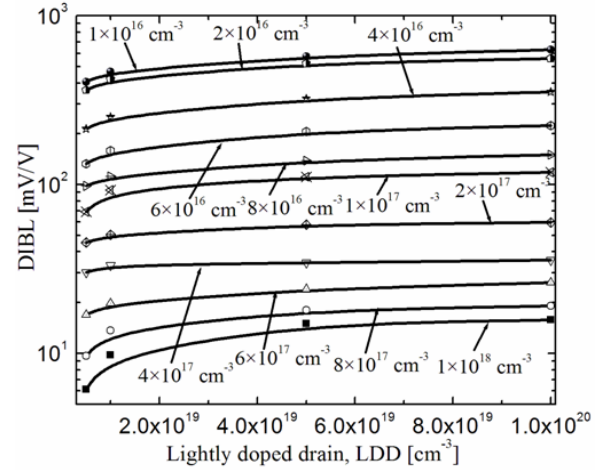


Figure 7: Drain induced barrier lowering (DIBL) of 100 nm vertical MOSFETs as a function of lightly doped drain (LDD) for different body doping values.

Figure 7 shows drain induced barrier lowering (DIBL) of 100 nm vertical MOSFETs as a function of lightly doped drain (LDD) for different body doping values. DIBL characteristics of 100nm vertical MOSFETs at different LDD doping are similar to the sub-threshold slopes at different LDD doping. For body doping of $1 \times 10^{18} / \text{cm}^3$ DIBL values are 6, 10, 15 and 17 mV/V for LDD doping of $5 \times 10^{18} / \text{cm}^3$, $1 \times 10^{19} / \text{cm}^3$, $5 \times 10^{19} / \text{cm}^3$ and $1 \times 10^{20} / \text{cm}^3$ respectively. This result again implying a mild degradation of the sub-threshold characteristics of 100 nm vertical MOSFETs with increasing LDD doping in a relatively a heavy body doping of $1 \times 10^{18} / \text{cm}^3$. For the lowest body doping of $1 \times 10^{16} / \text{cm}^3$ DIBL values are found to be 406, 466, 574 and 630 mV/V for LDD doping of $5 \times 10^{18} / \text{cm}^3$, $1 \times 10^{19} / \text{cm}^3$, $5 \times 10^{19} / \text{cm}^3$ and $1 \times 10^{20} / \text{cm}^3$ respectively. This result again exhibits that DIBL degradation of 100 nm vertical MOSFET is more prominent at the body doping $1 \times 10^{16} / \text{cm}^3$ in comparison to the body doping $1 \times 10^{18} / \text{cm}^3$.

Figure 8 shows drive current of 100 nm vertical MOSFETs as a function of lightly doped drain (LDD) for different body doping values. The drive current are shown for $V_{DS}=1.5\text{V}$ and

for 1 V of gate voltage over drive. In general the current drive of 100 nm vertical MOSFET follows the trend of sub-threshold characteristics at different LDD and body doping values which increases with increasing LDD doping at each body doping.

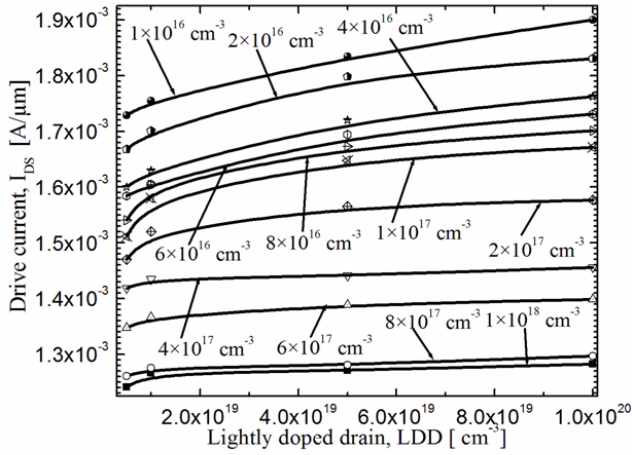


Figure 8: Drive current of 100 nm vertical MOSFETs as a function of lightly doped drain (LDD) for different body doping values.

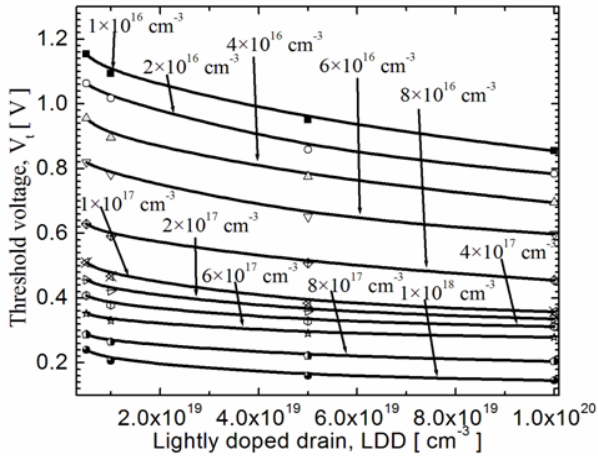


Figure 9: Threshold voltage (V_T) of 100 nm vertical MOSFETs as a function of lightly doped drain (LDD) for different body doping values.

Finally, figure 9 shows threshold voltage (V_T) of 100 nm vertical MOSFETs as a function of lightly doped drain (LDD) for different body doping values. For a body doping of $1 \times 10^{18} / \text{cm}^3$ V_T values are found to be 1.15, 1.1, 0.95 and 0.85V for LDD doping of $5 \times 10^{18} / \text{cm}^3$, $1 \times 10^{19} / \text{cm}^3$, $5 \times 10^{19} / \text{cm}^3$ and $1 \times 10^{20} / \text{cm}^3$ respectively exhibiting a decrease of V_T with increasing LDD doping in 100 nm vertical MOSFETs. This trend is found for each body doping values and for the lowest body doping of $1 \times 10^{16} / \text{cm}^3$ V_T values are found to be 0.24, 0.20, 0.16 and 0.14V respectively for LDD doping of $5 \times 10^{18} / \text{cm}^3$, $1 \times 10^{19} / \text{cm}^3$, $5 \times 10^{19} / \text{cm}^3$ and $1 \times 10^{20} / \text{cm}^3$. Unlike sub-threshold slope and DIBL characteristics the threshold voltage (V_T) change with LDD doping is more prominent for body doping of $1 \times 10^{18} / \text{cm}^3$ in comparison to the body doping of $1 \times 10^{16} / \text{cm}^3$.

IV. DISCUSSION

The electrical results in section 3 exhibits some non-trivial effects of LDD doping on the electrical characteristics of CMOS compatible vertical MOSFETs such as sub-threshold slope, drain induced barrier lowering (DIBL), threshold voltage, and drive current. An increase of the drive current was

observed with the increase of LDD doping for any fixed body doping. As the source to drain series resistance of the device is reduced due to increase of LDD doping, the drive current is expected to increase. In addition to the source/ drain series resistances, effective channel length of device also decreases with the increase of LDD doping which also increase the drive current. The increase in the drive current with the decrease of the body doping can be explained by the onset of reduction of threshold voltages and significant reduction of the channel lengths with the decrease of body doping. The degradation of sub-threshold characteristics with the increase of LDD doping is also due to the similar phenomenon of effective channel length reduction with the increase of LDD doping for any fixed body doping. However, this effect is more prominent at low body doping values as effective channel length reduction with the increase of LDD doping is expected to be high. Similar phenomenon can be attributed to the reduction of threshold voltages with the increase of LDD doping. Based on the aforementioned analysis it can be decided that the body doping of 100 nm vertical MOSFETs should be around $6 \times 10^{17} / \text{cm}^3$ to $1 \times 10^{18} / \text{cm}^3$ with an appropriate anneal to deliver a LDD doping of $5 \times 10^{19} / \text{cm}^3$ for a good sub-threshold characteristics with drive currents around.

V. CONCLUSION

We have investigated the effect of the variation of LDD doping in CMOS compatible vertical MOSFETs architecture. It is found that with the increase of LDD doping drive current of vertical MOSFET increases whereas sub-threshold performance is degraded. The degradation of sub-threshold performance is found to be more prominent at low body doping values. In addition to this threshold voltage of vertical MOSFETs are found to decrease with the increase of LDD doping. These effects are explained by the reduction of the effective channel lengths and decrease in the source/ drain series resistances with the increase of LDD doping values. These results are very significant for choosing appropriate body doping and LDD doping values for fabricating 100nm CMOS compatible vertical MOSFETs.

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Numerical Analysis of Performance Improvement of Modified ITO Based Organic Light Emitting Diode

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Abstract— The effects of indium tin oxide (ITO) based anode work function modification on charge injection at anode/organic layer interface and charge transport into the organic layers of organic light emitting diode (OLED) have been investigated quantitatively through this numerical simulation work. The improvement of device performances due to this anode surface modification also explained in this work in terms of simulated current density-voltage-luminance (J-V-L) characteristic, efficiency and recombination zone.

Keywords— work function; indium tin oxide (ITO); numerical simulation; charge injection; recombination rate

I. INTRODUCTION

In the field of displays and solid state lightings technologies, multilayer organic light emitting diodes (OLEDs) have been covered most of the area over the other lighting sources like light emitting diodes (LEDs) and liquid crystal displays (LCDs) because of their fast turn on response time ($<1\mu s$), higher refresh rates, low turn on voltage, low input power (portable use), wide viewing angle and low cost [1, 2]. Among these advantages, low efficiency and short lifetime are the common drawbacks of OLEDs which can be improved using different kind of multilayer [2] and mixed host emissive layers based OLEDs [3, 4]. However, reduction of energy barrier (increase the charge injection) between the anode/organic layer interface using the modification of surface work function of indium tin oxide (ITO) which is commonly used as anode in most optoelectronic devices, is the recent development to increase the OLED efficiency and life time [5, 6]. Therefore, fabrication of different complex multilayer OLED to improve the device performance and lifetime can be avoided easily using the modified ITO in simpler single or bi-layer device structure.

O₂ and Cl₂ plasma treatment, post annealing treatments and UV-ozone treatments have been widely used as ITO work function modification techniques [7, 8]. However, still there is energy barrier for hole injection in the ITO/organic interface after applying these surface modification techniques. Recently, H.-W. Lu, C.-W. Huang, P.-C. Kao, S.-Y. Chu, and Y.-D. Juang. [9] reported simple surface modification

techniques of ITO by soaking in 5 wt% and 10 wt% concentration of caesium fluoride (CsF) solution and subsequent in UV-ozone treatment where the work function of ITO modified from normal work function of 4.78 eV to 5.01 eV, 5.11 eV and 4.91 eV correspondingly. In that work, they explained the effects of ITO surface modification on the J-V-L and surface morphology through the qualitative study.

Though, it is possible to explain some effects of OLED qualitatively by the systematic study of the device performance but not possible to explain the effect quantitatively [10]. Motivating from that point of view, a complete numerical simulation to analyze the effects of modified ITO work function on the ITO/CsF/NPB/Alq₃/LiF/Al based bilayer OLED [9] performance have been presented in this work using the ‘SETFOS’ (Semiconducting Thin Film Optics Simulation Software) simulator. Also, the effects of ITO properties after surface modification on hole injection and internal device physics explained quantitatively from the simulation of charge density and mobility, distribution of electric field, carriers exciton density and recombination zone.

II. NUMERICAL MODELING

Using a numerical simulator called ‘SETFOS’ which cover all key physical processes in OLEDs, effects of anode work function modification on Alq₃ based bilayer OLED performance have been simulated in this work. Drift-diffusion model considered as simulation model which is integrated with charge injection, transport, exciton density and recombination equations. The electric field distribution in the device and the time evaluation of the system calculated by the Poisson and the continuity equations as presented in “(1),” and “(2),” correspondingly.

$$\frac{dE}{dx} = \frac{q}{\epsilon\epsilon_0} (p - n) \quad (1)$$

$$\frac{dn}{dt} = \frac{dJ_n}{qdx} - R - T \quad (2)$$

where, E is the electric field inside the OLED, q the elementary charge, ϵ the dielectric permittivity of the organic materials, ϵ_0 the dielectric constant, p and n are the hole density and the electron density respectively, J_n the electron current density, R the recombination rate and T the trapping rate for free carriers. The current density composed of a drift and diffusion carriers, and the recombination rate calculated using “(3),” and “(4),” respectively.

$$J_n = qp\mu_n E + qD_n \frac{dn}{dx} \quad (3)$$

$$r = \frac{q}{\epsilon\epsilon_0} (\mu_p + \mu_n) \quad (4)$$

Here, μ_p and μ_n are the hole and electron mobility respectively, D_n the diffusion constant. The carrier mobility in the organic materials calculated by the conventional electric field dependent Poole-Frenkel mobility expressed as:

$$\mu(E) = \mu_0 \exp(\gamma^* \sqrt{E}) \quad (5)$$

where, $\mu(E)$ is the charge carrier mobility under electric field (E), μ_0 is the mobility under zero electric field and γ is the Poole-Frenkel field dependent factor.

III. SIMULATION PARAMETERS FOR BI-LAYERS OLED

The structure used in this ‘SETFOS’ simulator is ITO/CsF/NPB (40 nm)/Alq₃ (60nm)/LiF/Al. As input parameters, the HOMO and LUMO levels of NPB (Alq₃) are 5.4 eV (5.7 eV) and 2.3 eV (3.0 eV) and the work function of LiF/Al is 3.0 eV [3, 9]. The work function of ITO is 4.78 eV at normal condition which is changed to 5.01 eV and 5.11 eV after soaking in 5 wt% and 10 wt% CsF solution respectively. When CsF thin film deposited (thermal evaporation technique) on ITO as buffer layer using UV-ozone treatment then the work function of ITO become 4.91 eV. All of these work functions of ITO measured using UV-photoelectron spectrometer [9]. The mobility (electron and hole) parameters for organic materials (NPB and Alq₃) those are determined by time of flight (TOF) measurement technique used in the simulation [11, 12]. Considering these four different work functions of ITO, the performance of bi-layers OLED has been investigated and explained in the next section.

