Dear friend,

It is nice that you are planning to attend the Antenna and Propagation Symposium (APSYM 2004) at Cochin University of Science and Technology - Department of Electronics. I welcome you warm to this important event.

APSYM 2004 is the 9th one in the series, which we started in 1986. A chronological listing of the earlier APSYMs is given below. Over this period, the papers and talks presented at APSYM 2004 are scheduled to be presented during APSYM 2004. The APSYM 2004 organizing committee have planned and conducted the technical programmes with a number of invited talks by eminent scientists in the field.

Chronology of APSYMs

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<td>APSYM-2006</td>
<td>Dec. 12-14, 2006</td>
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Proceedings of the earlier symposia are available with organizers of APSYM 2004. Those who are interested may please contact them.

Once again I extend a warm heartiest welcome to all the authors, delegates and distinguished guests of APSYM 2004.

Signed - 22
December 01, 2004

Prof. R.G. Nair
Our Tribute to Great Pioneers

JAMES CLERK MAXWELL. The founder of Electromagnetic theory of Radiation. His theoretical prediction of the existence of Electric and Magnetic fields associated with wave propagation carrying energy of Electromagnetic nature was a breakthrough in the history of Science. A new era of Electromagnetism was thus opened by this great Scientist.

HEINRICH HERTZ. Experimentally demonstrated the generation, propagation and detection of electromagnetic waves. Thus he gave a firm experimental support for the theoretical conclusions drawn by James Clerk Maxwell.

JAGATISH CHANDRA BOSE. The First Indian Scientist who marked his footprints in the world of Electromagnetics. In fact, Bose generated millimeter waves using a circuit developed in his laboratory and used these waves for communication, much earlier than the western scientists. He also developed microwave antennas (horns) which are still considered to be ideal feeds.
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Dr. C.K. Anandhan
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Research Papers Published in APSYM 2004
December 21, Thursday
(10:30 a.m. to 5:00 p.m.)
KOMATIYA Auditorium
MORPHOTIP Antennas
Chair: Y. N. S. Jaikl, CEERE, PULAN.

1.1 Novel technique to couple energy from waveguide feed network to microstrip antenna.
Prasad Kumar, Debajyoti Basu Prasad, V. K. Rastogi, and A. Das.
Email: rastogiv@yahoo.com, dbasu@comsyst.in, das@comsyst.in.

1.2 Wide Band cylindrical shaped curve patch antenna.
School of Electronics and Communication Engineering, Christian College, Ernakulam, Kerala.
Email: rreddy@christuniversity.in, avb@christuniversity.in.

1.3 Broad Band Aperture coupled microstrip antenna.
Department of Electronics, VITM, Vellore.
Email: reddy@vit.ac.in, avb@vit.ac.in.

1.4 Empirical design considerations for waveguide coupled microstrip patch antenna.
R. S. N. S. Reddy and K. V. Anjaneyulu.
Microwave Research Group, Dept. of Physics, S.D.M. College of Engineering and Technology, Dharwad, Karnataka - 580002.
Email: anjaneyulu@gmail.com.

1.5 Radiation characteristic of patch microstrip antenna in anisotropic and non-isotropic cold plasma medium.
Department of Electrical Engineering, R. V. College, Bangalore.
Email: prasad@rvce.ac.in, psv@rvce.ac.in.

1.6 Fractal Patch Antenna for 2.4GHz Band.
Centre for Development of Telematics, Meena, Ranchi-83215.
Phone: 0654-2475132, Email: bhalani@cdt.gov.in, srengarajan@cdt.gov.in.

1.7 Even mode multiport network model for dual band rectangular microstrip antenna.
Department of Electrical Engineering, R. V. College, Bangalore.
Email: basus@rvce.ac.in, prasad@rvce.ac.in.

1.8 Smart Antenna application to a wireless thin-client network with integration framework.
A. S. N. S. Reddy.
Department of Electronics, Regional Institute of Technology, Mysore.
Email: ars@rit.ac.in, prasad@rit.ac.in.

2.1 Research Papers Published in APSYM 2004
December 21, Thursday
(10:30 a.m. to 5:00 p.m.)
KOMATIYA Auditorium
MORPHOTIP Antennas
Chair: Y. N. S. Jaikl, CEERE, PULAN.
Compact WideBand Antenna For BLUETOOTH™ Applications


KREML, Department of Electronics, Cusat University of<br>Science and Technology, Kottayam 686 022. arohit@ieee.org

Rectangular Microstrip Antenna on EDG Ground Plane with Unequal Orthogonal Periods


KREML, Department of Electronics, Cusat University of<br>Science and Technology, Kottayam 686 022. srmaven@ieee.org

December 27, Tuesday
(4.15 pm to 5.15 pm)

Antenna Session II

MICROWAVE PROPAGATION

C. D. Y. A. Almoder, University of Ceylon

Millimeter wave propagation through layers of sand and dust particles

Department of Physics, Tippu Sultan University, Kasargod

Electromagnetic radiations from the moon's surface originated from interactions of extraterrestrial neutrons with lunar regolith

Kalpana Roy Saha, Premanee Datta

Aeronautical Engineering Institute, Guwahati

Microwave Surface scattering model for moisture estimation in residue sensing
S.K. Sreelastor and Ambika Jain

Dept. of Physics, Govt. P. I. College, mthapal (Guj)

Microwave Imaging of Dielectric Wax Cylinders
G. Bhada, A.K. Pravoon Kumar, Anil Laxman, Mini Thomas, C.K. Amaranathan and S.I. Thather

Microwave Imaging and Material Research Laboratory, Department of Electronics, Pune University of Science and Technology, Chinch, Karad, India – 415052. wng@csat.ac.in

Superrate Loaded Metalic-Dielectric Structure based on Sierpinski Gasket for Backscattering Reduction

Amruth K. Krishnav, Thimmakkuty Mathew, C.K. Amaranathan, P. Mohanan, A. Vasandevan and S. Roy

KREML, Department of Electronics, Cusat University of Science and Technology, Kottayam 686 022. aramch@csat.ac.in
December 32, Wednesday
(10.00 a.m. 11.30 a.m.)

AEROSPACE SESSION III

HEROES OF RF ANTENNAS: II

Chair: Prof. C.S. Sudheer, National Institute of Technology, Bangalore

3.1 Development of a Phased Array Antenna on Silicon with DGS Phase Shifters.
    Tweedon J., Jose A., Abraham, Harigowda Y. and N.K. Menon
    Center for the Engineering of Electronic and Acoustical Materials and Devices, Pennsylvania State University, University Park, PA, USA

3.2 Design and Evaluation of Linearly Polarized Microstrip Antennas.
    Raj Kumar and Anil Tomar
    Indian Institute of Technology, New Delhi, India

3.3 L-Strip Fed Circular Microstrip Antenna
    B. Latha Kumari, Sreedhar K. Manohar, C. C. Ananthan, K. Vasudevan, and P. Mohan
    CREMA, Department of Electronics, Cochin University of Science and Technology, Cochin 682 022, india@cremee.org

3.4 Time Domain Analysis of Rectangular Microstrip Patch Antenna By Conformal FDTD Method
    Rama Prasad S. Murthy, C.K. Anandan, E. Vasudevan, and P. Mohan
    CREMA, Department of Electronics, Cochin University of Science and Technology, Cochin 682 022, drc@creme.org

3.5 Dual Band Semi-Circular L-Shaped Microstrip Antenna
    Dheep Das Kirthivas and K.P. Ray
    Research School of Engineering and Technology, Cochin 682 022, dskirthivas@gmail.com

3.6 Phase Shifterless, Log-Periodic, Slot-And, Leaky-Wave Antenna For Beam Steeping Application
    Elango Ashokkumar, Shekar S.V., C.K. Anandan, P. Mohan
    CREMA, Department of Electronics, Cochin University of Science and Technology, Cochin 682 022, vasudevan@cochin.ac.in

3.7 Electronically Reconfigurable Dual Frequency Microstrip Antenna Using Varactor Diode
    CREMA, Department of Electronics, Cochin University of Science and Technology, Cochin 682022.
3.4 A Cross-Slot aperture coupled circularly polarized antenna array for Broadband articulated Satellites
Sandeep K. Solanki and Anil Tyagi
Dept. of Electronics and communication Engineering, Indian Institute of Technology, Roorkee, IITrooerke374767.anil.s@iit.ac.in, sk_solanki@iit.ac.in

3.5 A Study on Compact Rectangular Microstrip Antenna with wide band
H. W. van"v, Sara Fatima Faruqui", P. V. Harapatra", "University Science Instrumentation Center, and "Dept. of Aeronautical Engineering, University,
Gulabgarh - 580 105, sm_faruqui23@rediffmail.com
"Dept. of Electronics Engineering, But a commutity college, UH, 84130, USA

December 12, Wednesday
(11:45 a.m. to 12:45 p.m.)
RESEARCH SESSION IV
MICROWAVE ANTENNAS-IV
Chair : Prof. G.P. Ghodake, Emeritus Professor, Delhi University.

4.1 CVarsatool FDTD design tool for RF and microwave applications
S. M. V. S. Vantu and B. M. Murty
Aerospace Electronics and Systems Division, National Aerospace Laboratories, Bangalore-560 017, smantu@nal.res.in, bmmurty@nal.res.in

4.2 Parallel implementation of the time dependent Maxwell equations using the NAL MKS Supercomputer
S. M. V. S. Vantu
Aerospace Electronics and Systems Division, National Aerospace Laboratories, Bangalore-560 017, smantu@nal.res.in

4.3 Gain-Radiation characteristics of dielectric loaded horn antenna
Post graduate department of physics, University of Kashmir, Srinagar 190 006 [JK]

4.4 Moment method approach to use wire antenna as a near field sensor of the waveguide radiator
Parameshwara and Ajay Chakravarty
Department of Electronics and Communication Engineering, Indian Institute of Technology, Kharagpur-721 302, pmshwara@iitkgp.ac.in

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5. Design and Development of secondary radar antennas for low range surveillance radar
Shubha, Ramesh A., A.K. Singh, Y. Mohan Ran and S. Srinivas
Electronics and Radar Development Establishment, Defence Research and Development Organization, Bangalore, India
"Centre for Radio Systems, Defence Research and Development Organization", Bangalore, upsingh@drdo.gov.in

6. Design of Microstrip Antenna using Ganetic Algorithm
Dimuthu C. Raman*, Shreyas S. Patil*, Pranathi Kandike*, Swapna Dew* and Dipak K. Gang
NITTR, Bhopal-462029, India Shreyas Patil, Bhopal, Bhopal, NITTR, Bhopal-462029, India, shreyas_patil@nitr.co.in

7. Shipboard Radar Ground to Air Combat Radar System: A Review
Aravind, A. K., Mohan Ran, V. Mohana Rao, V. Seetharam, K. V., S. V., and P. V. S. Meher
National Institute of Technology, Warangal, India, avinava@nirand.in

December 22, 2016
11:15 AM - 12:15 PM
MICROWAVE ANTENNAS - II
Chair: Dr. D. Sridharan, Professor, Director, NITC, Chennai

5.1 Characteristics of Airborne Radomes for Flat Panel Radar and their significance in fire control radars
P. Shrithapa, A. K. Singh and T. Mohan Ran
Electronics and Radar Development Establishment, Defence Research and Development Organization, C.P. Room, Bhopal, Bhopal, 462029, India, shrithapa2003@gmail.com

5.2 Broadband design techniques for choke flanged feed horns for Earth station antennas
A.K. Pathi
Antenna Systems Group, Space Applications Centre, Ahmedabad

5.3 Simulation model of a multi-carry wave division multiple access (MC-CDMA)
A. A. P. Chimote, R. U. Singh, M.K. and D. M. Jha
Dept. of Electronics, University of Science and Technology, Coimbatore 641212, Kerala

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Simulation model of a Hybrid WAG MC-CDMA-AI Scheme for wireless multimedia communication networks

Suresh T, Arul Kumar M S, Anurag Bhat P and V. Ramarajappa
Department of Electronics, Cochin University of Science and Technology, Cochin 682022, Kerala.

Rectangular Dielectric Resonator Antenna for 2.4 GHz Wi-Fi Applications
S. Murthy, Binu Pud, C.V. Aasandan, M. Anusossoulathorn, V. Chintalapudi and P. Mohanar
ORIGIN, Department of Electronics, Cochin University of Science and Technology, Cochin-682022. dminthar@ieee.org

Cylindrical Dielectric Resonator Antenna excited by Conductor-backed Coplanar Waveguide
Sume M S, Shrivastha K Raman, P V Allasion, M T Sebastian and P Mohanar
CIRCA, Department of Electrophysics, Cochin University of Science and Technology, Cochin 682022. dminthar@ieee.org

Compact Cylindrical Antenna for Active Antenna Applications
Joseph K K, Malim Joseph, C K Anandan, K Vasudevan and P Mohanar
Centre for Research in Electromagnetics and Antennas, Department of Electronics, Cochin University of Science and Technology, Cochin - 682022. dminthar@ieee.org

Au X Band Rectangular Waveguide Conjugated PIn Antenna For Beam-Forming Applications
Shinay Nair
Department of Electrical Engineering and Electronics, University of Liverpool, Brownlow Hill Liverpool, United Kingdom, L69 3GF, email: shinay@liv.ac.uk

December 23, Thursday
(10.15 a.m. to 11.15 a.m.)

MICROWAVE DEVICES

Chair: Prof. K.G. Nair, Emeritus Professor, Department of Electronics, CUSAT.

Design and Development of Harmonic mixer at Ka Band
Harish Nair Roy, Balathithu K, Siddharth and S.S. Senthil
Defence Electronics Application Laboratory, Defence Research and Development Organisation, DRDO, DRDO HAL, India

Ka-Band Frequency Tripler
Sanju Menon, R.K. Joshi and S.S. Senthil
Defence Electronics Application Laboratory, Defence Research and Development Organisation, DRDO, DRDO HAL, India

Image and Image
4.1 4-18 GHz Coplanar waveguide MEMS shunt switch
Kishorekumar, V. R. Pandey
Department of ECE, Orissa University of Technology, Bhubaneswar, India

4.2 Computer-Aided design procedure of microstrip non-linear circuits
N.M. Vyas and B.R. Gaur
Department of Electronics and Communication Engineering, Indian Institute of Technology, Kharagpur

4.3 Design and Analysis of a compact circular to rectangular waveguide transition at 94 GHz
I.R. Kumar, Prasad Reddy Tanne, Ashok Kumar, and Meenakshi Singh
Centre for Electronics Applications, University of Delhi, New Delhi, India

A Compact DMT Design For A Kabard Boundary Analyst
K. N. S. Sengar, R.C. Saini, S. N. Nair, and A. G. W. Leitner
ECE-RIAE, Indian Institute of Technology, New Delhi, India

5.1 Miniaturized Design of a Microwave Phase Shifter Using Modified Ground Coplanar Wave Structure
Microstrip Technology, Department of Electrical and Computer Engineering, Indian Institute of Technology, New Delhi

5.2 Microwave Applications
Claro, Prof. M.C. K. Chandan, and K.L. College of Engineering, Vijayawada

1.1 Dielectric behaviour of some mesogenic, photo-nematic and organic molecules
JSS College of Engineering, JSS University, Mysore

1.2 Technology for Land mine detection: a structured review
Dr. K.K. Nair
Defence Electronics Applications Laboratory, P.O. Box 54, Rail Net, Delhi 265-268-001
kgnair@defelab.com
7.3 An introduction to cellular neural network and its application
Sridhar Pattanaik1, R.K. Mohra2
1Department of Electronic Sciences, BHU, Varanasi, India, 276022.
2Department of Electrical Engineering, Indian Institute of Technology, Delhi, India, 110016.

7.4 Study of the microwave Dielectric and Absorption behavior of Mn-Zn ferrites nanocomposites at X-band
S.K. Sharma1, Ramendra Kumar Tiwari2
1Dept. of Physics, University of Allahabad, Allahabad, India, 211002.
2Dept. of Physics, Amity University, Noida, India, 201310.

7.5 Graphical User Interface (GUI): Adaptive Arrays
V.M. Panchapakesan and Y. Revin1
1University of California, Los Angeles, California, USA, 90095.

7.6 Effect of dopants on the microwave dielectric properties of SrLa2TiO6
R. Raghu and M. T. Sebastian
Aerospace Technology Division, Regional Research Laboratory, Thiruvananthapuram 695019, India.

7.7 Preparation and Dielectric Properties of Composites Substrates: With varying Filler Concentration
1Centre for Materials for Electronic Technology (C-MET), Department of Information Technology, Thiruvananthapuram 695019, India.
2Department of Electronic, Cochin University of Science and Technology, Cochin, Kerala 682022, India.

7.8 10-GHz Microwave diodes for microwave substrate applications
A. B. Fournirn, R. Mohanan1 and M. T. Sebastian
1Ceramics Technology Division, Regional Research Laboratory, Trivandrum 695019, India.

7.9 Cu2MgMn2Ti4O12 Ceramic Dielectric Resonators: For Microwave Telecommunication Applications
P.V. Gopalan1, R. Mohanan1 and M. T. Sebastian
1Ceramics Technology Division, Regional Research Laboratory, Thiruvananthapuram 695019, India.

7.10 Broadband Ultra High Strength Rods for Under Water Applications
B. Rameshchandra Rao, M. Dikshitha and R. Reddy Kumar
Defense Electronics Research Laboratory, DRDL, Hyderabad, 500 025, India.

8.1 Application of nanomaterials for defense and communication
Sridhar Pattanaik1, R.K. Mohra2
1Dept. of Electronic Sciences, BHU, Varanasi, India, 276022.
2Department of Electrical Engineering, Indian Institute of Technology, Delhi, India, 110016.
RESEARCH SESSION 1
Novel Technique to couple energy from waveguide feed network to microstrip antenna.

Praveenkumar, Debojyothi Choudhuri, V.V. Srinivasan, V.K. Lakshmeesha and S. Pal
Communication systems group, ISRO Satellite centre, Bangalore-560017
pkumar@isac.ernet.in, tljnt@isac.ernet.in

Wide Band cylindrical shaped curve patch antenna

A.K. Shrivastav\(^1\), Annapurna Das\(^2\), Sisir K. Das\(^3\)
\(^1\)School of Electronic and Communication Engineering, College of Engineering, Anna University, Guindy, Chennai-600 025. \(^2\)SAMEER Centre for Electromagnetics, 2\(^{nd}\) cross road, CIT campus, Taramani. Chennai-600113

Broad Band Aperture coupled microstrip antenna

T.J.R.G. Sankar, C. Dharma Raj and T. Balakrishnan\(^*\).
Department of ECE, College of Engineering, GITAM, Visakhapatnam-530045
\(^*\) Microwave Division, DRDO-CABS, Bangalore-560037 dharmarajc@yahoo.com

Empirical design consideration for waveguide coupled microstrip patch antenna

Nandakumar M. Shetti and N.H. Ayachit
Microwave Research Group, Dept. of Physics, S.D.M. College of Engineering and Technology, Dhavalagiri, Dharwad-580002. sdmengec@vsnl.com

Radiation characteristics of polygon microstrip antenna in anisotropic and isotropic cold plasma medium.

Sunil Kumar Jha, A. Sharma and C.S. Rai
Dept. of Physics, R.D.S College, Muzaffarpur

Fractal Patch Antenna for 2.4GHz Band

Vbha R. Gupta, Nisha Gupta
Birla Institute of Technology, Mesra, Ranchi-835215
vgupta@bitmesra.ac.in, ngupta@bitmesra.ac.in

Even mode multiprport network model for dual band rectangular microstrip antennas

Amit A. Deshmukh and Girish Kumar
Email:- Amitd@ee.iitb.ac.in, gkumar@ee.iitb.ac.in

Smart Antenna application to a wireless thin-client network with adaptation framework

Arpana B. Jinaga, VNR VJIET
Bachupally, Hyderabad-72 arpana.jinaga@gmail.com
Compact WideBand Antenna For BLUETOOTH Applications
CREMA, Department of Electronics, Cochin University of Science & Technology, Cochin 682 022. drmohan@ieee.org

Rectangular Microstrip Antenna on EBG Ground Plane with Unequal Orthogonal Periods
CREMA, Department of Electronics, Cochin University of Science & Technology, Cochin 682 022. drmohan@ieee.org
NOVEL TECHNIQUE TO COUPLE ENERGY FROM WAVEGUIDE FEED NETWORK TO MICROSTRIP ANTENNA

PRAVEEN KUMAR, DEBOJYOTI CHOUDHURI, VV SRINIVASAN, VK LAKSHMEESHA, S Pal FIEEE
Communication Systems Group ISRO Satellite Centre, Bangalore - 560017pkumar@isac.ernet.in, djyoti@isac.ernet.in

Abstract: A planar array antenna is designed and developed at Ku-band fed by a waveguide power divider to meet the high power handling requirement. One of the most critical parts in this antenna was the design of transition from waveguide network to microstrip patches. A 2x2 microstrip antenna array is used as building block of a 32x32 array. In this paper the design of waveguide to microstrip transition is presented.

INTRODUCTION

In high performance aircraft and spacecraft, where size, weight, cost, ease of installation and aerodynamic profiles are constraints low profile antennas may be required. To meet these requirements microstrip antennas are one of the best candidates. Major operational disadvantages of microstrip antenna are their low efficiency, lower power, poor polarization purity and very narrow frequency bandwidth. Techniques have been developed to overcome low efficiency, low bandwidth [1] and poor polarization purity of microstrip antenna [2]. To increase the power handling capability of microstrip antenna array can be feed by a waveguide feed network. For this type of arrays commonly used technique is slot coupling to microstrip antenna or antenna array used as a building block. In this paper a new type of technique to feed 2x2 microstrip array (used as a building block for 32x32 microstrip array) is presented.

DESIGN OF TRANSITION

In this technique conducting pin is used to couple energy from waveguide to 2x2 microstrip array. The waveguide feed network is fabricated by EDM method. A 0.3mm Aluminum sheet is used as a top cover and is bonded with the feed network. The radiating structure consists of three layers, the lower layer is RT Duroid6002 ($\varepsilon_r=2.94$, 30mils thick), upper layer is Glass-epoxy ($\varepsilon_r=4.55$, 10mils thick). A 2mm thick Rohacell is sandwiched between top and bottom layers. To get high cross-polarization values technique reported in [2] is used with stacking of the patches for broadband performance. Now there was a requirement to couple energy from waveguide to 2x2 patch antenna array (shown in figure 1) with broadband performance. To couple energy from waveguide network to this building block a pin coupling is used. One end of the pin was soldered to microstrip line and its other end was inserted inside the waveguide through a hole in the broad wall. Pins of different diameter were tried but a pin whose diameter was exactly equal to the width of the microstrip line gave best performance. Pin length was optimized for maximum coupling at the required frequency. 5mm clearance is given between pin and ground plane of microstrip antenna and a 2mm dia. hole is made in waveguide to insert pin into the waveguide. Commercially available FEM based EM simulation software package was used for the simulation of the transition. A hat was provided with the pin on the microstrip side for additional support and it was soldered with the microstrip line. 2x2 microstrip array with pin is shown in figure 2.
The simulated return loss of the transition is shown in Fig. 3. The material used for the pin is Beryllium copper and the pin is gold plated for better conductivity. 256 such pins were fabricated for the full array. A high level of soldering skill is required to solder these pins as soldering quality affects the performance of the transition. Electrical characterizations of the antenna were carried out and the measured results met the required specifications. High power test (400 Watts) was also carried out for four hours and no deviation in the electrical performance of the antenna was observed.

Simulated return loss is better than -20 dB over the required band of 500 MHz and -15 dB return loss bandwidth is approximately 1.5 GHz. To withstand high temperature SN63 solder material is used for soldering.
CONCLUSION

In this paper the technique to realize coupling from waveguide to microstrip through pin is explained. These pins are robust enough to handle high power.

ACKNOWLEDGEMENT

The authors like to thank A Suresh, ML. Subramanya, N Murthy and GNV Prasad for their valuable contributions in realizing the antenna.

REFERENCES

WIDE BAND CYLINDRICAL SHAPED CURVE PATCH ANTENNA

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²SAMEER-Centre for Electromagnetics
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Cylindrical shaped microstrip antenna having wide bandwidth is realized. The enhancement of bandwidth is achieved by aperture coupling concept. The aperture is cut on ground plane of feed network on a cylindrical shaped substrate. Radiating patch is kept on another cylindrical substrate, which is co-centering with the feed. The V.S.W.R. bandwidth is found out experimentally. The measured bandwidth for V.S.W.R < 2.0 is 17%.

INTRODUCTION

The microstrip antennas are in general narrow band antenna even though it has got many advantages such as conformability to the host body, over other type of antenna. Due to foresaid limitation many times its usefulness get restricted. Microstrip patch antenna on curve surface is analyze [2,3]. The bandwidth achieved is of the order of 1.1%.

Conventionally bandwidth of patch antenna in enhanced by increasing the substrate thickness [4,5]. In this way although bandwidth increases but at the same time surface wave power increases resulting in poor radiation efficiency. Thick substrate with microstrip feed will give rise to increased spurious radiation. Moreover, thick substrate cannot be made conformal to host body on which patch is mounted. In this present work cylindrical shaped substrate is achieve by using combination of thin substrate along with air layer. This is implemented by constructing a metal frame with cylindrical shape diameter. Inner side of all the four wall has grove. This is to hold the thin substrate and maintaining the required gap in between them. The gap is filled with air because of curved nature sufficient stiffness is obtain for thin substrate material. The structure of this antenna is shown in fig.1.

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BANDWIDTH CHARACTERISTICS

As shown in figure-1 rectangular (W x L) Patch is located at the bottom side of the patch substrate. Topside of it is acting as Radome, which protects the patch and feed network from environmental effect. Aperture excites the patch. The aperture is made at ground plane of feed network, which is placed below the antenna substrate at optimally required height. This gives an acceptable V.S.W.R. 2:1 for required bandwidth. Bandwidth of a patch antenna [5] can be expressed by

\[ BW = \frac{16}{37.92} \frac{p}{\varepsilon_r} \frac{1}{\varepsilon_r} \frac{h}{\lambda_0} \frac{W}{L} q \]  

(1)

Where,

- \( h \) Height of the substrate
- \( W \) Width of the patch
- \( L \) Length of the patch
- \( \lambda_0 \) Free space wavelength
- \( \varepsilon_r \) Dielectric constant of the substrate

(At present case it is to be taken as effective dielectric const.)

\[ p = 1 - 0.16605 \left( k_0 W \right)^2 + 0.02283 \left( k_0 W \right)^4 - 0.00914 \left( k_0 W \right)^6 \]

\[ q = 1 - \frac{1}{\varepsilon_r} + \frac{2}{\varepsilon_r^2} \frac{1}{5\varepsilon_r^2} \]

\( e_r = \) Radiation efficiency

As we can see that in equation (1) bandwidth is the function of height of the substrate in between the patch and aperture.

EXPERIMENTAL EVALUATION AND RESULTS

The antenna configuration shown in figure-1 is fabricated and evaluated for its V.S.W.R and radiation characteristics. The V.S.W.R measurement is carried out using
HP 8510B network analyzer. The V.S.W.R plot is shown in figure-2. It is noticed that overall V.S.W.R for required frequency 1560-1850MHz is within 2:1

![Figure-2: V.S.W.R plot of wideband cylindrical curve patch antenna](image)

Radiation pattern measurement is carried out at elevated far-field range at four frequency is 1575 MHz, 1700 MHz, 1775 MHz and 1850 MHz respectively. The radiation plots for above said frequencies are shown in figure-3 and 6.

![Figure-3: Radiation pattern plot at 1575 MHz](image)
It is noticed that all the radiation plot has regular pattern at the full right half and upto 40° in to the left half of the polar diagram. There are some irregularity at particular angular region higher than 40° at left side of the plot. The reason for this irregularity is investigated and it was attributed to reflection present in the test range and this was conformed by measuring the radiation pattern of wide band standard horn antenna. From the radiation pattern of standard horns it was noticed that horn radiation pattern is also perturbed at same frequency and same angular region. The radiation pattern of standard horn at one frequency is shown in figure-7.
The perturbed angular region is shown by dotted line. This conform the perturbation of radiation pattern of cylindrical shape antenna is due to reflection present in the range. concluded

**REFERENCE**


BROAD BAND APERTURE COUPLED MICROSTRIP ANTENNA

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Microwave Division, DRDO-CABS, Bangalore-560037.

Microstrip Antennas are very widely used due to that lightweight, simple geometry, inexpensive to fabricate and has very good dimensional stability. Microstrip Antenna can be fabricated to cover a broad frequency range of 100MHz to 100 GHz. The Antenna Size, shape, dielectric constant of substrate, and other physical parameters can be tailored according to the needs of the end user. Depending upon the characteristics required for a particular application, the Microstrip Antenna can be designed. Of the numerous types of Antenna designs, in the Aperture Coupled Microstrip Antenna has advantage of broad bandwidth. In this paper dual stacked Aperture Coupled Microstrip Antenna is considered and investigations are made to analyze the bandwidth of the considered antenna. The bandwidth of the antenna is determined practically with network analyzer for different slot widths and air gaps. The bandwidth of the antenna was observed to be very poor when there are no air gaps between the two layers. As the air gap length is increased the bandwidth improved and a bandwidth of about 42% can be achieved when the air gaps are of order of 5cm and above. These investigations are made in the S band frequency range.

INTRODUCTION

A Microstrip Antenna [1] is as shown in its simplest configuration in fig.1, the radiation pattern, bandwidth, and other parameters of the antenna depend upon the shape of the patch, dielectric constant of a substrate and thickness of the dielectric substrate. The material for the dielectric substrate is chosen as a compromise between dimension stability and loss in the substrate. Higher the dielectric constant (εr) value, higher the loss in the material and higher is the dimension stability of the material.

Microstrip Antennas are arrays of microstrip patches find very large applications due to light weight, low volume, low fabrication cost, low scattering, can be integrated with a MMIC, possibility antennas fabrications, simple matching network and etc. However, the limitation of these antennas are narrow bandwidth, lower gain, higher ohmic loss, complex feeding structure, fringing fields, low power handling capacity, not suitable to generate pencil beam, possibility of radiation from feeds and junction cross polarization products etc.

This paper discusses an antenna, which enhances the bandwidth of the Microstrip Antenna. Of the various possible configurations, Aperture Coupled Microstrip Antenna is considered as shown in fig.2.
The radiating microstrip patch element is etched on the top of the antenna substrate; in microstrip feed line is etched on the bottom of the feed substrate. The thickness and dielectric constants of substrate are chosen independently to optimize the radiation of the antenna. The antenna considered is dual stacked as shown in fig.2 A second layer with a slot cut in the ground plane is stacked above the feed substrate. This slot couples the energy to the rectangular patch antenna on the second layer. One more layer is stacked above the rectangular patch antenna to improve the directional characteristics of the antenna.

Dual stacked Aperture Coupled Microstrip Antenna is designed to be operated in the frequency range of S band. The center frequency is chosen to be 2.3 GHz, in the material of the substrate with εr 2.55 and thickness of the substrate is 62 Mils. The dimensions of the antenna designed are L=40.12mm, W=33.44mm, Lc=33.47mm and Ls=20.42mm.

**APERTURE COUPLED MICROSTRIP ANTENNA**

The coupling of aperture depends upon various dimensions and parameters of the substrate [2]. The input resistance as the rectangular aperture length increases since coupling between feed line and patch antenna increases, the input resistance as well as reactance increases width of the aperture. The coupling is maximum when the patch is centered over the aperture. The dielectric constant and substrate thickness effect the coupling. The coupling increases with dielectric constant and coupling decreases with substrate thickness.

The Bandwidth of the antenna can be increased by reducing the effective dielectric constant (εr ) of the antenna and also increasing the height of the antenna above ground plane. An improvement of 6 to 8% of bandwidth converted to single patch antenna can be achieved by introducing the air gaps between the patch and the ground plane. The Aperture can be of various shapes as reported in literature [3], the possible apertures are circular, rectangular, U-shaped, L-shaped, Bowtie, Dog-Bone, H-shaped and Hour glassed shape. As reported in the literature the Hourglass aperture as the maximum coupling and so is chosen in this analysis.

The length and width of the patch antenna are chosen for optimum radiation efficiency[1].

\[
W = \sqrt{h \lambda_d} \cdot \left( \ln \left( \frac{\lambda_d}{h} \right) - 1 \right)
\]

\[
L = \frac{c}{2 f_s \sqrt{\varepsilon_{re}}} - 2AL \quad \text{and} \quad Z_0 = \frac{\eta_0 \cdot h}{W \cdot \sqrt{\varepsilon_{re}}}
\]
Where \( W \) is the width of the patch, \( h \) is the substrate thickness, \( \Delta L \) is the substrate thickness, \( \lambda_d \) is the guided wavelength, \( \varepsilon_r \) is the effective dielectric constant and \( f_0 \) is the center frequency.

The location of the feed point changes the input impedance and is chosen such away that a perfect impedance matching is achieved with a 50 \( \Omega \) feed line. The input admittance can be determined at resonant frequency.

The length and width of the slot can be determined as

\[
Y_{in} = 2G_r \left\{ \frac{G_r^2 + B_r^2}{\cos^2 (\beta_g L_1)} + \frac{\sin^2 (\beta_g L_1)}{Y_0^2} - \frac{B_r \sin (2\beta_g L_1)}{Y_0} \right\}
\]

The impedance of the slot can be determined as

\[
Z_{0s} = 60 + 3.69 \sin ((\varepsilon_r - 2.22) - \pi/2.36) + 133.5 \ln (\lambda_0 \varepsilon_r) \sqrt{W_a} + 2.81 \left[ 1 + 0.011 (4.48 + \ln \varepsilon_r) \right] (W_a/h) \ln (100h/\lambda_0) + 131.1
\]

\[
(1.028 - \ln \varepsilon_r) \sqrt{(h/\lambda_0)} + 12.48 (1 + 0.18 \ln \varepsilon_r) (W_a/h) + \sqrt{(\varepsilon_r - 2.06 + 0.85 (W_a/h)^2)}
\]

RESULTS AND CONCLUSIONS

The antenna is fabricated with a resonant frequency 2.3 GHz with a slot dimensions as 20 x 8 mm. The experiments are conducted to determine the bandwidth of Dual Stacked Aperture Coupled Microstrip Antenna for different combinations of air gaps and slot width of the hour glass aperture. The results are fabricated in Table 1. It can be observed from the table that the bandwidth increases with increases in slot width as well as air gap width. Optimum performance for the bandwidth is achieved at \( h_1 = 6.8 \) cm, \( h_2 = 5.4 \) cm and \( S = 2 \) mm.
TABLE I

<table>
<thead>
<tr>
<th>h1(cm)</th>
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<td>37.0</td>
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<td>38.4</td>
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<td>1.1</td>
<td>3</td>
<td>40.0</td>
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</tbody>
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REFERENCES

EMPIRICAL DESIGN CONSIDERATION FOR WAVEGUIDE-COUPLED MICROSTRIP PATCH ANTENNA

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A novel technique has been designed to couple the advantages of both microstrip patch antenna and rectangular waveguide. Here patch antenna consists of two patches, one radiating patch which faces outside, and the other faces mouth of the rectangular waveguide. The one which faces waveguide plays a very important role in the study of radiation pattern. Before coming to this pattern which resembles the one which is proposed by J.C. Slater Series Resonator a lot of other patterns were studied. The series resonator type yields better results. For various lengths of a and b, the width and height of Series resonator, detailed study is made, and the formula proposed by J.C. Slater for series in resonator with a correction factor is presented here. Numbers of tests were conducted to verify the formula and with the help of this formula return loss graph is explained. Detail construction of waveguide-coupled microstrip patch antenna is presented along with the observed results: return loss, radiation pattern, polarization, bandwidth, beam width, main peak and side lobes.

INTRODUCTION:

In many wireless communication systems there is a requirement for low-profile antennas. The reason is that these antennas are less obstructive than traditionally used parabolic reflector. In addition, snow, rain, or wind has less effect on their performance. Microstrip patch antenna is an example of low profile antenna. In order to make this patch antenna an effective radiator, patch has to be suitably fed. Different methods can be used to achieve this goal. To couple power from the feed network to the patches different coupling mechanisms can be employed. One possible way is to use a direct connection between feed network and the patches by employing the edge feed method [1]. One disadvantage of this method is that the feed network appears in the same layer as in the radiating elements, and produces residual radiation. Alternative to the edge feed are co-axial probes and electromagnetically coupled microstrip lines. The coaxial probe requires a hybrid connection is unattractive in large arrays, as it is labor intensive. To implement the coupling mechanisms include proximity coupling and aperture coupling from microstrip or strip line to the patches. One draw back of the aperture coupled microstrip patch antenna is its backward radiation, due to slots (aperture) in the ground plane [2]. In addition to this all these methods of coupling suffer from poor power handling capacity and poor band width.

In this paper, a new design of waveguide coupled microstrip patch antenna is proposed, which incorporates attractive features of microstrip antenna like low in profile, light in weight, compact and conformable in structure, easy to fabricate and good features of waveguide, like high power handling capability, lower losses (resistive), frequency and band pass response stability, ease of tuning after manufacture and robustness. For the proposed patch earlier works were referred [4],[5],[6],[9],[11],[12]. The radiating patch in the proposed design is a simple circular patch. The proposed patch antenna has two patterns on either side. One is radiating patch which is circular which faces outside and the other which faces mouth of the waveguide plays an important role in the designing of the patch antenna. In this study the empirically designed formula is examined for various
patches. Waveguide facing patch pattern resembles the one that is proposed by J.C. Slatter for iris series ring resonator. The resonator shape is that of a rectangle with the sides, length $a'$, width $b'$ and in the present study, $a'$ and $b'$ are varied and return loss is studied. A design formula is proposed which is a corrected J.C. Slatter series ring resonator empirical formula and this formula is studied for all values of $a'$, $b'$ and detail study is made to verify this formula and correction factor is tabulated.

**ANTENNA CONFIGURATION:**

The waveguide coupled microstrip antenna to be studied is shown in the figure. Antenna consists of two sides, one radiating side that is shown in the figure as antenna side view. The other side of the antenna which faces the waveguide is shown in the figure as waveguide side view. The radiating patch is a circular patch. The waveguide facing patch pattern consists of rectangular ring which resembles iris resonator ring. The thickness of the rectangular patch is 0.215 mm from either side of the specified dimension in the figure (1).

**ANTENNA DESIGNING**

Operating frequency is the most important parameter in the antenna designing. The first step of a design is to calculate patch parameters with the help of the resonant frequency formula, but in the literature designing formulas are not there for such kind of waveguide coupled microstrip antennas. So we started from J.C. Slater formula [3] for iris series resonator. The formula is given below.

$$
f_r = \frac{c}{2a} \sqrt{1 - \left(\frac{b'}{b}\right)^2} \quad \ldots 1
$$

- $c = \text{velocity of light}$
- $a' = \text{waveguide facing rectangular patch length}$
- $b' = \text{rectangular patch width}$
- $f_r = \text{resonant frequency}$
- $a = \text{waveguide broadside breadth}$
- $b = \text{Width of the waveguide}$
This if the formula for iris series resonator, but this formula fails to give results for all values of \(a'\) and \(b'\). So the formula is modified with a correction factor 'A' so that \((a'/a)^2\) is always greater than \((b'/b)^2\). Now the modified formula is given by

\[
\frac{c}{2a} \sqrt{1 - \left(\frac{b'}{b}\right)^2 - \left(\frac{a'}{a}\right)^2} \tag{2}
\]

The correction factor 'A' is calculated for different values of \(a'\) and \(b'\) for observed frequency and graph is plotted against \((a'/a)^2\). The graph is shown in the figure (2).

The rectangular resonant length \(a'\) is to be considered first along with resonant frequency and after calculating \(a'/a\), the correction factor 'A' can be had from the above table. The resonator width can be calculated from the formula given here with \(b' = \) rectangular resonator width = waveguide width. \(A = \) Correction factor, \(f_r = \) resonant frequency, \(c = \) velocity of light

\[
\frac{A \left(\frac{a'}{a}\right)^2 \left(\frac{2af_r}{c}\right)^2 - 1}{\left(\frac{2af_r}{c}\right)^2 - 1} \tag{3}
\]
With the help of this graph a table 1 is formulated and is presented here.
Table presenting correction factor for various values of (a'/a)

<table>
<thead>
<tr>
<th>(a'/a)</th>
<th>Correction Factor 'A'</th>
<th>(a'/a)</th>
<th>Correction Factor 'A'</th>
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<td>0.950</td>
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</table>

**EXPERIMENT**

In this experiment X-band microwave bench setup is used and 12 different frequencies obtained by adjusting micrometer of gun diode [8]. These settings of micrometer along with the frequency are given below in table 2.

<table>
<thead>
<tr>
<th>Micrometer reading</th>
<th>Frequency (GHz)</th>
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<th>Frequency (GHz)</th>
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</tr>
<tr>
<td>11.53cm</td>
<td>10.746</td>
<td>13.6cm</td>
<td>10.102</td>
<td>19.45cm</td>
<td>9.499</td>
</tr>
</tbody>
</table>

For reflection coefficient measurement the reflectometer [7] (3dB direction coupler) is used and proper correction measures were taken for unequal strength of the detector diode [10]. Horn antenna mounted on the arm of the spectrometer was used for measuring radiation strength in the E-plane for 0-180° from the center position of the patch.

**RESULT AND ANALYSIS**

Here, we considered two patch antennas for the verification of the designed formula and along with this other parameters such as radiation pattern, polarization

**CASE 1:**

The waveguide-coupled microstrip patch antenna is as shown in the figure (1). The dimension of the rectangular resonator is a' = 5.7mm and b' = 4.14mm. These dimensions are
based on assumption that in TE_{10} mode electric field in the waveguide is of the shape of half cycle of sine wave. The resonator rectangles taken for study are formed by different positions on the sine wave. The observed resonant frequency is 11.114GHz where as the calculated frequency is 9.6063GHz.

The calculated value is 13.59% less than observed frequency and the return loss graph exhibits four peaks. These peaks are due to the finite thickness of the rectangular patch. The thickness of the sides of the rectangular patch is 0.215 mm. After considering the thickness, four rectangles are formed by inner and outer dimensions of the rectangle. These four rectangles are analyzed and the results are tabulated in table 3.

<table>
<thead>
<tr>
<th>The rectangular sides Considered</th>
<th>a' Breadth Of Rectangle</th>
<th>b' Width Of Rectangle</th>
<th>f', Calculated</th>
<th>f', Observed</th>
<th>Error %</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inner Outer</td>
<td>5.27mm</td>
<td>4.57mm</td>
<td>10.842GHz</td>
<td>11.114GHz</td>
<td>2.44</td>
</tr>
<tr>
<td>Inner Inner</td>
<td>5.27mm</td>
<td>3.71mm</td>
<td>9.687GHz</td>
<td>10.525GHz</td>
<td>7.86</td>
</tr>
<tr>
<td>Outer Outer</td>
<td>6.13mm</td>
<td>4.57mm</td>
<td>9.113GHz</td>
<td>10.288GHz</td>
<td>11.42</td>
</tr>
<tr>
<td>Outer Inner</td>
<td>6.13mm</td>
<td>3.71mm</td>
<td>8.793GHz</td>
<td>9.9GHz</td>
<td>11.18</td>
</tr>
<tr>
<td>centre centre</td>
<td>5.7mm</td>
<td>4.14mm</td>
<td>9.6034GHz</td>
<td>11.114GHz</td>
<td>13.59</td>
</tr>
</tbody>
</table>

The table clearly explains the observed return loss graph. For this waveguide coupled microstrip antenna the other parameters are given below.

CASE 2:

To verify the formula a different pattern thickness patch antenna is considered here. Here the dimension of the sides of the rectangular patch is 0.7mm. The rectangle of length 6.24mm and breadth 4.655mm gives the resonance at 8.9739GHz which is of 19.25% error. Again, considering the thickness, four rectangles are formed by inner and outer dimensions of the rectangle. These four rectangles are analyzed and the results are tabulated in table 4.
The other specification of the other patch is:
- Resonant Frequency: 11.114GHz
- Return loss: -30.36dB
- Band width: 1.52%
- Gain: 15.9db
- Polarization: vertical

![RETURN LOSS GRAPH](image)

**TABLE 4**

<table>
<thead>
<tr>
<th>The rectangular sides Considered</th>
<th>a' (Breadth Of Rectangle)</th>
<th>b' (Width Of Rectangle)</th>
<th>( \bar{f} ) (Calculated)</th>
<th>( \bar{f} ) (Observed)</th>
<th>Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>Breadth</td>
<td>Outer 5.47mm</td>
<td>5.39mm</td>
<td>11.325GHz</td>
<td>11.114GHz</td>
<td>1.8</td>
</tr>
<tr>
<td>Inner</td>
<td>5.47mm</td>
<td>3.39mm</td>
<td>9.9GHz</td>
<td>10.525GHz</td>
<td>5.9</td>
</tr>
<tr>
<td>Outer</td>
<td>7.01mm</td>
<td>3.39mm</td>
<td>8.25GHz</td>
<td>9.9GHz</td>
<td>16.6</td>
</tr>
<tr>
<td>Outer Inner</td>
<td>7.01mm</td>
<td>3.39mm</td>
<td>7.9GHz</td>
<td>9.499GHz</td>
<td>16.8</td>
</tr>
<tr>
<td>Centre</td>
<td>6.24mm</td>
<td>4.65mm</td>
<td>8.9735GHz</td>
<td>11.114GHz</td>
<td>19.25</td>
</tr>
</tbody>
</table>

Like this totally 20 other WGCMSA with the same a' and b' and 7 with different a' and b' were checked. The results of the entire waveguide coupled microstrip patch antenna and those calculated using empirical formula are found well within the acceptable range of observed frequency.

**CONCLUSION**

A comprehensive investigation into the waveguide coupled microstrip patch antenna has been performed for various dimensions of rectangular patch, which faces the waveguide. The main goal was to find a formula for designing the antenna but in literature such formula was not available, except the one proposed by J.C. Slatter for series iris resonator. So that formula with a correction factor is made use of in our present work and it is tested for 27 patch antennas. To make the designing easy, correction factor table is presented here. A result shows that due to the finite thickness of the rectangular resonator the inner breadth and the outer width of the rectangle is responsible for the resonant frequency. This formula holds good even for zero values of b'.

**ACKNOWLEDGEMENT**

Authors thank the Management and Principal of S.D.M. college of engineering and technology Dharwad for all the support and encouragement shown. Authors also thank Dr. D.K. Deshpande for his valuable suggestions in carrying out this work.
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RADIATION CHARACTERISTICS OF POLYGON MICROSTRIP ANTENNA IN ANISOTROPIC AND ISOTROPIC COLD PLASMA MEDIUM

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Abstract: This paper deals with the analysis of polygon microstrip antenna in anisotropic and isotropic cold plasma media. For this purpose, first a polygon microstrip antenna is transformed into a unit circular antenna using the technique of conformal mapping. The outside area of a polygon is transformed into the outside area of a circle and the circular boundary problem has been solved. It has been found that the presence of plasma augments the radiation characteristics to a great extent in both the plasma media. The directivity of antenna is also found to increase with increasing value of plasma frequency in both the plasma media. This indicates that antenna having low permittivity substrate has large scanning area.

INTRODUCTION

Recently considerable interest has been devoted to estimate the influence of anisotropic and isotropic cold plasma on the radiation characteristics and directivity of antenna. Due to numerous advantages of microstrip antenna over conventional microwave antenna, now a days microstrip antenna has got its use in various systems such as satellite, missiles, tanks and other strategic applications in defense. The light weight, low volume, low cost, planers configuration and compatibility with integrated circuit have incited the research world over to pay attention in this subject which has resulted into establishing the microstrip antenna as a separate entity in the broad field of microwave antenna. In spite of all these advantages of this antenna very few definitive data are found regarding the effect of plasma medium on the various aspects of such antenna. The plasma medium is encountered during the voyage of space shuttles, may modify the properties of these antennas. Therefore, in the present paper an attempt has been confined to estimate the influence of plasma medium on the different aspects of antenna. The details of entire investigations are given in the different sections of this paper.

THEORETICAL CONSIDERATION

In order to obtain the transformation of a polygon microstrip antenna in circular antenna, let us consider a polygon in the w-plane having vertices at w1, w2, ..., wn, with corresponding interior angles α1, α2, ..., αn respectively. Let the points w1, w2, ..., wn map respectively onto points x1, x2, ..., xn. On the real axis of the Z-plane, then a transformation which maps the interior R of a polygon Fig 1 of the W-plane onto the upper half R' of the Z-plane can be given by the relation

\[
\frac{dw}{dz} = A(z - x_1)^\alpha_1(z - x_2)^\alpha_2 \cdots (z - x_n)^\alpha_n
\]

or,

\[
w = A(z - x_1)^{\alpha_1}(z - x_2)^{\alpha_2} \cdots (z - x_n)^{\alpha_n} + B \cdots (2)
\]

Where A and B are complex constants determine the size, orientation and position of the polygon. The value of A is given as,
Where \( C \) is another constant putting the value of \( C \) in equation (2) we get

\[
m = C(x^2 - y^2)^{-\frac{1}{2}} \left( r^2 + x^2 \right)^{\frac{1}{2}} \left( \frac{x}{x^2 + y^2} \right)^{\frac{1}{2}}
\]

Further a transformation which maps the upper half of the \( z \) plane into the unit circle in the \( \xi \) plane can be given as,

\[
\xi = \frac{1}{2} \left( 1 + \frac{x}{r^2} \right)
\]

let \( \frac{x}{r^2} \) map into \( \xi \), respectively on the unit circle then one has.

Combining the equations (6), (7), and (2) solving one has,

\[
\alpha = \frac{1}{2} \left( 1 + \frac{x}{r^2} \right)
\]

dividing equation (8) by \(-\pi\) we have,

\[
\frac{1}{2} \left( 1 + \frac{x}{r^2} \right) \cdot \frac{\partial \alpha}{\partial \xi} = 0
\]

Thus the required transformation is obtained as,

\[
w = A \left( \xi, \eta \right)^{\frac{1}{2}} \left( \xi^2 + \eta^2 \right)^{-\frac{1}{2}} \left( \frac{\partial \alpha}{\partial \xi} \right) \left( \frac{\partial \alpha}{\partial \eta} \right)
\]

Where \( A \) is an arbitrary constant and given by the relation,

\[
A = \frac{C}{\left( -2 \pi \right)}
\]

Further let,

\[
\alpha = \frac{1}{2} \left( 1 + \frac{x}{r^2} \right)
\]

Then the wave equation on the \( w \)-plane will be written as,

\[
\frac{\partial^2 E}{\partial u^2} + \frac{\partial^2 E}{\partial v^2} = \left( \frac{\partial^2 E}{\partial x^2} + \frac{\partial^2 E}{\partial y^2} \right) \left( \frac{\partial \alpha}{\partial u} \right)^2 + \left( \frac{\partial \alpha}{\partial v} \right)^2
\]

Where \( x, y, u, v \) are the co-ordinates of points in the \( w \)-plane. Also

\[
\frac{\partial^2 E}{\partial u^2} + \frac{\partial^2 E}{\partial v^2} = \left( \frac{\partial^2 E}{\partial x^2} + \frac{\partial^2 E}{\partial y^2} \right) \left( \frac{\partial \alpha}{\partial u} \right)^2 + \left( \frac{\partial \alpha}{\partial v} \right)^2
\]

Also

\[
\frac{\partial^2 E}{\partial u^2} + \frac{\partial^2 E}{\partial v^2} = \left( \frac{\partial^2 E}{\partial x^2} + \frac{\partial^2 E}{\partial y^2} \right) \left( \frac{\partial \alpha}{\partial u} \right)^2 + \left( \frac{\partial \alpha}{\partial v} \right)^2
\]

Thus we have

In the similar fashion one gets

\[
\frac{\partial^2 E}{\partial u^2} + \frac{\partial^2 E}{\partial v^2} = \left( \frac{\partial^2 E}{\partial x^2} + \frac{\partial^2 E}{\partial y^2} \right) \left( \frac{\partial \alpha}{\partial u} \right)^2 + \left( \frac{\partial \alpha}{\partial v} \right)^2
\]

Combining equations (14) and (15) and using the Cauchy-Riemann conditions one has,

\[
\frac{\partial^2 E}{\partial u^2} + \frac{\partial^2 E}{\partial v^2} = \left( \frac{\partial^2 E}{\partial x^2} + \frac{\partial^2 E}{\partial y^2} \right) \left( \frac{\partial \alpha}{\partial u} \right)^2 + \left( \frac{\partial \alpha}{\partial v} \right)^2
\]

and
Now the equation (12) may be modified as...

\[ \frac{\partial E}{\partial x} \nabla^2 + \frac{\partial E}{\partial y} \nabla^2 + \frac{1}{\varepsilon(\lambda)} \left( \frac{\partial E}{\partial z} \nabla^2 + \frac{\gamma E}{\varepsilon(\lambda)} \right) = 0 \]

Where \((x, y, z)\) axe the coordinates of the points in \(\varepsilon\)-plane. If feed line is considered as a microstrip line as shown in Fig. 2 and electric field \(E\) has only 1-component then excited current on the x-axis will be written as

\[ j = \dot{\phi} \delta t \delta \theta (q, \theta) \]

Where

\[ \dot{\phi} = \psi - \phi - \phi = \psi - \phi \]

and \(\dot{\phi}\) is the unit vector on the x-axis. Then equation (18) with excited current may be modified as

\[ \frac{\partial E}{\partial x} \nabla^2 + \frac{\partial E}{\partial y} \nabla^2 + \frac{1}{\varepsilon(\lambda)} \left( \frac{\partial E}{\partial z} \nabla^2 + \frac{\gamma E}{\varepsilon(\lambda)} \right) = 0 \]

Now we will adopt the cylindrical coordinates \(r, \theta\) to get the value of electromagnetic field which as follows

\[ \begin{align*}
\vec{E} &= E_r \hat{\hat{r}} + E_{\theta} \hat{\hat{\theta}} + E_z \hat{\hat{z}} \\
\vec{B} &= B_r \hat{\hat{r}} - B_{\theta} \hat{\hat{\theta}} + B_z \hat{\hat{z}}
\end{align*} \]

Where \(F\) and \(A\) are the electric and magnetic vector potentials respectively. Further one can write the value of \(E_\theta\) and \(E_\phi\).

\[ E_\theta = \frac{j \rho \mu_0}{4\pi} \frac{\cos \phi}{r} \left[ (K_1(p, \rho \phi) \cos (\rho \phi) + J_1(p, \rho \phi) \sin (\rho \phi) \right] \]

\[ E_\phi = \frac{j \rho \mu_0}{4\pi} \frac{\sin \phi}{r} \left[ (K_1(p, \rho \phi) \sin (\rho \phi) + J_1(p, \rho \phi) \cos (\rho \phi) \right] \]

In the similar fashion the components of magnetic field can be obtained as

\[ H_r = \dot{\phi} \frac{2a^2}{\pi} \sum_{n=1}^{\infty} \frac{2\delta}{\pi} \left( \frac{2\delta}{\kappa(n, a)} \sin (\kappa(n, a) \pi a) \right) \]

\[ H_\theta = 0 \]

\[ H_\phi = \dot{\phi} (\pi a)^2 \sum_{n=1}^{\infty} \frac{\sin (\kappa(n, a) \pi a)}{\kappa(n, a)} \left( \frac{\sin (\kappa(n, a) \pi a)}{\kappa(n, a)} \right) \]

\[ \sum_{n=1}^{\infty} \frac{1}{(\pi a)^2} \eta(n, \kappa(n, a)) \left( \frac{\sin (\kappa(n, a) \pi a)}{\kappa(n, a)} \right) \]

ANALYSIS OF POLYGON PATCH IN ANISOTROPIC AND ISOTROPIC COLD PLASMA MEDIUM

As we know that the anisotropic plasma is obtained when the electron plasma is treated...
by d.c magnetic field, i.e when d.c magnetic field \( B_0 \) applied in electron plasma.

Therefore for such medium without any collision the relative permittivity \( (\varepsilon_{rp}) \) is given by the relation,

\[
\begin{bmatrix}
\varepsilon_{rp}
\end{bmatrix}
\begin{bmatrix}
\varepsilon_0
\end{bmatrix} = \begin{bmatrix}
1 & 0
0 & 0
\end{bmatrix}
\]

Where,

\[
\varepsilon_{rp} = \varepsilon_0 \left( \frac{\varepsilon_0}{\varepsilon_0} \right)^{1/2} \left( \frac{\varepsilon_0}{\varepsilon_0} \right)^{1/2}
\]

Now from equation (30) we have the value of anisotropic plasma medium can be given as

\[
\varepsilon_{an} = \frac{(\omega_0)^2}{(\omega_0)^2 + \omega^2}
\]

Now for \( \omega = \omega_0 \) the plasma medium is known as isotropic cold plasma. Hence the modified value of relative permittivity for such plasma medium is given as

Then equation (29) is modified as,

\[
\begin{bmatrix}
1 & 0
0 & 1
\end{bmatrix}
\]

Therefore the value of the propagation constant \( (\gamma_p) \) in anisotropic and isotropic cold plasma medium can be given as respectively,

\[
\gamma_p = \frac{2\pi}{\lambda_{an}}
\]

and

\[
\gamma_p = \frac{2\pi}{\lambda_{in}}
\]

Hence the value of \( E_\theta, E_\phi, H_\theta \) and \( H_\phi \) for anisotropic plasma medium and isotropic cold plasma medium can be given as respectively. For anisotropic cold plasma medium,

\[
E_\theta = \frac{\rho_0 \mu}{4\pi} \left\{ \frac{K_1(\rho, \phi) \cos(\phi')}{\rho d\phi d\rho} \right\}
\]

\[
K_1(\rho, \phi) \cos(\phi') + \lambda_{an} \sin(\phi')
\]

For isotropic cold plasma medium,

\[
E_\theta = \frac{\rho_0 \mu}{4\pi} \left\{ \frac{K_1(\rho, \phi) \cos(\phi')}{\rho d\phi d\rho} \right\}
\]

\[
K_1(\rho, \phi) \cos(\phi') + \lambda_{in} \sin(\phi')
\]
For isotropic cold plasma medium

\[
E_r = \frac{-j\mu_0 \epsilon_0 \omega^2 \mu J}{4\pi} \int \int \frac{2\delta}{k^2 + \pi^2 a^2} e^{i(k_x x + k_y y + k_z z + \phi)} d\rho d\phi
\]

\[
E_\phi = \frac{-j\mu_0 \epsilon_0 \omega^2 \mu J}{4\pi} \int \int \frac{2\delta}{k^2 + \pi^2 a^2} e^{i(k_x x + k_y y + k_z z + \phi)} d\rho d\phi
\]

\[
H_z = \frac{\omega^2 \mu \epsilon_0 J}{2\pi} \int \int \frac{2\delta}{k^2 + \pi^2 a^2} e^{i(k_x x + k_y y + k_z z + \phi)} d\rho d\phi
\]

\[
H_y = \frac{\omega^2 \mu \epsilon_0 J}{2\pi} \int \int \frac{2\delta}{k^2 + \pi^2 a^2} e^{i(k_x x + k_y y + k_z z + \phi)} d\rho d\phi
\]

**DIRECTIVITY**

The directivity for the free space is given by the relation

\[
D_o = \frac{2\pi}{G_{rad}}
\]

Where \( G_{rad} \) = radiation conductance

The value of directivity (Do) in anisotropic arid isotropic cold plasma medium can be modified respectively as,

\[
D_{\eta_1} = (\eta_1, a, \eta, \omega, \epsilon_0, \mu_0, J_0, \sigma_0, \alpha, \beta, \delta)
\]

**NUMERICAL COMPUTATION**

In order to obtain the value of \( \frac{\mu_0}{\epsilon_0}, K_B, B, \) and directivity indifferently plasma media the computational works were done using equations (33), (34), (35), (36), (37), (38), (39), (40), (421) and (43) respectively

\[
for \, \delta = 2\pi, K_B = 1.84, 0, 10, 20, 50, ...
\]

\[
\eta_0 = 0.1, 0.2, 0.3, 0.4, 0.5, ...
\]
DISCUSSION OF RESULTS

The examination of E-plane radiation pattern for polygon microstrip antenna in anisotropic and isotropic cold plasma medium reveals that the presence of such plasma media enhances the radiation characteristics of antenna to a significant value. The radiation becomes more omni directional in half space with increasing value of the $\omega_1$. At the value of $\omega_1$ greater than $\omega_0$, the radiation can be start to flap within certain solid angle. This indicates that the directivity of antenna can be controlled by controlling the operating frequency. It is also observed that the plasma medium has more significant effect on directivity of the antenna, than radiation characteristics. As the value of d.c. magnetic field increases, the radiation becomes less omni directional in anisotropic plasma medium. It may therefore be concluded that there is a significant change in the magnitude of radiation field intensity in plasma medium than free space.

REFERENCES

Fractal Patch Antenna for 2.4 GHz Band

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Abstract: Fractals are space-filling contours and can be viewed as a viable miniaturization technique to efficiently pack electrically large features into small areas. In this paper, a novel fractal microstrip patch antenna is proposed for wireless communication at 2.4 GHz band. An advantage of this antenna is that it provides an optimized antenna area resulting in substantial reduction in size of a typical microstrip square patch designed at the same frequency of operation. The characteristics of the proposed antenna are determined in terms of Gain, return loss and VSWR bandwidth using the IE3D full wave simulator. It is found that the patch area of the proposed antenna is lowered by about 54% than that of conventional square microstrip antenna.

INTRODUCTION

As communication devices become smaller due to greater integration of electronics, the antenna becomes a significantly larger part of the overall package volume. This results in a demand for similar reduction in antenna size.

In existing built in antenna schemes much attention has been paid to planar structures such as microstrip aperture coupled, and slot antennas. These planar configurations though light weight, low cost, low profile, conformable, reliable, reproducible, easy to fabricate and suitable for integration with solid state devices, but exhibit excessive size for new wireless applications. Additional effort is needed to further reduce actual geometries in order to obtain miniaturized design. Different solutions [1-3] for reducing size include using high dielectric constant material, shorting planes or surface etching etc.

Consequently, the present work focuses on developing antenna geometries small enough to be easily accommodated in modern wireless communication terminals and devices. Fractal patch antennas have been studied in the past incorporating the properties of the Koch island fractal geometry and behavior at the fundamental mode and the existence of high-order modes exhibiting localized current density distribution is discussed.

In the present work, a square microstrip patch antenna and a fractal microstrip patch antenna are analyzed using IE3D, a commercial simulator, based on method of moment (MOM) and compared. The proposed antenna is compact, having a patch area less than that of a conventional square microstrip patch antenna fabricated on the same substrate and resonating at the same frequency. These antennas can find application in the WLAN 802.11b communication standard operating at 2.4 GHz. The fractal antenna elements can also find applications in linear phased arrays, to realize more elements tightly packed into a linear array. This configuration of linear array further reduces the amount of mutual coupling between the elements which otherwise would lead to a degradation in the radiation pattern. Fractal Generation

The fractal is generated by applying an iterative sequence to the starting structure. The starting structure in the present case is a square patch (28.2mm x 28.2mm). Fig.1 shows the fractal patch antenna generations up to four iterations keeping the electrical length constant.
As evident from this figure, the area occupied by the fractal patch decreases with the increase in iteration.

ANTENNA DESIGN

Fig. 2 shows the conventional square microstrip patch antenna with patch dimension L x L as 28.2 mm x 28.2 mm. The patch is printed on a dielectric substrate FR-4 (εr = 4.4) with a backplane conductor to form a microstrip antenna. The thickness of the substrate is assumed to be 1.6 mm. The antenna is probe fed, which is the most widely used feeding method in microstrip antenna. The coaxial feed position is determined to give optimal matching at 4.8 mm. The patch is found to resonate at 2.46 GHz. The figure 3 illustrates the variation of the conventional square microstrip patch antenna under investigation. The proposed patch antenna is fractal patch configuration after third iteration, printed on the same substrate. The effective size of the square patch is found to be reduced by 54% at a given frequency of operation. Next, the new feed position is determined by the iterative method for best possible impedance matching as 2.8 mm. The return loss, VSWR and gain are plotted and compared in Fig. 4, Fig. 5, and Fig. 6.

One problem associated with the proposed configuration of the patch is the diminishing antenna gain. However, incorporating an active microstrip fractal patch configuration can improve the gain significantly.

CONCLUSION

A microstrip fractal patch is proposed at 2.4 GHz band. The proposed patch configuration reduces the conventional square patch dimension by 54%, at the particular frequency of operation. This antenna can be used suitably in the WLAN 802.11b communication standard operating in 2.4 GHz band and Fractal Patch antenna elements also find application in linear phased arrays.

REFERENCES

Fig. 1 Fractal patch antenna generation

Fig. 2 Conventional square patch antenna

Fig. 3 Fractal Patch Antenna
Fig. 4 Return Loss vs. Frequency comparison

Fig. 5 VSWR vs. Frequency comparison

Fig. 6 Gain vs. Frequency comparison
EVEN MODE MULTI-PORT NETWORK MODEL FOR DUAL BAND RECTANGULAR MICROSTRIP ANTENNAS

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Abstract: The even mode multi-port network model of rectangular microstrip antenna is discussed, which reduces the computation time by more than half. This concept is extended to analyze dual band E-shaped MSA and U-slot RMSA.

INTRODUCTION

The multi-port network model (MNM) for variations in rectangular microstrip antenna (RMSA) is reported [1]. However, for complex shape antennas, the computation time increases due to increased number of segments and ports. If the MN model of the configuration is symmetrical with respect to the feed point axis, then its even mode equivalent has been used to reduce the computation time [2]. In this paper, the even mode analysis for the RMSA is discussed. This concept is extended to analyze dual band slotted RMSAs like E-shaped microstrip antenna (MSA) and U-slot RMSA.

EVEN MODE ANALYSIS OF RMSA

A RMSA and its segmented network using MNM are shown in Fig. 1 (a, b). For the MNM, \( N_a \) and \( N_b \) ports are taken along the non-radiating and radiating edges, respectively. The RMSA is loaded with the sum of radiation and surface wave conductances on both the radiating and non-radiating edges. The size of the Z-matrix for RMSA is \( 2(N_a+N_b) \times 2(N_a+N_b) \), corresponding to the connected ports. The segmentation method is used to evaluate the input impedance.

The configuration is symmetrical with respect to the feed point axis. Therefore only half portion is solved to find the input impedance. The even mode network of the RMSA is given in Fig. 1 (c). The width of the RMSA is reduced by half, which reduces the number of ports along the radiating edge by half thereby maintaining the same port width. The ports are taken along the two radiating edges and one non-radiating edge. No ports are taken along the edge where the feed is present. This reduces the size of the Z-matrix to \( (N_a+N_b) \times (N_a+N_b) \). The values of effective dielectric constant, loss tangent, and the total conductance at each port are kept the same. They are not recalculated using the reduced width as the width is reduced only for the computational simplicity. The input impedance of even mode network is divided by two to obtain the input impedance of the entire network.

For RMSA with length \( L = 6 \) cm and width \( W = 4 \) cm on RT-Duroid substrate (\( \varepsilon_r = 2.33, \tan \delta = 0.001 \)), with \( N_a = 14, N_b = 8 \), and feed point at \( x_f = 0.7 \) cm, the input impedance and the resonance frequency using MNM are \( (42.12 - j0.5) \Omega \) and 1.617 GHz, respectively. For the even mode network, these values are \( (42.12 - j0.7) \Omega \) and 1.617 GHz. The order of matrix size for the MNM is \( 44 \times 44 \) and for the even mode network it is \( 22 \times 22 \). The impedance and frequency are very close to each other in the two methods and the size of the Z-matrix is reduced by 3 times for even mode method, which reduces the computing time by more than half. Thus the above method is very useful for the complex shape MSAs, which involves larger number of segments and ports. MNM and even mode analysis of dual band slotted RMSAs is presented in the following section.

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DUAL BAND E-SHAPED MSA

An E-shaped MSA with dimensional details is shown in Fig. 2 (a). The dual band response is due to the patch (TM_{10}) and slot mode. The slot length is approximately 2/4 at the desired resonance frequency. The dual band response was optimized with the slots separated by 2Y_s = 6 cm from each other and feed point located at x_f = 2.5 cm from the center of the patch, which is inside the slotted area. The length and width of the slots are l = 4.0 and w = 0.2 cm, respectively. For the glass epoxy substrate (\varepsilon_r = 4.3, h = 0.159 cm, tan \delta = 0.02) the input impedance and VSWR plots using IE3D are shown in Fig. 2 (c) [3]. The dual frequencies are 736 and 892 MHz. The bandwidth at the two frequencies is 14 and 15 MHz. The second frequency is close to the resonance frequency of un-slotted RMSA.

The E-shaped MSA is analyzed using MNM as shown in Fig. 2 (b). The MSA is divided into four segments. The number of ports along the segmented edge is selected such that at least a single port is present along the slot width. As w = 0.2 cm, therefore for the second segment fifty ports (10 / 0.2 = 50) are taken along the segmented edge. The input impedance and VSWR plots using MNM are given in Fig. 2 (c), which shows dual band response at 710 and 909 MHz. The bandwidth at the two frequencies is 11 and 16 MHz. The voltage and field distributions along the patch obtained using MNM at the two resonant frequencies is shown in Fig. 3 (a – d). At the first frequency a half wavelength variation is observed across each of the slots indicating that the first frequency is governed by the slot mode whereas for the second frequency a distribution similar to that of the un-slotted RMSA is present. The radiation pattern at first frequency shows higher cross-polarization level due to the asymmetric voltage distribution around the slotted patch and is in the broadside direction with lower cross-polar level at the second frequency.

For a segmented network of E-shaped MSA, the Z-matrix size is 196 x 196. This increases the computation time for a single frequency to 135 seconds. Hence to reduce the time, an even mode equivalent of E-shaped MSA is analyzed as shown in Fig. 4. The even mode network reduces the matrix size to 98 x 98, which has reduced the computation time to 53 seconds. The input impedance and VSWR plots and the voltage distributions at the two resonant frequencies have similar variation as that given by MNM. Similarly, a dual band U-slot RMSA is analyzed using MNM and even mode network as shown in Fig. 5 (a – c).

CONCLUSION

The even mode network reduces the computation time by more than half and hence is an effective tool for analyzing MSAs with larger number of segments. This concept has been used to analyze various dual band slotted RMSAs like E-shaped MSA and U-slot RMSA.
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Fig. 1 (a) RMSA its (b) segmented network and its (c) even mode network

Fig. 2 (a) E-shaped MSA its (b) MN model and its (c) input impedance and VSWR plots, (—) MNM, (—-) IE3D
Fig. 3 Voltage and field distribution for F-shaped MSA at (a, b) f; and (c, d) f

Fig. 5 (a) U-slot RMSA its (b) MN model and its (c) even mode network
SMART ANTENNA APPLICATION TO A WIRELESS THIN-CLIENT NETWORK WITH ADAPTATION FRAMEWORK

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ABSTRACT: This paper proposes and analyses the applicability of the smart antennas to the wireless thin-client (TC) network with an adaptation framework. This framework has the capability to compress the bandwidth (BW) as per the requirements of the TCs. It is modified to adapt also to the requirements of smart antennas. The algorithms Constant Modulus Algorithm (CMA), and Total Least Squares - Estimation of Signal Parameters via Rotational invariance Techniques (TLS-ESPRIT) are proposed for beam forming and DOA estimation respectively. The hardware approach chosen for the implementation of the proposed algorithms for smart antennas is the Digital Beam forming technique.

INTRODUCTION

Thin-Client (TC) computing has brought about a higher level of centralized control to the desktop. A major force behind the implementation of TCs is the low Total Cost of Ownership (TCO) it offers. A wireless TC network basically requires a server, a transceiver, and a TC. So far only omni directional antennas have been used for wireless communications in TC networks. In this paper, the application of a smart antenna has been proposed for a "TC-server based Wireless network," and also the modification of the adaptation framework [1] to adapt to the smart antenna is explained. In Smart Antenna systems, knowledge is gained and applied via algorithms processed by a Digital Signal Processor (DSP). In this paper, the algorithms have been carefully chosen, so that they comply with all the required conditions and produce desired results. The hardware implementation of this framework with smart antennas requires the use of Digital beam forming methods.