IV. SIMULATION RESULTS AND DISCUSSIONS

A. ELECTRICAL CHARACTERISTICS

The current density-voltage-luminance (J-V-L) curves as shown in “Fig. 1,” indicated that with the increase of work function at anode side (ITO) both the current density and luminance increased due to higher hole injection rate. In both experimental [9] and simulation work, the OLEDs based on modified ITO soaked in 5 wt% and 10 wt% CsF solution, and UV-ozone treated CsF buffer layer not only showed the higher values of current density and luminance, but also reduced the turn-on voltage of OLEDs specially after modification of ITO work function (5.11 eV) with 10 wt% CsF solution compared with OLED based on unmodified ITO work function (4.78 eV).

eV). The modified ITO with 10 wt% CsF solution also give the higher electroluminescence (EL.) efficiency (γ) than the other modification process as presented in “Fig. 2,”. These performance improvements of bilayer OLED using ITO surface modification technique by CsF solution can be explained from the charge density, electric field, exciton density and recombination zone.

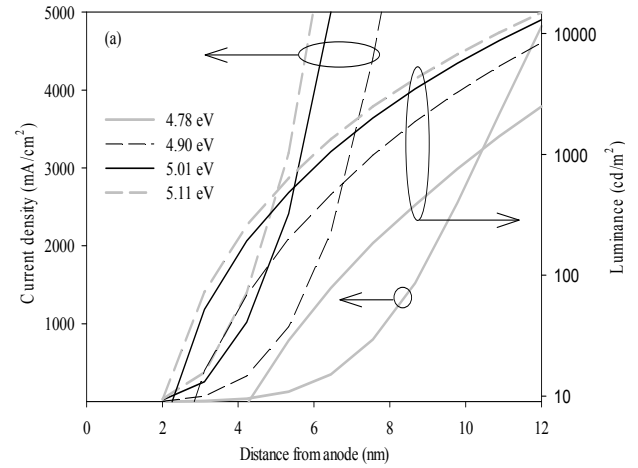


Fig. 1. J-V-L Characteristics of OLEDs where the work function of ITO changed from 4.78eV (normal) to 5.11 eV (after modification)

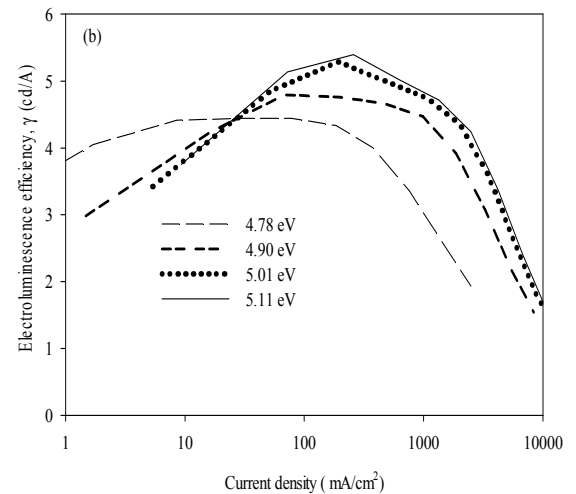


Fig. 2. Electroluminescence efficiency-vs-current density curves of unmodified and modified ITO based OLEDs

B. EFFECT OF MODIFIED ITO ON CHARGE DENSITY AND RECOMBINATION

In normal condition, the interfacial energy barrier between ITO (4.78 eV) and NPB (HOMO=5.4 eV) is 0.62 eV which is reduced to 0.29 eV by modifying the ITO surface with 10 wt% CsF solution (5.11 eV). Therefore, more than 2 times reduction of interfacial energy barrier increased the hole injection rate in the anode side significantly. Hence, the hole mobility increased near the organic/organic interface and also in Alq₃ layer with ITO surface modification techniques at 6V as shown in “Fig. 3,”. Modification of ITO work function specially at 5.11 eV (at applied voltage of 6V) also increased the hole density around 2 orders higher than the normal work function of 4.78 eV which observed in “Fig. 4,”. As a result, the electric field along the NPB layer increased until organic/organic interface due to this improved work function which correspondingly increased the exciton rate of carriers at that interface as presented in “Fig. 5,”. The most important evidence of device performance improvement by surface modification of ITO shown in “Fig. 6,” where the recombination rate increased with respect to ITO work function improvements over the active device area.

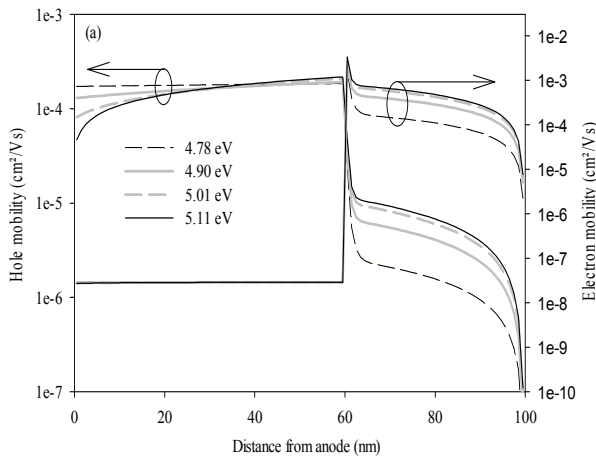


Fig. 3. Electron and hole mobilities with different work functions of ITO based OLEDs at 6V

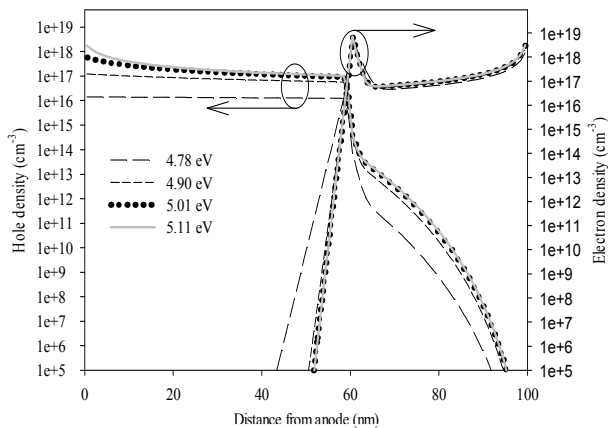


Fig. 4. Electron and hole densities with different work functions of ITO based OLEDs at 6V

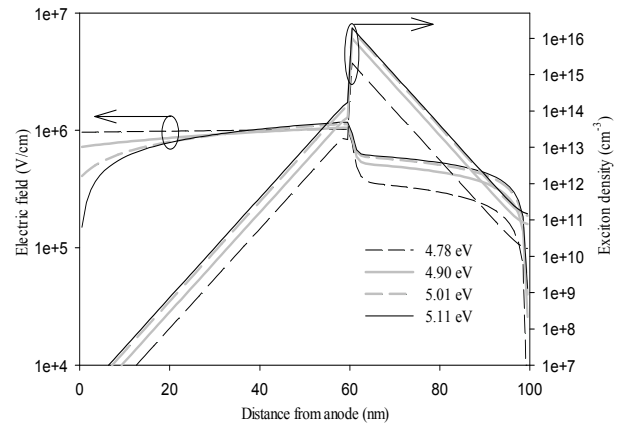


Fig. 5. Electric field distribution and exciton densities inside the OLED with respect to different work functions of ITO at 6V

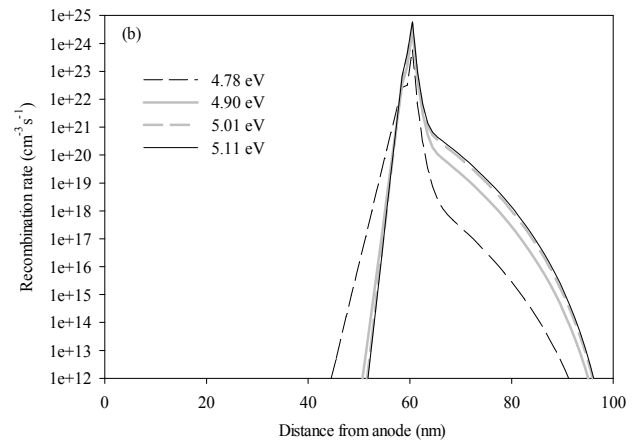


Fig. 6. Recombination zone inside the OLED with respect to different work functions of ITO at 6V

C. DISCUSSIONS

Using numerical simulation, the effects of different work functions of ITO (resulting from surface modification processes) on electrical characteristics, efficiency and recombination rate have been compared in TABLE I. Among the different ITO surface modification processes, ITO soaking in different concentrations of CsF solution method improved the device efficiency around 1.2 times than the unmodified ITO (normal work function of 4.78 eV) based OLED whereas UV-ozone treatment method improved device efficiency around one time. In the J-V-L characteristics curve, OLED with modified ITO showed the better performance over OLED with unmodified ITO at 6V. The increase of work function of ITO (from 4.78eV to 5.11 eV) with these surface modification methods reduced the Fermi levels between the ITO and NPB which significantly enhanced the hole injection. Hence, the hole density and mobility increased in the NPB layer which effect the recombination rate at the organic/organic interface. Under the active area, the recombination rate of OLED with ITO modification by soaking in 10 wt% CsF solution, increased around one order higher than the unmodified ITO based OLED.

TABLE I. COMPARISON OF OLEDS PERFORMANCES WITH AND WITHOUT ITO MODIFICATION

Improvement Terms		Work function of ITO (eV)			
		Unmodified	After Surface Modification		
		4.78	4.91	5.01	5.11
At 6V	Current density (mA/cm ²)	13	104	3648	13722
	Luminance (cd/m ²)	54	321	761	1083
	Recombination rate under the effective area (cm ⁻³ s ⁻¹)	6.51×10 ²³	3.16×10 ²⁴	5.61×10 ²⁴	6.61×10 ²⁴
Max. EL Efficiency, γ (cd/A)		4.4	4.8	5.3	5.4

V. CONCLUSIONS

In this ‘SETFOS’ based numerical simulation work, it is clearly investigated that the performance of bilayer OLED can be improved significantly in terms of J-V-L characteristic and efficiency curves using ITO surface modification techniques. Moreover, the impact of ITO work function modification on hole injection into the ITO/organic layer interface explained from the simulated charge density, mobility and recombination zone. Due to ITO surface modification, the work function improved from 4.78 eV to 5.11 eV which reduced the hole injection barrier at anode/organic layer interface and correspondingly increased the charge injection, transport and recombination rate under the active area. Therefore, significant improvement of modified ITO based OLED performances have been observed in this quantitative work by numerical simulation which also able to compare with experimental work [9].

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Characteristics of Triple Material Gate AlGa_N/Ga_N High Electron Mobility Transistor

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Abstract—A two dimensional simulation study has been carried out to investigate the characteristics of a newly proposed AlGa_N/Ga_N High Electron Mobility Transistor (HEMT) device with triple material gate (TMG) structure. Surface potential and electric field distribution along the device channel are comprehensively studied and compared with those of dual material gate (DMG) and single material gate (SMG) HEMT. TMG HEMT shows better suppression of drain induced barrier lowering (DIBL) and hot carrier effect (HCE) due to the formation of two steps screening effect in channel potential and having less amount of peak electric field at the drain end. Effect of different device parameters e.g. individual channel length related to different gate material, total channel length and externally applied voltage on the surface potential and electric field are also explored and proper choice of the device parameters and bias voltages is optimized to improve device performance, achieve high speed and reduce short channel effects (SCE) effectively.

Keywords—HEMT; TMG; DIBL; Short channel effect; Hot carrier effect

I. INTRODUCTION

High Electron Mobility Transistor (HEMT) technology is revealing the prospects for high speed low power very large scale integration. Numerous exciting features like fast switching speed, less power consumption, comparatively easy fabrication process and superior performance in high frequency application have made HEMT an excellent device to be used in microwave analog circuits for mobile communication [1], very high speed integrated circuits (VHSIC) and micro and millimeter wave frequencies for low noise amplification purpose [2-4]. HEMT operates by forming a channel consisted of two dimensional electron gas (2DEG) at the heterointerface of a heavily doped wide bandgap and a lightly doped narrow bandgap material [5]. AlGa_N/Ga_N HEMT are perfect among various types of materials used to fabricate HEMT for high power and high temperature devices because of large energy bandgap, high electron mobility, high saturation velocity, low thermal impedance and high breakdown field.

Recent study has revealed that various short channel effects and degradation of carrier transport efficiency have become predominant with the continuous effort of

downscaling the channel lengths of conventional HEMTs into nanometer regime to achieve high speed, enhanced performance and high packing density [6,7]. Reduction in channel length results in extra control of drain over the channel charge in comparison with gate terminal because of more amount of penetration of electric flux from the drain into the channel. The lesser amount of gate control over the channel leads to different short channel effects like drain induced barrier lowering (DIBL) and gate leakage current (hot carrier effect) resulting in degradation of device reliability. Various approaches have been proposed to minimize the short channel effects due to continuous downscaling of the device dimensions and ensure better performance at high speed. One efficient approach is to form the gate contact using more than one material with different work functions which results in step function in the surface potential profile and leads to screening the effect of drain potential on the source-channel barrier of the device. This approach was first proposed by Long et al. [8,9] for MOSFETs and the new structure was Dual Material Gate (DMG) MOSFET. In the DMG MOSFET, the gate contact is comprised of two materials with different work functions connected laterally. The metal gate with higher work function is control gate while the other gate with lower work function is called screen gate [9]. This structure provides better suppression of SCEs because of having a step shape surface potential and improvement in carrier transport velocity as a result of higher average electric field in the channel.

Inspired by this structure, Kumar et al. utilized this approach and proposed a new structure named Dual Material Gate HEMT (DMG HEMT) [10]. Better performance in terms of reduction of SCEs and increase of carrier transport efficiency in DMG HEMT compared to conventional single material gate HEMT (SMG HEMT) has been investigated. In recent times, some theoretical, experimental [11-14] and simulation works are published on the Triple Material Gate (TMG) structure for MOSFET. TMG structure has two steps in the surface potential profile compared to one step in DMG structure which results in better screening of the effects of drain voltage on the channel. Moreover, TMG structure has less amount of peak

electric field at the drain end than the corresponding DMG structure and reduces the hot carrier effect efficiently. But, till now, no model has been developed to study the effect of TMG structure in HEMTs. In this work, we have developed a two dimensional model to explore the characteristics of a AlGaIn/GaN HEMT device with Triple Material Gate structure (TMG HEMT). Simulated surface potential profile and electric field distribution along the channel position for this new structure have been presented. Variation of the channel potential and electric field with different device parameters like length of each metal gate, total channel length and drain to source voltage are studied thoroughly and compared to those of DMG and SMG HEMT to recognize the better performance in terms of suppression of SCEs and HCEs.

II. DEVICE STRUCTURE

The schematic structure of a TMG $\text{Al}_{0.3}\text{Ga}_{0.7}\text{N}/\text{GaN}$ HEMT is shown in Fig. 1. Three metals, M1, M2 and M3 having different work functions, ϕ_{M1} , ϕ_{M2} and ϕ_{M3} and lengths of $L1$, $L2$ and $L3$ respectively are connected laterally to form the gate electrode with a length of L . Work functions and gate lengths of the metals are chosen such that $\phi_{M1} > \phi_{M2} > \phi_{M3}$ and $L1+L2+L3 = L$. The leftmost gate material in the figure with the highest work function (ϕ_{M1}) is known as the control gate which is placed near the source end, the middle material with the intermediate work function (ϕ_{M2}) is termed as the first screen gate and the rightmost gate metal positioned near the drain end having the lowest work function (ϕ_{M3}) is called the second screen gate. In our simulation, ϕ_{M1} , ϕ_{M2} and ϕ_{M3} are taken as 5.2 V, 4.7 V and 4.4 V respectively for TMG HEMT; for DMG HEMT $\phi_{M1} = 4.7$ V and $\phi_{M2} = 4.4$ V and in case of SMG HEMT, $\phi_M = 4.4$ V. Uniform doping concentration of $N_d^+ = 10^{25} \text{ m}^{-3}$ is used for the source and drain. AlGaIn barrier layer is also uniformly doped at doping concentration of $N_d = 10^{23} \text{ m}^{-3}$ and has a thickness of $d_d = 25$ nm. An undoped AlGaIn spacer layer of thickness of $d_i = 5$ nm is included in

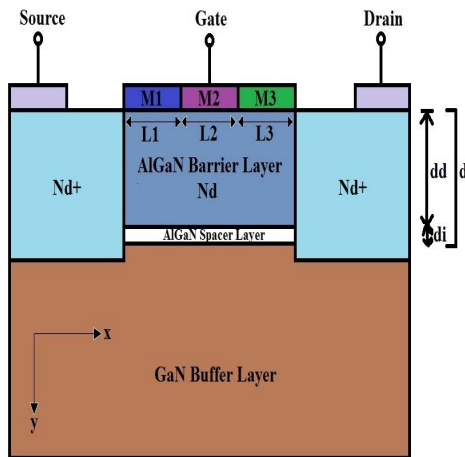


Fig. 1. Cross sectional view of a Triple Material Gate $\text{Al}_{0.3}\text{Ga}_{0.7}\text{N}/\text{GaN}$ HEMT.

between AlGaIn barrier layer and GaN buffer layer to reduce impurity scattering and increase carrier mobility. The directions of x and y axis are shown in the figure. $y=0$ location corresponds to the front interface where $d=d_d+d_i$ is the thickness of the total AlGaIn layer. The thickness of the GaN buffer layer is taken to be $0.2 \mu\text{m}$. The total channel length is set to be 120 nm . We investigate the different characteristics of the device structure using the 2-D device simulation software ATLAS [15].

III. RESULTS AND DISCUSSION

Fig. 2 shows the surface potential of different structures along the position of the channel. It is evident from the figure that TMG HEMT has the smallest slope of the surface potential curve among the three at the drain end which symbolizes that the TMG structure suffers from least amount of hot carrier effect. There are two steps in surface potential profile of TMG compared to only one step and no step in DMG and SMG HEMT respectively. The steps occur at the interface of the two different metal gates because of their work function variation. More step changes confirm better screening of the area under the control gate (channel region under M1) from the drain to source voltage, V_{ds} . For the TMG HEMT, M2 and M3 gates work as first and second screen gate respectively (whereas there is only one screen gate M2 for DMG HEMT) and extra V_{ds} is absorbed by the screen gates beyond the saturation point. So there is a significant reduction in DIBL effect in TMG structure because of very little impact of drain voltage on drain current after saturation.

Fig. 3 shows the variation of surface potential for three different combinations of $L1$, $L2$ and $L3$ keeping the total gate length, L ($L1+L2+L3$) constant. The gate length under the control gate, $L1$ is 20 nm , 40 nm and 60 nm while the ratios of $L1:L2:L3$ are equal to $1:2:3$, $2:3:1$ and $3:2:1$ respectively. It is evident from the figure that with the

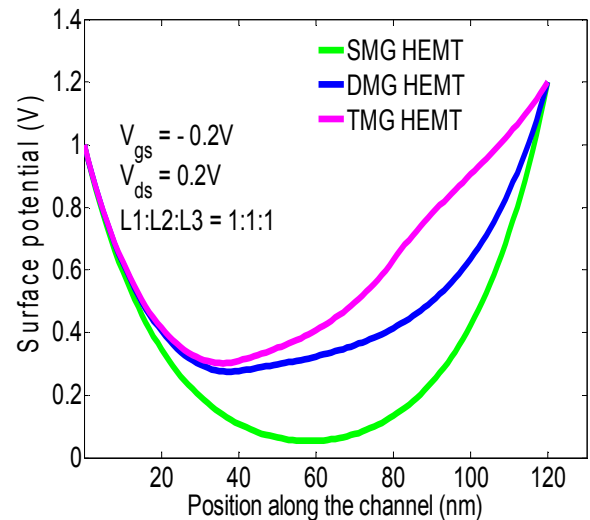


Fig. 2. Surface potential variation along the channel position (TMG HEMT compared with DMG and SMG HEMT).

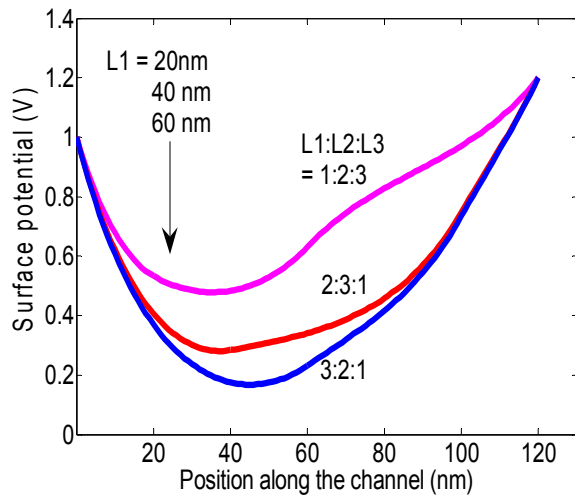


Fig. 3. Surface potential versus channel position in TMG HEMT for different ratios of L1, L2 and L3.

decrease of L1, the position of the minimum surface potential shifts gradually towards the source. It results in a more uniform distribution of electric field in the channel as the peak electric field is also shifted towards the source end. Hot carrier effect is minimized with the reduction of L1 due to this uniformity of electric field profile. The figure also reveals that minimum channel potential increases when L1 gets reduced. Increase of channel potential minima causes lowering of potential barrier and increase of current. DIBL also increases with the reduction of the source-channel barrier. Essentially, the decrease of L1 means the reduction of the control on the channel by the gate metal having larger work function and therefore different short channel effects like DIBL becomes larger which was also true for some other structures like triple material surrounding gate MOSFET [16] and tri material gate FinFET [17].

Fig. 4 demonstrates the effect of channel length L on the surface potential variation along the channel position.

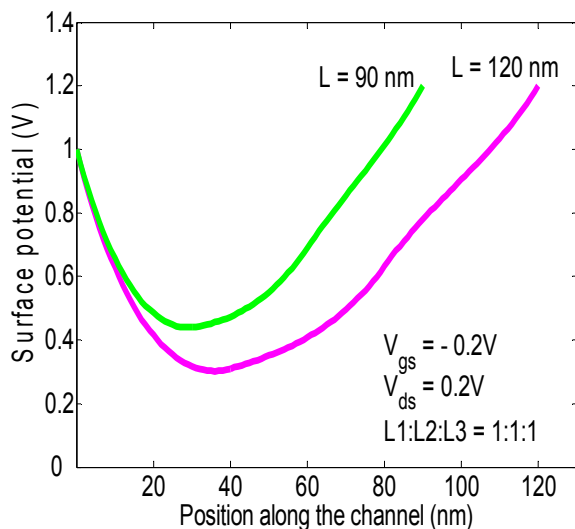


Fig. 4. Surface potential along the position of the channel in TMG HEMT for total channel lengths of 90nm and 120nm.

Minimum channel potential rises and position of the minima moves towards the source end with the decrease of the channel length. Thus the channel barrier becomes smaller and short channel effects become significant.

Fig. 5 exhibits surface potential variation in TMG structure along the channel position as a function of drain to source voltages. Minimum value of surface potential increases with the increase in V_{ds} which leads to the reduction in potential barrier and enhancement of carrier transport efficiency.

Fig. 6 shows the electric field profile of TMG, DMG and SMG HEMTs along the position of the channel. Electric field profile of TMG HEMT exhibits two peaks whereas there appears a single peak and no peak for DMG and SMG HEMT, respectively. The peaks actually occur at the interface of two gate metals with different work functions which increase the average electric field in the channel.

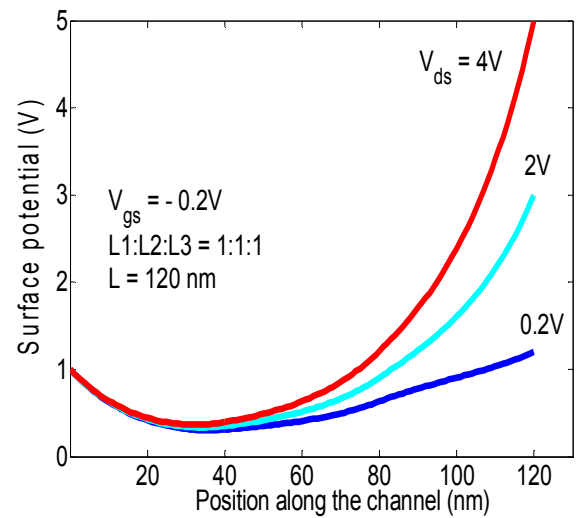


Fig. 5. Surface potential along the position of the channel in TMG HEMT for three distinct drain to source voltages, $V_{ds} = 0.2V$, $2V$ and $4V$.

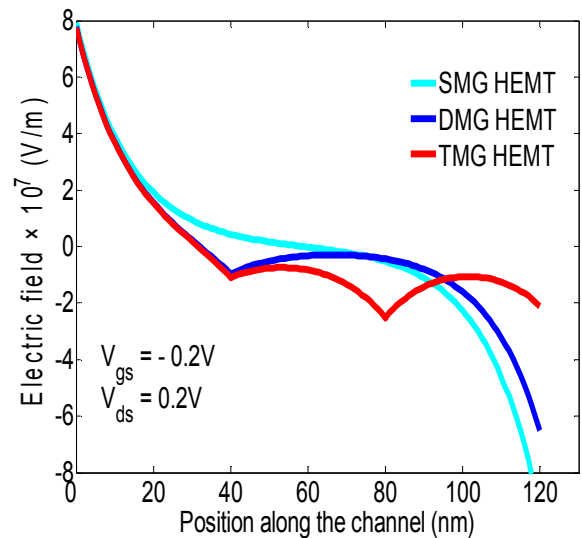


Fig. 6. Electric field distribution along the channel position for SMG, DMG and TMG HEMT at $V_{gs} = -0.2V$ and $V_{ds} = 0.2V$.

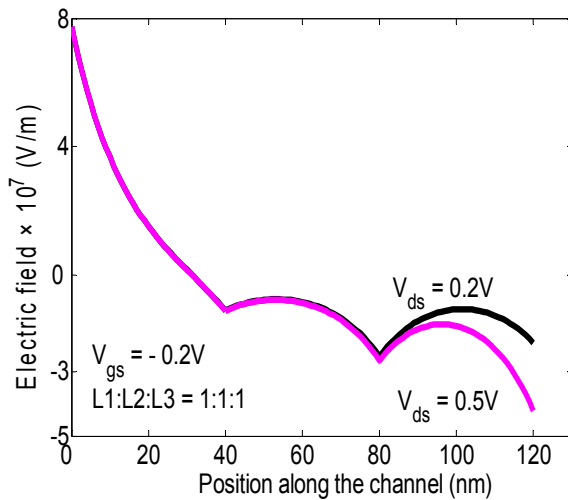


Fig. 7. Electric field variation as a function of channel position of TMG HEMT structure for two different drain to source potential, V_{ds} of 0.2V and 0.5V.

Electron transport velocity is increased because of this enhancement of average field. Moreover, lifetime of the device is increased because of the existence of the electric field peak underneath the gate. Another fact, clearly visible from the figure is that, the peak electric field at the drain end has the smallest value for TMG HEMT compared to DMG and SMG HEMT. Therefore, TMG structure suffers from least amount of hot carrier effects.

Fig. 7 shows electric field variation against channel position for different drain to source voltages, V_{ds} . The plot reveals that peak electric field at the drain end increases with the rise of V_{ds} resulting in boost of hot carrier effect.