THE TC NETWORK

NETWORK DESCRIPTION: The concept of TC computing has taken a significant position in today's business computing networks. In general, TCs are low power handheld devices, which deal with the challenges of multipath fading, polarization mismatch and interference, which tend to degrade the quality of transmission. All the processing takes place at the server; the TC just needs to have an access to the server. TCs have to rely on their connections to the server for applications since they do not have the ability to run off-the-shelf programs independently. A wireless mobile network has a possibility of the loss of connectivity. Such networks need automatic address assignment, to be operational. Most of the data transferred is multimedia, which needs a broadband of frequencies; the average data transfer rate for a normal network being around 10Mb/s. The frequency ranges of the signals used being from 2.4 GHz to 2.483 GHz, which are determined by the IEEE 802.11 standards for Personal Communication Systems (PCS). The type of digital modulation used is the ideal Binary Phase Shift Keying (BPSK), whose waveform always has a constant envelope. The server is stationary and is located at the base-station and the TCs are mobile, the area of coverage having a few hundred meters of radius.
**THE ADAPTATION FRAMEWORK:** Latency is added to the signal during the transmission in this network. So, tradeoff decisions have to be made amongst bandwidth (BW), latency, error rate, and usage cost. The TC system dynamically changes the type or level of compression of data or BW used to transmit screen updates to a client over a wireless link. Battery energy usage is optimized by choosing the encoding schemes with low computational complexity [1]. Let us consider a TC network where B is link BW in bps, \( \alpha \) is the compression ratio, \( D(\alpha) \) is the decoding rate and \( 1/\mu \) is the average rectangle size in bits/rectangle, where the rectangle size is the size of the pixel on the screen. The average total latency which is the sum of the latencies of the server, communication channel and the transceiver of the TC assuming that \( B/\alpha \gg a \) and \( D(\alpha) \gg a \) where ‘a’ is the arrival rate in rectangles per second; is given as [1],

\[
T_{\text{total}} = \frac{1}{\mu} \left[ \frac{1}{B} + \frac{1}{D(\alpha)} \right] = \frac{1}{\mu} \frac{1}{BW_{\text{virtual}}} \quad \text{(1)}
\]

Therefore, we have

\[
BW_{\text{virtual}} = \frac{B \cdot D(\alpha)}{\alpha D(\alpha) + B} \quad \text{(2)}
\]

Here, the \( BW_{\text{virtual}} \) is a virtual bandwidth that results due to the compression. Changing the value of \( \alpha \) as per the requirements of the TC optimizes \( BW_{\text{virtual}} \). To this adaptation network, a fuzzy-null based proxy adaptation is used. The TC specifies the target BW as QDo, where Q is the target factor and Do is the best-case scenario decoding rate. \( BW_{\text{virtual}} \) is made almost equal to the target. The error signal is given as,

\[
Error = BW_{\text{virtual}} - QD_0 \quad \text{(3)}
\]

This error signal is used to drive the fuzzy engine that outputs a new value for the compression ratio \( \alpha \). The Fig.1 represents the selection of the optimal point where the tradeoff is made between the decoding rate and the virtual bandwidth, the point being the intersection of the two plots. The total energy consumed by a single screen rectangle is given as,

\[
E_{\text{total}} = \frac{k_c}{\mu D(\alpha)} + k_d \frac{\alpha}{\mu B} \quad \text{(4)}
\]

Here, ‘\( k_c \)’ is energy cost per unit time for processor and ‘\( k_d \)’ is the same for the transceiver in receiving mode. This shows that power is also made adaptive as a function of \( \alpha \). So, power usage is optimized based on the available battery life. The fuzzy network used can be designed such that it detects the context information of the smart antenna by using the fuzzy inference engine of the proxy adaptation and changes the BW accordingly. Here some parameters related to the smart antenna are added to the fuzzy rule base. This relieves the transceiver from the use of broadband frequencies for multimedia transmission.

**THE NEED FOR A SMART ANTENNA**

Smart antennas are applied for a transceiver where power is lost due the use of omni directional antennas and where the signals are received from a number of sources (for the server).
The array processing in smart antennas involves the manipulation of signals induced on elements in such a way that the nulls are steered to reduce the interference to a maximum extent.

SMART ANTENNA APPLICATION TO THE TC NETWORK:

The smart antenna is an array of antennas arranged in a specific manner. Constant Modulus Algorithm (CMA) uses the information collected by the elements of the array to calculate the weights of the signals. The weights calculated are used to find the steering vectors required to find the Direction Of Arrival (DOA) of the signals. Once DOA is estimated, the radiation patterns are steered in the required directions and the nulls in the direction of the unwanted signals. There is an N dimensional space, which is divided into 'Signal subspace' and 'Noise subspace', which are orthogonal to each other. M steering vectors associated with M sources span signal subspace. Based on the eigen decomposition of the sample covariance matrix of the array-received signals, the noise subspace is associated with the lowest eigen values (noise variance) are orthogonal to the signal subspace corresponding to the actual DOAs of arriving signals. A beamformer performs spatial filtering to separate signals that have overlapping frequency content but originate from different spatial locations. The DOA algorithm supplies the information to the beamformer, which orients the maximum radiation pattern toward the Signal

![Diagram](image)

Of Interest (SOI) and rejects the interference, which is the Signal Not Of Interest (SNOI), by placing nulls toward their direction [5] as shown in Fig 2.

ADAPTIVE BEAMFORMING: Array signal processing involves the manipulation of signals induced on the elements of an array. The beamforming is a process where the optimal
weights of signals are estimated by using adaptive algorithms, which use the information derived from the array output signals. For beamforming, factors like convergence speed, mutual coupling, interferences from other sources, etc are considered. The convergence speed of the algorithm refers to the speed with which the mean of the estimated weights (ensemble average of many trials) approaches the optimal weight. There is a large correlation between signals due to the use of similar or same applications of the TCs on the server. Multipath fading is present to a small extent. Since the signals have a constant modulus envelope due to the BPSK modulation technique, CMA is proposed for beamforming. CMA is an iterative gradient-based algorithm. It is capable of maximizing the carrier to noise ratio (CNR), and strongly rejects the interference signals. In practice, the correlation matrices of signals and the noise, R and Rn respectively are not available and they have to be approximated using the available information [3]. The array weights are estimated based on the DOA estimates, array signals and array outputs. These iterations continue till the optimal solution is obtained. They are calculated as,

\[ w(n+1) = w(n) - \mu g(w(n)) \]  

where, \( \mu \) is the gradient step size, which decides the convergence speed, \( w(n) \) is the array weight vector at the instant n, and \( g(w(n)) \) is the unbiased estimate of the gradient of the mean squared error or the mean output power. If \( y(n) \) is the output of the beamformer when operating with weights \( w(n) \) at the nth iteration given as

\[ y(n) = w^H(n) * x(n+1) \]  

where \( w^H(n) \) is the complex conjugate transpose of the vector matrix. If \( y_o \) is the desired amplitude in the absence of interference, the cumulative mean square error at the nth iteration (cost function) is minimized as

\[ J(n) = \frac{1}{2} \| y(n) - y_o \|^2 \]  

The cost function is minimized for the solution of the optimum location of the server.

The gradient is estimated by [3],

\[ g(w(n)) = 2 * \epsilon(n) * x(n+1) \]  

where the error signal \( \epsilon(n) \) is given as,

\[ \epsilon(n) = (y(n) - y_o)^H * y(n) \]

where \( y(n) \) is the array signal vector at the time instant n. The weight update equation used for this is,

\[ w(n+1) = w(n) - 2 * \mu \epsilon(n) * x(n+1) \]

where \( \epsilon(n) \) is the array signal vector at the time instant n. From this estimate of the gradient, the weights are estimated in iterations till the optimal weight is obtained. Once the optimal weight is obtained, the beamformer directs the maxima towards SOI and nulls towards SNOI, which are calculated using the algorithm of DOA estimation.

**DOA ESTIMATION:** The weights achieved from the CMA algorithm are used to provide substantial information for finding the DOA. We use the Total Least Squares - Estimation of Signal Parameters via Rotational Invariance Techniques (TLS-ESPRIT) for the DOA estimation. Consider a ‘Uniformly spaced Linear Array’ (ULA) of ‘N’ omnidirectional elements with a spacing of \( \lambda/2 \), where \( \lambda \) is the wavelength. DOA is calculated in terms of the angle with which the signal from the terminal arrives with respect to a reference direction. When calculating DOA, factors like correlation between signals from different sources, multipath fading, latency of the algorithm, etc; are considered. The number of directional sources is estimated by using either of Akaike’s information criterion (AIC) or minimum description length (MDL) methods. The ESPRIT divides the ULA into 2 sub arrays (imaginary) such that the first K elements (K<N, K>\lambda/2) form the first sub array and the last K elements form the second sub array and the array elements are used as pairs. x1(t) and y1(t) are the signals received by the 1st pair and x(t) and y(t) are the signals received by the first and the second sub-arrays respectively. These signals are incorporated with a time delay between
each other, which are represented in terms of phase difference as \[3\].

\[
y(t) = x(t) \exp(j 2 \pi \Lambda \cos \theta) \quad \ldots (11)
\]

where \(\Lambda = \lambda / 2\); \(\lambda\) is the wavelength of the signal received; \(\theta\) is the angle of arrival of that particular source. In terms of the steering vectors and noise induced in the arrays, \(x(t)\) and \(y(t)\) are represented as:

\[
x(t) = [A]s(t) + n_s(t) \\
y(t) = [A][\Phi]s(t) + n_t(t) \quad \ldots (12)
\]

where \([A]_{KxM}\) is the steering vector matrix, \([\Phi]_{MxM}\) is the matrix whose diagonal elements is given as \(\Phi_{mm} = \exp(j 2 \pi \Lambda \cos \theta_m)\), \(s(t)\) is the source signal, and \(n_s(t), n_t(t)\) are the noises induced on the two sub arrays. The measurements made from the array elements are used to find out the array correlation matrices \(R_{XX}\) and \(R_{YY}\). \([Ux]\), \([Uy]\) denote 2 \(KxM\) matrices for each of the 2 sub arrays respectively, with their columns denoting the \(M\) eigenvectors corresponding to the largest eigenvalues of \(R_{XX}\) and \(R_{YY}\) respectively. From these, we form a \(2Mx2M\) matrix given as,

\[
\begin{bmatrix}
U_x'' \\
U_y''
\end{bmatrix}
\times
\begin{bmatrix}
U_x' \\
U_y'
\end{bmatrix}
\ldots (13)
\]

where \([U_x'']\) and \([U_y'']\) are complex conjugate transpose matrices of \([Ux]\) and \([Uy]\) respectively, and then find the Eigen vectors of this matrix which form the columns of a matrix \([V]_{2Mx2M}\) given as,

\[
V = 
\begin{pmatrix}
V_{11} & V_{12} \\
V_{21} & V_{22}
\end{pmatrix}
\ldots (14)
\]

where \(V_{11}, V_{12}, V_{21}\), & \(V_{22}\) are sub-matrices of the order \(MxM\). From this matrix, we find the value of \(V_{11}V_{12}^{-1}\) and calculate the Eigen values \(\lambda_m\), \(m=1, 2, \ldots, M\) of the new matrix. The angle of arrival \(\theta_m\) is estimated using,

\[
\theta_m = \cos \left( \frac{\text{Arg} (\lambda_m)}{\pi \lambda} \right), \quad m=1, 2, \ldots, M. \ldots (15)
\]

where, \(\lambda\) is the wavelength of the signal. Here the angles and delays can be estimated independently of each other.

![Figure 3](image)

However this does not give a pairing of angles corresponding to delays and might result in poor resolution for closely spaced angles and delays. Smoothing techniques \([6]\) can be applied to the TLS-ESPRIT to improve the resolution. The results achieved by this algorithm are used as inputs to form the radiation patterns in the given directions in the required form. Consider a case of one signal source and one noise source with an adaptive array of say 5 elements with 2 sub arrays of 3 elements each as shown in fig.3. The weights of the signals received are calculated using the CMA algorithm. The steering vector associated with the direction \((\Phi, \theta)\) is given as

\[
\text{(Equation)}
\]
\[ s = [\exp(j2\pi f_0 t_1(\Phi_1, \theta_1)), \ldots, \exp(j2\pi f_0 t_N(\Phi_1, \theta_1))] \]  \text{(16)}

where, \( f_0 \) is the fundamental frequency, \( t_n(\Phi_1, \theta_1) \) is the time taken by a plane wave to arrive from the \( i \text{th} \) source in the direction \((\Phi_1, \theta_1)\) which is the direction of the \( i \text{th} \) source and measured from the \( n \text{th} \) element. Therefore the Steering vector matrix is, \( A = \begin{bmatrix} s_1 & s_2 & \cdots & s_M \end{bmatrix} \), \( M \) being the number of directional sources. The elements of the matrix \([A]\) can be found out using the weights derived by the CMA algorithm. Here there are only \( N=5 \) elements and \( M=2 \) sources as shown in the Fig.3. Thus the TLS-ESPRIT algorithm is utilized in this environment to reduce the constraints of the array shape and to estimate the nearest possible angles of arrival.

**SMART ANTENNA APPROACH FOR HARDWARE**

The Digital Signal processing with Digital beamforming is proposed as the optimal solution for the hardware implementation of the described TC network. This approach is better understood from the Fig.4. Computation intensive and throughput intensive tasks are separated. Advanced adaptive algorithms can be applied with a good flexibility. In this approach, Signal Processing and beamforming both are done using a Digital Signal Processor (DSP). Digital beamforming provides high flexibility, low throughput and allows programming an advanced algorithm. The signal information received is fed to the Signal Processor, which performs the assigned algorithm on the information and gives the required output for beamforming. The filters perform the required tasks and the DSP gives the required output in the digital format. Signal phase and amplitude can be restored in DSP if the spatial sampling is fast enough. The TC network data throughput required is less than 10-Mb/s due to the virtual bandwidth implementation by the adaptation framework. For a typical DSP system, the maximum data throughput allowable is around 10-Mb/s, thus making the system efficient.

![Figure 4.](image)

**CONCLUSIONS**

Analysis of the application of an adaptation framework to a wireless TC-server based network is presented in this paper. A modification to this framework to adapt to the smart antenna requirements is proposed. The smart antenna application to the server side of this network is presented. Two algorithms, CMA for beamforming and TLS-ESPRIT for DOA estimation have been proposed and analyzed for this network. The server side smart antenna
approach for hardware implementation i.e. Digital Processing and Digital Beamforming is discussed. The modification of adaptation framework results in real BW reduction, latency improvement, and reduction in power consumption at the terminals. Use of smart antennas increases the capacity of the transmission saving valuable spectrum, supports increasing data rates, extends range coverage, and reduces multipath fading, co-channel interferences, system complexity and cost, Bit Error Rate (BER), and outage probability and improves the factors like convergence speed, QoS, power control, data throughputs and battery life in TCs. As a result of the hardware approach chosen, the data throughput requirements are reduced. The method of analysis and the algorithms presented establish the firm demand and supremacy of a smart antenna for practical implementation of wireless TC networks with the associated merits.

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COMPACT WIDEBAND ANTENNA FOR BLUETOOTH™ APPLICATIONS


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A novel low profile wideband antenna of dimensions 15mm x 14.5mm x 1.6mm resonating at 2.47 GHz with a 2.1 VSWR bandwidth of 18.6% and an average gain of 5 dBi is proposed. The antenna can be directly excited using a 50 Ω coaxial probe. Details of the antenna prototype and experimental results are presented and discussed.

INTRODUCTION

Recent developments in wireless communication demand low profile antennas with good reflection characteristics, gain, wide impedance bandwidth and radiation coverage. Several antennas have already been reported [1-3] to suit the above requirement. But, the inherent narrow bandwidth and broadside radiation characteristics of a microstrip patch antenna, limits its application in this area. Most of the compact antennas exhibit poor radiation efficiency. A 50 ohm coaxial line fed narrow flat plate antenna with an inverted L slot [4] operating in 2.4 GHz band (2400 - 2485 MHz) gives a gain of 3.6 dBi. In this paper a novel compact wideband antenna resonating at 2.4GHz exhibiting an impedance bandwidth of 18.6% and an average gain of 5dBi is proposed.

ANTENNA CONFIGURATION

Figure 1 shows the geometry of the proposed antenna fabricated on a substrate of dielectric constant, εr = 4.7 and thickness, h = 1.6 mm. It consists of a small rectangular portion of length L and width W. Two identical arms of dimensions a and b are placed symmetrically on one of the longer sides of the rectangle, at an offset s from either corners. The optimum feed point of the antenna is on the shorter edge of the rectangle at a distance d as shown in the figure. The dimensions of the geometry were experimentally tuned to achieve resonance in the required band.

![Fig.1 Geometry of the proposed antenna](image)

a. Top view  b. Side view

L=15mm, W=4.9mm, h=1.6mm,

s=1.9mm, \( a=0.3mm, b=9.6mm, \)

\( d=0.3mm \)
RESULTS

The return loss characteristics of the proposed antenna are shown in Figure 2. The antenna resonating at 2.47 GHz exhibits -10 dB return loss from 2260 MHz to 2720 MHz, covering the 2.4 GHz band. The radiation patterns at 2.47 GHz in the yz, xz and xy planes are shown in Figures 3a, 3b and 3c respectively. The patterns are nearly omni directional in the elevation planes (xz and yz planes). The antenna shows similar radiation patterns in the entire band with an average gain of 5dBi. At 2.485 GHz a peak gain of 6 dBi is observed as shown in Figure.

---

Fig. 2 Reflection characteristics

Fig. 3 Radiation patterns at 2.47 GHz
a. yz plane  b. xz plane  c. xy plane
Length of arm 2 is used to tune the resonant frequency and arm1 to achieve better impedance matching. Effect of varying the arm length on resonant frequency is shown in figure 5.

The proposed configuration with an effective area of 218 sq mm gives an area reduction of ~73% with respect to the circular patch resonating at the same frequency. From the experimental studies it is found that the design parameters of the antenna are $L=0.125 \lambda_0$, $W=0.0403 \lambda_0$, $a=0.079 \lambda_0$, $b=0.00247 \lambda_0$, $d=0.0156 \lambda_0$ and $s=0.00247 \lambda_0$, where $\lambda_0$ is the free space wavelength at the operating frequency.

CONCLUSION

A new compact wideband antenna with a simple geometry is proposed and experimentally verified. The antenna exhibits moderate gain and omnidirectional radiation in
the 2.4 GHz band. The antenna can be directly excited through a
simple probe extending from the circuit board, allowing its easy integration with Bluetooth™ enabled devices.

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RECTANGULAR MICROSTRIP ANTENNA ON EBG GROUND PLANE WITH UNEQUAL ORTHOGONAL PERIODS

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Electromagnetic Band Gap (EBG) substrate of unequal orthogonal period in the ground plane of a dual-band Rectangular Microstrip Antenna is studied in this paper. The two operating frequency of the Rectangular Microstrip Antenna offered better performance in comparison with the conventional antenna.

INTRODUCTION

Electromagnetic Band Gap (EBG) structures are periodic, composite, metallic or dielectric structures that exhibits pass and stop band in their frequency response. They forbid the propagation of electromagnetic waves whose frequency is with in the stop band, so termed as Electromagnetic Band Gap. This property of EBG structures to confine, manipulate or guide electromagnetic waves through periodic composite structures can be used in many novel microwave applications. EBG structures find applications as filters in microstrip lines [1] and as substrates for printed antenna structures [2, 3, and 4].

In this paper, experimental investigation on an EBG structure of unequal orthogonal periods is presented. The proposed EBG structure has different stop bands in the two orthogonal directions. Here, the surface wave excitation of Rectangular Microstrip Antenna (RMSA) in the two operating frequencies is suppressed by the two orthogonal periods thus enhancing the antenna performance for both the frequencies.

ANTENNA DESIGN AND EXPERIMENTAL RESULTS:

![Figure 1](image)

Figure 1: Geometry of the proposed EBG backed antenna

Dual band RMSA on EBG ground plane is shown in Figure 1. RMSA of dimension L X W is etched on a dielectric slab of thickness \( h \) and dielectric constant \( \varepsilon_r \).
For a stop band to be centered at \( f \), \( \beta d = \pi \), where ‘\( \beta \)’ is the propagation constant and ‘\( d \)’ is the period. The EBG structure consists of 2x2 circular slots of \( r = 0.75 \text{ cm} \) in the ground plane with orthogonal periods \( d_1 \) and \( d_2 \). These periods are selected such that the stop band center frequency of each period is equal to the operating frequency of the RMSA in that direction.

Three feed points are selected along the diagonal of the RMSA. Feed points \( f_1 \) and \( f_3 \) excites both \( TM_{10} \) and \( TM_{01} \) mode frequencies of the antenna while \( f_2 \) merges both the frequencies resulting in a wide band.

RMSA with \( L \times W = 4 \times 3 \text{ cm}^2 \) is fabricated on a dielectric substrate with \( h = 0.16 \text{ cm} \) and \( \varepsilon_r = 4.28 \). This antenna with metallic ground plane when coaxially fed gave resonance at 2.55 GHz and 1.95 GHz. When EBG ground plane with \( d_1 = 4\text{cm} \) and \( d_2 = 3\text{cm} \) was incorporated the antenna performance was enhanced. The feed point has considerable influence in the reflection characteristics. When the antenna was coaxially fed at \( f_1 = (1.5, 1) \) and \( f_1 = (2.5, 2) \) dual frequencies are excited at 1.925 GHz and 2.45 GHz with a gain of 8.15 dBi and 9.1 dBi respectively. When the feed point was at \( f_2 = (2, 1.5) \) the two orthogonal frequencies merge to give a wide bandwidth at 2.175 GHz. The return loss of the antenna is shown in Figure 2. This wide band is obtained with a gain of 8.17 dBi at 2.175 GHz. Gain of the EBG backed antenna is shown in Figure 3.

![Figure 2. Return Loss variation with frequency](image)

![Figure 3. Variation of gain with frequency](image)

At the feed location \( f_2 \), antenna radiates in the band 2.1 to 2.3 GHz with excellent cross-polar level. The radiation patterns of the antenna at 2.175 GHz is shown in Figure 4. HPBW of the antenna is 80° and 84° in the E and H plane respectively. The antenna offers a cross-polar level better than -27 dB in both the principal planes.

**CONCLUSION:**

An EBG ground plane with unequal orthogonal periods has been proposed and studied for RMSA. Effect of feed location on antenna reflection characteristics is studied. Depending on the feed position two orthogonal frequencies can be excited with gain enhancement or a single frequency with wide band and improved gain can be obtained.
ACKNOWLEDGEMENT

Sreedevi K Meen is grateful to Council of Scientific and Industrial Research, Government of India, New Delhi for the Senior Research Fellowship.

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RESEARCH SESSION 2
2.1 Millimeter wave propagation through layers of sand and dust particles
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2.2 Electromagnetic radiation from the moon's surface originating from interactions of extragalactic neutrinos with lunar regolith
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2.4 Microwave Imaging of Dielectric Wax Cylinders
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MILLIMETER WAVE PROPAGATION THROUGH LAYERS OF SAND AND DUST PARTICLES

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Abstract: In the present paper theoretical investigation has been carried out to evaluate the effects of the layers of sand, dust particles on the propagation of millimeter wave. For this purpose the layers of sand, silt and clay particles have been considered. The expression of effective propagation constant has been utilized to get the value of normalized phase velocity, loss tangent and related attenuation for low frequency and high frequency approximation. It has been found that the normalized phase velocity and attenuation all increase with increasing frequency for different values of fractional volume. Further the loss tangent first increases with increasing fractional volume and then starts to decrease with increasing fractional volume.

INTRODUCTION

Recently considerable interest has been devoted to estimate the influence of the layers of dust particles on the propagation characteristics of millimeter wave [1-3]. It may be emphasized that while the millimeter waves are allowed to pass through the medium having such layers the phase and amplitude of wave are affected by the dust particles, cause the attenuation of the wave. Further the normalized phase velocity is also affected due to presence of such layers in medium. Therefore in present paper an attempt has been confined to evaluate the effects of such layers on normalized phase velocity and loss tangent of millimeter wave. The details of entire investigations are given in following sections of the paper.

THEORETICAL CONSIDERATION

In order to obtain normalized phase velocity \( v_p \), effective loss tangent \( \tan \theta \), and attenuation constant \( a \) of millimeter wave from the layers of sand, silt and clay, the profile structure of these constituents in atmosphere must be taken into account, which is shown in fig.1. Two cases have been taken separately.

CASE 1 LOW FREQUENCY APPROXIMATION

Now, the effective propagation constant for low frequency approximation may be given as
\[ k = k' \left( \frac{2k'y}{1 - k'y} \right) \]

where

- \( k' \) is the propagation constant,
- \( \gamma \) is the ratio of spherical particle size,
- \( n_s \) is the number of sand dust particles in the unit volume of atmosphere.

Now we have,

\[ n = \frac{6k'\gamma(1 - \gamma)}{2(1 - k'y)(1 - 2\gamma)} \]

Separating real and imaginary parts of the propagation constant, one has

\[ a = \frac{1 - \left( \frac{1}{2} \right)^2 \gamma}{1 - \gamma} \]
\[ b = \frac{1 - \left( \frac{1}{2} \right)^2 \gamma^2}{1 - \gamma} \]

Now the normalized phase velocity \( V_p_h \) is given by

\[ \frac{v}{k'} = \frac{1}{b} \]

Similarly the effective loss tangent may be obtained as

\[ \tan \delta_p = 2k'/k' \]
Combining equations (5), (4), (3) and (6) one has

\[
\frac{V_{xy}}{C_{y}} = \frac{1}{\left(1 + \frac{1}{2} \left(1 + \frac{1}{2}\right)\right)}
\]

and

\[
\frac{1}{a} \ln \frac{\beta}{a} = \frac{2 \pi \lambda \omega_a^2 (1 + 1)^3}{(1 + \frac{1}{2} \lambda)}
\]

**CASE II HIGH FREQUENCY APPROXIMATION**

In this case the expression for normalized phase velocity and effective mass suggest can be obtained from the relation given as

\[
K_v = k \cdot \left(\frac{\omega_a}{\omega_o}\right)^2 \left(1 - \omega_o^2 \right) + \left(\frac{\omega_o}{\omega_o}\right)^2 + \left(\frac{\omega_o}{\omega_o}\right)^2 \cdot \cdot \cdot (9)
\]

Where \( k, M \) and \( N \) are the coefficients give the amplitudes of the field distribution on spherical dust particles.

Thus we have

\[
\begin{align*}
T^M_1 & = \frac{\omega_o}{\omega_o} \left(1 - \omega_o^2 \right) \cdot \cdot \cdot (11) \\
\epsilon & = \frac{\omega_o}{\omega_o} \left(1 - \omega_o^2 \right) \cdot \cdot \cdot (12) \\
\frac{3}{5} \epsilon & = \frac{\omega_o}{\omega_o} \left(1 - \omega_o^2 \right) \cdot \cdot \cdot (13)
\end{align*}
\]

Where

\[
\rho = \frac{2 \pi a^3}{\lambda}
\]

Now for small value of \( \rho \) and finite \( \epsilon \) only the first term of equation (11) is important
\[ T_1^{(N)} = \left[ \frac{k}{\delta_2} \right] \frac{\left( \frac{\delta_2}{\delta_2 + \delta_3} \right)^2}{\left( \frac{\delta_2}{\delta_2 + \delta_3} + 1 \right)^2} \delta^3 \]  

Putting the values of \( T_1^M \) and \( T_1^S \) in eq(9) one has

\[ k_x = k \cdot \frac{2 \pi n c_0}{\nu} \left\{ \frac{\delta_1}{\delta_2 + \delta_3} \right\} \left( \frac{\delta_2}{\delta_2 + \delta_3} + 1 \right)^2 \left( \frac{\delta_2}{\delta_2 + \delta_3} \right)^4 \]

Putting the value of \( c_0 \) in equation (15) one has

\[ k_x = k \cdot \frac{2 \pi n c_0}{\nu} \left\{ \frac{\delta_1}{\delta_2 + \delta_3} \right\} \left( \frac{\delta_2}{\delta_2 + \delta_3} + 1 \right)^2 \left( \frac{\delta_2}{\delta_2 + \delta_3} \right)^4 \]

Therefore for high approximation the phase velocity and loss tangent can be given as

\[ V_{ph} = \frac{k}{\nu} \left( \frac{\delta_1}{\delta_2 + \delta_3} + 1 \right)^2 \left( \frac{\delta_2}{\delta_2 + \delta_3} \right)^4 \]

and

\[ \tan \delta_{\nu} = \frac{k}{\nu} \left( \frac{\delta_1}{\delta_2 + \delta_3} + 1 \right)^2 \left( \frac{\delta_2}{\delta_2 + \delta_3} \right)^4 \]

Now the attenuation constant \( \alpha \) for medium can be given as

\[ \alpha = \frac{\pi}{k} \tan \delta_{\nu} \]

is dB/Km. The attenuation constant \( \alpha \) can be written as

\[ \alpha = 8.68 \times \frac{\pi}{k} \tan \delta_{\nu} \]

**NUMERICAL COMPUTATION:**

In order to obtain the value of normalized phase velocity and effective loss tangent for low frequency and high frequency approximation, the computational work has been done using equations (15) and (16) respectively. Further the value of attenuation constant \( \alpha \) has been calculated using equation 18. The results thus obtained are shown plotted in form
DISCUSSION OF RESULTS

The examination of Fig. 2 indicates that normalized phase velocity first decreases with increasing frequency and then oscillates as frequency increases. This oscillation corroborates the characteristics of resonant scattering at high frequency. Further from Fig 3 it is found that normalized phase velocity decreases with fractional volume for both low approximation and high approximation of frequency. This is according to the fact that increasing fractional volume increases the particle density in atmosphere. It is also observed from fig. 4 that loss tangent first increases rapidly with increasing frequency for different value of fractional volume and then saturates at high frequencies. This is due to the fact that increasing frequency enhances the real and imaginary parts of complex permittivity of dust particles, also at low frequency the absorption dominates over scattering. Further from fig 5 it is found that the loss tangent first increases with increasing value of fractional volume then saturates and start to decrease with increasing value of fractional volume. This is according to the fact that at higher value of the fractional volume the scattering by dust particles dominates over the absorption of incident wave. The variation of attenuation constant with frequency has been shown in fig 6 where it is found that attenuation constant increases with increasing frequency.

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ELECTROMAGNETIC RADIATIONS FROM THE MOON'S SURFACE ORIGINATED FROM INTERACTIONS OF EXTRA-GALACTIC NEUTRINOS WITH THE LUNAR REGOLITH.

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INTRODUCTION

Electromagnetic radiations (optical Cherenkov radiation and radioemission) associated with Extreme Air Showers (EAS) initiated as a result of interaction of particles from galactic as well as extra-galactic Cosmic Rays (CR) sources with the earth's atmosphere was first detected by Jelley et al. in 1963 (1). The theoretical as well as experimental aspects of the whole spectrum of this radiation from ~300 kHz to ~50 MHz have been studied extensively by different research groups all over the globe (2) and this, in turn, has opened a new branch “Ultra high energy Neutrino Astronomy”. Detection of UHE neutrinos is expected to play an important role in understanding a number of cosmologically significant phenomena. Active Galactic Nuclei (AGN) may be copious producers of such neutrinos. Topological defects or cosmic strings will also lead to UHE neutrino production.

G. A. Askaryan, a remarkable theorist in the field of Cosmic Rays, predicted a negative charge imbalance in the EAS (3) which gives rise to coherent Cherenkov Radiation (CRR) at HF - VHF band of radiofrequency. For LF - MF band, the most probable production mechanism is the coherent Transition Radiation (TR14). TR is produced when relativistic charged particles cross an interface of two media of different dielectric constant. An UHE neutrino entering the lunar regolith can initiate an electron-photon cascade and hence emit CRR. Detail theoretical (5) as well as experimental investigations (6) are going on for the CRR emitted from the moon's regolith.

The first attempt to detect the CRR pulses from the moon with the help of earth based detectors was made by T. Hanks et. al. in 1996. They were using the Parkes 64 metre radio telescope with a wideband (1180-1662 MHz) receiver and high time resolution (of few ms) oscilloscope, but till now no genuine pulse is detected. Another attempt was made by Gehrmann et. al in 1999 using the NASA Goldstone 70m and 34m antennas. The observations were made at S-band (2.2 - 2.3 GHz). An L-band (1.6 GHz) was used at 70m antenna simultaneously with S-band receiver. No pulses of expected duration are observed till date.

The negative results of these two experiments are followed by these two excellent laboratory experiments by UCLA/NASA group of Physics which confirms the negative charge excess mechanism as the origin of CRR pulses. At present, Kalpatri 6m radio telescope is one of the most suitable tool for searching the CR pulses from the moon surface.

The charged particles of the electron-photon cascade, initiated in the lunar regolith while crossing the moon vacuum interface, can emit TR. The possibility of detection of radio emission from the lunar regolith with the help of apparatus at the moon's surface was 81
In this paper, a method is outlined for detection of TR pulses, with the help of moon based detectors, from lunar vacuum interface excited by UHE neutrinos (from GR sources) interacting with the lunar regolith.

Method: When a charged particle moving uniformly in a medium enters another medium, radiation is emitted in the forward as well as backward direction. This radiation is called annihilation radiation. When charged particles of electron-photon cascade initiated by cosmic rays cross the lunar vacuum interface, the phenomenon of TR must occur. For the 1-F-1-MF cascade, all the charged particles of the shower may be assumed, for mathematical convenience, to be concentrated in a point instead of distribution over a region and only the excess negative charge, in effect, will contribute to the TR. For a particle of charge $e$ moving with relativistic velocity $v$ along the $z$ axis and crossing the interface at $z = 0, t = 0$, the radiation field in the vacuum is given by (4).

$$\vec{E}_{\text{vac}} = \frac{eN_{\nu}}{2\pi^2} \frac{\gamma \kappa}{\gamma^2 - 1} \cos 0$$

(1)

For a neutrino induced shower in the lunar regolith, the magnitude of the horizontal component of the field is

$$E_{\text{reg}} = \frac{eN_{\nu}}{2\pi^2} \frac{\gamma \kappa}{\gamma^2 - 1} \cos 0$$

(2)

And the magnitude of the vertical component of the field is

$$E_{\text{reg}} = \frac{eN_{\nu}}{2\pi^2} \frac{\gamma \kappa}{\gamma^2 - 1} \cos 0$$

(3)

Where $N_{\nu}$ is the number of neutrinos in the electron-photon cascade at the interface.

- $eN_{\nu}$ = Excess negative charge
- $\kappa = \alpha / c = 2m_v/c$ = wave number,
- $\kappa'_i = \frac{e^2}{\varepsilon_0} \kappa \chi_i \kappa'_i$ = $\chi_i = \varepsilon_i / \varepsilon_0$

$\varepsilon_0$, $\varepsilon_i$ are dielectric constant of the medium of the lunar regolith and vacuum respectively.

$\mu_0$, $\mu_i$ are permeability of the medium of the lunar regolith and vacuum.

$$\zeta = 1 - \mu_0 / \mu_i$$

$$\tan 0 = Z/R$$

$Z$ = height of the antennas above the moon's surface

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EXPERIMENTAL METHOD

Inspired by the existing methods (8) of registering RCR with earth-based radio telescope, a method proposed for registering TR pulses by erecting antenna at the lunar surface is outlined below:

i) Radio antennas (a minimum of four) of frequency (<1 MHz) are to be erected at the lunar surface to register the TR pulses. Output of radio channels is to be applied to a coincidence circuit and the output of the coincidence circuit is to be taken to trigger the recording system employed for determining the pulse height of individual radio channels.

ii) Time delay between the two pulses is to be determined and comparing this with theoretical value estimated on the basis of atmospheric dispersion, gain inness of the registered pulses can be established.

iii) \( x_{0}, y_{0} \) and core co-ordinates \( x_{N}, y_{N} \), are to be determined by the method of steepest descent as well as by the Newton's method, employing equations (1) & (2).

RESULTS

Preliminary investigations on frequency spectrum estimated with the help of equation (2) is given in Fig. 1.
DISCUSSION

Registration of radio pulses at the earth’s surface suffers from atmospheric noise as well as noises from radio, TV transmission etc. Moreover, there are noise pulses from radioemission associated with EAS initiated by CR particles. However, the moon’s surface is free from all these noises, even the EAS noise pulses, since the moon does not have any atmosphere for initiation of EAS. It is seen from equations (2) & (3), that the field strength scales linearly with the negative charge excess and hence, with the neutrino energy. If a number of antennas are installed on the moon’s surface and field strengths of each of these antenna systems are measured taking coincidences, then from equations (2) & (3), $c, N$ and core co-ordinates can be calculated for each neutrino event. Knowing $c N$, depth of first interaction can be estimated from a fig similar to Fig. 3.

CONCLUSION

Detection of TR pulses at the moon’s surface and at the earth’s surface will give information on characteristics of high energy neutrinos coming from CR source outside our galaxy as well as behaviour of these high energy neutrinos in the moon’s regolith. Close collaboration among radio astronomers from different countries would help to progress further in the field of UHE neutrino astronomy.

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MICROWAVE SURFACE SCATTERING MODEL FOR MOISTURE ESTIMATION IN REMOTE SENSING

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The scattering coefficient is an important parameter in determining soil moisture content. The knowledge of the variability of the surface soil moisture is important for agriculture and remote sensing. In the present paper an attempt has been made to develop a theoretical model for estimation of scattering coefficient using measured dielectric data (S.K. Srivastava et al 2004) both for vertical and horizontal polarization. The results obtained in this model is compared with earlier developed model REM. SPOM’s PC model.

INTRODUCTION

Soil moisture is important to the hydrology research for partitioning rainfall into runoff and infiltration components as well as separating incoming solar radiation into latent and sensible heat. Recently, soil moisture data have a great potential for providing Areal estimates of soil moisture [1]. Although remote sensing of soil moisture can be accomplished to some degree or other by all regions of the electromagnetic spectrum, only the microwave region offers truly quantitative measurements (Dingman and Garney 1995) because the primary physical property that affects the measurements is directly dependent on the amount of water present in the soil. Passive microwave remote sensing employs measurement of the thermal emission from the soil to determine the moisture content in the surface layer of the soil. It relies on the fact that the emissivity (ε), scattering coefficient (σ) at microwave, wavelength is a function of the dielectric constant of the soil water mixture and thus the soil moisture. The emissivity and scattering coefficient of soil at microwave, wavelength is affected by factors such as soil texture, surface roughness and vegetation cover. The emissivity and scattering coefficients of the soil also vary with different moisture contents. Therefore the knowledge of the warming and scattering coefficient of the soil is useful for the efficient use of soil, as this would be very helpful for building microwave instruments for applicable in agriculture [2].