IV. CONCLUSION

In this work, we have developed a simulation model of triple material gate HEMT and explored the device characteristics and its performance. The developed structure is compared with DMG and SMG HEMT in terms of reduction of DIBL effect and hot carrier effect. It is found that TMG HEMT provides enhancements and excels the other two structures in various aspects e.g., increased average electric field in the channel, better carrier transport efficiency, decreased DIBL effect and HCE. Effect of combinations of various channel length ratios under different gate material on the surface potential has been investigated and it is shown that the length of the control gate (L1) should be larger to reduce DIBL effect. The paper also presents the suitable device parameters to improve the performance of the proposed structure.

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Electronic Structure of Stacking Dependent Tri-layer Graphene on Cu Substrate

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Abstract—Free-standing tri-layer graphene shows stacking dependent band gap when external bias is applied. In this work, we have made an attempt to observe the electronic structure of tri-layer graphene with different stacking on Cu (111) substrate without applying any external bias. We have performed density functional theory (DFT) calculation and observe that band gap occurs and doping takes place due to the interaction of graphene-metal surface. We have also calculated density of states for different stacking systems. We observe that the band gap and doping of graphene on substrate largely depend on the stacking system and such study would be beneficial for material science as well as electronic industry.

Index Terms—Band-Structure, Cu (111) Substrate, Density of States, Mexican Hat, Tri-layer Graphene.

I. INTRODUCTION

Due to high conductivity of graphene, it has always been systematically studied with care for novel applications. The adsorption of graphene over metal is being studied from different perspectives. Interesting properties have always been found when graphene is placed over metal. Single layer graphene (SLG) has been studied in details both experimentally and using simulation techniques. A study of SLG on Cu (111) substrate has shown the interface structure of the system [1]. But when we move beyond the field of mono-layer graphene, an important thing is to be considered called stacking. In case of bi-layer graphene, we know only about Bernal Stacking (AB). With suitable Electric field, tunable band gap can be introduced in bi-layer graphene [2]. But when we move to tri-layer graphene, there come two types of stacking called Bernal stacking and rhombohedral stacking [3]. In Bernal stacking of tri-layer graphene (ABA), the atoms at uppermost layer are actually on top of the bottom layer. On the other hand, in rhombohedral stacking (ABC), we have to consider the center point of the hexagons in the bottom layer [4]. One sub-lattice of the uppermost layer is made sure to be above the center of the bottom layer hexagons. The presence of twist is seen in case of bi-layer and tri-layer graphene [5].

Experimental result using infrared absorption spectroscopy shows that there is a good amount of dependence of stacking in the electronic band-structure of tri-layer graphene [6].

Applying transverse electric field, it has been shown that we get an asymmetry band gap in case of bi-layer graphene. The structure is called “Mexican hat” structure [7]. Similar structure is also found using perpendicular electric field experimentally in tri-layer graphene. Tri-layer graphene itself is a semi-metal [8]. Hence stacking dependence of band gap in tri-layer graphene should be interesting.

Bernal stacking (ABA) in tri-layer graphene does not include any significant band gap in comparison to rhombohedral (ABC) stacking when external electric field is applied [9]. This phenomenon occurs because of weak and strong dimer in Bernal and rhombohedral stacking, respectively [10]. But external electric field was applied in each of the mentioned cases. The adsorption of single layer graphene (SLG) over different metal substrates (Al,Cu,Ag,Au,Pt) have been studied with their (111) plane using various systematic simulations [11]. From the simulation results of single layer graphene (SLG), we have come to know about the band-structure of each particular case. The Dirac cone at K point in case of these metal (111) substrates, stay intact but with different amount of doping induced for different substrates. Because of these substrates in case of few layer graphene, an internal electric field is introduced on the interface of graphene-metal and we expect similar “Mexican hat” structure in the $E - k$ diagram (band-structure). As we have dealt with Cu substrate, a question may arise for simulating with only Cu (111) plane as substrate in spite of significant works of graphene with Cu (100) plane. The answer is simple. If we speak from the perspective of experimental work, we find that Cu (111) film ensures uniform orientation of graphene [12]. If we speak from the perspective of simulation works, our DFT calculation with tri-layer graphene on Cu (111) plane is easily relaxed and convergence takes place without much difficulty. Reason behind it is the stable fitting of graphene on substrate with small amount of mismatch. Since Cu (111) structure itself creates a number of hexagons to fit the hexagonal graphene with small mismatch percentage, it is obviously reasonable to use as substrate.

In this work, we focus on different stacking of tri-layer graphene on Cu (111) substrate and study the electronic band-

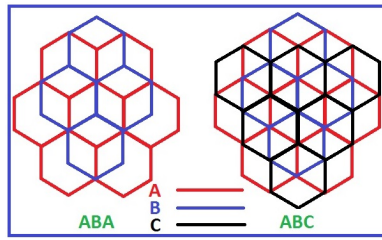


Fig. 1: Tri-layer graphene structure stacking

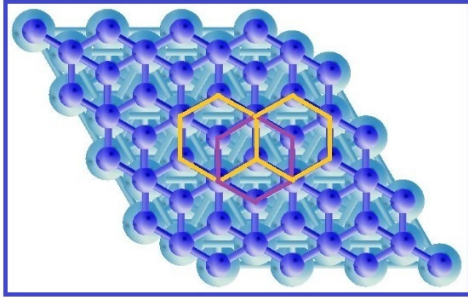


Fig. 2: Bernal stacking of tri-layer graphene on Cu (111) substrate

structure of Bernal (ABA, ACA) and rhombohedral (ABC, ACB) stacking combinations on substrate. Our results show the stacking dependent band gap in tri-layer graphene on substrate. We also analyze the doping phenomenon as well as the density of states for each of the system.

II. GEOMETRY

Cu (111) crystal is an FCC structure with one Cu atom surrounded by 6 Cu atoms just as a coin can be surrounded by similar 6 coins. In Bernal (ABA) stacking of tri-layer graphene, layer 1 and layer 2 remain a little bit twisted but the third layer is without any twist to the 1st layer. On the other hand rhombohedral (ABC) stacking makes sure all three layers are twisted to each other (Fig.1). The spacing between the substrate and the closest graphene layer usually is between 2.1 to 3.8 angstrom [13]. For Cu (111) substrate, the relaxed spacing has been around 2.25 angstrom. Tri-layer graphene with Bernal and rhombohedral stacking on Cu (111) substrate are shown in Fig.2 and Fig.3.

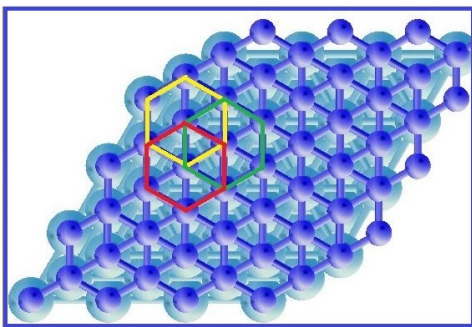


Fig. 3: Rhombohedral stacking of tri-layer graphene on Cu (111) substrate

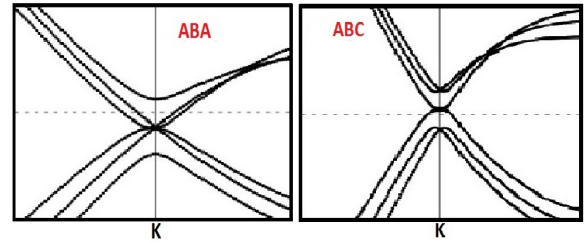


Fig. 4: ABA and ABC stacking of tri-layer graphene (without substrate) band plot, zoomed near K point region

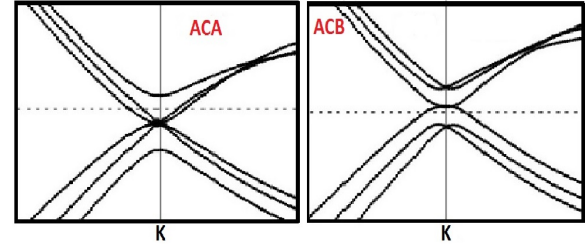


Fig. 5: ACA and ACB stacking of tri-layer graphene (without substrate) band plot, zoomed near K point region

III. SIMULATION METHOD

The simulation method used in this paper includes basically density functional theorem (DFT) with local density approximation (LDA). We have used Quantum Espresso to perform first principle calculation. We have considered 5 layers of Cu (111) substrate with graphene tri-layer on top of it considering the stacking issue. The system relaxation was followed by a self-consistent field (scf) calculation. After that we have calculated band-structure energy data for equally spaced k point samples.

IV. RESULTS AND DISCUSSION

Two types of Bernal and two types of rhombohedral structures are analyzed systematically using DFT calculation. We have relaxed each of the system and then performed a self-consistent field (scf) calculation and finally calculated energy values for different k points to plot the $E - k$ diagram that is the band-structure. All the four stacking structured systems are very much close in energy i.e. stability. Fig.4 and Fig.5 show the four stacking structures of tri-layer graphene before introducing metal substrate. And the notable thing is that no significant band gap is available.

Fig.6 shows the band-structure of a stable Bernal stacked system with ABA stacking. We can see from the figure that at K point, in stead of a Dirac cone we have found a “Mexican hat” structure due to the internally generated electric field. A little energy gap is also notable here along with high n-doping. Fig.7 shows the electronic band-structure of ABC rhombohedral stacking of the tri-layer graphene on Cu (111). The $E - k$ diagram clearly puts the fact in front of us that at K point, a significant amount of energy gap is induced along with a relatively lower but still high n-doping.

ACA stacking (Fig.8) is another Bernal stacking with similar energy gap at K point as the previous (ABA) Bernal one

TABLE I: Energy value, K point energy gap, per unit area binding energy, interface metal-graphene (d_i) and average graphene-graphene (d_g) layer distance of different stacking of tri-layer graphene on Cu(111) substrate

Stacking	Energy(Ry)	K point energy gap (meV)	Per unit area Binding Energy (meV)	d_i (Å)	d_g (Å)
ABA	-507.64450442	149.8	-26.12	2.266	3.383
ABC	-507.64448488	409	-26.16	2.267	3.382
ACA	-507.64473906	126.5	-26.69	2.267	3.383
ACB	-507.64472184	403.5	-26.73	2.266	3.383

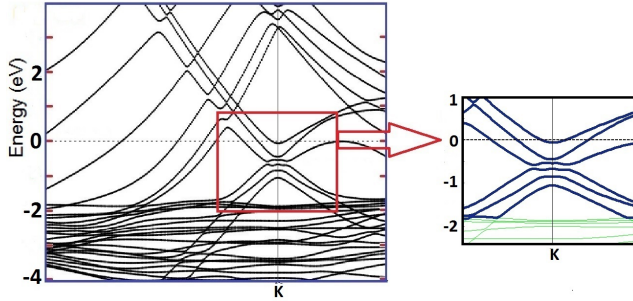


Fig. 6: Band-structure of ABA Bernal stacked tri-layer graphene on Cu (111) substrate near K point (more zoomed view is shown in the right side)

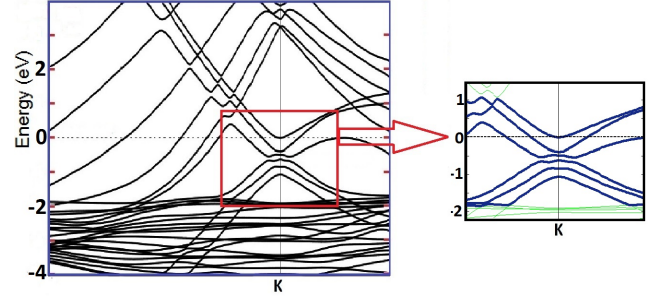


Fig. 8: Band-structure of ACA Bernal stacked tri-layer graphene on Cu (111) substrate near K point (more zoomed view is shown in the right side)

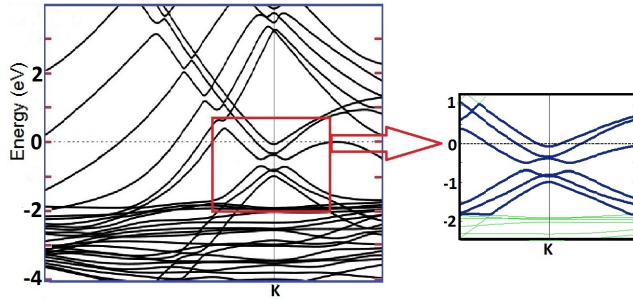


Fig. 7: Band-structure of ABC rhombohedral stacked tri-layer graphene on Cu (111) substrate near K point (more zoomed view is shown in the right side)

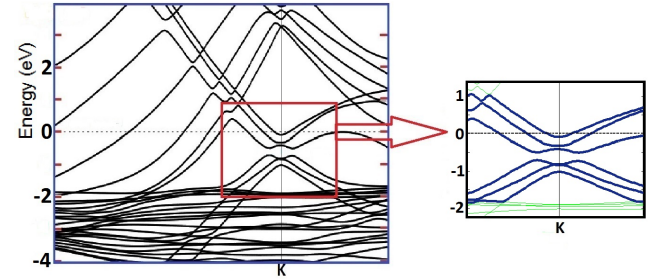


Fig. 9: Band-structure of ACB rhombohedral stacked tri-layer graphene on Cu (111) substrate near K point (more zoomed view is shown in the right side)

and it also includes high n-doping. ACB stacking structure (Fig.9) is also a rhombohedral structure and includes quite notable energy gap at K point along with n-doping. K point energy gap in rhombohedral stacking is still higher than that of Bernal stacking in spite of not applying any external bias. There is clearly an internally introduced electric field due to presence of substrate. Another explanation is the effect of dimer. From tight binding diagram [10], we can see in Bernal stacking weakly bonded dimer ensures a small band gap whereas in rhombohedral stacking, presence of a strongly bonded dimer ensures larger band gap.

Table I provides us with the system energy, K point energy gap and per unit area (area in square angstrom) binding energy. From the table, we observe that the most stable and thermodynamically favorable structure is ACA Bernal stacking on copper substrate since the energy value for this stacking is -507.64473906 Ry, the negative maximum considering five decimal point, although all the energy values are very close and rhombohedral stacking gives larger band gap. Binding energy for each of the stacking has been calculated following the equation $E_b = E_s - E_g - E_m$, where E_b is the binding

energy, E_s is the system energy, E_g is the only tri-layer graphene energy and E_m is the only metal energy i.e. the binding energy calculation includes the energy of the whole system as well as individuals. The interface metal-graphene distances (d_i) and average interlayer graphene distances (d_g) are also mentioned in the table. We have calculated the density of states (DOS) for each of these systems as shown in Fig.10. We find the shapes of the DOS plots are similar for each of the system with different peak values.

Density of state plots show us that at an energy value of around 2.5 eV, we get the highest peak value. So it can be said that peak values are maintaining same energy value in spite of different stacking. Stacking dependency in tri-layer graphene on Cu (111) substrate therefore gives us different density of states but corresponding energy value is unchanged.

V. CONCLUSION

In this work, we observe that suspended tri-layer graphene of different stacking do not have any notable band gap without any external bias. But tri-layers with same structure on Cu (111) substrate have induced energy gap at K point, large in rhombohedral stacking and small in Bernal stacking. Since

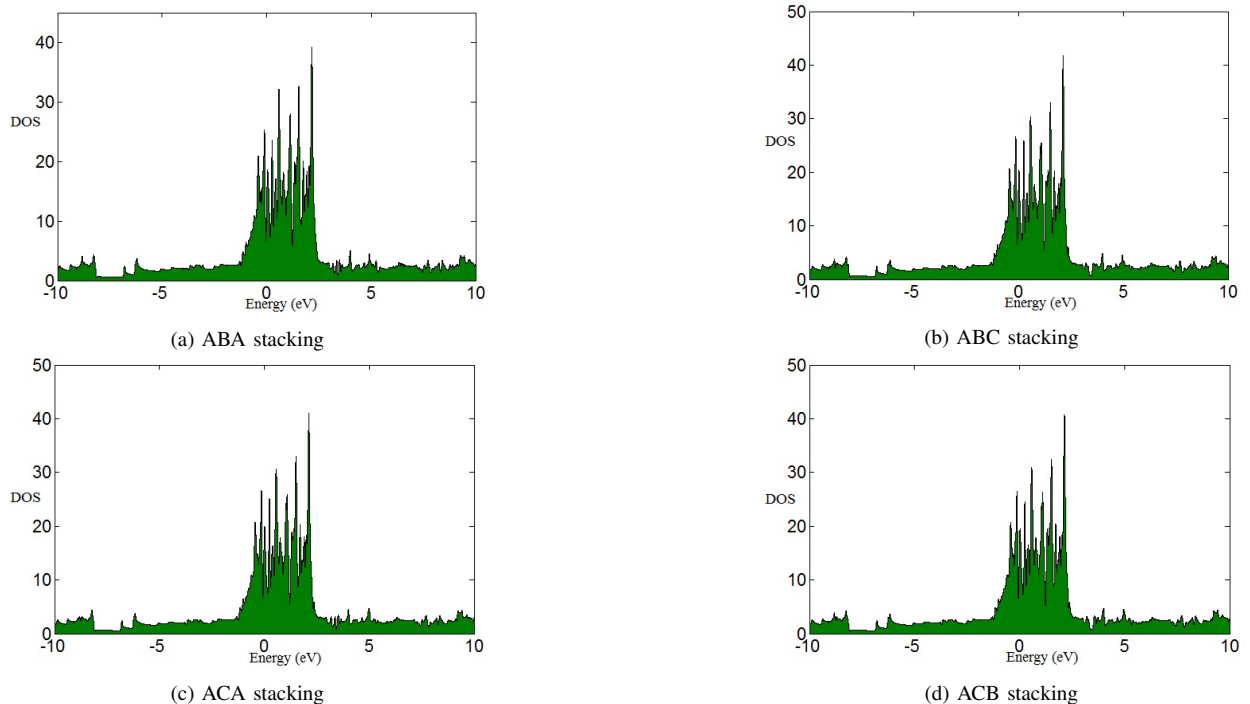


Fig. 10: Density of states of different stacking of tri-layer graphene on Cu (111) substrate

band gap is generated along with “Mexican hat” structure after introducing metal substrate, there has been a clear effect of internal electric field. This is due to Cu (111) surface with a work function of around 5.0 eV [14] and also a surface energy of 0.75 eV. Moreover structures with both Bernal and rhombohedral stacking include n-doping but Bernal stacking of tri-layer graphene has little more doping in comparison to rhombohedral stacking. The obtained results reveal that the band gap and doping of graphene on substrate is strongly dependent on the stacking type of tri-layer graphene.

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Intensity and Phase Noise of Semiconductor Lasers Operating in Single Mode

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Abstract— Semiconductor lasers often involve various noise and instability problems due to fluctuation of photon and carrier numbers. In this paper, we analyze the intensity and phase noise of semiconductor lasers operating in single mode. To overcome limitations of the small-signal analysis we use direct numerical integration of the self-consistent rate equations for photon number, phase and carrier density. Langevin noise sources for photon number and phase have been introduced to the rate equations to include fluctuations due to spontaneous emission and the process of carrier recombination. Rate equations are applied to 850-nm GaAs lasers. Fast Fourier Transform (FFT) has been used to calculate the frequency spectra of both intensity and phase noise. Noise characteristics for different injection currents have been demonstrated. Results show that intensity and phase noise decrease with the increase of the injection current density and linewidths were decreasing substantially with increasing injection current as well.

Keywords— Laser rate equation; quantum noise; intensity noise; phase noise; linewidth; noise spectrum

1. INTRODUCTION

In recent ages, semiconductor lasers have become pivotal elements in optoelectronics. They come up with sharp accuracy under sophisticated design that includes complex multi-level structure based on semiconductor gain media. The optical gain is usually achieved by stimulated emission at an interband transition under conditions of a high carrier density in the conduction band [1].

Semiconductor lasers are in widespread use today in many areas from CD, DVD and other forms of data storage through to communication links. Laser diodes offer cost effective and reliable means of generating coherent light suitable for these

applications. But the performance of laser diodes is largely affected by various noise and instability problems.

The term “noise” in semiconductor laser refers to random fluctuations of various output parameters due to quantum-mechanical effects along with many internal and external influences which limits the performance of the laser. Intensity noise and phase noise are two typical types of quantum noise, that correspond to intrinsic fluctuations in the photon number, carrier number and phase that are generated during the quantum interaction processes of the lasing field with the injected charge carriers [2] [3]. These intensity and phase noise are impossible to control in principle. It is, therefore, essential to analyze various types of noise in semiconductor lasers with a view to maximizing laser performance.

Therefore, to improve the performance of laser diodes, analysis of laser noise is important. Theoretical analysis can be made by solving of the laser rate equations including Langevin noise sources in order to generate the fluctuations [4]. Moreover, the application of direct numerical integration of the rate equations results in overcoming the limitations of small-signal analysis [5] [6]. Considering the existence of only the fundamental transverse mode, we have applied the rate equation model to 850 nm GaAs laser and observed the frequency spectra of intensity and phase noise following numerical simulation.

2. THE RATE EQUATION MODEL

A. Laser rate equation:

The rate equations of semiconductor lasers can be presented as follows: [7]

For photon numbers:

$$\frac{dS}{dt} = (G - G_{tho})S + \frac{a\xi N}{V} + F_S(t) \quad (1)$$

For photon phase:

$$\frac{d\theta}{dt} = \frac{\alpha a \xi}{2V} (N - \bar{N}) + F_\theta(t) \quad (2)$$

For injected carrier (electron) numbers:

$$\frac{dN}{dt} = -AS - \frac{N}{\tau_s} + \frac{I}{e} \quad (3)$$

In eqn. (1) G is the gain of single mode laser with wavelength λ , given by-

$$G = A - BS \quad (4)$$

G_{tho} is the threshold gain of solitary laser given by,

$$G_{tho} = \frac{c}{n_r} \left[k + \frac{1}{2L} \ln \frac{1}{R_f R_b} \right] \quad (5)$$

A is the linear gain which can be expressed as,

$$A = \frac{a \xi}{V} (N - N_g) \quad (6)$$

B is the self-suppression coefficient written as,

$$B = \frac{9}{4} \frac{\hbar \omega}{\epsilon_0 n_r^2} \left(\frac{\xi \tau_{in}}{\hbar V} \right)^2 a R_{cv}^2 (N - N_s) \quad (7)$$

Other parameters are: a =differential gain coefficient, ξ =field confinement factor, V =volume of the active region, L =length of the active region, α =linewidth enhancement factor, $\bar{N}$ =time average value of $N(t)$, τ_s =electron lifetime, I =injection current, e =electron charge, k =internal loss of the laser cavity, N_g =electron number at transparency, $\hbar$ =reduced Planck constant, τ_{in} =intraband relaxation time, R_{cv} =dipole moment, N_s =electron number characterizing self-suppression coefficient.

The functions $F_S(t)$ and $F_\theta(t)$ are Langevin noise sources for photon numbers and phase, respectively. These noise sources are given by,

$$F_S(t) = \sqrt{\frac{V_{SS}}{\Delta t}} g_S \quad (8)$$

$$\text{and } F_\theta(t) = \sqrt{\frac{V_{SS}}{\Delta t}} \frac{g_\theta}{2(S+1)} \quad (9)$$

where V_{SS} is the variance of autocorrelation.

$$V_{SS} = \left[\frac{a \xi}{V} (N + N_g) + G_{tho} \right] S + \frac{a \xi N}{V} \quad (10)$$

g_S and g_θ are the Gaussian random numbers in ranges of $-1 < g_S < +1$ and $-1 < g_\theta < +1$, and Δt = time-step of the calculation.

B. Output Power of Semiconductor Laser:

The output power $P(t)$ from the front facet of semiconductor lasers is given by,

$$P(t) = \frac{h \nu c}{2n_r L} \frac{\ln(1/(R_f R_b)) (1 - R_f)}{(1 - \sqrt{R_f R_b})(1 + \sqrt{R_f/R_b})} S(t) \quad (11)$$

where, c is the speed of light in vacuum, n_r is the refractive index of the active region, $h \nu$ is the photon

energy of the emitted light, R_f is the power reflectivity of the front facet and R_b is the power reflectivity of the back facet.

C. Noise and Spectral Linewidth Calculation:

In the previous calculations used by McCumber and Haug [1], [10], the time-fluctuating components are transformed into Fourier frequency components from which the relative intensity noise, phase noise and linewidth are calculated. But in our numerical approach in this paper RIN, FN and linewidth can be calculated from the optical power fluctuations of semiconductor laser.

$$\delta P(t_i) = P(t_i) - \bar{P} \quad (12)$$

where, $\bar{P}$ = time average optical power.

The power fluctuation can be obtained from the rate equations of photon number and carrier number by integrating them numerically and calculating the spectrum of intensity fluctuations.

The RIN and FN spectrum can be originally defined as the Fourier transform of the autocorrelation function written as,

$$RIN = \frac{1}{\bar{P}^2} \frac{\Delta t^2}{T} |FFT[\delta P(t_i)]|^2 \quad (13)$$

$$\text{and } FN = \frac{\Delta t^2}{T} |FFT[\Delta \nu(t_i)]|^2 \quad (14)$$

where, variation of optical phase is given as-

$$\Delta \nu(t_i) = \frac{1}{2\pi} \frac{d\theta}{dt} = \frac{1}{2\pi} \frac{\Delta \theta(t_i)}{\Delta t} \quad (15)$$

Laser linewidth is the full-width at half-maximum (FWHM) spectrum of the laser operating in single mode. It is determined from the low frequency component of the frequency or phase noise as-

$$\text{Laser Linewidth} = \Delta f = 4\pi FN|_{\omega=0} \quad (16)$$

3. NUMERICAL SIMULATION

Through numerical simulation, photon number $S(t)$, carrier number $N(t)$ and phase $\theta(t)$ have been determined with corresponding noise terms. Typical values of the parameters for 850-nm GaAs laser, used for numerical simulation are: $a = 2.75 \times 10^{-12} \text{ m}^3 \text{ s}^{-1}$, $|R_{cv}|^2 = 2.8 \times 10^{-57} \text{ C}^2 \text{ m}^2$, $\xi = 0.2$, $\tau_{in} = 0.1 \text{ ns}$, $\tau_s = 2.79 \text{ ns}$, $N_s = 1.7 \times 10^8$, $N_g = 2.1 \times 10^8$, $V = 100 \mu\text{m}^3$, $d = 0.11 \mu\text{m}$, $L = 300 \mu\text{m}$, $n_r = 3.6$, $k = 10 \text{ cm}^{-1}$, $R_f = 0.3$ and $R_b = 0.6$.

The numerical integration has been performed within a time interval of $\Delta t = 5 \text{ ps}$ by applying the

fourth-order Runge-Kutta algorithm. The significance of such small value of Δt is that it can generate noise sources that approximately describe a white noise spectrum up to a frequency of 200GHz ($=1/\Delta t$) [8]. We have expanded the integration to a time period as long as 10 μ s. This requires more than 1 million integration steps and helps us to find out the intensity noise as low as 100kHz.

Box-Muller transformation method has been used to generate the independent Gaussian variables g_s and g_0 after taking two uniformly distributed random variables u_1 and u_2 ranging between -1 and +1, which are necessary to achieve the proper Langevin noise source.

4. RESULTS AND DISCUSSION

A. Intensity Noise and Phase Noise:

The main objective of this study is to examine the frequency spectra of intensity noise and phase noise of GaAs laser operating in single mode. Eqns. (13) and (14), have been used to calculate the quantum relative intensity noise (RIN) and quantum frequency noise (FN), respectively. A comparative study has been made on the effect of injection current on RIN spectrum and FN spectrum by applying two different values of injection current; $I=1.3I_{th}$ and $I=1.7I_{th}$. The corresponding simulated spectra of RIN and FN are shown in Fig. 1(a) and (b), respectively.