In the present paper an attempt has been made to develop a theoretical model for estimation of emissivity and scattering coefficient at different moisture contents and frequency using measured dielectric data (S.K. Srivastava et al 2004) [3] both for vertical and horizontal polarization.

PPE OF SURFACES

A given surface may appear very rough to an optical wave but the same may appear very smooth to microwaves. The roughness of a random surface with one dimensional surface height profile, z(x) can be characterized in terms of the standard deviation of the surface height, surface correlation length 1 and the rms slope. The
scattering coefficient $e^{(0)}$ of a random surface can be expressed as a product of two junctions as follows

$$e^{(0)} = 1 - \left( 1 - e^{(0)} \right) \frac{r}{(1 + r)}$$

(1)

If $\sigma$ define the standard deviation and $t$ be correlation length of the surface. Then surface is described in terms of its rms slope $m$ as

$$e^{(0)} = \frac{\sigma}{(1 + t)}$$

(2)

where $\sigma$ is the standard deviation of the surface auto correlation junction evaluated at $x = 0$ for Gaussian correlation function $m = 1.414 \sigma / t$.

For estimating the scattering coefficient of soil surfaces one can use many approaches. Here the following approaches have been considered: Physical optics model, Geometrical optics model, Perturbation model (Paff and At (1982), Integral equation model (Fung et al. (92), (94)). The combination of the surface parameters with frequency range of the sensors allowed to cover a large part of validity range of the available electromagnetic models are given table 1.

<table>
<thead>
<tr>
<th>S.N.</th>
<th>Type of Model</th>
<th>Nature of Surfaces</th>
<th>Frequency Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Kirchhoff's Physical optics model (KO)</td>
<td>Medium rough surfaces</td>
<td>High frequencies regions</td>
</tr>
<tr>
<td>2.</td>
<td>Kirchhoff's Geometrical optics model (GO)</td>
<td>Undulating Surfaces</td>
<td>High frequencies regions</td>
</tr>
<tr>
<td>3.</td>
<td>Small Perturbation model (SPM)</td>
<td>Smooth Surfaces</td>
<td>Low frequencies</td>
</tr>
<tr>
<td>4.</td>
<td>Integral equation model (IEM)</td>
<td>Smooth surface</td>
<td>Low to medium frequencies</td>
</tr>
</tbody>
</table>

SCATTERING MODEL FOR SOIL SURFACES

The scattering from randomly rough dielectric surfaces can be estimated by theoretical models derived from the application of the theory of diffraction to surfaces. The models developed in this paper are based on a well known approximation. Small perturbation (SP), Kirchhoff Physic optics (KO), Kirchhoff Geometric (GO) and IEM (Integral equation model). The scattering coefficient can be expressed as
$$\sigma_m(0) = k^2(4 \pi \rho) \left| F_1(\alpha, \beta) \right|^2 \sigma \left[ (\cos \theta \cdot \cos \phi) - \left( \frac{1}{2} \right)^{\frac{3}{2}} \sigma \right]^{\frac{1}{2}} \frac{\sigma}{\sigma_m}$$

where $\alpha, \beta$ = vertical (Y) or horizontal polarization (X).

$$F_1(\alpha, \beta) = \sum_n \left( \frac{\sin \theta_n}{\sin \theta} \cdot \cos \phi \right) \int_0^{\pi} \left( 1 - \frac{\sin \theta}{\sin \theta_n} \right) \int_0^{2\pi} \left( 1 - \frac{\sin \phi}{\sin \phi_n} \right) \sigma \left( \cos \theta_n \cdot \cos \phi_n \right) \left( \frac{\sin \theta}{\sin \theta_n} \right)^{\frac{3}{2}} \frac{\sigma}{\sigma_m}$$

$$= \text{Fresnel reflection coefficient}, \ l = \text{surface roughness length}, \ a = \text{standard deviation}$$

$$R_0 = \frac{2 R_1}{\cos \theta} \cdot f_\alpha = -2 R_1 \cdot \cos \theta$$

$$R_m = \left( \sigma \cdot \cos \theta \right)^{\frac{3}{2}} / \left( \sigma_m \cdot \cos \theta \right)^{\frac{3}{2}}$$

RESULTS AND DISCUSSION

The computation of scattering coefficient of soils has been done with the help of basic computer programs and moisture contents both for vertical and horizontal polarization are shown in fig. 1-2. The variation of scattering coefficient with frequency both for vertical and horizontal polarization are shown in fig. 1. It is found that scattering coefficient increases in frequency and shown second order polynomial trend. The results of computation obtained in this model are compared with earlier developed IEM, SPM & PO model. It follows that the value of scattering coefficient found in this model are closer to other developed IEM, SPM & PO model. This verifies the accuracy of the theory developed. The value of scattering coefficient for vertical polarization is so that for horizontal polarization.

The variations of scattering coefficients with moisture contents have been shown in fig. 2. It shows that value of scattering-coefficient increases with moisture contents both for vertical and horizontal polarization. The values obtained in this model are greater IEM, SPM & PO model. As moisture contents increases the number of free water molecules is wet water
mixtures increases. The tree water molecules have higher dielectric constant compared to bound water molecules. Since scattering coefficient depend upon dielectric constant which will increases with dielectric constant.

REFERENCES

Fig. 1: Variation of scattering coefficients with frequency in X-band

Fig. 2: Variation of scattering coefficients with moisture content in X-band

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MICROWAVE IMAGING OF DIELECTRIC WAX CYLINDERS

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Abstract

Microwave tomographic imaging of dielectric wax cylinders having inclusions of permittivity variations is presented. Coplanar strip line fed bowtie antennas are used for both transmission and reception of microwave energy. The 2-D images reconstructed using the measured scattered fields show the dielectric contrast of the sample.

INTRODUCTION

Medical imaging plays a vital role in the field of clinical diagnosis of diseases whose symptoms are not externally visible. Various types of imaging techniques give maps of the different properties of the imaged object. The microwave tomographic imaging is a relatively new field of imaging which has put certain advantages over the conventional systems like X-ray imaging, Ultrasound imaging and MRI. Anomalies of the soft tissues cannot be accurately detected by any of these conventional systems. Even though microwave belongs to electromagnetic spectrum, it does not produce radiation hazards during imaging, as the energy required is low [1]. All tomographic systems for internal body imaging are based on the differentiation of tissue properties. In X-ray tomography, a tissue is differentiated based on density. In majority of the cases, tissue density does not depend on the tissue physiological state. In the microwave region, a tissue is differentiated based on its permittivity. Tissue properties are analyzed by means of complex electric field vector \( \vec{E} \) \( = E_r \hat{E} + i \times E_i \hat{E} \). Tissue dielectric properties in the microwave region depend upon molecular constituents, ion concentration, mobility concentration of free water and bound water and tissue temperature. In microwave tomography, there are extremely difficult problems connected with image reconstruction. The linear optical approximation, which is used in X-ray tomographic image reconstruction, cannot be used with microwave tomography [2]. The Maxwell's equations on their scalar approximations should be used for microwave tomographic image reconstruction.

The goal of microwave tomographic imaging is to exploit the contrast in dielectric properties profile of an object from the measurement of microwave energy scattered by the object [3]. The shape of the object and spatial distribution of the complex permittivity are obtained from the transmitted (transmit) and scattered (received) fields. The challenge of microwave tomography is that it involves the solution of an ill-conditioned nonlinear inverse scattering problem. Further, these problems require image reconstruction algorithms, which are
scattering problems. Further, these problems require image reconstruction algorithms, which are computationally intensive.

This paper presents microwave tomographic imaging of dielectric wax cylinders having inclinations of high dielectric permittivity. Coplanar strip line feed bowtie antennas are used for both transmission and reception of microwave energy. The 2-D images are reconstructed from the collected microwave data using distributed Born iterative conjugate gradient fast Fourier transform method (DIBICCGFFT). The results show the permittivity variations, which suggest the feasibility of using bowtie antennas for microwave medical imaging.

METHODOLOGY

A compact coplanar strip line feed bowtie antenna having unidirectional radiation patterns [4] is used for both illumination and reception of microwave signals. Dielectric wax cylinders of 
\( \varepsilon \text{ (real)} = 5 \pm 0.3 \) of various dimensions are illuminated by the antenna at 3 GHz. Notes are taken in samples and are filled with water of 
\( \varepsilon \text{ (real)} = 75.6 \) to provide dielectric contrast in the sample. The coupling medium selected here is air. For every 3° rotation of the sample, the receiving bowtie antenna takes the measurement in steps of 1°. All measurements are done using HP 8714ET network analyzer.

IMAGING TECHNIQUE

The most difficult part of microwave imaging is the reconstruction of the images in terms of dielectric parameters. The substantial multiple scattering effects inside the object are to be dealt with non-linear inverse scattering theories [2]. The scattering due to the inhomogeneity can be formulated in terms of the non-linear Fredholm integral equation

\[ E(\mathbf{r}) = \int \int \int d\mathbf{r}' \mathbf{g}(\mathbf{r}, \mathbf{r}') \partial \varepsilon(\mathbf{r}') E(\mathbf{r}') \]  

where \( \mathbf{g}(\mathbf{r}, \mathbf{r}') \) is the incident field in the background medium in the absence of the scatterer. The total field inside the VAT when the microwave signal incident, is denoted as \( E(\mathbf{r}) \) and \( \partial \varepsilon(\mathbf{r}') \) is the dielectric contrast. It stands for the dielectric contrast of the sample, and \( g(\mathbf{r}, \mathbf{r}') \) represents the green's function.

From equation (1), the scattered field can be written as

\[ E_{\text{scatt}}(\mathbf{r}) = \int d\mathbf{r}' \mathbf{g}(\mathbf{r}, \mathbf{r}') \partial \varepsilon(\mathbf{r}') E(\mathbf{r}') \]  

For an implied time dependence of \( \mathbf{g}(\mathbf{r}, \mathbf{r}') \), the green's function for a homogeneous background medium can be computed as

\[ \mathbf{g}(\mathbf{r}, \mathbf{r}') = \frac{j}{4} (\mathbf{H}(\mathbf{r}) \times \nabla) \]  

Equation (1) is the central equation used for both forward and inverse scattering solutions. The scattered field is a linear functional of \( \partial \varepsilon(\mathbf{r}) \) because \( E(\mathbf{r}) \) is in the integrand as also a function of \( \partial \varepsilon(\mathbf{r}) \). In the inverse scattering problem \( \partial \varepsilon(\mathbf{r}) \) has to be computed from.
the measurement of the scattered field outside the target. Hence this constitutes a nonlinear inverse scattering problem. This problem could be linearised by using the Born approximation by replacing the total field inside the scatterer as $E_0 k_0 r$. So equation (3) is modified as

$$\mu_0 (r') \mathbf{E}_0 (r') = \oint \mathbf{E}_0 (r') \cdot \frac{\delta k_0 (r') \delta E_0 (r') \cdot \mathbf{E}_0 (r')}{\delta k_0 (r') \delta E_0 (r')} dr'$$

Progressively better estimates of the permittivity can be obtained by successive iterations, till convergence is reached.

But Born approximation is poor when the scattered field is not weak compared with the incident field. Here the background medium is assumed to be homogeneous, so Green's function is known in the closed form. In Distorted Born Iterative Method (DBIM), the background medium is not assumed to be homogeneous; hence the Green’s function is updated in each iteration [5]. The integral equation is discretised using direct method and the resultant ill conditioned matrix is solved using conjugate gradient fast fourier transform (CGFFT). Due to computational complexity, the imaging area is restricted to 16x16 pixels. The sampling rate considered is 0.14 (resolution).

RESULTS AND DISCUSSIONS

2-D images of various dielectric wax samples reconstructed using DBIM-CGFFT from the scattering data collected by the bowtie antenna are shown in Table 1. The dielectric contrast of the different samples is clearly visible from the images.

ACKNOWLEDGEMENTS

Authors G. Bindu and Anil Zotaapun thankfully acknowledge CSIR for providing Senior Research Fellowships.

REFERENCES


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<table>
<thead>
<tr>
<th>Sample</th>
<th>Reconstructed images</th>
<th>Permittivity profile</th>
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<td>Wax cylinder of radius 6 cm, kept cent.</td>
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<td><img src="image2" alt="Permittivity Image" /></td>
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<tr>
<td>Wax cylinder of 2.2 cm kept off-centered</td>
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<td>Wax cylinder of radius 2.2 cm with a hole of 0.5 cm radius at the center. The hole is filled with water.</td>
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<tr>
<td>Wax cylinder of radius 2.2 cm with two holes of 0.1 cm, 0.5 cm each filled 1 cm apart. The holes are filled with water.</td>
<td><img src="image7" alt="Reconstructed Image" /></td>
<td><img src="image8" alt="Permittivity Image" /></td>
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</table>
SUPERSTRATE LOADED METAL-DIELECTRIC STRUCTURE BASED ON SIERPINSKI GASKET FOR BACKSCATTERING REDUCTION

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The scattering behaviour of a metal-dielectric structure based on Sierpinski gasket fractal geometry loaded with superstrate is presented. The proposed structure gives a reduction in backscattered power over a wide range of frequencies in X-band. The same structure when loaded with superstrate is found to give backscattering reduction in C-band. Frequency tuning effect is also achieved by an appreciable metal to the superstrate thickness. The reduction obtained is idealised for both TE and TM polarisations.

INTRODUCTION

The Sierpinski antenna was the first example of a multifractal antenna based on fractal geometry [1, 2]. These antennas keep the same behaviour at several bands in terms of input impedance and radiation patterns. It features as many folds as scale levels in the geometry, and its hyper-period matches the scale factor of a self-similar geometry. Other examples include the classical Sierpinski antenna representing a clear example of multifractal antenna geometry, where a new fractal can be introduced by adding new fractal elements. In addition by modifying the Sierpinski geometry fractal angle, the antenna pattern at different resonant can be tailored [3]. Further investigation on geometries implored by fractal structures shows that the classical Sierpinski gasket (1) is a special case of a wider set of structures which can be referred to as Pascal Sierpinski gasket [4]. Fractal structures based on Sierpinski gasket geometries are thus finding wide range of applications in the design of electromagnetic scattering structures [5]. The idea of using fractal based metalization in the design of metallic-dielectric structures has been reported recently in which reduction in backscattered power of 90 dB is obtained simultaneously for TE and TM polarisations using Sierpinski carpet fractal geometry [6].

In the present work, scattering behavior of a metal-dielectric structure based on Sierpinski gasket fractal geometry is envisaged. From the experimental results it is found that a reduction in backscattered power over a wide range of frequencies is achieved by loading superstrate of appropriate dielectric thickness over the same structure, the reduction in backscattered power is obtained in a lower frequency band. The reduction in backscattered power is achievable for both TE and TM polarisation.
DESIGN METHODOLOGY AND EXPERIMENTAL SETUP

The third iterative stage of the Siebert's gasket fractal geometry is shown in Fig. 1(a). The metal-dielectric structure used in the present work is designed by combining four third iterative Siebert's gasket triangles arranged to form a structure as shown in Fig. 1(b). The metal-dielectric structure is fabricated by photolithography metallization on a low loss dielectric substrate (εr = 2.58) of size 30 x 30 cm². Fig. 2 shows the geometrical layout of the superstrate loaded metal-dielectric structure.

![Diagram of metal-dielectric structure]

The metal-dielectric structure backed with a plate metallic plate is placed on a turntable in an anechoic chamber. The measurements are performed over C and X band frequencies in an anechoic chamber with the help of an HP 8510C vector network analyser. The structure is illuminated using a wide band horn antenna. The backscattered power over the entire band of frequencies is measured and compared with the back scattered power from a plane metallic plate of same dimensions. This is repeated for various dielectric thickness and the optimum thickness giving the minimum backscattered power is found out. The same measurement are done for structures loaded with superstrates of various thicknesses.

![Diagram of cross-sectional view of superstrate loaded metal-dielectric structure]

Fig. 2. Cross-sectional view of superstrate loaded metal-dielectric structure.
EXPERIMENTAL RESULTS

Fig. 3. Shows the variation of backscattered power with frequencies in X band without superstrate loading. The frequency at which the minimum backscattered power obtained depends on the substrate thickness \( h_s \), and a maximum reduction of 30dB is obtained at 8.9GHz for a dielectric thickness of height \( h_s = 5 \text{ mm} \).

![Graph showing backscattered power vs. frequency](image)

Fig. 3. Variation of relative backscattered power with frequency

Fig. 4. shows the variation in backscattered power with frequency of the structure loaded with superstrate of different thickness. It is found that a maximum reduction in backscattered power of 38 dB at 9.94 GHz is obtained for an optimum superstrate of thickness \( h_s = 4 \text{ mm} \). It is also observed that frequency tuning is possible in C-band by varying the thickness of the superstrate. The frequency tunability of the superstrate loaded metallo-dielectric structure is shown in Fig.5.

![Graph showing backscattered power vs. frequency](image)

Fig. 4. Variation of relative backscattered power with frequency by loading superstrate \( h_s = 5 \text{ mm} \)

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CONCLUSIONS

The structure with metallizations based on Sierpinski gasket fractal geometry is found to reduce backscattering over an appreciable range of frequencies. The response of the structure can be shifted to a lower frequency band by loading superstrate on the metallo-dielectric structure. It is also observed that loading superstrate gives frequency tunability effect over an appreciable range of frequencies. This property is observed for both TE and TM polarizations of the incident field, owing to the symmetrical nature of the geometry. This structure will have implications in the design of frequency selective surfaces and RCS reduction techniques.

REFERENCES

RESEARCH SESSION 3
A Cross-Slot aperture coupled circularly polarized antenna array for Briefcase telephone terminals

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A Study on Compact Reconfigurable Microstrip Antenna with Wide Band

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*University Science Instrumentation center and **Dept. of Aers. Electronics, Guwahati University, Guwahati- 781010, wanyen753@rediffmail.com, **Dept. of Electrical Engineering, Salt Lake community college, UT-AIR 84130, USA.
DEVELOPMENT OF A PHASED ARRAY ANTENNA ON SILICON WITH BST PHASE SHIFTERS

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The design and development of a four element phased array antenna monolithically implemented on silicon is presented in this paper. The array antenna designed for operation around 150 GHz consists of a monolithic 1 x 4 CPW fed network including power dividers, phase shifters, and DC blocks, and a monolithic set of microstrip patch radiators fed by coplanar waveguides (CPWs). The phase shifter based on thin barium strontium titanate (BST) film employs bilateral capacitor simple (BSC) CPW structures, resulting in 30°/dB at 25GHz. Future work to improve the beam steering capability of the array antenna is discussed.

INTRODUCTION

Demand for broadband wireless communication services has been increasing explosively, driving the surge of research and development activities for future wireless communication systems with higher data rates and improved functionality. It is expected that this demand will be fulfilled by realizing 4G mobile systems which could consist of a hybrid combination of different access technologies such as wireless local area networks (LANs), intelligent transport systems (ITSs) and high-altitude atmospheric platform station systems (HAAPS), as well as cellular phones [1]. However, these systems will require the aid of advanced array antenna technologies to operate efficiently, since array antennas have proven to play a key role in improving system performance by increasing channel capacity, steering multiple beams, and compensating for aperture distortion electronically [2].

The array system offers the unique capability of electronic scanning of the main beam. By changing the phase of the exciting currents in each element antenna of the array, the radiation patterns can be scanned through space. By this means, the beam can be very quickly scanned electronically and becomes capable of tracking fast-moving and multiple targets in a fashion which is impossible with a traditional rotating-dish antenna.

This work introduces a four element phase array antenna system with each radiating patch element fed by a barium strontium titanate (BST) phase shifter. The array antenna system was fabricated on a 400μm thick high resistivity silicon (1V-9000Ω/cm) surface-stabilized by polysilicon. With this
The BST phase shifters developed herein have adopted a bilateral coplanar waveguide (CPW) structure for ease of monolithic integration with control circuits as well as fabrication, as shown in Fig. 1 [4–6]. Since the bilateral CPW phase shifter consists of periodic capacitive loads (C_{load}) on a CPW (C_{pinc}) as shown in Fig. 1(b), the periodic loads in a unit cell can be analyzed by the periodic shunt circuit model [7]. To maximize the tunability of the BST film by confining electric fields near the surface region, the bilateral coplanar antenna (BCS) electrodes employed a very narrow width and spacing of 4.4 μm and 2 μm, respectively, resulting in high filling factors and low operating voltages. The critical dimensions of the BST phase shifters with a total length of 10 mm are listed in Table I referring to Fig. 1(b).
A BST thin film of 0.4 μm thickness with 50% dielectric tunability was RF-sputtered on SrO/Poly-SrO/BRS substrates. Fabrication of the BST phase shifter adopting BCS-CPW structures was completed by defining the metal patterns on BST-grown silicon substrate using conventional photolithography and electroplating techniques [4-6]. The BCS-CPW phase shifters were characterized with a HP8580C network analyzer and Cascade microwave station. First, scattering parameters were measured from 5 to 25 GHz at room temperature. As shown in Fig. 2, an insertion (S21) and a return loss (S11) of better than 5 dB and 20 dB, respectively, were obtained from 5 to 25 GHz. This low shunt conducting loss from silicon substrates is attributed to the existence of the polysilicon layer with large trap density, screening the surface charge accumulation [4].

The measured differential phase shifts of the phase shifter are shown in Fig. 3. DC bias voltages were applied up to 500 V between the ground and signal line through bias-T and DC blocks. It is easily seen that the BCS phase shifter exhibits high linearity for overall frequencies and continuous phase shifts of 0° - 150° at 25 GHz, corresponding to 30°/dB figure of merit. Based on this design of the phase shifter, we developed a four element phased array antenna system on silicon, which is designed to operate at 150 GHz.
The 15GHz array antenna system developed herein consists of a monolithic 1:4 CPW feed network including power dividers, BST phase shifters, and DC blocks, and a monolithic set of microstrip patch radiators fed by CPW, as shown in Fig. 4. Inter-element spacing is 11.25mm, corresponding to half free-space wavelength.

Monolithic implementation of the array started with the preparation of silicon substrates. High resistivity silicon (HRS) substrates were thoroughly cleaned through standard clean 1 (SC1) and standard clean 2 (SC2), followed by deposition of a 1μm-thick polysilicon layer and a 400μm-thick silicon oxide layer. BST thin film with a thickness of 0.4μm was then RF-sputtered on the substrates as presented above for construction of BST phase shifters. To minimize RF loss from the BST film, the BST thin film was etched out using hydrogenfluoric (HF) acid, except where the phase shifter circuits were to be built on. Gold metal deposited using an evaporator was then patterned by standard photolithography techniques, and electroplated up to 2μm thickness, in order to construct the four element phased array antenna circuits.

Prior to measuring the radiation patterns of the array antenna fabricated on a silicon substrate, S11 was measured to find the resonance frequency of the antenna. Fig. 5 shows the measured results for the S11 and input impedance drawn on a Smith chart. It is noted that this array antenna has a resonance frequency of 14.83 GHz with an excellent return loss (S11) value of around 32dB, and a 2.1 SWR bandwidth of 8.7%.
To realize beam steering of the array antennas, the wave phase arriving at the patch terminal needs to be controlled. This work is possible by varying the bias voltage applied to the phase shifters that are connected to each patch element, resulting in a change in phase velocity. Thus, the direction of the beam can be changed by controlling the relative phase relationships between the individual antenna elements using the BST phase shifters in our array antennas. The external voltages are applied through the bias shift connected directly to the BST phase shifter (see Fig. 4) using gold bond wires.

Fig. 6 shows the measured radiation patterns when the phase shifter circuits are biased. The voltages applied to each phase shifter were 0V, 80V, 150V, and 300V, which phase shifts of 0°, 30°, 60°, and 90° are expected at 1 GHz, respectively, according to Fig. 1. It is noted from Fig. 5 that a 90° phase tilt of the main beam on each side, corresponding to 70° of total scan, is obtained. Since this degree of steering capability is not yet sufficient for military or civil applications, further research should be concentrated on enhanced the steering capability, which would be accomplished by improving the performance of the
CONCLUSION

Monolithic implementation of a four element phased array antenna on silicon has been presented. It is expected that the monolithic integration on silicon could not only make the whole circuitry compact, but also reduce the cost by utilizing mature CMOS technology when developing large phased arrays.

Several works to improve the performance of phase shifters are underway, resulting in better beam steering capabilities of the array. This includes the modification of the bilateral structure as well as the optimization of the BST film growing condition. Efforts in the design modification will be made to confine high external electric fields more efficiently within the bilateral structure, making the contribution of BST films to dielectric healing much greater.

REFERENCES


A simple microstrip patch antenna is presented. The printed circular patch is etched out of the metal and supported at the center by strategic points by (metallic or nonmetallic) post. An antenna has been developed in 1. bandwidth with the center frequency of 1.81 GHz. The bandwidth of this antenna is observed around 11.11% at VSWR 2.1. The return loss of center frequency is around 30.4 dB. The bandwidth and efficiency of this antenna is higher than the conventional antennas. This technique is inexpensive and can be used to develop the various types of antennas and arrays.

INTRODUCTION

Microstrip antennas are used in a broad range of applications such as radars, telemetry, and navigation primarily due to their simplicity of fabrication, ease of production, low manufacturing cost and light weight. The mass problem of microstrip antennas is their very narrow bandwidth and poor efficiency [1]. A number of papers have appeared in technical literature on bandwidth enhancement of microstrip antennas [2, 3, 4]. The bandwidth of a microstrip antenna is typically 2% for h/L = 0.02. This limitation is mainly imposed by the presence of dielectric. The bandwidth of the microstrip antenna increases with substrate thickness h and lower dielectric constant i.e., ε < 2 because a lower value of ε and a thicker substrate will give rise to an increase (fitting) field and therefore a large value of radiated power P is realized.

The dielectric losses associated with the microstrip patch antenna limits the maximum efficiency that can be achieved. The impact of the dielectric substrate on the printed antenna performance was extensively treated in the literature [4,5]. In [6], the losses associated with the fed network radiation, surface waves and their effect in the overall efficiency of an antenna have been discussed.

In this paper, an microstrip antenna is proposed to improve the bandwidth and efficiency of the antenna. This type of approach gives various advantages over the conventional microstrip patch or other types of planar suspended microstrip patch. This antenna consists only the metallization part of the printed circuit suspended in air over the ground plane. This antenna can provide considerably higher efficiency because there are no dielectric losses and no surface wave losses. Fabrication cost is considerably lower than that of conventional microstrip. It can handle significantly higher power levels especially (CW). The overall advantages of this antenna include, very large VSWR bandwidth, freedom from dispersion and surface wave effects, higher efficiency, higher power handling capability and improve radiation pattern. The feed network in this antenna is purely TEM.
THE BASIC RADIATION ELEMENT

The microstrip patch radiating element suspended in air dielctric over the ground plane is shown in Fig. 3. It is a microstrip patch cut-out of sheet metal and supported in the center by a post (grounded). It is fed by microstrip line perpendicular to the ground plane. The other possible feeds can be parallel to the patch or oblique to ground plane. The main problem here is to control the width of the feeding transmission line. In the conventional printed microstrip antenna, the input impedance of the patch can be very high (or low), involving a very narrow (or wide) feeding line. Here, as shown in Fig. 1, the width of the feeding transmission line is tapered gradually and brought to a comfortable size at the proper height above the ground plane.

DESIGN

This patch antenna has been designed in square shape and is fed by a perpendicularly transmitted line. The patch is supported in the center by a plastic post of diameter 3.5 mm to avoid the effective crosstalk and microcirculation. The wide band microstrip simple-layer patch antenna is shown in Fig. 1. The patch antenna dimension is 71.12 mm 71.12 mm and suspended at 20 mm air height over the ground plane. The feeding strip is 18.5 mm wide and tapered down vertically 3.5 mm at 1.5 mm above the ground plane and is connected to SMA connector vertically. The impact on the input impedance was found negligible till the dihedral of the post does not exceed about 30% of the largest dimension of the patch even if the post were metallic. The size of the ground plane is taken 3.5 x 3.5. The design was also analyzed using simulation package (Microwave Office). The comparison between the measurements and software predictions are very closed. The VSAR of the antenna is shown in Fig. 2.

MEASUREMENT AND DISCUSSION

The radiation pattern of the wide-band patch is shown in Fig. 3. Fig. 3 shows the measured radiation pattern at 1.81 GHz (3.5 x 3.5 ground plane). The impact of the small ground plane is more pronounced. The return loss of the antenna can be seen in Fig. 2. The return loss at the center frequency is 30.4 dB. The bandwidth of this antenna at VSAR 2:1 is about 14.1%. The measured voltage standing wave ratio (VSWR) of the antenna is definitely exceeding the required personal communication system band. The pattern measured at 1.81 GHz exhibits a relatively low front to back ratio due to the size of the ground plane. The bandwidth enhancement of this antenna is due to various factors, i.e., the use of air as "dielectric Substrate" and second the large height of the patch element. Ground is located at 12 wavelengths away from the patch and simply acts as a reflector.

CONCLUSION

A microstrip antenna suspended in air dielectric over the ground is presented in this paper. The suspended dielectric substrate provide large bandwidth and higher efficiency compared to other types of conventional patches while considerable reducing the fabrication cost. This technique, for many applications, eliminates the need for parasitic elements and the dielectric substrates necessary to support them. Using this technique many antenna and arrays can be developed.
ACKNOWLEDGEMENT

Author expressing his gratitude to the Director and Dean of I.A.T., Pune to permit him to publish this work. Author is also thankful to Chairman of Electronic Faculty and Professor V. A. Deshmukh, Sr. Tc for constant encouragement.

REFERENCES


Fig. 1. Microstrip Antenna Suspended in air over Ground Plane
Fig. 2. Return Loss for Patch Antenna

Fig. 3. E & H-Plane Radiation Pattern
L-STRIP FED CIRCULAR MICROSTRIP ANTENNA

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A broad band L- strip fed Circular Microstrip antenna design is presented. The reflection and radiation characteristics of this antenna are studied experimentally.

INTRODUCTION

The major disadvantage of circular as well as rectangular microstrip patch antenna is its inherently narrow bandwidth. Much intensive research has been done in past years to develop bandwidth enhancement techniques. These techniques include use of thick substrates with a low dielectric constant [1], and stacked or co-planar parasitic patches [2]. The use of thick substrates introduces a large inductance due to increased length of the probe. Stacked patch geometry increases the complexity. The addition U-shaped slot and the use of L-probe [3, 4] are some other methods introduced for wide bandwidth. The limitations of these are they increase the volume of the antenna substantially. L-strip feed for bandwidth enhancement of rectangular patches has been reported recently [5]. In this paper we introduce a L-strip fed wideband circular microstrip patch antenna. This antenna has a maximum impedance bandwidth of 17% at 3.5GHz and 12.9% at 2.4GHz and 1.8GHz respectively, with a gain of 7.25dBi.

ANTENNA GEOMETRY

![Diagram of L-strip fed circular microstrip antenna](image)

Figure 1 Geometry of L-strip fed circular microstrip antenna

Circular patch antennas resonating at 3.5 GHz, 2.4GHz and 1.8GHz are fabricated on a substrate having dielectric constant $\varepsilon_r = 4.28$ and thickness $h = 0.16cm$. The antenna is excited by proximity method using L-shaped microstrip feed, fabricated on another substrate having the same dielectric constant and thickness. Dimensions of the circular patches used for study are $r = 1.22\, cm$, $1.78\, cm$ and $1.36\, cm$. The antenna geometry is illustrated in Figure 1.
EXPERIMENTAL RESULTS

The variation of return loss of different circular patch antennas as a function of frequency is studied for different feed length $S_2$ and feed segment length $S_4$. $S_1$ is varied from 0.334 to 1.75, and $S_3$ from 0.13 to 1.2, $S_2$, and $S_4$ are optimized for maximum bandwidth for all patches. A 2.5 VSWR bandwidth of 17% has been obtained for the circular patch resonating at 3.5 GHz. This is >5 times greater than the bandwidths obtained when the same circular patch is fed by simple microstrip line. Maximum bandwidth is obtained when $S_2$ and $S_3$ are 1.074a and 0.747a, respectively for the patches with a feed at $r = 1.22cm$. Circular patch resonating at 2.4 GHz offered a maximum bandwidth of 12.5%, for $S_2 = 0.94a$, and $S_3 = 0.47a$, combination. While the $S_2$ and $S_3$ combination of patch resonating at 1.8 GHz is 0.904a and 0.257a, respectively. The return loss characteristics of the antennas for optimum bandwidth problem are plotted along with the return loss characteristics when matched with single microstrip feed, and are shown in Figure 2. The bandwidth variations of the patches with $S_2$ for different $S_3$ are shown in the Figure 3. Gain plot of the optimized configuration of L-strip fed antenna is measured and is shown in Figure 4. The antenna has gain of 23.35 dBi at the centre frequency.

![Figure 2 Return loss variations of different L-strip fed circular patch antennas](image)

(a) $r = 1.22cm$ (b) $r = 1.30cm$ (c) $r = 2.3cm$
Figure 3 Bandwidth variation of different L-antenna fed circular microstrip antennas for different $S_1$: (a) $r = 1.22\text{cm}$, (b) $r = 1.78\text{cm}$, (c) $r = 2.36\text{mm}$.

Figure 4 Gain plot of the L-antenna fed circular microstrip patch antenna at the optimum bandwidth position.
The radiation patterns of the antenna configuration which offered maximum bandwidth at the resonant frequency is also presented. Figure 5 shows the E-plane and H-plane patterns of the antenna with the optimum bandwidth configuration. The half power beam width is found to be 100° and 74° in E-plane and H-plane respectively. The cross polarization level is better than −30dB in both the principal planes.

CONCLUSIONS

In this paper the characteristics of different wide band L-strip fed circular microstrip antennas are described. Bandwidth of the antenna is increased to 17% when excited with L-strip first. As this antenna has wide bandwidth, compact structure, it may find applications in wideband communication systems.

REFERENCES:
TIME DOMAIN ANALYSIS OF OCTAGONAL MICROSTRIP PATCH ANTENNA BY CONFORMAL FDTD METHOD

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ABSTRACT

Conformal finite difference time domain method is used to analyze the reflection and radiation characteristics of a dual port dual band octagonal Microstrip Patch Antenna. The theoretical results are compared against the experimental observations and IESD simulation results.

INTRODUCTION

Wireless communication is an area of great interest among researchers in view of its direct application in industry. Dual band antennas that can operate with good reflection and radiation characteristics in GSM1800 and Bluetooth (2.4 GHz) bands is a basic requirement in multifunction mobile devices. An Octagonal Microstrip Patch Antenna (OMPA) for mobile and Bluetooth applications has been proposed by Binu Paul et al. [1]. In the dual port Octagonal patch antenna, dual frequency operation in the required bands is achieved by tuning its dimensions. Good 2:1 VSWR characteristics, isolation between ports and reduced area are the striking features of the above antenna. In this paper conformal FDTD method [2] with Perfect Magnetic Conductor (PMC) applied along the plane of symmetry [3] is used to study the characteristics of the OMPA. The theoretical results are compared against experimental values. IESD simulation results and non-PMC conformal FDTD.

Fig. 1 Octagonal patch antenna configuration

Figure 1 shows the geometry of the octagonal patch antenna fabricated on substrate 1 (or H1), excited by electromagnetic coupling using two 50 Ω orthogonal microstrip feed lines fabricated on substrate 2 (or H2). The microstrip feed lines of lengths L1 and L2 are an offset
The PMC wall applied along the symmetry plane, as shown in Figure 2.a, helps in reducing the computational domain to half. The FDTD parameters shown in Table 1 were used in the Matlab® based code. For analyzing the antenna reflection characteristics, a Gaussian pulse of appropriate width and time delay is impressed into the computational domain at the feed point corresponding to the port under study. Figure 2.b illustrates the conformal FDTD approach employed in the theoretical investigations. The undamaged cell sites are chosen so that there are more than 20 cells per wavelength in all directions. The time step is chosen as 75% of Courant stability criterion. It is assumed that when \( \Delta x ( or \Delta y ) < 0.125 \gamma \Delta t \) (or 2\( \Delta x \)), \( \Delta x, \Delta y \geq 0.5 \) for a 50 \( \Omega \) resistance attributed to the source and detector probe pins ensures fast convergence.

The input impedance of the antenna is computed as ratio of the FFT of voltage derived from field values at the feed point, over the entire time steps, to the FFT of current at the same point, derived from the field values. Reflection Coefficient \( S_11 \) (in dB) is then computed. The isolation characteristics is calculated from the voltage induced on a 50 \( \Omega \) resistance at feed point 5, derived from the field values, and the voltage impressed at feed point 1.

A sinusoidal excitation is used to address the radiation problem. The excitation frequencies are chosen for Port 1 and 2 based on the reflection characteristics of the antenna. Iteration is carried out until the system achieves sinusoidal steady state. An aperture is defined in a layer above the patch surface, as shown in Figure 3.c. At each spatial point in the plane, the time independent flux harmonic coefficient is computed by sampling the tangential field components over N time steps corresponding to one period of the excitation. From the aperture
RESULTS

Figure 3a shows the theoretical, simulated and measured return loss characteristics of the antenna with the patch and feed lines fabricated on similar substrates (ε1=ε2=4.4, h=1.625 mm). The patch dimensions suitable for dual band operations in the GSM1800 and Bluetooth frequencies are a=24 mm, b=14.14 mm, w=10 mm and coplanar feed lengths and offset are P1=112 mm, P2=12 mm, d=5 mm. The close agreement observed between theoretical characteristics of the antenna computed with the PMC wall and in the full domain, and experimental results validate the FDTD approach used in the study. Figure 3b illustrates the isolation characteristics of the antenna. Experimentally observed gain of the antenna with respect to standard circular patches in the two operating bands is shown in Figure 3c.