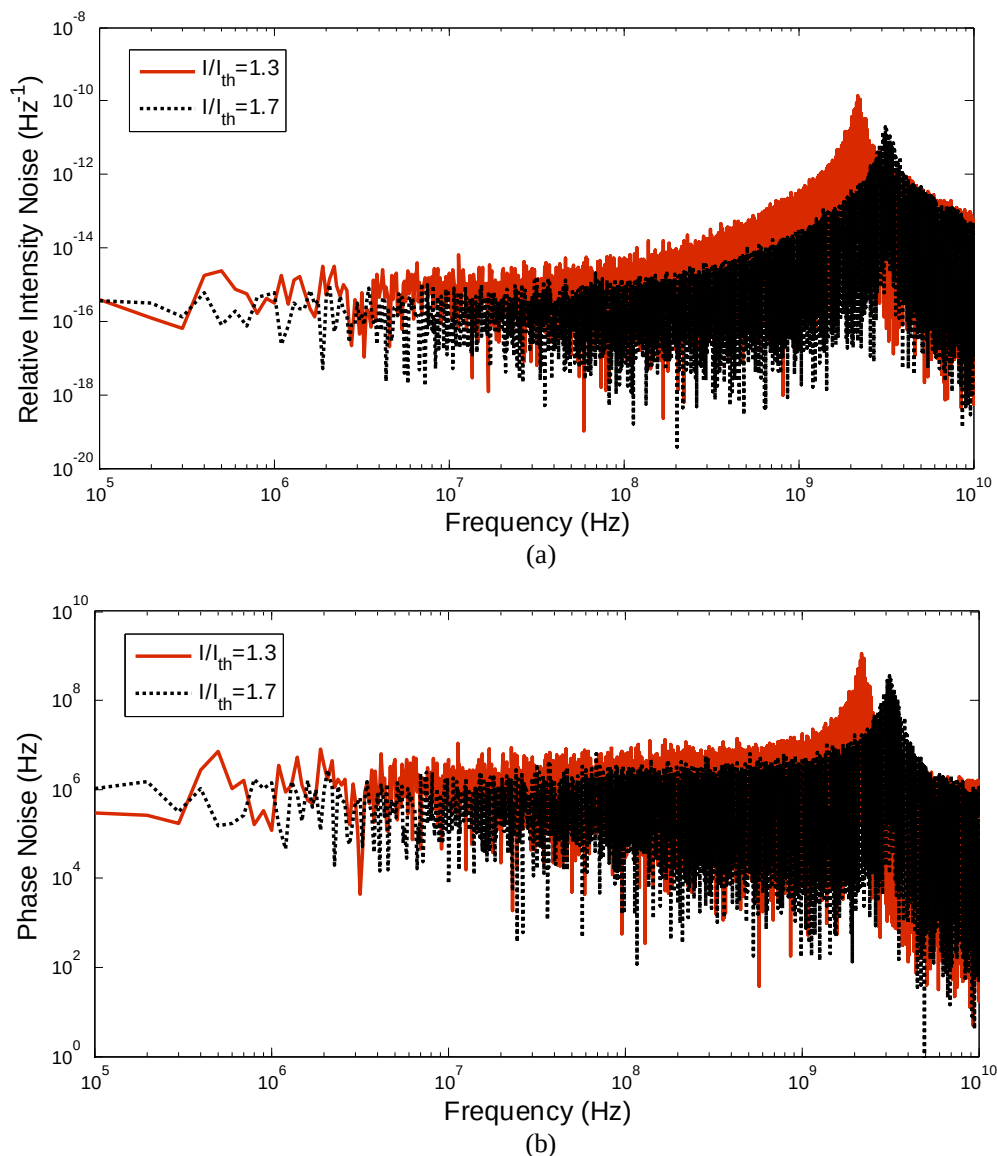


Fig. 1: (a) RIN spectrum for different values of injection current; (b) FN spectrum for different values of injection current

Figures show that, both intensity and phase noise were decreased with the increase of injection current density I . Figure also shows pronounced peaks for both RIN and FN around the relaxation oscillation frequencies that were experimentally detected in [11]. At low frequencies, the FN is flatter than RIN. These characteristics are in good agreement with those determined by the small-signal approximations described in [4]. At the same time, increasing the injection current results in shifting of the relaxation oscillation peaks towards higher frequencies. This is due to the fact that the repetition of fluctuations of both $\delta P(t)$ and $\Delta v(t)$ becomes faster with increasing I .

B. Laser Linewidth:

Laser linewidth can be calculated from eqn. (16) applying the condition of very low frequencies. At frequency as low as 100 kHz, we can approximately calculate Δf from the flatness of FN at the low frequency side.

At injection current $I=1.3I_{th}$, the value of $\Delta f=11.38$ MHz.

At injection current $I=1.7I_{th}$, the value of $\Delta f=10.75$ MHz.

These show the rapid narrowing of Δf with increasing I . The decrease of Δf with I matches the corresponding decrease of the FN reported in [4]. This result also agrees with the modified Schawlow-Townes relation that describes the dependence of the linewidth Δf on the injection current I [12].

5. CONCLUSION

Noise in semiconductor lasers is described in terms of photon number and phase fluctuations. We have investigated the intensity and phase noise of a 850-nm GaAs laser operating in single-mode through numerical simulation of the self-consistent laser rate equations that include Langevin noise sources for photon number and phase. Due to direct numerical integration of the rate equations, the limitations of the small signal analyses have been overcome. Langevin noise was generated systematically to introduce photon and carrier numbers fluctuations due to spontaneous emission and carrier recombination. The frequency spectra of RIN and FN have been illustrated with the help of FFT. The noise characteristics have been demonstrated under different injection current values and found to be decreasing with the increase of injection current density. Laser linewidths were also decreasing with the increase of carrier injection.

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Effects of the Field-dependence of the Carrier Mobility on the Base Transit Time of SiGe HBTs Having Non-uniformly Doped Base

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Abstract—This work develops an analytical model of the base transit time for the SiGe-heterojunction bipolar transistors (HBT) having non-uniformly and heavily doped base considering the field-dependence of the carrier mobility under moderate-level injection. The proposed model considers all the non-ideal effects found in the literature owing to the non-uniformity and the heavy levels of the base doping. The evolving mathematical intractability has been resolved by applying an exponential approximation technique and the concept of the perturbation theory. Simulation results of the proposed model for the exponentially doped base with three types of Ge-dosing profile (box, trapezoidal and triangular) show that the effects of the field-dependent carrier mobility on the base transit time is significant.

Index Terms—Base Transit Time, Field Dependence of Carrier Mobility, Ge-Dosing Profile, Moderate-level Injection, SiGe-HBT.

I. INTRODUCTION

Compared to the homojunction bipolar junction transistors (BJT), heterojunction bipolar transistors (HBT) have shown improved unity-power-gain-bandwidth and improved high frequency response and hence, draws the attentions of the researchers for high speed applications. Till to date, a lot of research works have been done in the literature on the III-V HBTs, especially on the AlGaAs-based HBTs. In the recent years, however, SiGe-based HBTs are proved to be their efficient alternative [1].

Of various components of the total transit time (τ_{ec}), which mainly characterizes the high-frequency performance of BJTs and HBTs, the most significant one is the base transit time (τ_B). This τ_B is the time required for the minority carriers to traverse through the base region and constitutes more than 70% of τ_{ec} [2], especially under the increased injection-level condition. Therefore, modern BJTs and HBTs use nonuniform and heavy-level base doping profile to introduce an electric field in the base to reduce τ_B and also, to increase the current handling capacity. However, for these devices, the analytical modeling of τ_B becomes complicated and mathematically intractable owing to the prevailing non-ideal effects such as high injection effects, bandgap narrowing effects, velocity saturation effects, the doping and field dependence of the carrier mobility and the recombination mechanism in the base.

In addition, for SiGe HBTs, this intractability increases for the presence of Ge-dosing profile, which is introduced to reduce the band-gap narrowing effects. Therefore, conventional models of τ_B for SiGe HBTs [3]–[8] neglected the recombination in the base and the field-dependence of the carrier mobility. However, including recombination in the base, Chang *et al.* [9] developed their analytical model for the uniformly doped base and recently, Mahfuz and Iqbal [10] developed a model for the nonuniformly-doped base. But, both these models neglected the field-dependency of the carrier mobility.

This work extends the work in [10] by incorporating the effect of the field-dependency of the carrier mobility by the appropriate and judicious application of the concept of the perturbation theory [11] and the exponential approximation technique [12]. The proposed model considers the base doping profile as exponential and chooses three types of Ge-dosing profile in the base, namely, box, trapezoidal and triangular profile.

II. ANALYSIS

A. Basic Equations

For an npn bipolar junction transistor (regardless of whether it is homo or hetero junction), the base transit time τ_B can be defined as:

$$\tau_B = \frac{q}{J_C} \int_0^{W_B} n(x) dx \quad (1)$$

where J_C is the collector current density and $n(x)$ is the minority electron concentration and the current density within the base of width W_B . In order to determine $n(x)$, the following pair of differential equations need to solved

$$J_n = qn(x)\mu_n E_n(x) - qD_n \frac{dn(x)}{dx} \quad (2)$$

$$\frac{dJ_n}{dx} = \frac{qn(x)}{\tau_n} \quad (3)$$

where μ_n , D_n , τ_n are the mobility, diffusivity and lifetime of electrons respectively and E_n is the electric field within the

base. The physical models for all these parameters are given as

$$\frac{E_n}{V_T} = \frac{d}{dx} \ln p = \frac{d}{dx} \ln n_{ie}^2 x \quad (4)$$

$$n_{ie}^2 = \gamma_r n_{io, Si}^2 e^{\frac{\Delta E_{gHD}(x) + \Delta E_{gGe}(x)}{kT}} \quad (5)$$

$$\gamma_r = e^{-sqrt5y_{av}} \quad (6)$$

$$\Delta E_{gHD} x = qV_{gHD} \ln \left(\frac{N_A}{N_{ref}} \right) \quad (7)$$

$$\Delta E_{gGe} x = qV_{gGe} y x \quad (8)$$

$$\mu_{n, SiGe} = \frac{v_{s, SiGe}}{a E} \frac{v_{s, SiGe}}{\mu_{n0, SiGe}} \quad (9)$$

$$v_{s, SiGe} = \frac{0.342}{0.342 y_{av} + 1} v_{s, Si} \quad (10)$$

$$\mu_{n0, SiGe} = b \mu_{n0, Si} \left[\frac{D_n}{V_T} \left(\frac{N_A x}{N_{m, ref}} \right)^{-\gamma_1} \right] \quad (11)$$

Eqn. (4) represents the electric field acting on the electrons in the base considering the bandgap narrowing effects [13] where V_T is the thermal voltage and n_{ie} is the effective intrinsic carrier concentration. This n_{ie} in SiGe can be modeled by the Eqn. (5) where γ_r is the ratio of the effective density of states in SiGe to that in Si [3], ΔE_{gHD} is the bandgap narrowing due to heavy doping effects modeled by [14] and ΔE_{gGe} is the bandgap narrowing due to presence of Ge which is assumed to have a linear dependence on Ge concentration [15]. Combining the Eqns. (5) to (8) gives

$$n_{ie}^2 = N_0 u^{\gamma_2} \quad (12)$$

where $N_0 = \gamma_r n_{io, Si}^2 \left(\frac{N_A(0)}{N_{ref}} \right)^{\gamma_{2, Si}} e^{\frac{V_{gGe} y_E}{V_T}}$ and $\gamma_2 = \gamma_{2, Si}$ with $\gamma_{2, Si} = \frac{qV_{gGe} y_C - y_E}{kT}$ and $u(x)$ represents the space-dependency of the base doping profile. The doping and the field dependency of the carrier mobility can be expressed by the empirical expression [6], [11] given by the Eqn. (9) where $a = 0.7743$ and $v_{s, SiGe}$ expressed by the Eqn. (10) [9] is the saturation velocity inside the SiGe alloy. The low-field doping dependent electron mobility in the presence of Ge ($\mu_{n0, SiGe}$) can be related with that in Si ($\mu_{n0, Si}$) through the expression given in the Eqn. (11) [3], [11] with $b = 1 - 3y_{av}$.

Exponential doping profile is chosen as the base doping profile in this work. This profile can be expressed as,

$$N_A x = N_A(0) e^{\frac{-\eta x}{W_B}} = N_A(0) u x \quad (13)$$

where, u represents the exponential variation of the doping profile and $\eta = \log \left[\frac{N_A(0)}{N_A(W_B)} \right]$ is the logarithmic slope of the doping profile. In order to represent all the three types of Ge-dosing profile, a generalized trapezoidal profile $y(x)$ as shown in Fig. (1) can be used, given by

$$y x = m_{Ge} x + y_E \left[\frac{y_C - y_E}{W_B} \right] x + y_E \quad (14)$$

where y_C, y_E is the Ge-fraction in the collector (emitter). This Eqn. (14) represents triangular and box profiles for $y_E = 0$ and for $y_E = y_C$ respectively [4]. Total Ge-content within

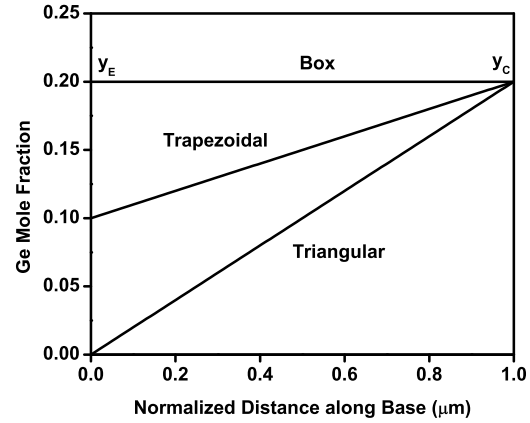


Fig. 1. Distribution of Ge content in Si on emitter side.

the base is referred to as Ge-dose and can be determined as [5] $y_D = y_{av} W_B$ where $y_{av} = \frac{y_C + y_E}{2}$.

III. MODEL DERIVATION

Differentiating Eqn. (2) with respect to x , combining with Eqn. (3) and then rearranging results in a second order differential equation (DE) of $n(x)$ as

$$\frac{d^2 n}{dx^2} \left[\frac{E_n}{V_T} \frac{d}{dx} \ln \left(\frac{1}{qD_n} \right) \right] \frac{dn}{dx} + \left[\frac{d}{dx} \left(\frac{E_n}{V_T} \right) \frac{E_n}{V_T} \frac{d}{dx} \ln \left(\frac{1}{qD_n} \right) + \frac{1}{\tau_n D_n} \right] n = 0 \quad (15)$$

with

$$\frac{E_n}{V_T} = \frac{d(n + N_A)}{dx} \frac{\eta \gamma_2}{W_B} \quad (16)$$

$$\frac{1}{qD_n} = \left(F_e u^{\gamma_1} \frac{\eta}{W_B} \frac{a_e}{q v_{s, SiGe}} \frac{E_n}{V_T} \right) \frac{n}{2n} \frac{N_A}{N_A} \quad (17)$$

where $F_e = \frac{1}{qD_n(0)} \left[\frac{N_A(0)}{N_{mref}} \right]^{\gamma_1}$. The non-linear DE [Eqn. (15)] can be made linear by using the concept of the perturbation theory [11] as

$$\frac{d^2 n}{dx^2} \left[\frac{E_{ne}}{V_T} \frac{d}{dx} \ln \left(\frac{1}{F_1 qD_n} \right) \right] \frac{dn}{dx} + \left[\frac{d}{dx} \left(\frac{E_{ne}}{V_T} \right) \frac{E_{ne}}{V_T} \frac{d}{dx} \ln \left(\frac{1}{qD_n F_1} \right) + \frac{1}{\tau_n D_n F_1} \right] n = 0 \quad (18)$$

where, the term $n = N_A$ can be approximated as $n_{l2} = N_A$, n_{l2} being the low-level-injection solution of the DE and

$$\frac{E_{ne}}{V_T} = \frac{\eta}{W_B} \frac{1}{2} \gamma_2 b_e \quad (19)$$

$$b_e = \frac{2}{\gamma_2} \frac{n_{l2}}{2n_{l2} N_A} \quad (20)$$

$$F_1 = \frac{2n_{l2} N_A}{n_{l2} N_A} \quad (21)$$

Making appropriate use of the exponential approximation technique to transform the complicated variable coefficients into simple exponentials [12] and using the following boundary conditions [11]:

$$n(0) = \frac{n_{ie}^2(0)}{N_A(0)} e^{\frac{V_{BE}}{V_T}} \quad \text{at } x = 0 \quad (22)$$

$$\frac{1}{2} \sqrt{\frac{1}{4} \left[\frac{N_A(0)}{n_{ie}(0)} \right]^2 e^{\frac{V_{BE}}{V_T}}} = \frac{J_n W_B}{q v_s} \quad \text{at } x = W_B \quad (23)$$

the injection-level dependent solution $n(x)$ of the DE given by Eqn. (18) can be obtained as

$$n(x) = u^{-\frac{b}{2}} \left(\frac{1}{q D_n} \right)^{\frac{1}{2}} C_{1l} I_{\beta_l} (b_{0l} x) + C_{2l} K_{\beta_l} (b_{0l} x) \quad (24)$$

under low injection and

$$n(x) = u^{-\frac{b}{2}} \left(\frac{1}{F_1} \frac{1}{q D_n} \right)^{\frac{1}{2}} C_1 J_{\beta} (b_0 x) + C_2 Y_{\beta} (b_0 x) \quad (25)$$

under intermediate injection, where $b = 1 + \gamma_2$, J_{β} , Y_{β} is the Bessel function of the first (second) kind, I_{β} , K_{β} is the modified Bessel function of the first (second) kind and β , b_0 , C_1 , C_2 are the constants. Plugging the Eqns. (25) and (2) into Eqn. (1), finally, gives the base transit time τ_B .

IV. RESULTS AND DISCUSSIONS

This section presents the simulation results of the proposed model considering the field-dependence of the carrier mobility in the base in comparison with the model developed in [10] that did not consider the same. The proposed model and the model in [10] are termed, hereafter, as the present and the previous model respectively.

In the high-field regime, the carrier mobility is a function of electric field and decreases with the increase of the electric field. Therefore, the electron mobility considering the field-dependence observed in Fig. (2) is found to be lower than that obtained when the field-dependence is neglected. Lower mobility causes the carrier flow to decrease and hence, results in lower current. This fact is evidenced from Fig. (3). When mobility decreases the probability of the recombination of electrons in the base increases, thereby, causing the electron concentration in the base to reduce. Therefore, a decrease in electron concentration for the present model in comparison with the previous model has been observed in Fig. (4).

When electrons in the base move towards the collector, the probability of the recombination increases which is further enhanced by the decreased mobility due to the field-dependence. Therefore, near the collector side, more carriers are lost [Fig. (4)], less current is observed [Fig. (3)] and lower mobility has been observed [Fig. (2)] for the present model in comparison with the previous model.

Figs. [(5) and (6)] show the variation of the collector current density, J_C and the base transit time, τ_B against the variation of the base-emitter voltage, V_{BE} . Since J_C is defined as $J_C = J_N W_B$, the reduction of J_C for the present model

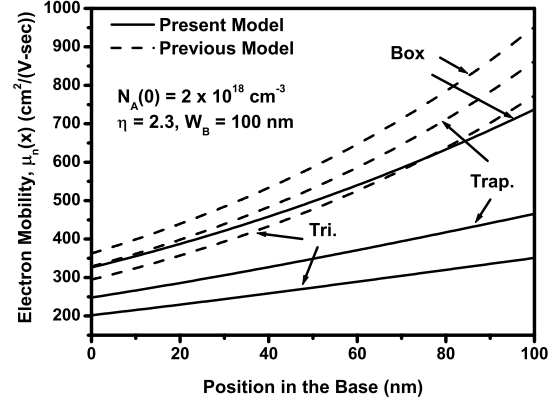


Fig. 2. Mobility profile in the base for different shape of Ge profiles with and without considering field-dependence for $V_{BE} = 0.9$ V.

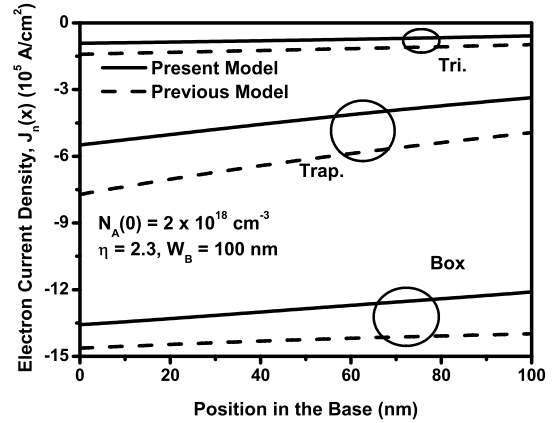


Fig. 3. Electron current density in the base for different shape of Ge profiles with and without considering field-dependence for $V_{BE} = 0.9$ V.

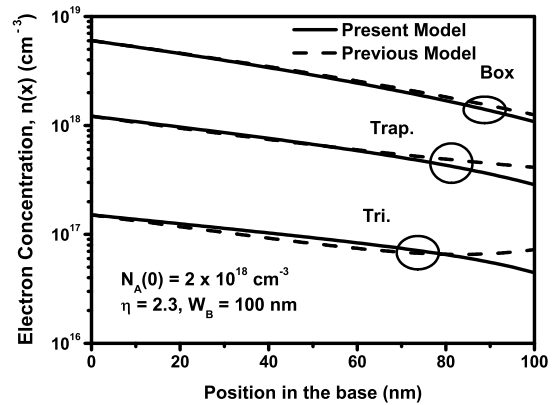


Fig. 4. Electron concentration in the base for different shape of Ge profiles with and without considering field-dependence for $V_{BE} = 0.9$ V.

is expected from the reduction of the electron current density near the collector side explained earlier [Fig. (5)]. The base transit time observed in Fig. (6) is higher for the present model than that for the previous model, since the carrier mobility owing to the field-dependency decreases in the present model [Fig. (2)]. From Fig. (6), it is also observed that the τ_B is the highest for the box dosing profile and the lowest for the triangular dosing profile. These results are in accordance with those in [16] which shows that, for the triangular dosing profile, the transit time of the minority electrons through the base is minimum.

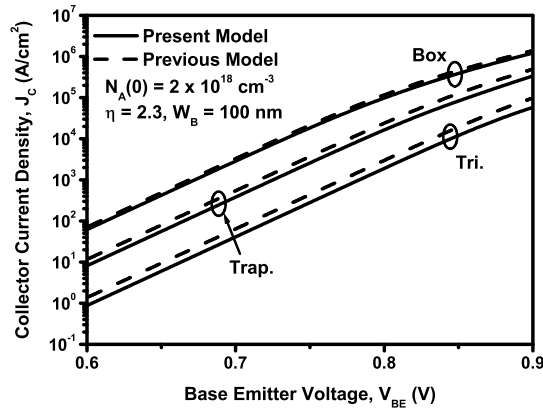


Fig. 5. Collector current density vs. base-emitter voltage for different shape of Ge profiles with and without considering field-dependence.

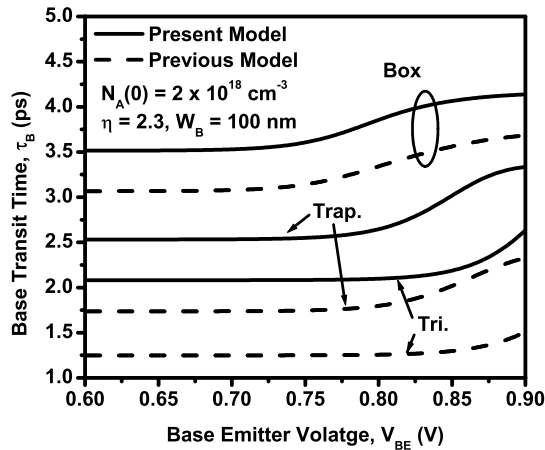


Fig. 6. Base transit time vs. base-emitter voltage for different shape of Ge profiles with and without considering field-dependence.

V. CONCLUSION

In this work, the base transit time considering the field-dependency of the carrier mobility is calculated for a SiGe HBT having heavily and exponentially doped base employing

a general type of trapezoidal Ge profile. The simulation results show that the base transit time increases owing to the consideration of the field dependence. This effect is the highest for the triangular dosing profile of Ge-content and the lowest for the box type.

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Analytical Modeling of J-V Characteristics of CIGS Based Thin Film Solar Cell Considering Voltage and Space Dependent Electric Field in the Absorber Layer

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Abstract— This paper focuses on developing an analytical model to analyze the current-voltage (J-V) characteristics of lightly doped CIGS-based thin film solar cell considering the series resistance and the voltage-space dependent electric field in the absorber layer. The mathematical intractability evolved for these considerations has been resolved by employing the concept of perturbation theory. Effects of various parameters such as intensity of light, electron lifetime and hole lifetime have been investigated using the developed model. Simulation results show that the proposed model is helpful in providing better physical insight.

Keywords—Thin Film Solar Cell, J-V Characteristics, CdS/CIGS Solar Cell, Perturbation Theory, n-p Heterojunction.

I. INTRODUCTION

Over the years, phenomenal success has occurred in the field of development of solar cells. Focusing on development in cost, safety, environment, efficiency and fabrication, solar cells are being switched from Silicon based homo-junction structure to thin film hetero-junction structures- mostly made from CdS/CIGS, CdS/CdTe and a-Si. Among those three, CIGS is demonstrably capable and versatile material [1] which achieved higher efficiency with a world record of 20.8% [2]. In the superstrate device structure based glass/SnO₂/CdS/CIGS/metal p-n junction hetero-junction thin film solar cells [3, 4], a very thin layer (0.1 μm) of CdS acts as a highly-doped n-type (emitter) and a comparatively thicker (few microns) CIGS layer acts as a lightly doped p-type absorber layer. Most of the light incident on a CIGS solar cell is absorbed in the first micrometer of the CIGS layer. The photo-generated carriers in this region are collected by the voltage-dependent built-in electric field in this layer [5, 6].

Researchers have faced numerous challenges in modelling the J-V characteristics of CdS/CIGS solar cells. Efforts have been made to develop numerical and analytical models [5-9]. Numerical iteration methods are time consuming, less helpful in providing physical insight of the device and sometimes unexpectedly erroneous. Analytical models, on the other hand, are better for getting physical insight. In [5, 6] analytical models have been developed with a number of fitting parameters such as maximum photocurrent, effective attenuation current etc. Moreover, these models assumed

uniform electric field in the absorber layer. Anjan *et al.* [7] developed an analytical model with a fewer number of fitting parameters but they assumed constant electric field in the absorber layer. Later, Mannan *et al.* [8] derived a model using voltage and space dependency of the electric field. Both the models in [7, 8] used iterative schemes to deduce the J_L-V_L characteristics curves owing to the mathematical intractability, although they derived the photo-generated currents analytically and hence, their models are treated as semi-analytical. Recently, Haque *et al.* [9] proposed a fully analytical model for CdS/CIGS based solar cell using the concept of perturbation theory, but their model considered the electric field in the absorber layer as uniform.

The focus of this paper is to overcome the drawbacks stated above. Here, a varying and load-voltage dependent electric field has been considered across the fully depleted absorber layer. Like [9], an approach using perturbation method has been employed to solve the intractable governing differential equations to avoid lengthy iteration process. In the following sections of this paper, the analytical procedures for determining J-V characteristics of a practical thin film CdS/CIGS solar cell have been stated stepwise and then simulation results based on that model have been produced to analyze the J-V characteristics with respect to different parameters, i.e., intensity, life time of holes and electrons.