The antenna offers linear orthogonal polarisation with broad radiation patterns in the E and H planes as shown in Figure 3d. In all cases the theoretical and experimental values show good agreement. Table 2 shows a comparison of the results obtained in the study of the Octagonal MPA for a feed length of 25 mm.
TABLE II: Comparison of results

<table>
<thead>
<tr>
<th>Port</th>
<th>f (GHz)</th>
<th>VSWR (Max)</th>
<th>VSWR (Min)</th>
<th>% Difference w/ w/o</th>
<th>Perp.</th>
<th>f (GHz)</th>
<th>VSWR (Max)</th>
<th>VSWR (Min)</th>
<th>% Difference w/ w/o</th>
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<tr>
<td>Port 1</td>
<td>1.87</td>
<td>3.04</td>
<td>0.99</td>
<td>7.0</td>
<td>1.84</td>
<td>3.27</td>
<td>0.94</td>
<td>1.84</td>
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<tr>
<td>Port 2</td>
<td>1.60</td>
<td>1.84</td>
<td>1.28</td>
<td>4.3</td>
<td>1.62</td>
<td>1.84</td>
<td>1.28</td>
<td>1.62</td>
<td>1.84</td>
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</table>

Figure 4 shows the variation in resonant frequency and % BW of the antenna as feed length is varied experimentally. It is observed that the F1-F2/F3-F4 antenna gives good reflection characteristics assuming the compactness of the antenna.

The field distributions in the patch at the two resonant frequencies, computed using the FDTD method is shown in Figure 5. It is observed that the 1.8 GHz band corresponds to TM11 mode and 2.4 GHz band corresponds to TM01 mode of the geometry.

**CONCLUSIONS**

The dual-band analysis of compact dual - polarized, dual band, electromagnetically coupled octagonal patch antenna using the conformal FDTD is presented. The antennas exhibit good isolation characteristics suitable for mobile and Bluetooth applications. The FMC approach used in the FDTD study helps in reducing the computational domain size and time. The antenna is suitable for mobile wireless gadgets.
REFERENCES


DITA/1. RAND SEMI-CIRCULAR I.- SLOTTED MICROSTRIP PATH ANTENNA.

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ABSTRACT
A novel dual frequency, rugged and compact coaxially fed microstrip patch antenna is proposed and the simulated results are presented in this paper. It has a semi-circular patch with an L shaped slot on an inverted dielectric substrate. The dielectric above the patch, in itself, works as a radome besides improving the ruggedness of the antenna. A parametric study of this configuration is carried out by varying the length and width of the slot and the probe diameter. The dual frequency ratio from about 1.54 to 1.70 is achieved by varying the slot parameters.

INTRODUCTION
Large bandwidth and good radiation characteristics are obtained using a probe-fed microstrip antenna (MSA) with an U-shaped slot etched on it [1-4]. The broad bandwidth arises because the substrate is relatively thick. The U slot introduces a capacitance in parallel with the probe. This has the effect of creating another resonance near the resonances of the patch. When these two resonances are nearby, the broadband response is achieved, whereas, dual frequency operation is obtained by separating these two resonances. Recently, a wide bandwidth has been obtained with a configuration with semi-circular patch with half U slot etched on it [5]. This configuration is more compact when compared to a circular MSA (CMSA) with a U slot for some frequency range as it has the area as compared to CMSA.

In this paper, dual frequency operation of this compact antenna is presented. The coaxially fed semi-circular patch, with an L shaped slot, is etched on an inverted dielectric substrate. The inverted dielectric forms the radome for the antenna. In addition to improving its structural robustness, the height of the wave permittivity substrate k=0.13 at 2 GHz. A parametric study based on Method of Moments based software [6] is presented to observe the change in the antenna’s behaviour as a result of the alterations of parameter such as slot length, slot width, probe position and probe radius.

ANTENNA GEOMETRY
The resonance frequencies of a semicircular patch with an L shaped slot (one end short, and other end open circuited) are same as those of a CMSA and a U slot (both end short circuited) respectively. A dual frequency operation is achieved by selecting the radius of the semi-circular patch and length of the slot cut in it. The optimum result is obtained when the
resonance frequency of the slot is higher than that of the semi-circular patch. The geometry of the proposed antenna is shown in Figure 1. The semi-circular patch with an L-shaped slot is printed on a dielectric substrate of thickness 1.6 mm and dielectric constant 2.2. For the present dual band configuration, the semi-circular patch is designed in L-band and length of the slot (quarter wavelength) is chosen for higher resonance frequency as compared to that of a patch [7]. The dual frequency is tuned by changing the dimension of the slot.

**PARAMETRIC STUDY**

All the parameters of the antenna are labelled in Figure 1. To get an insight of the working of this compact dual band MPA, configuration, one of the parameters, x, and the radius of the probe, R, varied as a time, keeping other remaining parameters fixed, and performance of two dual band frequencies are noticed. Dimensions of this compact antenna used in the parameter studies are given below:

<table>
<thead>
<tr>
<th>a (mm)</th>
<th>b (mm)</th>
<th>e (mm)</th>
<th>h (mm)</th>
<th>d (mm)</th>
<th>R (mm)</th>
<th>Probe Radii (mm)</th>
<th>off (mm)</th>
<th>h (mm)</th>
<th>Feed</th>
<th>Radiator Height (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>32.1</td>
<td>6</td>
<td>2.5</td>
<td>23.1</td>
<td>58</td>
<td>0.8</td>
<td>23.1</td>
<td>15</td>
<td>(1.2)</td>
<td>3.6</td>
</tr>
</tbody>
</table>

**Figure 1.** Geometry of a compact dual L-slotted semi-circular patch antenna with Radiator.
Figure 2. Variations of resonance frequencies with slot (a) length b and (b) width a.

Figure 2 (a) and (b) gives the variation of two frequencies of dual band operation ($f_1$ and $f_2$) respectively when plotted as a function of slot length and width. Table 1 gives the variation of two resonance frequencies ($f_1$ and $f_2$), bandwidth (BW; and $BW_2$) and return loss ($RL_1$ and $RL_2$) at these two frequencies when the probe radius is changed.

<table>
<thead>
<tr>
<th>Probe radius (mm)</th>
<th>$f_1$ (GHz)</th>
<th>$f_2$ (GHz)</th>
<th>BW (GHz)</th>
<th>$BW_2$ (GHz)</th>
<th>$RL_1$ (dB)</th>
<th>$RL_2$ (dB)</th>
</tr>
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<tbody>
<tr>
<td>0.4</td>
<td>1.79</td>
<td>2.62</td>
<td>0.05</td>
<td>2.98</td>
<td>1.19</td>
<td>1.42</td>
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<td>0.4</td>
<td>1.79</td>
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<td>2.98</td>
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<tr>
<td>1.0</td>
<td>2.2</td>
<td>3.12</td>
<td>0.10</td>
<td>3.2</td>
<td>1.89</td>
<td>2.08</td>
</tr>
<tr>
<td>1.2</td>
<td>2.2</td>
<td>3.12</td>
<td>0.10</td>
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<tr>
<td>1.5</td>
<td>2.2</td>
<td>3.12</td>
<td>0.10</td>
<td>3.2</td>
<td>1.89</td>
<td>2.08</td>
</tr>
</tbody>
</table>

Table 1: Effect of varying the probe radius

It is observed that when the length of the slot b is increased from 24.2mm to 40.2mm, the second resonance frequency $f_2$ decreases rapidly from 2.08 GHz to 1.35 GHz, whereas the first resonance frequency $f_1$ increases marginally from 1.22 GHz to 1.19 GHz. This is because $f_2$ depends primarily on the slot length and hence it decreases with increase in the slot length. The patch radius mainly decides $f_1$, therefore the increase in slot length has secondary effect on $f_1$.

The variation of the width of the patch has marginal effect on both the frequencies. By varying the probe radius from 0.6mm to 1.5mm, the second resonance frequency shifts from 1.97 GHz to 2.04 GHz while there is insignificant change in the first resonance frequency. The feed probe provides indefinite $50\Omega$ with the input impedance, which is normalized by a slot in the patch at $f_2$, and hence it affects $f_1$.

THEORETICAL AND EXPERIMENTAL RESULTS

Simulation studies on the dual band, coaxially fed semi-circular patch with an L shaped slot on an isotropic dielectric substrate, as shown in Figure 1, with $B = 30$ mm, $a = 10$ mm, $b =$
32.2 mm, ε = 6 mm, d = 2.5 mm, t = 3.1 mm, b=15mm, probe radius=1mm, feed at (1.2,2) and the radar εr=4.5, d=1.6 mm, dielectric constant = 2.2, was conducted. The variation of the theoretical Return Loss with frequency of the antenna is shown in Figure 3. The values of the dual band operation frequencies fL and fH are 1.22 GHz and 1.89 GHz. Theoretical radiation patterns in the E and H planes at both frequencies are in the broadside direction with directivity of 8 dB and 6 dB at fL and fH respectively.

One set of measurements are performed on a suspended dual band, coarsely fed dipole circular-in circumference patch with a L-shaped slot with R = 35 mm, a = 6 mm, b = 44.2 mm, ε = 6 mm, d = 2.5 mm, t = 15.1 mm, b=15mm, probe radius=1mm, feed at (0.8, 0.4). Figure 4 shows that the experimental results are in agreement with the theoretical results for the above mentioned structure.

CONCLUSIONS

A compact semi-circular with L-shaped slot MSA has been presented for dual frequency operation. The inverted dielectric substrate forms the radome for the antenna. Dual frequency ratio from about 1.54 to 1.70 is achieved by changing the slot dimensions and the position of the feed probe. This compact dual frequency antenna is an improved design in comparison with an air dielectric patch in terms of structural stability and ruggedness. This antenna can find applications in Radar and Communication systems such as Synthetic Aperture Radar (SAR), Global Positioning System (GPS), Personal Computer Networks (PCN), etc., which often requires a dual band operation.

Figure 3. Theoretical Return Loss of the compact dual band MSA with radome

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REFERENCES

5. A. Dotunsh and G. Kumar, personal communication.
PHASE SHIFTERLESS, LOG-PERIODIC, SLOT-LINE LEAKY- WAVE ANTENNA FOR BEAM STEERING APPLICATIONS

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A novel wideband, frequency scalable, log-periodic leaky-wave antenna is presented. The log periodic slot line structure results wide bandwidth of about 25.19% and successful suppression of the dominant mode. Dual period beam pattern in the H-plane of the log-periodic leaky wave antenna is frequency scalable and also scalable at fixed frequency by reacting loading of the radiating slots of the antenna. A maximum beam steaming of about 14 degree is observed experimentally by changing the loaded reactance at the radiating slots which results in improved band width of about 33.3 %

INTRODUCTION

Frequency scanning is an effective technique for providing antenna beam steering which has been implemented with microstrip technology in recent years [1 - 3]. Frequency scanning can be a cost effective alternative to phase scanning in applications like low cost radars, imaging and side looking sensors in automobiles because phase shifters and associated active elements are not required to steer the antenna beam [4]. The leaky-wave antenna has the property of narrow beam in the H-Plane which make it suitable for integrated array applications. A leaky wave antenna that can be scanned at a fixed frequency is often preferably to a frequency- scanned one in certain applications where narrow frequency band is available [5].

Here we presents the outcome of the experimental study of log-periodic leaky-wave antennas (Figure 1) which has the excellent advantage of wide bandwidth, dual period beam and frequency scanning capability. The leaky-wave antenna has a dual narrow beam in the H-Plane which can also be scanned at a fixed frequency with simple excitation. The log-periodic construction of the leaky slots give wide bandwidth of about 25.19% and thus more frequency scalability. The reactive loading enable beam steering at fixed frequency by changing the capacitance values included at the leaky slots. The design has been successfully implemented and proved with passive components but can be modified with varactors for beam steering by dc bias tuning.

DESIGN

The proposed leaky-wave antenna geometry is shown in Figure 1. Log-Periodically separated leaky slot lines are drawn using CAD tools and etched on a substrate of dielectric constant $\varepsilon_r = 4.4$ and thickness $d = 1.6$ mm. The design ratio (4) of the log Periodic structure is suited for various widths and optimized for 0.8 with wide bandwidth. The leaky slots of the antenna is loaded with SMD capacitors which results beam steering at fixed frequency by
varying the capacitor values. The antenna is electromagnetically coupled using a 50 Ω microstrip feed line.

**EXPERIMENTAL RESULTS**

The wideband, beam-steering, leaky-wave antenna is studied experimentally using Vector Network analyser HP8510C. The log-periodic structure is optimized for bandwidth for different values of design ratio (t) from 0.75 to 0.9 and maximum -10 dB bandwidth of 35.19 is obtained for t = 0.8. The measured return loss characteristics of the unloaded leaky-wave antenna with optimized design ratio (t = 0.8) gives a -10dB bandwidth of 35.19 % in the band 4.79GHz to 6.095GHz. Frequency scanning behavior is observed when the frequency is varied from 4.03GHz to 5.81GHz and a maximum beam steering of about 14 degree is obtained which is shown in figure 2.

![Figure 2 The Frequency scanned H-plane pattern of the Log-periodic leaky-wave antenna. The beam is scanned for an angle of about 14 degree when frequency is varied from 4.03GHz to 5.81GHz; 4.93GHz; 5.81GHz. The reactive loading in the log-periodic slot lines resulted beam steering at fixed](image-url)
frequency. The capacitor is placed in an optimum position which results uncontaminous beam steering with wide bandwidth. An improved -1dB bandwidth of 33.3 % is obtained by reactive loading, the return loss characteristics of a capacitor loaded leaky-wave antenna with design ratio, \( t = 0.8 \) and \( C \approx 2 \text{pF} \) is shown in Figure 3.

![Figure 3](image)

Measured H-plane patterns for different values of reactive loads of the log periodic leaky wave antenna, shown in figure 4, gives a maximum steering angle of 15.3 degree when the capacitance is changed from 2pF to 33pF at a fixed frequency 5 GHz.

![Figure 4](image)

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CONCLUSION

The work reported presents a wideband, phase shiftless microstrip leaky-wave antenna which can be electronically steerable in a wide band of about 38.3 % bandwidth. The dual pencil beam can either be steerable by reactive loading at the leaky-when at fixed frequency or by varying the operating frequency in the wide operating band. Varactors or switching diodes can be used as reactive loads and can find applications in side looking radars for automotive sensors or such low cost tracking.

REFERENCES

ELECTRONICALLY RECONFIGURABLE DUAL FREQUENCY MICROSTRIP ANTENNA USING VARACTOR DIODE

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This paper presents a novel electronically reconfigurable single feed dual frequency dual polarized microstrip antenna loaded with a hexagonal slot. A varactor diode embedded in the extended arm of the hexagonal slot is used to tune the two operating frequencies with a frequency ratio varying in the range 1:23 to 1:28, without much affecting the radiation characteristics and gain. Desired operating frequency values can be obtained electronically by varying the bias voltage of the varactor diode. This design has an added advantage of size reduction up to 52.6% and 21.9% for the two resonant frequencies compared to standard rectangular patches. This design also gives a considerable bandwidth which almost maintains in the entire tuning range with a maximum of 2.15% with 3.02% for the two operating frequencies.

INTRODUCTION

Dual frequency dual polarized antennas are creating great interest due to their wide applications in satellite base communication, mobile satellite personal communication systems and air route surveillance. They can be applied in a satellite communication system to confine frequency reuse for doubling the system capacity. In mobile communication systems they can be used to obtain polarization diversity for good performance of reception and transmission or to integrate the receiving and transmitting functions into one antenna for reducing the antenna size. Reconfigurable microstrip antennas can cover these multiple functions with the use of a single antenna aperture thus increasing the antenna efficiency and signal processing speed while maintaining high degree of flexibility [1-3]. Moreover, tuning gives the freedom of selecting the operating frequencies of the antenna with desired frequency ratio value, without altering the original design. Thus, all the above mentioned applications need a compact reconfigurable microstrip antenna with better bandwidth and easy tuning capabilities. Here we demonstrate a novel compact single feed reconfigurable dual frequency square microstrip patch antenna loaded with a hexagonal slot and a varactor embedded extended arm for tuning the resonant frequencies. The varactor diode placed in the protruding slot arm helps to control the current density of the TM01 mode which is strongly modified due to the presence of the slot arm. A highly simplified biasing circuitry adds to the compactness of the antenna design. Also the proposed design provides a greater gain and reduction and better bandwidth compared to conventional rectangular patch antennas. The attractive feature of the antenna is its flexibility in selecting the desired operating frequency with without affecting the good return loss characteristics and bandwidth of the two resonant modes. The antenna shows almost identical

I/1
radiation patterns for the two resonant modes in the center tuning voltage range of the varactor diode used. Details of the antenna design for the two resonant modes excited and experimental results are presented.

**ANTENNA STRUCTURE AND DESIGN**

The reconfigurable antenna geometry is illustrated in the Fig. 1. A square microstrip patch antenna with side dimension L is fabricated on a substrate of thickness h and relative permittivity εr. A hexagonal slot of side parameters a1 and a2 with a slot semi-length b, and width w, extending up to the edge of the square patch is placed at its center. A varactor diode D is positioned at the slot center.8.5 mm below the square patch side to tune the frequency ratio of the two excited resonant frequencies with good matching. A capacitor C and a narrow slit are used to achieve good DC isolation, DC bias voltage is supplied from a battery, through two oblique isolators. The antenna is electromagnetically coupled using a microstrip line as shown in Figure 1.

![Antenna Diagram](image)

**Figure 1. Geometry of the proposed reconfigurable antenna.**

The fundamental resonant frequency of the conventional square microstrip patch antennas without slot is 1.885GHz. By loading the hexagonal slot along the operating frequency can be lowered to 1.74GHz, results in greater area reduction. The extended slot arm greatly modifies the TM0n resonant mode current density, which can be controlled using the varactor diode by changing the applied reverse bias voltage. This makes the resonant frequency of the slot arm considerable increases the effective lengths of the two excited resonant modes, TM0n and TM2n, and the excited patch surface current densities are perturbed in such a way that these two modes can be excited for dual frequency operation with a single feed. When the varactor is in the off state, the currents have to flow through the capacitor C, with an increased current path resulting in an increase in the effective length of the TM0n mode giving a high frequency ratio. The varactor loaded across the protruding slot provides variable capacitive loadings to the slot arm. The junction capacitance of the varactor varies against the
reverse bias voltage applied and these different capacitive loadings correspond to different electrical lengths and thus different resonant frequencies.

EXPERIMENTAL RESULTS

The proposed reconfigurable antenna is tested using a HP 8510C Vector Network Analyzer. When the protruding slot arms are absent, the antenna shows a single resonant frequency at 1.24GHz, much lower than the fundamental resonant frequency (1.984GHz) of the unslotted square patch antenna. When the varactor diode is on, the antenna resonates at 1.59GHz and 2.0125GHz with a frequency ratio 1.38. The reconfigurable antenna was then electronically tuned with a reverse bias voltage applied across the diode. When the bias voltage varied from 0 to -30V, the operating frequency ratio is found to decrease to 1.23. Figure 2, shows the measured return loss (S11), of the antenna for different bias voltages of the varactor diode with dimensions: E = 4mm, l1 = l2 = 0.8cm, l3 = 1.4cm, w1 = 0.1cm, h = 0.16cm and f0 = 3.98 GHz. The proposed design has a good matching for the two frequencies in all the tuning voltage range, well below -10dB, with a single feed position. The new design provides an increased extraction of 2.26% for the first resonant frequency and 21.9% for the second, compared to standard rectangular patch antenna. Bandwidths up to 2.15% and 3.02% respectively, have been obtained in the two modes in different bias voltage conditions.

Figure 2. Measured return loss of the antenna for different varactor reverse bias voltages. (L = 4cm, h = h1 = h2 = 0.8cm, l1 = 1.4cm, w1 = 0.1cm, h = 0.16cm and f0 = 3.98 GHz).

The polarization planes of the two resonant frequencies are mutually orthogonal in both states of the pin diode. The E and H-field radiation patterns are measured for different bias voltages and displayed in Figure 4. All the patterns show similar broadside radiation characteristics with good cross polarization levels.
CONCLUSION

A novel electronically reconfigurable dual frequency, dual polarized microstrip patch antenna with a varactor diode control is reported in this paper. The antenna uses a highly simplified circuitry without any transmission lines, for tuning the operating frequencies. The frequency ratio of the antenna varies in the range of 1.23 to 1.28 with a good band width of 2.15% and 3.02% for the two resonant modes. The added advantage of the antenna is that it has a high slice reduction of 52.6% and 21.9% for the two operating frequencies compared to standard rectangular patches, without much reduction in gain. The frequency tuning provided fine tuning and switching over a wide frequency range as is required for many applications.

REFERENCES:
A CROSS-SLOT APEXURE-COUPLED CIRCULARLY POLARIZED ANTENNA ARRAY FOR BRIEFCASE SATPHONE TERMINALS

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A cross-slot aperture-coupled circularly polarized microstrip antenna and its array implementation operating at 2.45 GHz has been designed and investigated. A detailed simulation study has been carried out with a view to optimize various geometrical parameters. Hardware implementation results show that the configuration is suitable for designing light-weight and cost-effective antennas for mobile satellite communication.

INTRODUCTION

Circularly polarized (CP) microstrip antennas are used in several commercial applications, such as, mobile satellite communication, direct broadcast satellite system remote sensing, WLANs, etc. [1] - [3]. A systematic study was carried out on a cross-slot, aperture-coupled, circularly polarized microstrip antenna, which is well suited for low-cost, light-weight, circularly polarized array design for applications, such as in satellite briefcase terminals. The antenna was simulated and optimized using Zeland's HFSS simulation software and its properties were verified experimentally. Next, the prototype antenna element was used for the design of a 4-element circularly polarized array. In the following sections, we present the simulation and experimental results on the cross-slot antenna and its array implementation.

ANTENNA GEOMETRY

Fig. 1 shows the geometry of the antenna under investigation. It consists of two substrate layers, each with a nearly silver path etched on the lower surface of the top substrate and a diagonal microstrip line feed etched on the lower side of bottom substrate.
The cross slot is embedded on the upper surface of the bottom substrate. For bandwidth enhancement, either a foam or air substrate has to be used. In the present case, we considered an air substrate and the air-gap is maintained between the two substrates with the help of Teflon spacers. The diagonal microstrip line feed excites two orthogonal modes with relative phase of \(-45^\circ\) and \(+45^\circ\), which results in phase quadrature between the two modes. The resonant frequency of one of the modes is set below the desired frequency and that of other mode is slightly above the desired frequency. The antenna is excited at a frequency of 2.45 GHz in between the resonant frequencies of these two modes, such that the magnitude of the two excited modes are equal. These two conditions are sufficient to yield CP. The length of the microstrip beyond the center of the slot acts as an open-circuited stub and plays a crucial role in the optimization of antenna performance. The diagonal feed line is bent near the edge of the substrate so that it terminates at one of the sides, instead of at the corner to ease in connector soldering and proper chamfering is done at the bend to minimize reflections.

Evidently, there are a number of parameters such as the height of the air gap, \(a\) mm lengths of the slot, \(a\) mm length, patch dimensions etc., which have to adjusted for the optimum performance. Since this optimization is not possible by experimental means, Zeland's IE3D simulation software was used to obtain a prototype design.

RESULTS

For both the substrates, ARIFON clad with \(\varepsilon_r = 2.17\), \(\tan \delta = 0.0022\), and thickness = 1.55 mm was used. Extensive simulation runs were taken to study the effect of the remaining parameters, namely, the air gap slot lengths and widths, patch dimensions, and the length of matching stub. These parameters were varied one at a time keeping the others constant. The goal was to achieve maximum impedance (VSWR 2:1) and the axial-ratio (AR) bandwidth. It was found that the height of the air gap and the stub length have profound influence on the VSWR and AR bandwidth. Every time a dimensional parameter is changed, the optimum stub length is different. Various dimensions of the optimized antenna are shown in Fig. 2.

![Diagram](image)

Fig. 2. Dimensions of the optimized simulated patch antenna (\(d_1 = 1.35 \text{ mm}, e_c = 2.17, d_2 = 3.0 \text{ mm}, e_r = 1.0, d_3 = 1.55 \text{ mm}, e_c = 2.17, W_p = 48.0 \text{ mm}, l_p = 51.2 \text{ mm}, W_F = 4.8 \text{ mm, } l_F = 21 \text{ mm, } W_a = 1.8 \text{ mm, } L_a = 12 \text{ mm})
The optimized antenna was fabricated and extensively tested in the laboratory. Fig. 3 shows the VSWR plot measured on a HP8720B Network Analyzer and the axial ratio plot enclosed at the center frequency in an anechoic chamber. It is seen that the antenna gives a VSWR bandwidth of about 8.2% as compared to 8.6% predicted by simulation results. Further, the purity of circular polarization is found to be reasonably good. A more detailed comparison of the simulated and experimental results is given in Table 1.

![Fig. 3: VSWR (2.3) plot and axial ratio polar plot at 1.45 GHz](image)

**Table 1: Comparison of experimental and simulated results for cross-dipole antenna**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Simulation Results</th>
<th>Experimental Results</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Minimum value</td>
<td>Center frequency (GHz)</td>
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<tr>
<td>VSWR</td>
<td>1.1</td>
<td>2.45</td>
</tr>
<tr>
<td></td>
<td>2.2</td>
<td>2.45</td>
</tr>
<tr>
<td>Gain (dB)</td>
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<td>Center frequency (GHz)</td>
</tr>
<tr>
<td></td>
<td>9.2</td>
<td>2.45</td>
</tr>
</tbody>
</table>

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The optimized cross-slot element was used to implement a 4-element circularly polarized array. A sequential rotation technique was used as it gives a better axial ratio bandwidth \[4\]. The layout of the elements and the corresponding feed structure are shown in Fig. 4 and 5.

Fig. 4. Layout of a 2x2 array (a) Radiating Patch (b) Coupling Holes (dimensions in mm)

![Diagram of 2x2 array layout](image)

Fig. 5. Layout of the feed network (All dimensions in mm)

The array was first simulated in IE3D and then fabricated and tested in the laboratory. A comparison of the simulated and experimental results is given in Table 2, where a very good agreement can be seen. It is observed that the array implementation gives a considerable improvement in both the VSWR and AR bandwidths, which are 15.2 % and 10 % (3-dB AR), respectively, as against the corresponding figures of 8.2 % and 1 % obtained with a single element. Further, the array has a maximum experimental gain of about 17.4 dB with a 15 % bandwidth for 3-dB reduction in gain. The radiation patterns were measured at several frequencies over the bandwidth of the array and were found to be reasonably constant over the entire range but are not presented here due to space limitations.
Table 2. Comparison of simulation and experimental results for a 2x2 array

<table>
<thead>
<tr>
<th>Parameter</th>
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<td><strong>VSWR</strong></td>
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<td><strong>Antenna</strong></td>
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</tr>
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</table>

CONCLUSION

A cross-antenna-coupled microstrip antenna and its array implementation has been reported, which is quite suitable for making lightweight and cost-effective multibeam terminal antennas.

REFERENCES

A STUDY ON COMPACT RECTANGULAR MICROSTRIP ANTENNA WITH WIDE-BAND

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ABSTRACT

A compact broad band rectangular microstrip antenna with narrow vertical slots and gap coupling has been discussed. The slots are embedded on the patch and only one patch is fed and other patches are parasitically coupled to obtain broader bandwidth without much increase in area.

INTRODUCTION

With the increasing requirements for personal and mobile communications, the demand for smaller and low profile antennas has brought the rectangular antenna to the forefront. Its additional to the requirement for compactness, the designs to achieve properly polarized radiation pattern, high gain and broadband operation are also of particular importance. Several techniques have been reported in the literature for obtaining the compact as well as broadband antennas ([1-4]. In these paper multisection gap-coupled configurations by using narrow vertical slots at 1/4 distance on the patch surface has been presented to realize the compact broadband operation. The present proposed broadband design has been simulated by using IE3D software and experimentally verified by network analyzer. The details of the antenna design and the obtained results for the antenna performance are presented and discussed.

ANTENNA DESIGN AND EXPERIMENTAL RESULTS

Fig. 1 shows the rectangular microstrip antenna (RMSA) with narrow vertical slots without gap coupling. The patch is having the dimensions of L x W and is printed on a glass epoxy substrate with substrate thickness of h and relative permittivity ɛr. The three slots are embedded at 1/4 distance and all have same length L1. All slots are having width of 1 mm. The patch is fed by a coaxial probe at distance d1 from centre. The resonance frequency and bandwidth are 1.68 GHz and 26 MHz (1.54%) respectively. The size reduction of this configuration is 29.38%, which is smaller than that of conventional RMSA [1] the same resonance frequency but with a smaller bandwidth. To improve the bandwidth of these slotted RMSA, multisection gap-coupled configurations by using two and three patch have been discussed.

Fig. 2 shows two gap-coupled slotted RMSA configurations. In this, one in the driven patch & another is parasitic patch with equal dimensions & the gap between the patch is "g".

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driven RMSA is fed with coaxial feed and the parasitic patch is placed along one of the radiating edges of the driven RMSA. Similarly, Fig. 5 shows the configuration of three gap-coupled slotted RMSA, in which only the central RMSA is fed using a coaxial probe, while the other two slotted patches are parasitically coupled along the radiating edges of the driven patch. Dimensions of all the patches, gap between the fed and parasitic patches, feed point location, position of slots, length of slots etc have been optimized to get broader bandwidth with maximum possible frequency reduction.

Fig. 1 Geometry of slotted RMSA

a=1.5 mm, b=1.6 mm, L=20 mm, W=20 mm, h=18 mm, w=1 mm, h=1.585 mm, r=0.6 mm
The proposed design with two gap-coupled and three gap-coupled slotted RMSA has been successfully implemented. The simulated results and measurement are in close agreement. The results for all the configurations are summarized in Table 1. From the results it is clear that the two gap-coupled slotted RMSA gives an experimental bandwidth of 55 MHz (2.26%) and three gap-coupled slotted RMSA gives a broader bandwidth of 77 MHz (4.5%).
Table 1: Comparison of gap-coupled Slotted RMSA configurations

<table>
<thead>
<tr>
<th>Prototype Antenna</th>
<th>Feed from center axial to gap in cm</th>
<th>Frequency (GHz)</th>
<th>Simulated (Exp)</th>
<th>Current Loss (dB)</th>
<th>Simulated (Exp)</th>
<th>Bandwidth Width (MHz, %)</th>
<th>Experimental Bandwidth Width (MHz, %)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conventional RMSA</td>
<td>5.00</td>
<td>2.367</td>
<td>2.379</td>
<td>-19.81</td>
<td>-23.818</td>
<td>50 2.18</td>
<td>58 2.19</td>
</tr>
<tr>
<td>Slotted RMSA (Without Gap Filling)</td>
<td>2.56</td>
<td>1.624</td>
<td>1.601</td>
<td>-19.89</td>
<td>-14.799</td>
<td>10 1.85</td>
<td>16 1.55</td>
</tr>
<tr>
<td>Two gap-coupled slotted RMSA</td>
<td>14.00</td>
<td>1.633</td>
<td>1.700</td>
<td>-20.25</td>
<td>-22.579</td>
<td>22 3.14</td>
<td>51 3.26</td>
</tr>
<tr>
<td>Three gap-coupled slotted RMSA</td>
<td>2.20</td>
<td>1.636</td>
<td>1.708</td>
<td>-14.56</td>
<td>-18.779</td>
<td>64 5.94</td>
<td>27 6.00</td>
</tr>
</tbody>
</table>

Also it has been observed that all configurations exhibit broadband radiation. Fig. 4 shows the impedance and VSWR plots and Fig. 5 shows the H-Plane & E-Plane radiation patterns for three gap-coupled slotted RMSA.

CONCLUSION

Various configurations of Gap-coupled slotted RMSA's yielding broadband and size reduction have been presented. By using periodic patches on both sides of the driven patch the broader bandwidth has been real. All broadband radiation patterns have been observed.

Fig. 4 | input impedance
Fig. 5 Radiation patterns for three-gap coupled slotted RMSA.

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S.M. Valtheparan and K.N. Murty
Aerospace Electronics and Systems Division, National Aerospace Laboratories, Bangalore-560 017, valthen@nal.res.in.

4.2 Parallel implementation of the three dependent Maxwell equations using the MACH VISHW Supercomputer
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4.3 Gain-Radiation characteristics of dielectric loaded horn antennas
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4.4 Moment method approach to use wire antenna as a near field sensor of the waveguide radiating Paramesh and Ajay Chakravarty
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4.5 Design and Development of secondary radar antenna for long range surveillance radar
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4.6 Design of Microstrip Antenna Using Genetic Algorithm
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4.7 Sr$_2$Na$_5$Nb$_2$O$_9$: Ceramic Dielectric Resonator Antennas.
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4.8 Conformal FDTD Modeling of Circular Microstrip Antenna
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Proceedings of APSYM 2004, Dec. 21-23, Dept. of Electronics, CUSAT, Cochin, INDIA.
VERSATILE FETID DESIGN TOOL FOR RF AND MICROWAVE APPLICATIONS

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Abstract: Traditionally, electronic design at high frequencies, notably at RF and microwave follows an iterative procedure, requiring the construction and evaluation of many prototype models. Unfortunately, the construction and evaluation of prototypes can be labor intensive, time consuming, and expensive. At this juncture, the electronics industry is under pressure from the marketplace to design better, less expensive products in less time. Segments of the electronics industry have responded to this pressure with some success, by developing software tools and techniques that assist designers in their work. It appears that for RF designers to be able to keep pace with the rest of industry, such tools and techniques will need to be developed to assist in the task of RF design. As an alternative to the constrast and real prototype models, this paper considers the development of specific tools that will build and evaluate the mathematical models of prototype on a given computer system.

A well-known technique called the Finite Difference Time Domain technique has been selected to perform the evaluation. It is hoped that such an implementation will not only provide a performance for use in RF and microwave applications but in related areas of optics and acoustics as well.

INTRODUCTION

The increased complexity of present day electronic systems calls for improved modeling strategies at early stages of the design process. Traditionally below RF and microwave frequencies, circuit designer have relied on the use of SPICE (Simulation Program with Integrated Circuit Emphasis) based computer software, a circuit-based approach which has been extended to take into account the interaction of electromagnetic fields between components and metallic structures in the PCB assembly under a quasi-static assumption. By introducing single and coupled transmission line models into SPICE and using proper models, the approach has considered the effect of electromagnetic fields interaction up to operating frequency of GHz. However, for very high frequency system where quasi-static no longer applies, this approach has not been found to be accurate. Nor has it been efficient when the metallic structures in the PCB are tightly packed and the equivalent circuit for the system is large.

A variety of full wave models to account for the interactions and coupling have been proposed in the frequency domain which include the method of moments (MoM) and its variants including the boundary element method (BEM), the Finite element method (FEM) and the partial element equivalent circuit (PEEC) approach.  

The discussion of these methods is beyond the scope of the paper and the interested in
referred to review publications by Hayes and Bajer [1], Rauhut [2] and Mitra and Gordon [3]. When short time-scale signals exist, the high frequency effects are no longer negligible and consequently frequency dependent analysis based on MoM and FEM have been developed and are presented in publications [4] and [5]. However, these high frequency extensions often require major reworking of the low-frequency numerical approaches adhered to earlier and it is common to use completely separate software packages for the quasi-static (electro- and magnetic) and frequency dependent field solutions.

This constraint has led to the exploration of direct time domain solutions of the Maxwell's equations and the Finite Difference Time Domain (FDTD) [6] method has been the first technique in this field and has remained the subject of continuous development.

We discuss here key aspects of the FDTD method's application to high speed digital and microwave circuits, for mathematical prototyping for a number of cases. The common thread among all the examples is that the solution quantities are obtained through the FDTD method and used to calculate the voltage and current values at certain predefined points.

**METHODOLOGY**

The tools developed in the present work are based on a set of codes principally using the FDTD method. Several investigators have reported the method of analysis using FDTD method. The algorithm is quite straightforward to implement and the details of the procedure have been left out here as it has appeared in numerous publications [6, 10]. It may be mentioned however here, that the primary focus of the paper is not just the field solution, but also the development of equivalent circuit models within the FDTD framework and prediction of solutions for their electrical performance for the complete network. The method of characteristics is used for this in conjunction with the FDTD model to represent the signals as forward and backward wave components. For lossless lines the method yields an exact solution, since signals are unchanged as they propagate, and the line is represented simply by a time delay. At the ends of the line, the 'travelling' signals are forcing boundary for the circuit element (or network) terminated at a fast or driver to the line. The network solution provides the departing signals that propagate from the line ends back the other direction.

**APPLICATIONS**

A general FDTD based time-stepping network code has been written and validated to work at RF and Higher frequencies. The input to the code is a text file in ASCII format containing the components, connections, drivers and output files of the network. The tools built in have features to handle propagation in multi-conductor transmission lines, transmission line networks with arbitrary loads, Cross talk simulation, skin effect simulations, microstrip antennas etc. We use a few examples to illustrate them although they may not be fully illustrative about the full potential of the code developed:

- **Multi-conductor transmission lines**: Electronic sub-systems on aircraft, missiles and ground electronic systems are generally connected by closely coupled multi-conductor cables. These multi-conductor cables are generally made of conductors with different insulating materials resulting in an inhomogeneous cross...
section media. Similarly traces embedded in a layered circuit board are characterized by lines with different propagation speeds and modes. In general, a discrete number of modes are present. The number of modes, in general, are equal to the number of traces and some may be equal to each other. In two-work all the propagation modes on each line are identified and determined first. The terminal voltages and currents at the ends of the network are calculated using the FDTD method with the developed generalized network code and validated first with the results of the SPICE Model as given in [10].

Results were in good agreement and not shown for brevity of the paper here. Figure 1(a) shows the far end terminal voltages at the ends of the transmission line for the geometry shown in figure 1(a) when taking a Digital Fourier Transform of the time domain values. All the cross talk in this case is the termination cross talk because the line is lossless in this model.

\[ E \text{ (mV) to } \text{ (dB)} \]

\[ V \text{ (V) to } \text{ (dB)} \]

\[ \text{Fig 1 Multiconductor Transmission Line Geometry and Far End Frequency Response} \]

\( b) \text{ Transmission/Microstrip Lines.} \) The modeling of conducting sheets and transmission lines with finite electrical conductivity play an importance in the prediction of losses in microwave circuits. Furthermore, the accurate characterization of shielding surfaces is of great interest to the electromagnetic compatibility problems to study the shielding integrity of enclosures. From the standpoint of computational efficiency, the FDTD Method is perhaps becoming the most commonly method to address these. Full wave methods like the FEM could provide the desired accuracy, but the computational expense and complexity of these methods is considerably greater than the FDTD models. Surface impedance boundary conditions (SIBC) [3] are incorporated in the network FDTD model to characterize the lossy behavior. The conducting region is replaced by an equivalent surface and SIBCs are applied locally. The time domain approximations of the convolution integrals, which are arrived at during the calculations, are solved efficiently using a polynomial approximation of the impedance [8]. Figure 2(a) and 2(b) show the near and far end voltages obtained for the same geometry as in Fig 1a but with a trace resistance of 100 ohm/s. Figure 2(a) gives only the frequency domain responses. Next we compute the code developed with published experimental results [9] for the shielding effectiveness SE measured in screened rooms with either a 100 GHz antenna(100-1000 MHz), a log periodic antenna(200-10000 MHz) or a striplines(10-300 MHz). The field at various points in the enclosure is reported to be obtained with a short wire in the lid, coaxial
A multi-conductor transmission line network: A multi-conductor transmission line network can be considered to be made up of parts in which the preposition of voltage an current cannot be neglected and parts in which this can be disregarded. The former are the transmission lines or tubes and the latter are represented by lumped circuits. Such are called junctions. A junction consists of two or more tubes. A termination is a particular junction with only one tube connected. Representing the signals into forward and backward traveling components uses a characteristic named approach. At the ends of the line, "serving" signals are forcing functions for the network (circuit elements) attached as a load or as a driver to the line. The network solution then provides the "departing" signals that propagate from the line ends back the tube. Figure 4 shows the network responses for a typical configuration for study and implementation of the code. There have been compared with analytical solutions and have been found to be in good correlation with the calculated values.
External field effects on Transmission lines: Electromagnetic fields from distant sources such as transmitters, lightning discharges and an electromagnetic pulse (EMP), as well as nearby sources such as arcane 5X relay contacts can induce undesired signals at the endpoints of transmission lines which may cause detrimental effects in those termination networks. The incident field effects are modeled as distributed sources along the transmission lines so that the telegrapher’s equations are modified. Solutions for this modified telegraphers equations with distributed sources exist in both the time and frequency domains and in recent times the modeling approach in the time domain has been commonly the FDTD method. We present here validated typical results with [10] for the incident field case for the structure shown in figure 5 A wire of radius $r_w=0.1$ mm and length $L=1$ m is suspended at a height $h=2$ cm above an infinite ground plane. The termination are resistive with $R_t=500$ ohms and $R_r=1000$ ohms. A uniform plane wave incident from above has a $V=1$ Amplitude and a rise time $\tau_r = 50\mu s$. The developed codes using our network model (fig 7) were in excellent agreement with published results [10, p.465].
CONCLUSIONS:

In this paper we have illustrated the application of a general purpose three dimensional Maxwell solver to build and validate mathematical models of prototype RF and microwave networks. Although a variety of techniques are available for mathematical modeling, the approach used in this paper allows the advantage of modeling at both high and low frequencies and arbitrary geometries with complex configurations.

![Diagram of a transmission line](image)

Fig. 5 Two conductor Transmission Line (ground included) and computation and prediction of input and output of line with [10, pp465].

REFERENCES:


PARALLEL IMPLEMENTATION OF THE TIME-DEPENDENT MAXWELL EQUATIONS USING THE NAL MK5 SUPERCOMPUTER

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Abstract: An in-house state of art developed time-dependent Maxwell equations FDTD code is ported to multicomputers using the Message Passing Interface, and ANSI C programming language. The ported software is tested on the 32-node MK5 NAL Flux solver system. The parallel and the computing efficiency of the models developed are assessed. A previously unobtainable geometry typified by NAL developed SARS aircraft is simulated by the parallel code. The parallel code is validated with the sequential one for a microstrip printed circuit antenna geometry configuration reported in literature.

INTRODUCTION

Since the inception of the finite-difference time domain (FDTD) method for solving electromagnetic problems in 1966 [1] the algorithm has been successfully applied to solving problems ranging from ground-propagating radar systems, for landmine detection, multi-frequency antenna design, for wireless communications and high propagation in plasma structures [2,3,4]. FDTD computation involves a time-based leapfrog method where updating equations alternate between electric field and magnetic field calculations. To do so, Maxwell’s equations are transformed into their vector, differential form into difference equations. For each time step, x, y, and z components of the E field are evaluated at the time-step. Similarly, the three H-field components are evaluated between time-steps requiring a total of 6N calculations.

The FDTD algorithm is computationally intensive, due to the sheer number of calculations required for each time step. The problem domain is decomposed into a uniform cubic lattice with the dimensions of each unit cell constrained based on the desired accuracy. Typically, a unit cell has dimensions of \( \frac{1}{20} \) to \( \frac{1}{100} \) of the smallest free-space wavelength in the problem domain. Thus the program memory requirements are proportional to \( n^3 \) where \( n \) is the length of each grid dimension.

A constraint is also placed in the time domain such that the largest time-step for numerical stability must satisfy the Courant condition where \( v \) is the minimum propagation velocity in any material in the problem domain:

\[
\Delta t \leq \frac{1}{v} \left( \frac{\Delta x}{2} + \frac{\Delta y}{2} + \frac{\Delta z}{2} \right)
\]

Combining these conditions scales the runtime by a factor of \( n^3 \) using the conventional approach. The FDTD algorithm is essentially data-parallel in nature and parallelization of the FDTD method using distributed computing was pioneered by Varadarajan and Mitra [5] who suggested rules and tolerances for implementation on
a cluster of computers and emphasized having a high computational/communication ratio for good scalability of the FTIDT algorithm. They used the Parallel Virtual Machine (PVM) 3.2 message-passing protocol over TCP/IP on a cluster of eight HP-735 computers. Since then, many problems have been solved using FTIDT on various hardware platforms, processor core configurations and software.

In this paper, an in-house state of art developed time-dependent Maxwell equations FTIDT code is tested in supercomputers using the Message Passing Interface, and ANSI C programming language. The parallel code is tested in the 32-node MK-3 NAL FlexSolve system. The parallel and the computing efficiency of the models developed are assessed. A previously unobtainable geometry simplified by NAL developed SASA aircrafts is simulated by the parallel code. The parallel code is validated with the sequential one for a microstrip printed circuit antenna geometry configuration reported in literature.

MK-5 ARCHITECTURE USED

The NAL MK-5 distributed architecture used has specification shown in TABLE 1. Pentium III III is used as processing element. The operating system used on MK-5 is Linux. Distributed memory architecture is used, each PE has its own local memory module of 256MB. Inter Processor Communication (IPC) card provides the high-speed communication link between PE's and communication switch. It consists of FPGA and DPM. FPAG is used to interface to signals to signals of DPM. DPM has unique feature of being accessible at both of its ports. Each DPM has 128-KB memory buffer, we can carry out both read and write operations.

<table>
<thead>
<tr>
<th>PE</th>
<th>Processor</th>
<th>Memory</th>
</tr>
</thead>
<tbody>
<tr>
<td>80</td>
<td>INTEL PENTIUM III Processors at 445Mhz</td>
<td>1GB per PE, 8 bit EPROM and 4-6 GB Disk size</td>
</tr>
</tbody>
</table>

Sustained GFLOPS simulataneously, 8251A USART (Universal Synchronous Asynchronous Receiver Transmitter) is used for testing the interface and verifying the data transmission.

TABLE 1 MK-5 ARCHITECTURE USED

<table>
<thead>
<tr>
<th>Communication</th>
<th>Distributed Memory</th>
</tr>
</thead>
<tbody>
<tr>
<td>Message Passing Architecture</td>
<td>High Speed Communication Switch</td>
</tr>
</tbody>
</table>

PARALLELIZATION PROCESS

The present approach uses the simple method of parallelization by using spatial decomposition, splitting the problem domain across several processors. The problem domain is evenly partitioned across the maximum size of the chosen volume. Each processor reads the input file to determine the material properties of each of the unit cells. Additional memory is allocated for each processor to store values for communication between nodes. The main loop proceeds by computing data then sending updated data values to processors responsible for adjacent sub-domains as

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I illustrated in Fig 2 and 3 in a one dimensional code.

As shown in fig 2, the Hz update at the border cell of each processor except
the last processor (rank = m-1), requires the Ey value at the border x=0 of the
processor with rank one more than the current processor’s rank. Therefore each
processor except the first (rank = 0), has to send its Ey value to the next processor
with rank one less than the current processor’s rank. And each processor except the
last processor (rank = m-1), has to receive the value of Ey from the processor with
rank one more than the current processor’s rank. This can be clearly understood from
figure 3. Values at the boundaries are updated using a set of special update equations
[6]

![Diagram](image)

**Figure 2** Communicating Hz value required for updating Ey at the border cell of the PE’s

![Diagram](image)

**Figure 3** Communicating Ey value required for updating Hz at the border cells of the PE’s

**RESULTS AND TABULATIONS**

The test geometry chosen is the same as reported in literature [7] and is shown
in figure 4. The developed three dimensional parallel code is executed on different
processors and the run time is compared with the sequential code. The parameters
related to scalability such as speed-up, efficiency are calculated and the graphs of
number of processor versus run time and number of processors versus speed-up are
plotted.

As can be seen from figure 4 good correlation is obtained between the
CONCLUSIONS

The availability of larger cache and main memory as well as low cost computing has aided in the acceptance of FDTD as a viable numerical algorithm. MPI is a widely embraced standard, the operations only occur at initialization, the methods are scalable and extensible, and there is no impact due to run-time load balancing to yield an improvement in overall performance. This is demonstrated in the above.
REFERENCES


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Abstract: The gain radiation characteristics of the standard rectangular horn antenna used extensively as microwave antenna in the 5-20 GHz frequency bands can be significantly improved by loading the horn antennas with dielectric antennas. In X-band, wedge waveguide antenna loaded inside a nilaiscular horn antenna increases the gain between 2-4 dBi and the 3-dB beam width decreases by 47u.

INTRODUCTION

A Horn antenna is a fundamental and standard antenna utilized extensively in the microwave frequencies. Round or rectangular shapes of horn antennas are common but at C, X and Ka microwave bands respectively, the rectangular horn antennas radiating in the TE10 mode is considered a standard antenna as it has fairly high gain (18-22 dBi) with half-power (-3 dBi) beam width between 17-16 degrees and relatively few side-lobes. The polyhedral or the dielectric antennas are Worth versus rectangular microwave antenna that radiates in the end-fire mode and is also characterized by high gain (between 14-20 dBi) and sharp -3 dB beam width (-8-16 degrees). The materials that have proved excellent gain-radiation characteristics in the C, X and Ka bands are Teflon, Polyethylene and Polypropylene respectively. The present study describes the influence of loading circular and rectangular dielectric rod antennas inside the throat cavity of the X-band standard horn antennas using WR-90 waveguide flange in order to improve the gain and beam width performance of standard rectangular horn antennas. During the study it has been observed that wedge waveguide dielectric antennas loaded inside the horn antennas has superior performance as compared to the circular or rectangular dielectric antenna in the X-band frequencies.

EXPERIMENTAL SETUP

The experimental set-up for the studies comprises of X-band Gunn Oscillator, calibrated attenuator, frequency meter/absorption type) by directional coupler and the X-band horn antenna in the transmission mode. The transmitted power is monitored on the 3-dB coupler of directional coupler connected to microwave thermistor mount and the power meter. On the receiver side, the antennas and/or investigation are mounted on a rotating calibrated stand for radiation pattern study at a distance of 3m from the transmitting antenna. The antenna power is again measured through X-band thermistor mount connected to precision microwave power meter.

EXPERIMENTAL RESULTS

The gain and radiation characteristics of the following antennas were measured:

- Standard X-band Horn antennas

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The configuration of the three types of dielectric rod antennas is shown in Fig. 1. The standard horn antenna has dimensions as per WR 90 waveguide specifications and has a gain of 19.5 dB at 10.8 GHz.

Fig. 2 shows the gain-radiation characteristics of the various antennas as mentioned above. It was experimentally observed that the horn antenna loaded with a

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wedge waveguide dielectric antenna results in better gain and radiation characteristics with an overall advantage of approximately 2-dB over the entire X-band frequencies. The VSWR in the entire range is satisfactory, ranging from 1.5 to 1.9 in the entire frequency band.

**CONCLUSION**

A wedge waveguide loaded inside the cavity throat of standard X-band horn antenna realizes a gain that is ~1.5 dB greater than that of the Horn antenna itself and the ~3-dB beam width radiation characteristics improves by 2.4 degrees in the 8.5-11.5 GHz frequencies.

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MOMENT METHOD APPROACH TO USE WIRE ANTENNA AS A NEAR FIELD SENSOR OF THE WAVEGUIDE RADIATOR

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Abstract: The Method of Moment analysis is applied to radiating aperture in an infinite ground plane excited by a rectangular waveguide in the presence of dipole antenna at the near field is presented. From the field existing at the aperture fed by a waveguide, the radiated field at the dipole is evaluated using the plane wave spectrum approach[1]. It is assumed here that, the dipole is thin and that induced current flows only in the axial direction. The scattered field is expressed as an integral over the surface current and is fed to an integral equation.

INTRODUCTION

The radiating waveguide is a fundamental electromagnetic structure, and one about which a great deal is known. With the realization of large-scale arrays, the subject of waveguide radiation and mutual coupling has aroused renewed interest. Theoretical analysis using Variational formulation for a radiating window has been presented by Das [2]. A Moment Method formulation for the problem of radiating rectangular aperture in an infinite ground plane in the presence of dipole antenna at the near field is presented. The Harrington’s general method of analysis [4] [5] is used.

The boundary conditions for the tangential component of the magnetic field are formulated, taking into effect of multiple reflections between waveguide radiator and dipole antenna. The interior region of the hollow wave can be regarded as a circular waveguide, and the field there can be expanded in a series of waveguide modes [3] beyond the cutoff. In this way, it is shown that the total field essentially vanishes in the interior region.

The Waveguide feeding the rectangular aperture is excited in the dominant TE_{10} mode. In the present analysis, the following assumptions are made:

1. The ground plane is infinite in extent.
2. The x-component of the electric field at the plate of the aperture is ignored.
3. Only the x-component of the magnetic field at the aperture plane is considered.
4. The electric field E_x in the aperture varies only in the y-direction and is constant in the z-direction.

The incident magnetic field at the aperture for the dominant TE_{10} mode is given by:

\[ H_{x0} = -j \omega \varepsilon_0 c \frac{z^2}{2z^2} e^{-j \beta z} \]

(1)

and the circular field at the radiating aperture is described by:

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\[ E(x, y, z) = \sum_{n} E_n(x, y) \]

Where the basis functions \( E_n \) are defined by:

\[
E_n = \begin{cases} 
\sin \left( \frac{\pi x}{2a} \right) & \text{if } -a \leq x \leq a \\
0 & \text{elsewhere} 
\end{cases}
\]

The in-component of the magnetic field at the aperture is obtained as:

\[
H_\phi = \frac{\mu_0}{\pi} \sum_{n} \frac{1}{r_n} \sin \left( \frac{\pi x}{2a} \right) \left( \frac{\cos(k_n a)}{k_n^2} \right) \]

The internally scattered magnetic field is obtained as:

\[
H_\phi = \frac{\mu_0}{\pi} \sum_{n} \sum_{l=0}^{\infty} \sin \left( \frac{\pi x}{2a} \right) \left( \frac{n=0}{n=0} \right)
\]

The scattered magnetic field at the waveguide aperture is obtained as:

\[
B = \frac{1}{\kappa} \int_{-\infty}^{\infty} e^{-j\kappa y} \left( -j\omega \cos \phi \right) (1 + j\beta \phi) dy
\]

The boundary condition for the surface of the dipole

\[ E_\phi + \sqrt{\mu_0} \]
RESULTS:

The wire antenna at the near field of the waveguide radiator acts as a scatterer and the reflection coefficient of the waveguide aperture increases. The reflection coefficient is calculated using Method of Moment and it is compared with the reflection coefficient when there is no wire antenna, it is obtained as shown in Fig. 2. In the second step, moving the wire antenna in the electric field direction and computed power at different positions and presented as shown in Fig. 3. The power acquired by the wire antenna is maximum, when it is at the center. The waveguide is designed for s-band, and the wire antenna half-wave dipole and is designed for 9 GHz.

Fig. 2 Variation of absolute reflection coefficient with frequency when the wire antenna is at 1.0 cm from the aperture.
Fig. 3. Variation of the received power with the position of wire antenna.

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DESIGN & DEVELOPMENT OF SECONDARY RADAR ANTENNA FOR LONG RANGE SURVEILLANCE RADAR

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Abstract: A low-cost, lightweight rigid planar single patch monopole feed horn aperture coupled inverted microstrip square patch secondary antenna operating at L band has been designed and developed for the IFF system of long range surveillance radar. The sides of the patch formed the integral radome and the backup metal reflector plate provided the mechanical effect to the antenna pedestal. The sides were covered with FRP layer to prevent moisture from entering the antenna. A return loss of less than -12dB and isolation between sum and difference ports less than -25dB is obtained. The side lobe level in azimuth was less than 22dB and the null depth less than -40 dB at 1.09GHz has been achieved at design frequency.

INTRODUCTION

The main electrical specifications that the IFF antenna has to fulfill are prescribed by the international civil aviation organization (ICAO) the IFF system adopted by its advisory circular on SSRs (Amendment — 4). The antenna used in such a system needs a wide angle radiation pattern in the vertical plane and a monopole pattern on the horizontal plane. Microstrip antennas are well suited for secondary radar applications because of its inherent advantages of low profile, lightweight and easy fabrication. Aperture coupled monopole antenna arrays are extremely advantageous since the feed network and the antenna element are all on the same plane. Hence two layer design, aperture coupled patch antennas present optimal radiation.

In this paper, the design, fabrication and testing of wide band horn aperture coupled inverted square patch array with single plate monopole configuration backed by a metal plate to reduce back radiation and then to act as a mechanical backup structure for fixing onto the antenna pedestal is described. The performance of the inverted patch formed the integral radome thereby eliminating the need for additional radome structure. The sides are covered with FRP layer to prevent moisture from entering the antenna. The IFF antenna developed is suitable for all environmental conditions.

ANTENNA ARCHITECTURE AND DESIGN

The present planar antenna consists of half horn glass aperture coupled 16 (4X4) inverted microstrip square patch radiating element fed by an E-plane microstrip power distribution network, consisting of a corporate power divider network. The

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instrumental comparator circuit for the sum and difference signal extraction. As this antenna has to operate at 1.03 GHz (during transmit) and 1.09 GHz (during receive), the impedance bandwidth of the antenna should cover both the frequencies i.e., 1.02 - 1.1 GHz which is approximately 7.5 %. The structure of the Wilkinson antenna coupled through a hourglass shaped aperture fed by a microstrip line feed network as shown in Fig 3 (topped view) is most suitable for this application. The patch is kept inset so that the substrate itself can act as an integral radome, thereby avoiding a requirement of putting separate radome. The hourglass shaped aperture is also used to improve the coupling and the resonant impedance (1-5). An hourglass shaped aperture uses the features of both dog bone (coupling factor twice and input impedance three times as that of rectangular aperture) and broad (increased coupling factor) shaped apertures without any sharp edges and has maximum coupling factor. The square patch antenna is printed on a RO 4003 substrate of permittivity 2.2 and thickness 8.79 mm. In order to achieve the required impedance bandwidth of an air gap of 14 mm is introduced. As the four sides were covered with FRP with the square patch was designed taking into consideration the dielectric constant of FRP. Experimentally it has been found that the square patch alone is required for this band of operation in 107 MHz. It is fed through a hourglass shaped aperture by a 50-ohm microstrip line terminated by a low impedance open circuit stub of length 13 mm and impedance 2.1 ohms. The antenna element was optimized using commercial software for its optimum performance. A close agreement between simulated and measured results has been found.

**Substrate**

![Diagram](image)

**ARRAY FEED NETWORK CONFIGURATION**

As per the secondary order requirements a 2 X 8 way corporate feed network is designed in micro strip line to get an omnidirectional beam width of 11° and elevation beam width of 33°. The microstrip feed network is designed on RO 4003 substrate of thickness 0.79 mm and permittivity 2.2. Since the entire array in 1.6 x 1.6, the feed network and patch array was divided into two symmetric halves. It is connected to the single plane monopolar comparator made in strips through coastal vias. The monopolar comparator is a hybrid type and its layout is shown in Fig 2b in order to reduce the back radiation (7) from the aperture a metal plate is kept at a distance of 54 mm from the feed layer. This distance is found experimentally so that
the reflected radiation from the slot should add up thereby improving the radiation efficiency.

Fig 2. Monopulse comparator circuit

ARRAY ASSEMBLY

The distance between the feed network and antenna layer and feed network and metal plate was maintained at 14mm and 24mm respectively with Teflon spacers. The antenna was enclosed on the four sides with FRP. The single plane monopulse comparator was fixed to metal back plate. A 1.6 m metal plate was used as the back plate to make the antenna into one single unit by combining the two symmetric halves of the patch antenna array. The complete antenna structure is shown in figure 3. The entire structure becomes a single unit with mechanical fixture to the antenna potential becoming an integral part of the antenna.

Fig 3. Aperture coupled patch antennas for secondary radar applications

The lensed glass shaped aperture coupled square patch gave excellent radiation efficiency with a gain of 18dB, A VSWR of 1.8 and isolation between 5.5MHz and frequency port less than -22 dB is obtained at 1.03GHz and 1.09GHz as shown in Fig 4 and 5. A 5km pattern of 11° beam-width with side lobe level of -22 dB in zenith is achieved for both the frequencies and null depth of less than 40 dB has been obtained at 1.09 GHz (receive frequency) as shown in Fig 6. The elevation beam-width achieved is 33° and the cross polarization has been found to be less than -25 dB.

CONCLUSION

The lightweight lensed glass shaped aperture coupled inverted square patch antennas array with integral radome and mechanical fixture is extremely suitable for radar applications. This method can be applied for higher frequency band arrays with the necessary changes in the feed network and the element. The plastic radome and FRP enclosure protects the antenna from moisture absorption and other similar issues.
environmental hazards.

![Figure 1](image1.png)

Figure 1: Beam loss at 1.003GHz, 44 dB at 1.0GHz.

![Figure 2](image2.png)

Fig. 2. Insertion between man and difference port -22 dB at 1.0GHz. Null depth at 1.0GHz is 20 dB. Null depth of -40 dB at 1.0GHz.

ACKNOWLEDGEMENT

Authors wish to thank Dr. D. C. Powel, Se '9G', Divisional Officer, Radar C and Mr. K. U. Limace, Se '89, Director, LEDE for their continued guidance and their permission to publish the present work.

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DESIGN OF MICROSTRIP ANTENNA USING GENETIC ALGORITHM

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Abstract: How a discrete GA has been applied for the design of microstrip antenna of different configuration is presented in this paper. Comparison is made with experimental results and that of calculation using GA. GA has been applied in one of the case studies in unconventional way to calculate dimension of rectangular microstrip antenna on FR4 substrate.

INTRODUCTION

Since the paradigm shift of computational cost from high to low, many computational techniques are used now a days. CAD techniques made the electromagnetic calculations much interesting, accurate and highly visible. Genetic Algorithm (GA) is used basically as an optimization technique. GA is a search technique based on biological genetics. In recent years, GA is gaining popularity in electromagnetic applications, where generally, the number of variables are more, for their easy searching process, global optimality, independent of non-linear space and probabilistic nature. GA is capable of optimizing non-linear multi-modal functions of many variables. They require no derivative information and they robustly find global or very strong local optima. Numerical experiments indicate that using GA, good solutions to difficult antennas can be obtained quickly, even in time comparable to that taken by analytical methods such as steepness descent. The details of GA and its applications in electromagnetics can be found in [1]-[3]. The expression for resonant frequency acts as limiting function for GA. In an intuitive way, GA chooses optimal dimension which closely matches experimental resonant frequency.

APPLICATION OF GA TO MICROSTRIP ANTENNA

A. DESIGN OF CIRCULAR MICROSTRIP ANTENNA ON THIN SUBSTRATE

The resonant frequency of circular microstrip antenna from [3] is taken as fitness function of GA. The details of implementation technique can be found from [4]. The comparisons of results of GA with experimental are listed in Table I for seven different fabricated circular microstrip antennas. The optimized radii obtained using GA are in good agreement with the experimental results as listed in column 'V' of the Table. The symbols represent the usual notations.
Table 1. Comparison of Results between GA and Exp

<table>
<thead>
<tr>
<th>Patch</th>
<th>No.</th>
<th>f_c (GHz)</th>
<th>h (mm)</th>
<th>GA (cm²)</th>
<th>Exp (cm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>6.0</td>
<td>2.55</td>
<td>3.0</td>
<td>31.142</td>
<td>8.975</td>
</tr>
<tr>
<td>2</td>
<td>6.0</td>
<td>2.20</td>
<td>9.07</td>
<td>19.148</td>
<td>19.346</td>
</tr>
<tr>
<td>3</td>
<td>3.97</td>
<td>2.22</td>
<td>0.79</td>
<td>24.298</td>
<td>24.057</td>
</tr>
</tbody>
</table>

B. DESIGN OF RECTANGULAR MICROSTRIP ANTENNA ON THICK SUBSTRATE

As there is no accurate theoretical formula to calculate resonant frequency of a rectangular microstrip antenna on thick substrate, a trained ANN is used as an expression of its resonant frequency. The work is published in [5]. The present case seems to be exciting. It is noticeable to watch how GA is applied to combination with ANN. The present technique makes GA an universal optimization technique.

These inputs to GA are listed in column II, III and IV. Columns V and VI show the experimental dimensions while GA-ANN based dimensions are listed in columns VI and VII of the table.

Table 2. Comparison of Results between GA-ANN and Exp

<table>
<thead>
<tr>
<th>Patch No.</th>
<th>f_c (GHz)</th>
<th>h (nm)</th>
<th>L (mm)</th>
<th>Exp (mm²)</th>
<th>GA-ANN (mm²)</th>
<th>Exp (mm²)</th>
<th>GA-ANN (mm²)</th>
<th>Exp (mm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.42</td>
<td>2.55</td>
<td>4.78</td>
<td>15.287</td>
<td>18.0</td>
<td>10.060</td>
<td></td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>0.46</td>
<td>2.55</td>
<td>5.25</td>
<td>18.2</td>
<td>20.248</td>
<td>9.74</td>
<td>9.391</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>0.98</td>
<td>2.55</td>
<td>9.52</td>
<td>28.2</td>
<td>31.286</td>
<td>7.75</td>
<td>7.752</td>
<td></td>
</tr>
</tbody>
</table>

C. DESIGN OF RECTANGULAR MICROSTRIP ANTENNA ON THIN SUBSTRATE

Resonant frequency, dielectric constant e_r, and thickness of the substrate h are given as inputs to GA, which gives the optimized lengths L and widths W. The dimensions obtained using GA are in good agreement with the experimental results, as listed in Table 3. Details are published in [6].

Table 3. Dimensions of Thin Substrate Rectangular Microstrip Antenna

<table>
<thead>
<tr>
<th>Patch No.</th>
<th>f_c (GHz)</th>
<th>h (nm)</th>
<th>L (mm)</th>
<th>W (mm)</th>
<th>Exp (mm²)</th>
<th>GA (mm²)</th>
<th>W (mm)</th>
<th>Exp (mm²)</th>
<th>GA (mm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.42</td>
<td>2.55</td>
<td>3.0</td>
<td>4.0</td>
<td>4.97</td>
<td>8.975</td>
<td>4.0</td>
<td>4.97</td>
<td>8.975</td>
</tr>
<tr>
<td>3</td>
<td>3.97</td>
<td>2.22</td>
<td>0.79</td>
<td>24.298</td>
<td>24.057</td>
<td>24.298</td>
<td>24.057</td>
<td>24.298</td>
<td>24.057</td>
</tr>
</tbody>
</table>

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D. DESIGN OF TRIANGULAR MICROSTRIP ANTENNA

The fitness function used in GA to optimize the triangular patch is taken from [7]. The resonance frequency of the triangular patch antenna[7] is given by:

\[ f_{res} = \frac{c}{2\pi \sqrt{\mu \epsilon}} \left( \frac{m^2 + n^2 + m \cdot n}{m^2 + n^2} \right)^{\frac{1}{2}} \]  

Equation (1) is used as the fitness function since it is more accurate as compared to earlier empirical formula. The variable to be optimized is \( r \). The population size is taken 20 individuals, and 200 generations are produced. The crossover probability is set at 0.75, while the probability of mutation is equal to 0.01. Resonant frequency, \( f_0 \), dielectric constant \( \epsilon_r \) and thickness of the substrate \( t \) are given as inputs to GA, which gives the optimized side lengths of the triangle antenna. The optimized side lengths \( r \) obtained using GA are in good agreement with the experimental results as listed in column \( V \) of the Table.

<table>
<thead>
<tr>
<th>Patch No.</th>
<th>( x )</th>
<th>( y )</th>
<th>( H )</th>
<th>( r )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>4.3</td>
<td>19.5</td>
<td>0.07</td>
<td>4.88</td>
</tr>
<tr>
<td>2</td>
<td>4.7</td>
<td>2.22</td>
<td>0.078</td>
<td>4.86</td>
</tr>
<tr>
<td>3</td>
<td>4.8</td>
<td>2.22</td>
<td>0.079</td>
<td>4.86</td>
</tr>
<tr>
<td>4</td>
<td>4.85</td>
<td>4.5</td>
<td>0.179</td>
<td>4.65</td>
</tr>
</tbody>
</table>

CONCLUSION

Design of Microstrip Antennas by using GA is interesting since it gives an optimal configuration. The accuracy of the design directly depends on the accuracy of the fitness function chosen. Thus, a highly selected fitness function in GA will give much accurate result. When there is no any accurate formula, ANN can solve such problems easily by training sufficient experimental data and which in turn can act as a fitness function for optimal design. The same has been implemented to design microstrip antenna on thick substrate. Application of GA to microstrip antenna design seems to be an accurate, computationally simple and cost-effective method.

ACKNOWLEDGEMENT

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SRx-xNaNbO3 CERAMIC DIELECTRIC RESONATOR ANTENNAS

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Abstract: Cylindrical dielectric resonator antennas made from ceramic dielectric material SRx-xNaNbO3 are reported. SRx=0.25Na0.75NbO3 dielectric resonator with co-axial feed were used for construction of antennas. The reflection characteristics, resonant frequency, bandwidth and radiation pattern are measured for the DR excited in TEM01 mode. The DR antenna shows wide bandwidth, high radiation efficiency and small size. Further advantages are low dielectric loss, good temperature stability of the ceramic material and ease of excitation.

INTRODUCTION

Dielectric resonant antennas are resonant volumes of dielectric material. When a dielectric resonator is not entirely enclosed by a conducting boundary, it can radiate, and so it becomes an antenna. The shape of the radiation pattern is a function of the resonator shape. This structure holds promise for the upper millimetre wave frequency bands [1]. This type of radiator has several features, which can be exploited over previously developed antennas. Being a resonant structure, higher efficiencies may be expected, provided that suitable material with low dielectric loss is utilized. Control of the radiated fields may also prove to be simpler with the main lobe of radiation remaining normal to the ground plane of the structure with variations in geometry and frequency.

A dielectric resonator is a piece of high dielectric constant material, usually in the geometry of cylindrical discs. The dielectric element functions as a resonator because of the internal reflections of electromagnetic waves at the high dielectric constant material-boundary. This results in confinement of energy within, and in the vicinity of, the dielectric material, which, therefore, forms a resonant structure. Dielectric resonators can replace traditional, high-Q waveguide cavity resonators in most applications, especially in microwave integrated circuit (MIC) and monolithic microwave integrated circuit (MMIC) structures [2]. The resonator is small, lightweight, high-Q, temperature stable, low-cost and easy to use.

Open dielectric resonators (DR) fabricated out of low dielectric loss materials have been proposed as high radiation efficiency antenna elements [3]. Recently this type of antenna has been analyzed rigorously by a method of moment (MoM) [4], FDTD analysis [5]. Further two five element linear aperture-coupled DR antenna arrays excited in the HE11b mode have been theoretically modeled and experimentally implemented.

In the present study, cylindrical dielectric resonator antenna made of SR0.25NaNbO3 ceramics (SNN) were studied. The resonant frequency, return loss,
radiation pattern and bandwidth of the antenna have been measured.

EXPERIMENTAL RESULTS:

$\text{Sr}_{1-x}\text{Ca}_{x}\text{Nb}_2\text{O}_6$ dielectric resonators are prepared and cylinders were machined with different odd ratios. The radius and thickness are chosen so that the dominant mode $\text{HE}_{11m}$ would be in $\text{S}$-band. The DR antenna is fed by the inner conductor of a co-axial line, which extends beyond the ground plane, a distance $l$ penetrates in to the resonator. The probe is located off centre, close to the periphery of the resonator. Excitation of the antenna may be achieved by a co-axial probe inserted into, or next to, the dielectric body, or by a microstrip under the ground plate with an aperture through it under the dielectric. Figure 1 shows the model of DR antenna mounted on a ground plane, excited in the $\text{HE}_{11m}$ mode.

Fields are excited inside the resonator and will draw most power from the feed when excited in a resonant mode. It is this field which radiates. Modes exist that radiate an omnidirectional pattern in azimuth. For higher orders, the pure transverse electric or transverse magnetic fields cannot exist, so that both electric and magnetic field must have non-vanishing longitudinal components. Such modes are called hybrid transverse-magnetic (HEM), the lowest of them being HEM$_{10}$. The HEM$_{11}$ mode is the fundamental mode for the cylinder. This mode may be, excited by an off-axis probe.

**Figure 1.** Schematic diagram of the DR antenna excited in the HEM$_{11}$ mode.

The dielectric resonator (DR) antenna considered has parameters $a = 10.4$, $a = 8$ mm, $b = 6$ mm, $d = 5$ mm, and $l = 4$ mm. The resonant frequencies and return losses are plotted in Figure 2. From the plot it is observed that the antenna has a bandwidth of 673 MHz and corresponding resonant frequency at 1855 MHz. The radiation patterns of the DR antenna are measured using network analyser and horn antenna. The radiation patterns were normalized with respect to their maximum values. Figure 3 shows the co-polarization and cross-polarization radiation patterns of the dielectric resonator antenna.

The radiation occurs mainly in the broadside direction and it is linearly polarized. The pattern is similar to that of a magnetic dipole. The fields radiate with a figure of eight pattern produced from the magnetic dipole distribution within. The probe can be thought of as a small segment from a vertically oriented current carrying loop. A magnetic field exists in the horizontal plane around this probe and also in the opposite direction around the image of this probe, diametrically opposite in the DRA. This arrangement produces the field that might be expected from two small diametrically opposite segments taken from a loop antenna, and is often called a magnetic dipole. This gives a similar radiation pattern; in azimuth, to a loop antenna.
The high radiation efficiency of DR antennas is achieved because there are no inherent conductor losses in the dielectric resonator. When the relationship between the parameters of the resonator and the input frequency are right, then the dielectric will resonate, i.e., we will observe a displacement current standing wave pattern corresponding to the voltage standing-wave pattern found in a conducting antenna. As Maxwell’s equations indicate, this displacement current wave will, just like the voltage wave, produce electromagnetic radiation. The radiation efficiency of coaxial probe excited DR antenna was measured to be as high as 98% [3]. The radiation pattern of HEM11 mode looks ideally like a pattern of the half wave dipole parallel to the ground plane. In practice, the feeding mechanism may excite more than one mode, so that the pattern will not look like the ideal one [6, 7].