II. ANALYSIS

The required equations to determine the photo generated current in the lightly-doped absorber layer of an n⁺-p junction based CdS/CIGS thin film solar cell are as follows

$$J_n(x) = qn(x)\mu_n E(x) \quad (1)$$

$$J_p(x) = qp(x)\mu_p E(x) \quad (2)$$

$$\frac{dJ_n(x)}{dx} = \frac{qn(x)}{\tau_n} - q\alpha(\lambda)G_0(\lambda)e^{-\alpha(\lambda)x} \quad (3)$$

$$\frac{dJ_p(x)}{dx} = -\frac{qp(x)}{\tau_p} + q\alpha(\lambda)G_0(\lambda)e^{-\alpha(\lambda)x} \quad (4)$$

where J_n(x) and J_p(x) stand for electron current density and hole current density, n(x) and p(x) for electron density and hole density and E(x) for the electric field at any point x (Fig.

1) respectively, μ_n for electron mobility, μ_p for hole mobility, τ_n and τ_p for the life time of electron and hole respectively, α for wavelength-dependent absorption coefficient of the material and G_0 for the optical generation rate given by

$$G_0(\lambda) = (1 - R(\lambda)) \frac{\lambda I_0(\lambda)}{hc} \quad (5)$$

Where h is Planck's constant, c is the velocity of light, $I_0(\lambda)$ is the intensity of light, $R(\lambda)$ is the reflection coefficient and λ is the wavelength of the incident light.

Though the photo generated current (J_{ph}) at emitter region (CdS) is caused by the absorptions of highly energetic photons, this current can be considered negligible because the width of depletion region at highly doped emitter side is too small hence its diffusion length. According to thin film hetero-junction structure of CdS/CIGS the base (CIGS) region is lightly doped. So, the entire absorber layer is depleted as shown in Fig. (1).

For intrinsic or very lightly doped (N_{a1}) absorber, the electric field can be approximated as constant. Previous analysts [5-7, 9] have used this approximation in their analytical approach; but in practical cases for a little increase in the absorber doping results in a linearly-decreasing electric field as shown in Fig. (1). However, the electric field in the depletion region close to the emitter side is considered negligible due to its very narrow width.

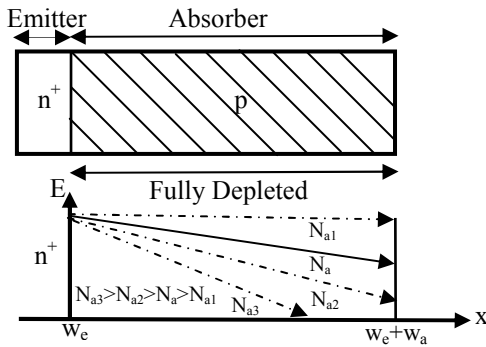


Fig. 1. Electric Field (E) Profile

Electric field $E(x)$ for Fig. (1) can be expressed using Poisson's equation as follows:

$$E(x) = \frac{V_{bi} - (V_L - J_L R_s)}{W_a} + \frac{qN_a}{\epsilon} \left(x - \frac{W_a}{2}\right) \quad (6)$$

Where W_a is the width of absorber region in the CIGS layer, V_L is the load voltage, R_s is the equivalent series resistance for the device, ϵ is the permittivity of absorber material (CIGS), V_{bi} is the built in voltage due to electric field.

A. Derivation of the model

J_{ph} is the sum of $J_n(x)$ and $J_p(x)$ at any point x [Fig. (1)] in the absorber layer. Since electric field is the dominant factor for current conduction in the depletion region, diffusion component of these currents are negligible.

Combining Eqns. (1) and (3) and Eqns. (2) and (4) gives the first order differential equations of $n(x)$ and $p(x)$ in the absorber layer as

$$\frac{dn}{dx} + \left[\frac{1}{E(x)} \frac{dE}{dx} - \frac{1}{\tau_n \mu_n E(x)} \right] n(x) = - \frac{\alpha(\lambda) G_0(\lambda)}{\mu_n E(x)} e^{-\alpha x} \quad (7)$$

$$\frac{dp}{dx} + \left[\frac{1}{E(x)} \frac{dE}{dx} + \frac{1}{\tau_p \mu_p E(x)} \right] p(x) = \frac{\alpha(\lambda) G_0(\lambda)}{\mu_p E(x)} e^{-\alpha x} \quad (8)$$

Since electrons and holes are swept out by the electric field from the absorber layer in the opposite direction, it can be assumed that $p(W_e) \approx 0$ and $n(W_e + W_a) \approx 0$. Using these boundary conditions, the solutions for the electron and hole concentration can be obtained as

$$n(x) = \frac{\alpha(\lambda) G_0(\lambda)}{\mu_n IF_n(x)} \int_{W_e+x}^{W_e+W_a} e^{-(\alpha+f_n(x))} dx \quad (9)$$

$$p(x) = \frac{\alpha(\lambda) G_0(\lambda)}{\mu_p IF_p(x)} \int_{W_e}^{W_e+x} e^{-(\alpha+f_p(x))} dx \quad (10)$$

Where, $IF_n(x) = E e^{-f_n(x)}$, $IF_p(x) = E e^{-f_p(x)}$,

$$f_n(x) = \int \frac{dx}{\mu_n \tau_n E(x)}, \quad f_p(x) = \int \frac{dx}{\mu_p \tau_p E(x)}$$

Now putting the values of $n(x)$ and $p(x)$ from Eqns. (9) and (10) into Eqns. (1) and (2) gives the electron and hole current densities as

$$J_n(x) = q G_0(\lambda) \alpha(\lambda) e^{f_n(x)} \int_{W_e+x}^{W_e+W_a} e^{-(\alpha+f_n(x))} dx \quad (11)$$

$$J_p(x) = q G_0(\lambda) \alpha(\lambda) e^{f_p(x)} \int_{W_e}^{W_e+x} e^{-(\alpha+f_p(x))} dx \quad (12)$$

Performing integration for the entire absorber region, the average current density for electron and hole can be deduced as

$$J_{n,avg} = \frac{1}{W_a} \int_0^{W_a} J_n(x) dx \quad (13)$$

$$J_{p,avg} = \frac{1}{W_a} \int_0^{W_a} J_p(x) dx \quad (14)$$

Finally, the photo-generated current equation can be obtained as

$$J_{ph} = J_{n,avg} + J_{p,avg} \quad (15)$$

Current density at load (J_L) considering the series resistance R_s can be derived as follows

$$J_L(V_L) = -J_{ph}(V_L) + J_0(V_L) \left(e^{\frac{V_L - J_L R_s}{nV_T}} - 1 \right) \quad (16)$$

Where, n is the ideality factor, V_L is the load voltage, V_t is the thermal voltage, J_{ph} is the photogenerated current density and J_0 is the dark saturation current density. Since the band-gap of CdS is quite high ($E_g = 2.42$ eV), the dark saturation current in the emitter layer caused by diffusion is negligible and hence, J_0 is dominated by the recombination mechanism in the absorber layer. Therefore, J_0 can be given by [8]

$$J_0(V_L) = \frac{qn_i w_a}{2\sqrt{\tau_n \tau_p}} \quad (17)$$

Where, n_i is intrinsic carrier concentration. From a close look on Eqn. (16) it is revealed that the load current (J_L) at any load voltage (V_L) can only be determined using numerical methods due to the presence of unknown J_L on both sides. However, using the concept of the perturbation theory this J_L can be determined analytically. According to this theory, if J_L at any V_L is known then, the same at $V_L + \delta V_L$ can be determined using this value, provided that $\delta V_L \rightarrow 0$. Proceeding in the same manner currents at $V_L + 2\delta V_L$, $V_L + 3\delta V_L$ can be determined from the current values at $V_L + \delta V_L$, $V_L + 2\delta V_L$ respectively. Photo-generated current calculated without considering R_s can be used as the initial guess of this process.

III. RESULTS AND DISCUSSIONS

This section presents and discusses the simulation results using the proposed model. The incident photon flux is the air mass (AM) 1.5 global spectrum and obtained from the ASTM G-173-03 standard [10]. The absorption coefficients for CdS and CIGS are obtained from the absorption curves in Ref. [11]. The thickness of CdS and CIGS layers are 0.1 and 2 μm respectively.

The current-voltage (J_L - V_L) characteristics curve simulated using the proposed model has been shown in Fig. (2). It is noteworthy that the open circuit voltage (0.63 v) and the short circuit current density (31.72 mA-cm⁻²) obtained in the proposed model are in well agreement with that obtained in

the work of Hegedus *et al.* (Fig.8 of [4]), for the same parameters: $R_s = 0.2 \Omega$ and $n=1.5$ and $J_0=2 \times 10^{-6}$ mA cm⁻².

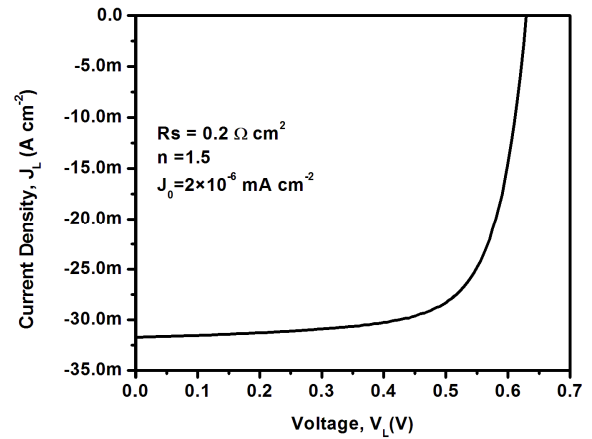


Fig. 2. J_L - V_L Characteristics of the proposed model

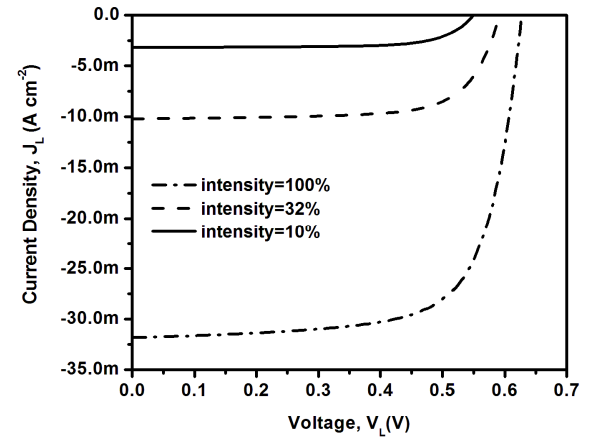


Fig. 3. J_L - V_L characteristics curve against the variation of light intensity (I)

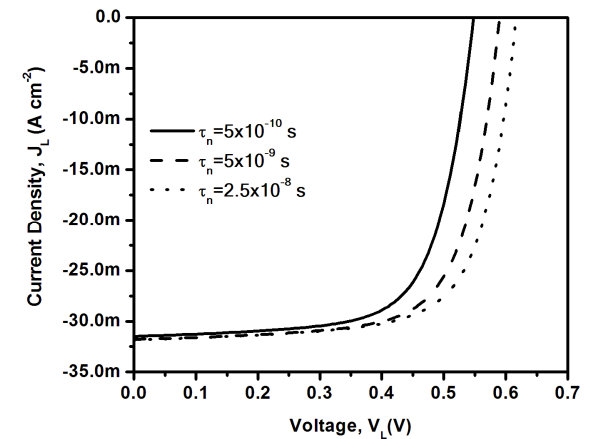


Fig. 4. J_L - V_L characteristics curve against the variation of electron lifetime (τ_n).

Fig. (3) shows the currents for different intensities of light. Besides depicting the fact that the current increases as light intensity increases, the curves also tell that both short-circuit current and open circuit voltage increase with the increase in light intensity. These results can be explained as follows. When light intensity increases, the photo-generation current and hence, the short circuit current increases due to increased absorption. On the other hand, increase of intensity has very little effect on the recombination of photogenerated carriers, and as a result, the dark saturation current increases negligibly. This means that the increase of J_{ph} supersedes that of J_0 due to increase of intensity and hence, the open circuit voltage increases.

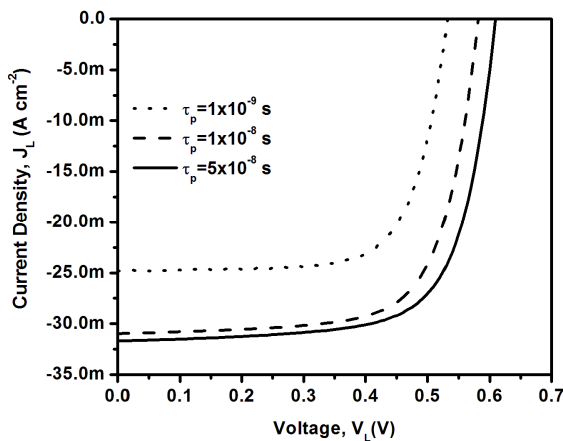


Fig. 5. J_L - V_L characteristics curve against the variation of hole lifetime (τ_p)

J_L - V_L curves for different lifetimes for electrons (τ_n) [Fig. (3)] and holes (τ_p) [Fig. (4)] reveal that J_L increases, as expected, as any of τ_n or τ_p increases. Also, the impact of change in τ_p is greater than that in τ_n . The reason behind this lies in the concept of absorption of photons, most of which is absorbed near the emitter side of the n-p junction. Since the fully depleted p-region is much longer than the n-region, it can be argued that the most of the absorption and hence the generation of electron-hole pairs is occurred in a region that is more distant from bottom contact than from the top contact [Fig. (1)]. The electrons are driven towards the top and the holes towards the opposite direction by the built-in electric field. Because the holes have to travel a longer way than the electrons do before they can find their way out of the cell and hence, holes have to face more recombination than the electrons, the impact of change in τ_p is greater than that in τ_n .

IV. CONCLUSION

A fully analytical model of current-voltage characteristics considering the series resistance has been developed in this work for a CdS/CIGS based thin film solar cells using the concept of perturbation theory. This model incorporates both the voltage and the space dependency of the electric field in

the absorber layer. The proposed model shows that the current density derived using the model increases with the increase of the light intensity and is more sensitive to the hole-transport in comparison with the electron transport. Therefore, the proposed model is helpful for the future analysts in their quests for physical realization of solar cells.

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