The microwave properties such as complex permittivity, temperature coefficient of resonant frequency of SNN resonators were studied. The dielectric constant and temperature coefficient of resonant frequency depend mainly on the composition of the material. However, the dissipation factor varies from sample to sample in the same composition, indicating that the loss factor is sensitive to slight differences in crystallographic structure and microstructure of the material. One of the loss mechanisms in dielectric materials is the resistance of electric dipole in the medium and will absorb power from the exciting electromagnetic field and the permittivity of the material is reduced.

![Figure 2. Reflection (S21) characteristics of the dielectric antenna made of SNN.](image)

![Figure 3. Radiation pattern of SNN sample (a) Co-polar, (b) Cross-polar.](image)

CONCLUSION

Cylindrical dielectric resonator antenna using SNN:Co0.5Ni0.5Cu0.5 crystals were made. The resonant frequency, return loss, radiation pattern and bandwidth were measured for co-axial feed configuration. The radiation characteristics show a broadside co-planar pattern. The antennas is having a bandwidth of 673 MHz at the resonant frequency 1855 MHz. The material exhibits excellent characteristics suitable for use as dielectric resonators and DR antennas.

REFERENCES

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CONFOMAL FDTD MODELLING OF CIRCULAR MICROSTRIP ANTENNA

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Abstract: A new algorithm for Conformal Finite Difference Time Domain (CFDTD) modelling and analysis of curved edge Microstrip Patch Antennas (MPAs) by superimposing suitable Rectangular MPAs is presented. It has the advantage of using the simple, well developed and proven FDTD algorithms for Rectangular MPA with simple modifications. It offers wide flexibility in design, modeling, and analysis of arbitrary-shaped MPAs. This new technique is applied to an electromagnetically fed Circular MPA. The computed results match with the experimental observations and theoretical data from literature.

INTRODUCTION

FDTD method [1-7] is widely used in the study of MPAs because of its flexibility and versatility, especially in the recent wake of large computational capability and memory availability. By suitable selection of the Yee cells and Courant criterion, Conventional FDTD can be used to give excellent performance in the case of Rectangular MPA. However, the algorithm causes errors while modelling the curved edges, as in Circular MPA. These inaccuracies are mainly due to the stair casing approximation. In order to minimize the error, a filter mesh is needed which can be demanding in terms of CPU time and memory. To overcome these difficulties, several conformal FDTD (C-FDTD) methods have been proposed [3]. However, most of these techniques require complex mesh generation and often suffer from the instability problems.

This paper proposes a robust FDTD technique, with simple modifications of the Cartesian type of FDTD. A multiple number of rectangular patches of appropriate sizes are superimposed to achieve the closest approximation to the geometry under study. Here there is an added advantage of course of finer meshing depending upon the geometry. In this paper, Circular patch antenna fabricated on a standard FR4 substrate is studied using the proposed algorithm.

ANTENNA GEOMETRY

Figure 1 shows the layout of the Circular MPA under study. The CPA with radius r=25mm, is etched on FR4 substrate of dielectric constant εr=4.4 and thickness h=4mm. A 50 Ohm Microstrip feed line, fabricated on a similar substrate, is used to excite the patch through electromagnetic coupling. The experimentally optimized feed length and feed offset from the geometrical center of the patch are F1=20mm and F2=5.5mm respectively. The substrate dimensions L×W are 75mm×72.5mm as shown. The experimental observations are taken using HP 8510C.
network analyzer.

![Diagram of a network analyzer](image)

**Figure**: Geometry of the proposed MPA

**THEORETICAL INVESTIGATIONS BASED ON FDTD**

Any arbitrary shaped MPA can be visualized as an array of multiple rectangular patches of appropriate dimensions. FDTD run for the compact rectangular Microstrip Patch Antenna is then performed in the entire computational domain. For the CPA under study, 12 rectangles of suitable dimensions are chosen for moderate accuracy and computation time. The computational domain dimensions are $115 \times 144 \times 20$ with grid dimensions $\Delta x = \Delta y = 1.167 \text{ mm}$ and $\Delta z = 0.6 \text{ mm}$. A Gaussian pulse of half width 15 ps and time delay 45 ps is launched into the computational domain. The Electric ($E$) and Magnetic ($H$) fields in the computational domain are updated based on the FDTD algorithm. The iterations are carried out for 10000 time steps. No instability is observed when the time steps are increased to 20000. Figure 2 shows the Voltage and Current variation at the observation point within the domain, over 5000 time steps. The Input impedance of the MPA in computed as ratio of the FFT of voltage derived from $E$ field values at the observation point, over the same time steps, to the FFT of current at the same point, derived from the $H$ field values. Reflection Coefficient $S_{11}$ (in dB) is then computed.
RESULTS

The experimental and theoretical Return loss characteristics in the two lower order modes of the CPA is shown in Figure 3. Good agreement is observed between the results as illustrated in Table 1. The antenna resonates at 1.99GHz with a ±3dBWR band width of 80MHz. The numerically computed resonance is at 1.892GHz with a band width of 122MHz.

CONCLUSIONS

A novel FDTD method suitable for analyzing arbitrary shaped MPA is proposed. Results of computation show good agreement with the experimental observation of the CPA. The algorithm is fast and employs the Cartesian type FDTD with simple modifications.
Table I. Comparison of Reflection Characteristics

<table>
<thead>
<tr>
<th></th>
<th>Expt</th>
<th>FDTD</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant Frequency (GHz)</td>
<td>1.7</td>
<td>1.952</td>
</tr>
<tr>
<td>% error between FDTD and expt</td>
<td>&lt;0.5%</td>
<td></td>
</tr>
<tr>
<td>2:1 VSWR Band (GHz)</td>
<td>1.86</td>
<td>1.83</td>
</tr>
<tr>
<td></td>
<td>1.94</td>
<td>1.953</td>
</tr>
<tr>
<td>2:1 VSWR Bandwidth (MHz)</td>
<td>80</td>
<td>173</td>
</tr>
<tr>
<td>% BW</td>
<td>4.2</td>
<td>6.9</td>
</tr>
</tbody>
</table>

Figure 4. Computed reflection characteristics of the PA illustrating the higher order modes

REFERENCES


RESEARCH SESSION 5
CHARACTERIZATION OF AIRBORNE RADOMES IN PLANAR NEAR FIELD TEST RANGE AND ITS SIGNIFICANCE IN FIRE CONTROL RADARS

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ABSTRACT: Modern day fire control radars require a very efficient antenna with very low side lobe levels and high radiation efficiency. Though antennas can be realized for its optimum performance, a poorly designed radome can degrade its performance significantly. The degradation occurs both because of losses in radome and distortion in the antenna pattern as a result of the deformation of the effective illumination pattern. Hence the characterization of radome, apart from its design issues, is a critical aspect for qualifying it for final use. Present paper describes a technique for characterizing airborne radomes using a planar near field measurement system & its significance in radar system performance.

INTRODUCTION

Radomes are dielectric structures that are used to protect antennas from the environment. It can alter the performance of the radar system significantly, if not designed properly. They are required not to exceed some maximum degradation in the electromagnetic performance of the antennas under operation, for some applications, the design of the radome is more complex than the design of the antennas to be protected. The shape of the radomes used in airborne applications involves conformal shapes, which helps in streamlining the body to improve aerodynamic profile and to achieve supersonic speeds. When an electromagnetic wave is incident on an interface between two homogeneous dielectric regions of the radome, part of the energy is reflected and part of it is refracted. This results in beam deflections, transmission loss, depolarization and distortion of the output waveform resulting in degradation of the antenna pattern and generation of spurious side lobes levels in the free space antenna pattern. The extent of distortion of the radiation pattern depends on the shape and material characteristics of the radome. Also with the presence of metal inserts like lightning strippers, pilot tubes and external antistatic coatings, the radome transparency reduces sufficiently. The radome effect in radar performance can be significant, if the radome is very aerodynamic, as with the case of cone shaped radomes in fire control radars. Incidence angle of the beam on the cone shaped radome surface is the function of the scan angle. Polarization of the incident beam is also dependent on the scan angle. Hence the radome has to be characterized over a wide range of azimuth and elevation combination of the scan angles over a desired band of frequencies. Numerical techniques can then be used to interpolate/extrapolate the intermediate values of scan angles and frequency points from the limited collected data available from limited scan angle combination. In the present paper a
methodology for experimentally characterizing a radome using near-field technique is described. Test results of airborne radome for fire control radar along with the critical issues of radome characterization are discussed.

**IMPORTANT OF RADOME CHARACTERIZATIONS**

Generally the radome measurements are meant for characterization of radome and the study of various radome parameters is required to ensure the radome behavior for electrical transparency. Ideally in representing these parameters or radar operation plays a vital role in development of airborne radomes. It may be observed that airborne radome must meet for the aerodynamic properties thus the electromagnetic properties due to its shape. Radome has to be characterized over a wide range of the scan angle & over a band of frequencies. Introduction of cone shaped radome in the nose of an aircraft introduces pattern distortion of the radiated field. Radome has to be characterized for variation transmission loss, VSWR, SSB beam width, significant side lobe levels, beam pointing error, flash lobes, RMS & Inter-constant plane side lobe levels, null depth & null positioning. Measurements of all these parameters are essential in order to ensure the efficient radar performance in the presence of radome. Fire control radar operate with the mono pulsed for the target tracking. The null pointing error caused due to the presence of radome has to be corrected using aberration correction tables.

**TECHNIQUES FOR RADOME CHARACTERIZATION USING PLANAR NEAR FIELD TEST RANGE**

The radome to be characterized is to be mounted on a platform, which should give sufficient clearance from the ground for required coverage angle in elevation plane. The two-axis scanner, which is required for mechanically rotating antennas, is mounted on a mounting platform structure. The antenna is in turn mounted on the scanner. It has to be ensured that the antenna aperture and the scanning plane are parallel when the antenna is scanned to bore sight. The scanner axis and the antenna axis should be aligned when radome is mounted on the setup. The path losses are calibrated and the good signal to noise ratio is ensured at the receiver end. Figure 1 shows the setup of the radome measurement in the planar near field.

![Radome measurement setup](image)

**Test chamber Sampling space is the area where probe samples the significant energy to calculate for field pattern over desired angular coverage. The size of the next window, which is required to calculate desired coverage angle varied for varying lengths of the radomes for typical radar frequency of operation is represented in the figure 2. The size of the scan window, which is required to confine for field pattern is + 10° angular coverage requires chamber size of the order of 30 meters in two scanning planes. The energy in the sampling space has to be sampled at an interval <20. The size of the sampling space and the**

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time required to sample this space is proportional to the length of the
radome to be characterized. Hence the characterization of radome in plane
Fig. 2 near fixed setup become more complex and time consuming for required
number of scan angles and frequencies. The size of the scan window is calculated by
following expression:

\[
S = A \times \pi \times (d + 1) \times \tan\Phi \quad \text{meters}
\]

Where,
- \( A \) = Aperture size of the antenna… meters
- \( L \) = length of the radome… meters
- \( \Phi \) = required angular coverage… deg.
- \( d \) = separation between probe and radome… meters.

In a planar near field measurement setup, the near-field data will be sampled
over a scan window, which is parallel to the antenna aperture as shown in Fig. 3. The
nature of the wave front setup due to the antenna, when antenna is scanned to bore
sight is shown. The transformed far field pattern calculated for sampled data and the
true far field pattern are similar in this case of measurement when antenna is scanned
to an angle \( \Phi \), nature of the wave front is shown in Fig. 4. It can be seen that, the
scanning plane in Fig. 4 makes an angle \( \Phi \) with the antenna aperture. Incident wave
front at the scanning plane makes an angle \( \Phi \) and introduces an amplitude and phase
ear in sampled data. It may be noted that amplitude distribution under these
additional taper in one side of the scan plane, which leads to inaccurate measurement
of free space radiation parameters. Addition of two-axis compensatory scanner at the
base of the radome mounting setup is essential to overcome the limitation of the
planar near field chamber for the radome evaluation. When antenna is scanned to \( \Phi \),
the corresponding compensatory axis must be rotated by \( \Phi \) to bring antenna aperture
parallel to scanning plane. The antenna aperture is made parallel to the scanning
plane, for all combinations of the scan angles.

TEST RESULTS AND DISCUSSION

CRITICAL ZONES OF RADOME

The radome is divided into different zones or regions, according to the
functional requirements of the radar. Zone A in Fig. 2 is a critical area, where
measurement at 1" or better resolution is needed for critical fire control modes of
the radar operation and accuracy of correction is significant for this region of the radome.
The region can extend up to 15" Zone.
Radome aberration is defined as the deflection of the elevation axis towards a target from the true geometric axis, due to refraction of the radar beam in the presence of radome. Generally, these aberration tables are defined in terms of correction factors in azimuth and elevation of the null. These radome aberration correction tables are represented as an average correction factor over a radar body. Radome aberration correction tables are stored in a radar computer as a look-up table and are referred when radar operation. For an example, if the correction factors were not incorporated for radome, which introduces a null drift of 5 milli-radians in one plane, the fixed missile for a 10 km target misses by 50 meters.

EXPERIMENTAL RESULTS

A case study of cone shaped radome most suitable for fire control radar was taken and subjected to rigorous characterization for various radiation parameters. Transmission loss of an antenna in the presence of radome over wide scan angles is plotted in Fig 6. Beam pointing errors over the frequency band is plotted for selected set of scan angles is shown in Fig 7. Scattering of incident energy in space can be noted in the presence of radome from Fig 8 and Fig 9. Variation of SIR for idealized is plotted across limited scan angles in Fig 10. Using all these measured data a correction table can easily be made and used for enhanced radar performance.
CONCLUSION:

Methods for characterisation of airborne radomes in planar near field facilities are discussed. The short falls and techniques to overcome these shortfalls are described in detail. The cone shaped radome shown in Fig. 1 has been characterised in a planar near field facility. Radiation parameters like contumacy loss, beam width, beam-pointing errors, and side lobe level variations are studied.

ACKNOWLEDGEMENT:

Authors are thankful for encouragement and support received from Dr. D. C. Pasley, Dgr, officer Radre.C. They are also thankful to Director IEBE for his continued support and permission to publish this work.
BROAD BAND DESIGN TECHNIQUES FOR CHOKE FLANGED FEED HORN FOR EARTH STATION ANTENNAS

A.H. PATEL
Antenna Systems Group, Space Applications Centre AHMEDABAD

ABSTRACT: This paper describes a systematic technique for the design of a broadband choke-flanged feed horn developed for an earth station antenna operating in Normal- and Extended-C band of satellite communication to cover a wide range of satellites (Galileo) beam symmetry, low side-lobe as well as low cross-polar performance over an almost octave bandwidth with the design goals for the design presented here. The design parameters have been discussed thoroughly to make it attractive for the design of the broad for medium sized reflector antennas in earth station terminals. The design, width and number of chokes for this design were optimally designed for radiation beam symmetry, cross-polar discrimination and low side-lobe performance over a 1:3 bandwidth. The ridge geometry has been optimized to control feed V.S.W.R. and radiation characteristics. The measured radiation characteristics and cross-pol, performance are presented. The integrated test performances of feed with a 2.5-meter reflector is reported. Various feed models were developed to study the effect of ridge and slot-length on feed horn performance and the results are summarized in last part of the paper.

INTRODUCTION

There has been a considerable interest and effort for the past several years in the development of dual-band feed horns for earth station antennas in satellite communications. A high degree of polarization purity over two discrete and widely separated frequency bands for two receive functions has been the need of the hour in efficient and attractive satellite communications. The INIA series of antennas are equipped with Normal- and Extended-C band since last 20 years. The frequency band covered is 3700 MHz to 7025 MHz for receive and transmit purposes. The Corrugated Horn, though an efficient primary radiator, is not attractive as a feed horn in conventional parabolic reflector antennas. The cross-polar requirement of a typical 4.5-m road-transportable terminal, for width of fringes indicates -equipped work was carried out. It was 25 dB. Hence a design based on a choke-flanged feed horn using wide band technique was adopted for this purpose. The fabrication of the feed was carried out in-house and then fully evaluated in a Anechoic Chamber for its radiation characteristics of bandwidth, polarization purity and side-lobe performance. The secondary performance of integrated feed with 2.5m reflector was finally carried out at the Space Application Centre using 300-meter elevated out door antenna test range.

DESIGN PARAMETERS

The horn design commences with the throat section which must be excited by TE11 mode in circular wave guide. The waveguide diameter selected is actually cut-off between the 20mA frequency band operating near cut-off wavelength and higher.
frequency was extending near higher-order mode region the the
accommodation of the TM11 to 3075 MHz frequency bands. The generation of higher-
order mode TM11 which usually causes depolarization and asymmetry in beamwidth,
is precluded. Assuming that only TE11 mode propagates in the circular waveguide,
the input radius, a, is given by

\[ a = \frac{\lambda_0}{2\pi} \sqrt{\frac{2}{\epsilon}} \]

Under these design criteria, the dimension a, is still near to cut-off region for
lowest frequency band used for receive function. The waveguide propagation
condition at the lowest band can be improved by introducing a pair of elastically
opposite ridges of about 0.014 square dimension across the horn circular section in the
plane of polarization of receiving horn.

The depth of the choke is chosen to be effective so that no surface current is
supported at the aperture in the E-plane. This will in turn, control the losses symmetry
in E- and H-planes and also the generation of x-polar components. The depth of choke d,
is chosen as per

\[ d = \frac{\lambda}{4} \]

When \( \lambda \) is wavelength at lowest frequency, \( d \) is wavelength at highest frequency

Chokes with smaller widths are more frequency sensitive. The field in wider slits is highly localized from the diffraction on the adjacent edge but gives larger
blockage. An empirical value of 0.12 to 0.15 slit width is thus a judicious choice.
The numbers of chokes in this design is three, which is optimum for beam roll-off and x-
pol performance. The larger number of slot gives larger blockage without any significant
improvement in performance.

The aperture diameter of the horn is designed from waveguide's requirement
over the operating frequency band for given \( F/D \) ratio of the reflector. The feed
illumination taper on the reflector periphery decides the overall maximum efficiency and
sidelobe performance. The optimum value of edge-taper is desirable between 10 to
15 dB for large bandwidth of 2:1 to maintain gain-factor value greater than 0.6. The
edge taper is a function of aperture diameter, which is controlled by generation of higher-
order modes over the given bandwidth. A trade-off value has to be taken for
optimum performance over the total operating band.

TEST PERFORMANCE AND ANALYSIS

The performance of the horn was evaluated for return loss, cross-polarization,
beam symmetry and sidelobe level (see Table 1.) The integrated feed performance
with a 2.5m reflector of 0.4 F/D was also carried out on an antenna test range. This
antenna was selected for the feed evaluation as the actual 4.2m antenna could not be
localized in the test range. However, both antennas had the same F/D to ensure the
validity of results. The measured x-pol and antenna efficiency performance for the
antenna is summarized in Table-1.
<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>5-dB Beamwidth (degree)</th>
<th>Cross-polar Performance dB</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>E-plane</td>
<td>H-plane</td>
</tr>
<tr>
<td>3.500</td>
<td>57°</td>
<td>58°</td>
</tr>
<tr>
<td>4.600</td>
<td>77°</td>
<td>76°</td>
</tr>
<tr>
<td>7.622</td>
<td>85°</td>
<td>85°</td>
</tr>
</tbody>
</table>

Table 2: E-PLANE V. H-PLANE OF CROSS-POLARIZATION BEAMWIDTHS FOR 91.6° ELEVATION ANGLE

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>5-dB Beamwidth (degree)</th>
<th>Cross-polar Performance dB</th>
</tr>
</thead>
<tbody>
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</tr>
<tr>
<td>7.622</td>
<td>85°</td>
<td>85°</td>
</tr>
</tbody>
</table>

Table 3: SUMMARIZED TEST PERFORMANCE OF INTEGRATED FEEDS WITH 91.6° ELEVATION ANGLE

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Gain (%)</th>
<th>X-pol DB</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.500</td>
<td>59.3</td>
<td>-36.0</td>
</tr>
<tr>
<td>4.600</td>
<td>51.0</td>
<td>-31.4</td>
</tr>
<tr>
<td>6.600</td>
<td>43.7</td>
<td>-50.8</td>
</tr>
<tr>
<td>8.600</td>
<td>42.3</td>
<td>-50.8</td>
</tr>
</tbody>
</table>

Measured return loss was typically better than 16 dB at the 8 GHz band and better than 18 dB at 11 GHz. The measured 40-dB beamwidths in E, H & 45° planes is within ± 1.5° for any frequency of 8-80 GHz feed (Table-1). This shows that the cross-polarization is highly symmetric in any plane. The cross-polarization gain is achieved at low angle to 100°H (Table-2). This indicates that the horn cross-sectional dimensions and choke-change geometry have good control over suppression of cross-polarized components. The measured gain and x-pol of antennas indicate that the overall efficiency of the order of 80% and x-pol performance nearly steep -36 dB at 11 GHz can be achieved with a carefully designed choke-change horn.

EFFECT OF NOISE ON BORN PERFORMANCE

Beamwidth: Beamwidth is wider than about 1 dB at 8 GHz and 2 dB at 11 GHz due to unattended angles of 64° with ridge.

X-pol: Fig. 1 depicts marginal change on X-pol with ridge at 8 GHz band, and deterioration of about 2 dB at 6.875 GHz. Fig. 2 depicts marginal change in X-pol with ridge at 11 GHz band. But improved by about

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EFFECT OF SLOT DEPTH

No effect on HI plane even for changing slot depths from 20 mm to 25 mm. Pattern became narrow in E plane with decrease in depth. But this effect is insignificant when the aperture diameter of the horn is large.

CONCLUSION

A high performance choke-flanged horn for 1:1.5 bandwidth could be evolved by using broadside design technique. The design is thus simple and has potential application in satellite communication earth station antennas for INSAT series of satellites. It can be made further attractive through its fabrication simplicity, low mass and cost effectiveness when required in larger quantities.

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REFERENCES

SIMULATION MODEL OF A MULTI-CARRIER CODE DIVISION MULTIPLEXS ACCESS (MC-CDMA) MAC PROTOCOL FOR WIRELESS MULTIMEDIA COMMUNICATION NETWORKS

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ARSTRAC is a research paper that is designed and developed by the students of M.C-CDMA over DS-CDMA and it is called as MC-CDMA MAC protocol. This protocol ensures higher channel utilization, maximum system capacity and less BER and thus it can support very much for multimedia traffic such as video, data, and text. The entire protocol is designed and developed using MATLAB software in WINDOWS environment. The input parameters to the program are: number of mobile users, bit rate, bandwidth, PM codes and channel characteristics. The results obtained from the program are: BER performance in combined AVGN and Rayleigh fading channel for different types of codes.

INTRODUCTION

The field of Wireless communication is growing rapidly as a result of advancements in digital communications, mobility, and semiconductor technology. Multiple Carrier Code Division Multiple Access (MC-CDMA) MAC protocol provides different channels to different users, thus can increase the advantages to accommodate more number of users in a communication system with less BER and multi path fading. MC-CDMA system has newly received considerable concentration. WIRELESS LAN IEEE 802.11 STANDARDS

The IEEE 802.11 standard specifies a basis for wireless LANs. The MAC protocol defined in the IEEE 802.11 draft is sophisticated and allows for considerable flexibility. Two primary topologies are supported by the IEEE 802.11 standard: one in which the station access the backbone network (distribution system in IEEE 802.11 nomenclature) via access points (i.e., base stations) and another one in which a group of stations communicates directly with each other in an ad hoc network, independent of any infrastructure or base stations. The first topology is useful for providing wireless coverage of buildings or campuses areas by deploying multiple access points whose radio coverage areas overlap to provide complete coverage. The IEEE 802.11 draft standard provides for three different types of physical layers to be used: 2.4 GHz ISM band frequency hopping (FH) spread spectrum radio, 2.4 GHz EM band direct sequence (DS) spread spectrum radio and radio Infrared (IR) light, Two different data rates (1 Mbps and 2 Mbps) for the above physical layer are used. The important characteristics of the IEEE 802.11 MAC protocol which are likely to remain unaltered in the final standard are its ability to support the access point - oriented and ad hoc networking topologies, both asynchronous and time - critical traffic and power management.

PROPOSED RESEARCH WORK

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This paper is focused on the design and development of the Multi Carrier Code Division Multiple Access (MC-CDMA). MC-CDMA (Multiple Access) protocol for wireless multimedia communication networks. The protocol is very useful to increase the performance of data transmission such as BER, number of users. By giving different combinations of input parameters such as maximum number of users, their data rate, modulation, PN codes and element characteristics, the performance of the MC protocol can be analyzed.

**CODE DIVISION MULTIPLE ACCESS (CDMA)**

We consider a 1/FN-SS (Frequency Division Multiple Access) multiple access system with BPSK signaling and a correlation receiver. There are M users in the system. Received signal of a user i at the base station is given by:

\[ r(t) = \sum_{i=1}^{M} \sqrt{E_i} \cdot-\frac{1}{\sqrt{T_c}} \cdot d(t) \cdot \sin(\omega_c t) \cdot \cos(\theta_i(t)) \]

where \( E_i \) is the transmitted energy per bit of the signal, \( T_c \) is the time duration of the data bit, and \( \omega_c = \frac{2\pi f_c}{T_c} \) represents the data signal of each user, \( \\theta_i(t) \) is the phase. \( d(t) \) is the data signal of each user, \( \omega_c = \frac{2\pi f_c}{T_c} \) represents the data signal of each user, \( \\theta_i(t) \) is the phase. The total number of chips per bit, \( T_c \), is given by \( T_c = \frac{1}{f_c} \), where \( f_c \) is the carrier frequency. The channel is modeled in the figure below. We consider a single cell system with M users with perfect power control, i.e., all users assigned to a base station should have the same received power at the base station. The symbol \( X(t) \) represents the addition of the data of every user in the medium. The data signal is given by:

\[ d(t) = \sum_{m=-\infty}^{\infty} s(t-mT) \cdot \delta(t-mT) \]

where \( s(t) \) is an infinite sequence of data and \( P(t) \) is a spreading sequence with unit amplitude and duration. The spreading signal is given by:

\[ s(t) = \sum_{m=-\infty}^{\infty} s(t-mT) \cdot \delta(t-mT) \]

where \( m \) is the AWGN with two side-power spectral density \( \frac{1}{2} \). The desired signal and the interference in the multiple access interference (MAI).
PERFORMANCE ANALYSIS WITH GAUSSIAN APPROXIMATIONS

With perfect power control, all users assigned to a base station should have the same received power at the base station. We consider here a single cell system with $M$ users. In a DS-CDMA system (see Fig. 1), the number of users that the cell can support is limited by the total received interference at the base station and it will vary with time. All users share the same frequency spectrum and experience interference from all other active users throughout. Given a quality of service requirement, the Eb/No ratio for user $i$ is given by:

$$\frac{E_b}{N_0} = \frac{P_i}{\sum_{j \neq i} P_j}$$

and the probability of error (BER) for user $i$ in CDMA environment with 3PSK modulation is given by

$$P_e = \left(1 - \frac{1}{2} \right) \left(1 - \sum_{k=0}^{\infty} \left(\frac{1}{4}ight)^{k} \frac{E_b}{N_0} k\right)$$

where $P_i$ is the received power level of user $i$ at the base station and $P_a$ is that of nth user.

The binary data sources corresponding to different users generate antipodal signals. The spreading may be done using different code for different users and the modulation employed is BPSK. The signals interfere with each other, creating multiple access interference (MAI). Noise is assumed to be AWGN. The detector is of 4th correlation type. The theoretical probability of error is given by

$$P_e = \frac{1}{2} \left(1 - \sum_{k=0}^{\infty} \left(\frac{1}{4}ight)^{k} \frac{E_b}{N_0} k\right)$$

where $E_b$ is the energy per bit and $N_0/2$ is the noise power spectral density, $M$ is the number of users, and $L_e$ is the processing gain (chip length). The first term in the denominator accounts for MAI.
The graph gives the performance of the DS-CDMA communication system. The expression for probability of error is evaluated by using MATLAB and plotted in Figure 2 as a function of Eb/N0, where Eb/N0 is displayed on a logarithmic scale (10log10(Eb/N0)) as expected the probability of error decreases exponentially as Eb/N0 increases.

The graph gives the performance of the DS-CDMA communication system from Number of users Vs Bit Error Rate. The expression for probability of error is evaluated by using MATLAB and plotted versus Number of users in Figure 3 for various values of Eb/N0 as expected the probability of error increases exponentially as Number of users increases.

MULTI CARRIER CODE DIVISION MULTIPLE ACCESS

The figure shows a schematic diagram of a Multi Carrier Code Division Multiple Access for wireless multimedia communication network. In the transmitter side, each user data is spreading using their Pseudo random code and modulate with their assigned carrier frequency and transmitted. In the medium all user data get added in time and that signal is reached at the receiver end. At the receiver, the signal is despread and demodulated to get the desired data.

We consider a Multi Carrier Direct Sequence Spread Spectrum multiple access
A system with BPSK signalling and a correlation receiver. There are M users in the system. Received signal of a Mth user at the base station \( s_m(t) = \sum_{n=-\infty}^{\infty} d_m(t-nT_s) \) where \( d_m(t) \) is the transmitted data signal for user \( m \), \( T_s \) is the symbol duration and \( d_m(t) \) is a PN (pseudo-random) sequence of user \( m \). The modulation waveform of each user is \( s_m(t) \). \( T_c \) is the duration of each chip in the PN sequence, \( f_c \) is the chip rate (in chips/sec) and \( F_b \) is the data rate (in bits/sec). We assume that \( T_c \) is independent for each user and that users with perfect power control, all users assigned to the same base station should have the same received power at the base station. The symbol \( \sum_{n=-\infty}^{\infty} d_m(t-nT_s) \) represents the addition of the data of all users in medium.

The data signal is given by
\[
d_m(t) = \sum_{n=-\infty}^{\infty} d_m(t-nT_s) \quad \text{where} \quad d_m(t) = \sum_{n=-\infty}^{\infty} c_{m,n} \delta(t-nT_s) \]

The spreading signal is \( s_m(t) \) where \( c_{m,n} \) is independent for each user and that users with perfect power control, all users assigned to the same base station should have the same received power at the base station.

\[
s_m(t) = \sum_{n=-\infty}^{\infty} c_{m,n} \delta(t-nT_s) \quad \text{where} \quad c_{m,n} = \text{independent for each user.}
\]

Assume that we receive signal from user \( m \) of a M user system is given by
\[
r(t) = \sum_{n=-\infty}^{\infty} s_m(t-nT_s) = \sum_{n=-\infty}^{\infty} d_m(t-nT_s) + \text{AWGN with two sided power spectral density} \quad \sigma^2 \quad \text{and} \quad \text{MAI}.
\]

![Figure 4](image-url) Possible implementation of an Multi-Carrier spread-spectrum transmitter and receiver in an AWGN channel. Each bit is transmitted over \( N \) different subcarriers. Each subcarrier has its own phase offset, determined by the spreading code.
DS-CDMA/MC-CDMA MAC PROTOCOLS

In DS-CDMA technique, the frequency spectrum of a data-signal is spread using a code uncorrelated with that signal. As a result the bandwidth occupancy is much higher than required. The codes used for spreading have low cross-correlation values and are unique to every user. This is the reason that a receiver, which has knowledge about the code of the intended transmitter, is capable of selecting the desired signal.

A multicarrier CDMA system, in which a data sequence multiplied by a spreading sequence modulates M carriers, rather than a single carrier. The receiver provides a correlator for each carrier, and the outputs of the correlators are combined to yield a processing gain comparable to that of a single carrier DS system. This type of system has the following advantages. First, a multicarrier CDMA system is robust to multipath fading. Second, a multicarrier system has a narrowband interference suppression effect. Third, a lower chip rate is required, since, in a multicarrier CDMA system with M carriers, the entire bandwidth of the system is divided into M equi-width frequency bands, and thus each carrier frequency is modulated by a spreading sequence with a chip duration which is M times as long as that of a single carrier system. In other words, a multicarrier system requires a lower speed, parallel-type of signal processing, in contrast to a fast, serial-type of signal processing in a single carrier RAKE receiver. This, in turn, might be helpful for use with a low power consumption device. Finally, a multicarrier system does not require a contiguous frequency band, so that very large spread bandwidths are achievable. More specifically, in this proposal, we combine multicarrier DS CDMA with single-user multiple access in interference suppression. The combination of these two techniques will result in a highly flexible design that will exhibit robustness to both narrowband interference and multiple access interference.

SIMULATION AND RESULTS

A DS-CDMA DIGITAL COMMUNICATION SYSTEM

The binary data (+1 or -1) signals are generated randomly for each user. Each signal is spread using Maximum length and Gold codes to get chips. (Note that different user is given a different code). Then they are converted into sinusoidal form at a particular Eb/No at the modulator. It is represented by the samples taken at regular intervals of time for simulation purpose. All the signals are added in parallel at the channel, creating multiple access interference (MAI). To each sample the Gaussian noise is added. The signals are de-spreaded at the receiver by multiplying with the corresponding code. The correlation detection is done by multiplying each de-spreaded signal with the same carrier and summing the samples over each bit duration. The result is compared with the original data to find the number of errors. The probability of error is found for different Eb/N0s (different variances) and also for different sets of users. The simulated probability of error is plotted against Eb/No in dB or the number of users. Curve fitting is done to get a smooth graph. The graph is almost coinciding with the theoretical one.
A MC-CDMA DIGITAL COMMUNICATION SYSTEM

The binary data (+1 or -1) signals are generated randomly for each user. Each signal is spread using Maximum Length and Gold codes to get chips. (Note that different users are given a different code). Then they are converted into different sinusoidal form for different users at a particular E_b/N_0 at the modulator. It is represented by the samples taken at regular intervals of time for simulation purpose. All the signals are added in parallel at the channel, creating multiple access interference (MAI). To each sample the Gaussian noise is added. The signals are despread at the receiver by multiplying with the corresponding code. The correlation detection is done by multiplying each despreaded signal with the same carrier of each user and summing the samples over each bit duration. The result is compared with the original data to find the number of errors. The probability of error is found for different E_b/N_0 (different variances) and also for different sets of users. The simulated probability of error is plotted against E_b/N_0 in dB or the number of users and can understand that MC-CDMA gives better performance than other communication systems.
Figure 10
Simulated BER/BER at three users Rake performance graph of MC-CDMA for different number of users with Gold code.

Figure 11
Number of Users Vs simulated Bit Error Rate performance graph of MC-CDMA with Maximum Length ML codes for 100 user at 3dB.

Figure 12
Number of Users Vs simulated Bit Error Rate performance graph of MC-CDMA with Gold code for different 800 user at 3dB.

CONCLUSION

In this paper, the MC-CDMA/DS-CDMA MIMO protocols of a multi user system is studied. Simulations are done using MATLAB. By changing the channel characteristics, the performance of the protocols are studied.

ACKNOWLEDGMENTS

The authors thank the reviewers of this paper, the technical chair and the organizers of APSYM 2004 Conference for their cooperation in printing this paper in their proceedings.
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SIMULATION MODEL OF A HYBRID MAC MC-CDMA/S-ALOHA PROTOCOL FOR WIRELESS MULTIMEDIA COMMUNICATION NETWORKS

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ABSTRACT: This research paper work is designed and developed by combining the merits of both Slotted ALOHA and MC-CDMA and it is called as MC-CDMA/S-ALOHA MAC protocol. This protocol develops better channel utilization, higher system capacity and supports multimedia traffic such as voice, video and text. The entire protocol is developed and tested using MATLAB software in Windows environment. The input parameters to the program are number of mobile users, their data rate, PN sequence, and channel characteristics. The result obtained from the program is the graph of throughput vs. offered load.

INTRODUCTION

The field of Wireless Local Area Networks (WLANs) is expanding rapidly as a result of advances in digital communications, portable, and semiconductor technology. The wireless MAC protocols used for LAN networks support multimedia applications like combined transmission of audio, video and text data. Facing the soaring demand for mobile computing and mobile communications, the use of smart cards in mobile communications systems has recently received considerable attention.

SYSTEM MODEL

Here for simplicity, we only focus on single cell systems as shown in figure 1. There are N mobile terminals sharing M possible, equal-capacity channels for transmission to one base station. Each mobile terminal can transmit a packet on a randomly selected channel and using a randomly selected PN code. Time is divided into fixed length slots; the transmission slot is equal to one slot.

According to this protocol, once a new packet is generated, the user transmits the packet at the start of the slot using a randomly selected channel and a randomly selected PN code. If two or more packets are transmitted simultaneously, using same channel and PN code, packet collision occurs. Here we also introduced stop-and -wait ARQ. The user whose transmission is unsuccessful will retransmit in future slot.
WIRELESS LOCAL AREA NETWORK (WLAN) STANDARDS

The IEEE 802.11 standard specifies a family of wireless LANs. The MAC protocol defined in the IEEE 802.11 draft is sophisticated and entails considerable complexity [1]. Two primary topologies are supported by the IEEE 802.11 standard: one in which the stations access the backbone network (distribution system in IEEE 802.11 nomenclature) via access points (i.e., base stations) and another in which a group of stations communicate directly with each other in an ad-hoc network, independent of any infrastructure or base stations. The first topology is useful for providing wireless coverage of buildings or campus areas by deploying multiple access points whose radio coverage areas overlap to provide complete coverage. The IEEE 802.11 draft standard provides for three different types of physical layers to be used: 2.4 GHz ISM band frequency hopping (FH) spread spectrum radio, 2.4 GHz ISM band direct sequence (DS) spread spectrum and radio frequency (RF) ethic. Two different data rates (1 Mbps and 2 Mbps) for the above physical layers are used. The important characteristics of the IEEE 802.11 MAC protocol which are likely to remain unchanged in the final standard are its ability to support the access point - oriented and ad-hoc networking topologies, asynchronous and time - critical traffic and power management.

PROPOSED RESEARCH WORK

This paper work is mainly focused on the simulation of hybrid MAC protocols. These protocols are very useful to obtain better performance of data transmission. By giving different combinations of input parameters such as maximum number of users, data rate, CDMA codes, power, the performance of the MAC protocol can be analyzed.

MAC PROTOCOLS

When a resource is shared among multiple independent stations or users, the need exists for a mechanism to control access to the resource, or in the absence of such a mechanism there must exist a course of action to be taken when two or more users attempt to acquire the resource at the same time. Such a resource, in the case of satellite communications, is the transponder. Because of the limited capacity of a satellite transponder it is necessary to utilize the bandwidth in an efficient manner while at the same time maximizing the data transmission capabilities of the independent stations subscribing to the satellites facilities. Multiaccess Control (MAC) protocols are designed for such a purpose. There are a number of MAC protocols which have been developed for various communications networks. Depending upon the type and limitations of the network, each protocol has its advantages and disadvantages as well as unique performance limitations. Much work
has been done developing theoretical formulas which describe the performance of many of these protocols in various controlled environments. Traffic assumptions typically implemented include Poisson input processes and uniform storms traffic. These assumptions become necessary in order the make the analysis tractable.

SLOTTED ALOHA

The slotted Aloha variation of the Aloha protocol is simply that of pure Aloha with a slot timer. The slot size equals $T$, the duration of packet transmission. There are restricted to start transmission of packets only in slot boundaries. Thus the transmission period is reduced to a single slot. In other words, a slot will not be successful if and only if only exactly one packet was scheduled for transmission sometime during the previous slot. The throughput is therefore a function of slots (or probability) in which a single packet is scheduled for transmission.

Figure 2 Collision mechanisms in Slotted CODE DIVISION MULTIPLE ACCESS (CDMA)

Code Division Multiple Access (CDMA) is a multiplexing technique where a number of users simultaneously and asynchronously access a channel by modulating and spreading their information-bearing signals with pre-assigned signature sequences. In this transmission technique, the frequency spectrum of a data signal is spread using a code uncorrelated with that signal. As a result the bandwidth occupancy is much higher than required. The codes used for spreading have low cross correlation values and are unique to every user. This is the reason that a receiver, which has knowledge about the code of the intended transmitter, is capable of selecting the desired signal.

Consider a 13-85 (Direct Sequence Spread Spectrum) multiple access system with BPSK signaling and correlation receiver as shown in Figure 3. There are M users in the system. The ith user power received at the base station $s_{ij}$ is:

$$s_{ij} = 2B_0 P_i (\tau - \tau_j) \exp (-\tau_j) \exp (j\theta_0 \tau_j),$$

where $P_i$ is transmitted power of the signal, $C_i$ ($1 - C_i$) represents the data signal, $\tau (\tau - \tau_j)$ represents PN (pseudo-random) sequence, and $\exp (-\tau_j)$ represents the modulating waveform. The $\tau_j$ is the random phase which is uniformly distributed over $[0, 2\pi]$, while $\theta_0$ is the random delay which is uniformly distributed over $[0, T_0]$. $T_0$ is the duration of the delay bit and $T_0$ is the duration of each chip in the PN sequence. The total number of chip per bit $N$ is given by $N = T_0 / T_0$ for where $T_0$ is the chip rate (in chips/second) and $T_0$ is the data bit rate (in bit/second) respectively [1].

Assume that $P_i$ is independent for each user and independent of $B_j$ and $\theta_j$. As shown in Figure 3, the data signal is given by $s_{ij} = \sum_{j=1}^{M} B_j P_i (\tau - \tau_j)$ where $B_j$ is

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\[ c = \{-1, +1\} \] is an infinite sequence of data bits and \( p(t) \) is a rectangular pulse wave form with unit rise time and duration \( T \). The spreading signal is given by:

\[
G(t) = \sum_{i=0}^{L-1} a_i p(t - iT)
\]

where \( a_i \) is \( -1, +1 \). Assume that we receive signals from user 0 and \( t = 0 \) and \( t = L \). When a single propagation path exists, the received signal is given by:

\[
x(t) = G(t) = \sum_{i=0}^{L-1} a_i p(t - iT)
\]

where \( x(t) \) is a 

**MC-CDMA/A-ALOHA Hybrid MAC Protocol**

This hybrid MAC protocol is a combination of Multi carrier CDMA and Slotted ALOHA. Before discussing about this combination its better to get some idea of multi carrier CDMA.

- **Figure 3:** MC-CDMA Transmitter - Receiver

Code Division Multiple Access (CDMA) is a FH or DS spread spectrum system in which a number of users in a network of spread spectrum signals communicate simultaneously, each operating over the same frequency band. In a CDMA system each user is given a distinct sequence. This sequence identifies the user. For example, if user 1 has a sequence \( S1 \) and user 2, a sequence \( S2 \), etc., then a receiver desiring to listen to user 1 will receive at its antenna all of the energy sent by all of the uses. However, after de-spreading user 1’s signal, it will see all the energy of user 1, but only a small fraction of the energy sent by users 2 and 3, for example.

In case of Multi carrier CDMA, the transmitter transmits the original data stream over different sub carriers using a given spreading code in frequency domain. In other words, a fraction of the symbol corresponding to a chip of the spreading code is transmitted through a different sub carrier. So each transmitter will have separate spreading code and modulation frequency. These modulated signals are then transmitted.

At the receiver the signal from each transmitter is demodulated with the same frequency (used in transmitter) and the demodulated signal is then decoded using the corresponding PN code. The schematic figure is shown above.

In MC-CDMA/A-ALOHA we combined the MC-CDMA with Slotted ALOHA. The following manner. The modulated signal that obtained at output of MC-CDMA transmitter is put into the channel in Slotted ALOHA manner. As we know the Slotted ALOHA shows better performance at lower arrival rates and MC-CDMA shows better performance at high arrival rates. So in this hybrid protocol, the performance is almost constant for any arrival rates.

**SIMULATION AND RESULTS**

Here we consider three different PN code and three different frequencies. When a system wants to communicate it randomly selects a PN code (out of available three) for spreading and a frequency (out of available three) for Modulation. The
packets are put in to the medium in slotted ALOHA manner. Whenever more than one user uses the same frequency and PN code in a single slot, a collision will occur. The graph plots in the X-axis, the offered load to the medium and in the Y-axis, the throughput.

In this simulation we consider the system model as shown above. Here we consider a single cell of radius 2000 meters. Each user is free to move anywhere within this cell. This simulation is working for a time period of 3600 micro seconds. During this period each mobile inside the cell is assigned three instances as by the exponential distribution and they will communicate as the instances arrive. After each 3600 micro seconds the arrival rate of packets (offered load to the channel) is changed by changing the mean of exponential distribution. The total attempts made by all the computers and the total successful packets are stored for each arrival rates. During this period of simulation (i.e. 3600 micro seconds) users are free to go out or come inside the cell. Only users within the cell are considered. The graph plots in the X-axis, the offered load to the channel and in the Y-axis, the throughput.

RESULTS OF MC-COMA IS ALOHA

Cell with 10 Users: Here the cell consists of ten users. The users can go out of the cell and come in the cell. The cell radius is 2000 meters. The result of the simulation is given in the figure 4.

Cell with 20 Users: Here the cell consists of twenty users. The users can go out of the cell and come in the cell. The cell radius is 2000 meters. The result of the simulation is given in the figure 5.
CONCLUSION

In this paper, the MC-CDMA/SALOHA MAC protocol of a wireless LAN is studied. Simulations are done using MATLAB. This simulation can be used for studying efficiency of MC-CDMA/S-ALOHA MAC protocol for any number of users. The advantage of using this protocol is that we can reduce the number of PN codes required for the MC-CDMA protocol. This protocol shows better performance as both lower and higher arrival rates.

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RECTANGULAR DIELECTRIC RESONATOR ANTENNA FOR 2.4 GHz Wi-LAN APPLICATIONS

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Abstract: A Microstrip line excited rectangular Dielectric Resonator Antenna (DRA) suitable for the 2.4 GHz ISM band Wi-LAN applications is described. Experimental results for the reflection and radiation characteristics are presented. The antenna exhibits 4.7% 2.1 GHz VSWR bandwidth, broad radiation patterns and a gain of +6.05 dB. Numerical analysis of the antenna using the FDTD method is also presented.

INTRODUCTION

Antennas for Wi-LAN applications ought to be compact, exhibit wide-impedance bandwidth and omni-directional radiation pattern. Dielectric Resonators (DRs) have proved themselves to be ideal candidates for antenna applications by virtue of their high radiation efficiency, flexible feed arrangements, simple geometry, small size and the ability to produce different radiation patterns using different modes [1]. However, the high Q factor restricts the bandwidth, which limits its usefulness as an antenna.

Though the rectangular shape has the advantage of decreased cross-polarisation over other shapes, among the various conventional geometries investigated till date, the Rectangular Dielectric Resonator is the one the least dealt with. The Microstrip feeding technique was applied to rectangular DRAs by Mouni et al. [2]. Experimental investigations were performed on a Microstrip line fed compact, high Q, possibility (εr = 45) Rectangular Dielectric Resonator Antenna (DRA) by S. M. Abdalla et al. [3]. To ascertain the influence of the feed line parameters on the reflection and radiation characteristics, this paper proposes a Microstrip line excited Rectangular Dielectric Resonator Antenna (DRA) suitable for the 2.4 GHz ISM band Wi-LAN applications.

ANTENNA GEOMETRY

The geometry of the proposed antenna configuration is shown in Figure 1. The Dielectric Resonator is excited directly by a 50 Ohm Microstrip Line of length L = 3 cm and width W = 0.3 cm. The feed line is fabricated on a substrate E - 10.1 cm. W = 0.19 cm with dielectric constant εr = 4.25 and thickness 0.14 mm, backed by a ground plane. The DR is proposed by the conventional solid-state ceramic route. Table 1. summarises the properties of the DR.
### Table 1. Microwave Dielectric properties of the Dielectric

<table>
<thead>
<tr>
<th>Composition</th>
<th>Dielectric Constant (εr)</th>
<th>Sintering temperature (°C)</th>
<th>Dimensions (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>BaTiO_3</td>
<td>34</td>
<td>1300, 1540 for 3 hours</td>
<td>0.85 x 1.2 x 1.87</td>
</tr>
</tbody>
</table>

**Figure 1. The Microstrip Line Fed Rectangular Dielectric Resonator Antenna**

**RESULTS AND DISCUSSION**

The reflection and radiation characteristics of the proposed antenna configuration are measured using HP 8513C Network Analyzer. The measured return loss is plotted in Figure 2. (a). The 2.3 dBiV WR bandwidth of the proposed antenna configuration excited at 2.455 GHz is 115 MHz. A 4.7% in the 2.39 GHz - 2.555 GHz band. The reflection characteristics of the antenna configuration under study has also computed theoretically using the FDTD method.

![Reflection characteristics of the proposed antenna configuration](image)

(a)  

![Gain of the proposed antenna configuration](image)

(b) Gain of the proposed antenna configuration

The code is written using MATLAB, a Gaussian voltage source with a source resistance of 500 Ω is used. A Gaussian pulse of half-width 7.15 μs and time delay 0-3 μs is applied at the source plane. The width of the pulse and the time delay are chosen to suit the frequency bands of interest. The time step used is Δt= 1.355 ps, satisfying the Courant stability condition. The cell size chosen is

![Figure 2](image)

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accommodate more than 70 cells per wavelength within the DR and the substrate is $a_1 = 0.5\text{mm}$, $a_2 = 0.5\text{mm}$ and $a_3 = 0.4\text{mm}$. Mu's First order Absorbing boundary conditions are used to terminate the computational domain, with 10 space steps distance from the object to the absorbing boundary in each direction. The iterations are performed for 5000 time steps to allow the input excitations to converge to zero within the computational domain. The input impedance of the antenna is computed as the ratio of the FFT of the voltage derived from E field values at the observation point over the entire time steps, to the FFT of the current at the same point, derived from the H field values. Reflection Coefficient S11 (in dB) is then computed. The measured gain characteristics of the DRA is shown in Figure 2(b). The antenna exhibits 6 dBi gain.

Figure 3. shows the measured radiation pattern of the antenna. The 3 dB beam width in the E and H plane are 92° and 117° respectively. The broad patterns indicate the suitability of the proposed antenna configuration for mobile communication applications. The intensity of the field components in the three directions suggest the presence of the TE(316) mode excited at 2.455 GHz with one half cycle variation in

![E-plane](image)

![H-plane](image)

Figure 3. Measured radiation pattern of the proposed antenna configuration at 2.455 GHz

- co-polar
- cross-polar

The field distribution within the Dielectric Resonator obtained through simulation using Microstripes® is shown in Figure 4.

![Field distribution](image)
CONCLUSIONS

A wide band Rectangular Dielectric Resonator Antenna suitable for W-LAN applications is presented. The antenna is compact, simple in geometry and easily excited by a 50Ω Microstrip Line at a frequency of 2.45 GHz. The proposed configuration exhibits a 2:1 VSWR bandwidth of 4.7 %. The co-polarization along the end-ads in the E-plane and H-plane is -23 dB and -14 dB respectively. The radiation patterns are broad in the principal planes. These features make the antenna highly suitable for W-LAN applications.

REFERENCES


A CYLINDRICAL DIELECTRIC RESONATOR ANTENNA EXCITED BY CONDUCTOR BACKED CO-PLANAR WAVEGUIDE

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Abstract: The radiation and resonance characteristics of cylindrical dielectric resonator antenna excited by a conductor backed coaxial waveguide (CB-CPW) is presented in this paper. This proposed antenna configuration offers a 2.1 VSWR bandwidth of 13% with consistent gain and cross-polar level.

INTRODUCTION

Dielectric resonator antennas (DRA) are increasingly popular as they have many advantages such as small size, low profile, and lightweight. However, DR is devoid of conductor losses and low efficiency due to surface wave excitation [1]. Fortunately, CPW-based waveguides (CPW) have also become very popular transmission lines used in high frequency [2] for advantages such as parallel and series interconnections of both active and passive components, low radiation leakage, low dispersion, and conductor losses up to millimeter wave frequencies. CPW is very suitable and competitive for active integrated antennas. The experimental results of cylindrical dielectric resonator excited using a CHCPW is presented in this paper.

ANTENNA DESIGN

Cylindrical DR with high dielectric constant, $e_r = 48$, Diameter D = 24.15 mm and height h = 6.88 mm prepared from low loss ceramic material Cr5Zn2O12 [3] as single phase by solid-state ceramic route is excited using a CB-CPW as is shown in Figure 1. The CB-CPW is fabricated on substrate of thickness $h = 1.6$ mm and permittivity $(e_r = 4.7)$ with $2G + S = 24.15$, the diameter of the DRA.

![Figure 1: Geometry of the cylindrical dielectric resonator antenna.](image)
The position of the DRA on the feed line is optimized for maximum bandwidth and gain. It is observed that at the optimum position the antenna offers a 2.35

bandwidth of 13% at 2.5 GHz. Figure 2 shows the variation of the return loss of the antenna at the optimum position. The gain of the proposed antenna configuration is measured using gain transfer method. The Cylindrical DRA offers a gain of 6.38 dB at 2.525 GHz. The operating range of the antenna is 350 MHz from 2.43 GHz to 2.765 GHz. The gain of

the antenna is shown in figure 3.

Figure 2: Variation of return loss with frequency

Figure 3: The gain of the DORA on CBCPW

\[ G=34.15 \text{ mm, } \theta_{a}=48, \theta=24.15 \text{ mm } H=6.81 \text{ mm and } t=1.6 \text{ mm } \epsilon=4.7. \]
The radiation patterns of cylindrical dielectric resonator antennas at resonant frequency are shown in figure 4. From the figure it can be inferred that antenna offers a cross-polar level better than -30 dB.

Figure 4: Radiation patterns of the Cylindrical DRA at 2.525 GHz.
20/48: 24.15 mm, ea = 48, d = 24.15 mm H = 0.81 mm and H = 1.6 mm e = 4.7.

CONCLUSIONS

Conductor Backed Coaxial Waveguide is demonstrated as an effective excitation technique for a Cylindrical Dielectric Resonator Antenna. Position of the Cylindrical DRA for maximum bandwidth, gain and cross-polar level is optimized on the CS-CPW.

REFERENCES

A new compact coplanar antenna configuration suitability for active antenna application is prepared. The proposed antenna exhibits an impedance bandwidth of 6% and average gain of 5.5 dB with good radiation coverage. A prototype of the proposed antenna is constructed and studied. The coplanar geometry of the antenna eliminates the need of via in active antenna design.

INTRODUCTION

Active integrated antennas is a growing area of research in recent years, as microwave integrated circuits and monolithic microwave integrated circuit technology have become more mature. Several active integrated antennas have been reported in literature [1-4]. Most attracting feature of a microstrip antenna is its planar geometry, and this feature enables the easy integration of passive and active microwave components to the antenna element. Several compact microstrip antennas have already been reported in literature [1-4]. But the major disadvantage of a microstrip antenna is its ground plane. The presence of ground plane limits the radiation coverage to the half space. Normally in microstrip active antennas the ground connections for the integrating circuits are achieved by using via to the ground plane from the top radiating patch through the substrate. Coplanar transmission line based circuit designs received more attention over microstrip line circuits recently due to its planar geometry with coplanar ground planes on the opposite layer [5]. A coplanar transmission line together with microstrip transmission line feed configuration having good radiation has been proposed in [6]. In this paper we present the experimental investigation of a compact coplanar antenna (CCPA) with excellent radiation coverage, comparatively good impedance bandwidth and gain at 1.8 GHz band. The excellent radiation and radiation characteristics together with its coplanar topology suggest it as an element for active antenna with more design freedom.

ANTENNA GEOMETRY AND DESIGN

Figure 1 shows the fundamental structure of the CCPA. The CCPA consists of three coplanar copper strips (two ground planes and a center conductor) without bottom ground plane. The linear strip is within the two ground conductors. The strip is etched on a substrate of dielectric constant $\varepsilon_r = 4.48$ and height $h = 1.6$ mm. Center conductor is having length $L = 0.038$ λ and width $W = 0.073$ λ, where λ is the free space wave length at the operating frequency. There are two radiating slots $sl1$ and $sl2$, which separates the central conductor from the two ground planes. Dimension of the slots are $L_s = 0.087$ λ and width $g = 0.063$ λ. An SMA connector
with two copper wires extending from the outer ground conductor of the SMA serves as the test fixture for the measurements. The antenna is fed through the radiating slots using a coaxial probe. The optimum feed point of the antenna is point B along the side a-b of the centre conductor. The other two patches are grounded exactly at the centre of L2 and L1 respectively as shown in Figure 1. Better input matching can be achieved by varying the feed point B along the side a-b of the centre conductor. The overall dimension of the present antenna is only L = 14.4 mm and W = 32.9 mm, where a conventional rectangular patch antenna resonating at 1.8 GHz has a dimension of L = 40 mm and W = 52 mm. Thus the antenna has 77.6% size reduction compared to a conventional rectangular patch antenna.

Figure 1. Proposed CPA geometry

RESULTS AND DISCUSSION

Figure 2 shows the experimental and simulated return loss of the proposed antenna. The antenna resonates at 1.79 GHz with 32 dB return loss, 2.4 VSWR band of the antenna is from 1.7535 GHz to 1.867 GHz (14 MHz). Figure 3 shows the measured E and H plane radiation patterns of the antenna at 1.79 GHz. The E-plane radiation pattern in Figure 5 (a) is figure of eight and the H-plane pattern is nearly circular as shown in Figure 3 (b). From the cross polar levels of the two principal plane patterns it is observed that the polarization purity of the antenna is poor. Figure 4 shows the measured antenna gain with in the 10-18 GHz return loss band. The antenna offers an average gain of 3.3 dBi in the operating band.
CONCLUSIONS

A compact coplanar antenna configuration suitable for active antenna application has been proposed and its characteristics have been studied by experiments. The proposed antenna has 77% area reduction compared to a
REFERENCES


AN X-HAND RECTANGULAR WAVEGUIDE CORRUGATED FIN ANTENNA FOR BEAM-SLITTING APPLICATIONS

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A phase-shifter free beam slitting device capable of one-dimensional beam-scan is designed using an x-hand wave-guide antenna array. The performance of the device operating at 10 GHz is simulated using an electromagnetic simulator HFSS and verified by experiment. Applications of such a waveguide device range from military through to commercial mobile wireless communications systems where directed beam-forming capabilities are needed. The variation in the beam slitting ability as a function of waveguide fin separation is also simulated.

INTRODUCTION

Several types of wave-guide antennas have been in wide-spread use for various communication applications in which fixed beams are generated [1]. Slotted waveguide techniques (such as T-slot, X-slot, and L-slot) and loopy wave-guide loaded rectangular waveguides are some of these [2]. A variety of approaches have been used to analyze mutual coupling in a finite array of rectangular waveguides arranged on a rectangular grid. In calculating the self and mutual admittances for mode coupling, a quadruple integral over the source and observer apertures is involved. Here in this paper a waveguide antenna operating at 10 GHz capable of one-dimensional beam slit is presented. The design is easy to fabricate and can be easily attached to feed assemblies.

PRINCIPLE OF OPERATION

A waveguide antenna apparatus operating in the x-hand is simulated using Ansoft's HFSS and its radiation properties are analyzed. It consists of a periodically positioned set of corrugated fins aligned along the ends of an open waveguide along a flat plane [4]. Through mutual coupling the non-fed waveguides constitute separate radiation elements and due to the interaction between these radiation elements, a directional beam can be formed in a particular direction.

The beam forming ability can be improved by changing the spacing between the elements without significantly affecting the gain. The basic design is depicted in figure 1.
The directed beam formation is also dependent on the distance from the feed point to the open end of the waveguide.

The corrugated fins have periodically changing heights and this causes the phase adjustment of the wave pattern in the radiating end accordingly, the direction of the beam to be transmitted.

A waveguide block followed by an array of corrugated fins constitutes the proposed design. As the design is a single positive antenna device, on account of the reciprocity principle the same properties apply to the receiving situation also. Hence in this paper only the transmitting situation will be referred to.

SIMULATION AND EXPERIMENTAL RESULTS

The waveguide apparatus is simulated using the Ansoft High Frequency Structure Simulator (HFSS) [5]. The return loss of the device was found to be better than 20 dB almost across the entire band for fixed fin spacing and is experimentally verified.

The corrugated waveguide structure is excited with a TM10 wave source. The gain of a standard horn antenna operating at this frequency range is calculated using standard equations [6]. The device provides maximum possible gain with the given Waveguide aperture dimensions (W=2.2 cm, B=1 cm, Standard X-band waveguide). For fabrication the value of the $g$ and $t$ were computed to be 0.07 λ and 0.2 λ, respectively. A ground plane minus the substrate forms the array structure. The heights of the fins were varied gradually in a specific proportion of the wavelength.

By analyzing the characteristics of the transmitted signal it was concluded that the mutual coupling is not purely depending on the spacing of the fins. So the simulations were repeated for various fin heights of $d$ by maintaining the gaps between the fins $g$ constant.

In the simulator the fin height $d$ was gradually increased in steps and it was seen that the wave guide with the fin height above a particular cut-off region is behaving as a beam-varying device and a beam lift of the order of 30° can be achieved at the optimum fin height of 1 cm. The measurement set-up for S21 is created.

At bore-sight the analyzer displays a low level signal as expected as the entire beam is redirected at a small angle. Now the receive antenna is gradually moved
along a circle containing the maximum of the beam from the waveguide a sensor until the network analyzer picks up the highest signal level. The angle of deviation from the bore-sight to this maximum point gives the measured beam tilting angle.

The received signal at the network analyzer for various angular positions is recorded as shown in figure 2 which is compared with simulated values. The simulation results also show that by slightly adjusting the spacing ‘g’ between the corrugated elements the beam forming direction can be varied.

![Figure 2: Measured and Simulated angular beam-tiltation](image)

**Fig. 3. Measured radiation patterns of the apparatus**

From the angular beam-tiltation analysis it was observed that the major beam is centered at 30° from bore-sight. The radiation pattern of the waveguide apparatus is also plotted by setting up test apparatus as the transmitter and a standard x-band horn antenna as the receiver. Before commencing the measurements the transmitter is aligned at 30° to the x-band horn so as to record the main beam at 0°. The resultant patterns for two spacings (g) are recorded and are shown in figure 3.
CONCLUSIONS

A waveguide antenna is designed and measured to study its beam redirection capabilities. The device, operating at 60 GHz, is having more than 15.9 MHz operational bandwidth. The edge diffraction from the device aperture is much lower, owing to the corrugated structure and the gradually increasing height of the bars. Single mode operation of the device enables the possibility of a polarizer to be integrated directly in to the structure. HFSS simulation predicts a 9.4 dB gain for the device, measured to be 7.3 dB. The beam-redirection capability of the apparatus makes it suitable for applications in mobile wireless communication systems for reducing multi-path fading effects.

ACKNOWLEDGEMENT

Thanks are due to Prof. V.P. Fursikov, Professor, Queens University of Belfast for the basic idea behind this waveguide apparatus during my research tenure at QUB under his supervision.

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RESEARCH SESSION 6
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5.1 Design and Development of Harmonic mixer at Ka Band
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Defence Electronics Applications Laboratory, Dibrugarh-786004. Phone: 2511762, Fax: 2511763
Second Generation Ka Band Converter by using MEMS

5.2 Ka-Band Frequency Tripler
Satish K. Jindal, S.K. Jindal and E.S. Saxena
Defence Electronics Applications Laboratory, Dibrugarh-786004. Phone: 2511762, Fax: 2511763

5.3 A 1.8-40GHz Coplanar waveguide MEMS chunt switches
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6.1 Computer-Aided design procedure of microwave mini linear circuits
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6.2 Design and Analysis of compact circular to rectangular waveguide transition at 94 GHz
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6.3 A Compact OM& Design For A Ka-band Biconic Antenna
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6.4 Miniaturized Unicaps of a Micromachined Phase Shifter Using Modified Ground Coplanar Wave Guide
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Proceedings of APSYM 2004, Dec. 21-23, Dept. of Electronics, CUSAT, Cochin, INDIA.
Design and development of an inexpensive, high power LO sources is one of the major challenges for researchers. The power available from solid-state sources drops off with the inverse square of frequency due to technological limitations in the material, and hence, at higher frequencies, noble LO power sources come in a much higher cost. Multipliers are generally used to overcome the above problem. Unfortunately, with increasing radio frequency multiplier becomes more and more complicated and expensive, and also noise behaviour of the local oscillator deteriorates.

Therefore, one of the main goals of mixer design has been the reduction of LO power requirements, with emphasis towards receiving configurations that permit harmonic mixing (2, 3, 4). The advantages of harmonic mixing include the diode's higher conversion loss compared to fundamental mixing when a pair of anti-parallel diodes is used in a mixer element. (1,2)

A diode directly coupled to the local oscillator is used. The mixer diode is a Schottky barrier diode which is directly coupled to the LO signal. The power available from solid-state sources drops off with the inverse square of frequency due to technological limitations in the material, and hence, at higher frequencies, noble LO power sources come in a much higher cost. Multipliers are generally used to overcome the above problem. Unfortunately, with increasing radio frequency multiplier becomes more and more complicated and expensive, and also noise behaviour of the local oscillator deteriorates.

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DESIGN APPROACH

The mixer is designed using an unparallel diode pair (DMK 2791) and matching stubs. The schematic is shown in Fig. 1.

Fig. 1 Schematic

(a) The short circuit stub of length $\lambda_g/4$ or $\lambda_i/4$ (right side of diode pair) is used to short the RF signal in the LO side while it presents open to LO signal, hence good LO/RF isolation is achieved. [4,5]

(b) The open stub of length $\lambda_g/4$ or $\lambda_i/4$ (left side of diode pair) is used to short the LO signal in the RF side while it presents open to RF signal. The IF port is coupled to the diodes via a $\lambda_i/4$ or $\lambda_R$ transmission line. If the IF port side of transmission line, there is a shunt $\lambda_i/4$ or $\lambda_R$ open circuit stub. The purpose of this network is to present open circuit to the RF signal so that IF port does not degrade conversion loss by loading the RF signals. The components used in this network are electrically small enough at IF to be essentially transparent at that frequency [4,5].

(c) The capacitor C1 passes the RF frequencies, but presents an open circuit to IF frequencies, so the IF signal is not loaded by the impedance of the RF port.

(d) To remove the 2X harmonic product $\lambda_i/4$ or $\lambda_R$ is included at the side of the diode pair from which the IF is extracted.

(e) LO and RF matching circuits are included to match the diode impedance at the LO and RF frequencies with the LO and RF source impedances respectively. These matching circuits are essential to achieve good VSWR at the RF and LO port, which in turn minimizes the conversion loss.
DESIGN OF MATCHING NETWORK AT RF AND LO

The impedance of the slide is given as:

\[ Z = R + jX \]

where \( R \) is resistance, \( X \) is reactance, \( RF \) is resonant frequency, \( C \) is capacitance, \( R_0 \) is RF source impedance, \( Z_0 \) is characteristic impedance of the line, \( Z_{in} \) is input impedance, and \( Z_{out} \) is output impedance.

SIMULATION

The circuit is simulated and optimized using ADS 2002/Agilent software with RF frequency = 18.05 GHz, LO frequency = 4.31 GHz, RF Power = 20 dBm, and LO power = 0 dBm. The desired IF frequency is 18 MHz (0.0018 GHz).

HARDWARE REALIZATION

The layout is generated from the schematic. The mixer circuit has been fabricated on 10 mil thick FR4 substrate. The fabricated hardware is shown in Fig. 3. Two LNAs have been used in RF and LO ports and SMA connector at the IF port. Two bandpass filters (4THA7 2790) and a capacitor have been used on the circuit for the transmission bandpass process.

Fig. 4 Spectrum at IF port

Fig. 5 CH300A Vs. 1 Opamp

Fig. 6 Layout of designed mixer

Fig. 7 Photograph of fabricated mixer
Test result for conversion loss and LO to RF isolation is shown below:

**Table 1**

<table>
<thead>
<tr>
<th>LO Frequency (GHz)</th>
<th>RF Frequency (GHz)</th>
<th>RF to LO Conversion Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>2</td>
<td>0</td>
</tr>
<tr>
<td>2</td>
<td>3</td>
<td>1</td>
</tr>
<tr>
<td>3</td>
<td>4</td>
<td>2</td>
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<tr>
<td>4</td>
<td>5</td>
<td>3</td>
</tr>
<tr>
<td>5</td>
<td>6</td>
<td>4</td>
</tr>
</tbody>
</table>

**LO:7 RF ISOLATION**

With Rf=7 dBm, the LO/RF at RF port was found to be -22.82 dBm hence LO to RF isolation of the circuit is 29.82 dB. The worst value of isolation was 24.3 dB at 19.86 GHz.

**CONCLUSION**

A harmonic mixer at Ka band has been fabricated on the RT-durid substrate for a conversion loss of 12.2 dB at 19.65 GHz with a bandwidth of 0.5 GHz. The mixer exhibits LO to RF isolation -30 dB and VSWR is less than 1.6 for the entire RF band. The difference between the test result (12.3 dB) and simulation result (9.49 dB) is mainly due to fabrication tolerances and bonding errors. This harmonic mixer was used for the development of phase locked oscillator at Ka-band frequency.

**ACKNOWLEDGEMENT**

The authors express their sincere gratitude and thanks to Mr. Asok Sen, Executive, DEAL for his kind permission to publish this paper. The authors are also grateful to MIT and CAMEL group for fabrication of the circuit. Finally, authors are highly obliged to Mr. A.K. Shukla and Mr. Ashok Kumar for giving proper guidance at every stage of development of the component.

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Abstract: This paper describes the design and development of a Kia-band frequency tripler (9.55GHz/9.56GHz) in microstrip line configuration using schottky barrier beam lead diodes. This circuit structure has been simplified and optimized using computer aided design tools (ADS from HP). The circuit has been realized on 10µm thick RT-Duriod (RJ=2.22) substrate. Experimental results show conversion loss of 15 dB and harmonic suppression greater than 50 dB including fundamental frequency.

INTRODUCTION

One of the most critical problems encountered by millimeter wave communication systems is the signal generation, as the complex source design must ensure a good stability and low phase noise level. An effective solution is to associate a low phase noise oscillator at a lower frequency with frequency multiplier. The present design uses diode as passive multiplier, which gives excellent performance in terms of phase purity. But the Field Effect Transistor (FET), High Electron Mobility Transistor (HEMT) can be used as an active multiplier which provide gain rather than loss and give more output power. Since the frequency conversion occurs by forcing these devices to operate in the triode nonlinear region and to generate harmonic signals, it is difficult to obtain relatively high output power at desired harmonic frequencies. Foster single-ended or hybrid architectures are typically used for multiplier design. Single-ended structures provide low cost and narrow bandwidth characteristics in a simple design but poor performance. In contrast, balanced or hybrid type multipliers with input or output baluns either broader bandwidth and higher output power than single-ended. However we are using single-ended type multiplier.

ANALYSIS AND DESIGN

The basic problem in designing a multiplier is to achieve sufficient suppression of unwanted harmonic signal at the output port. Circuits composed such as quarter wavelength stub can be used in the output port, which short the fundamental frequency to prevent it from appearing at the output port. Similarly using it in the input port which acts as short circuit for the fundamental frequency provides good isolation. The multiplication is done by and parallel combination of diodes as shown in Fig. 1. "High order" multiplier (i.e. N higer) can be very difficult to design for good performance, good isolation and many multipliers must be cascaded. Diodes have been traditionally employed for implementation of higher order multipliers.

Successful development of a tripler circuit hinges on proper design and careful modeling of the schottky diode, which determine the intrinsic efficiency conversion loss and the minimum output power that the tripler can achieve. The high frequency model of a schottky barrier diode consists of its in series with barrier resistance (Rb) and capacitors (Cj, C1, C2). The series resistance Rb consists of voltage dependent impact ionization, spreading resistance and contact resistance.

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A matched pair of GaAs schottky barrier beam lead diodes with series resistance ($R_{s}$) of 6 ohm and zero bias junction capacitance ($C_{J}$) of 0.07 picofarad have been used. To minimise the conversion loss, the diode impedance at input and output frequencies have been matched with 50 ohm filter. Fig. 3 shows the high frequency equivalent circuit of the diode.

Fig. 2 Equivalent circuit of the diode

From the equivalent circuit the diode impedance is given by

$$Z_{D} = R_{L} + j \omega C_{J}$$

Where $R_{L}$ = junction resistance, $R_{s}$ = series inductance, $C_{J}$ = junction capacitance. Substituting $R_{L}$=6 ohm and $C_{J}$=0.07 picofarad the diode impedance at input frequency 9.55 GHz in ZIN = (13.86+34.39j) ohm and at output frequency 28.05 GHz in ZOUT = (17.479-12.3j) ohm.

The diode equivalent impedance has come out one half of the individual diode impedance because of antiparallel diode pair used. Microwave matching network is designed to match diode impedance with 50 ohm impedance line using Smith chart.

To select third harmonic, band pass filter is used in the output which also gives extra suppression to unwanted frequency. A parallel coupled Butterworth band pass filter has been designed using insertion loss method. The specification for the filter design have been taken as center frequency 28.05 GHz, ripple in pass band ($S_{1}$) = 0.25 dB, bandwidth=1.5 GHz and...
CIRCUIT SIMULATION AND HARDWARE REALIZATION

The designed tripler circuit is simulated using nonlinear Harmonic Balance simulator (HB simulation) of Advanced design system. The simulation results with 0 dBm input power are given in the fig. 3. The result shows 13.98 dB conversion loss for third harmonic and above 100 dB for other harmonics. All the above designed components (matching circuit, hpf) are integrated on a single substrate along with diodes. Then, the tripler circuit has been fabricated on 10-mil thick RT duroid substrate. The realigned hardware is shown in fig. 5. SMA connector is used in the input and K-type connector in the output. Burned diodes have been bonded on the circuit by thermo sonic bonding technique.
EXPERIMENTAL RESULTS

For any variable-frequcy multiplier if power (Pm=0) is delivered to the sources, let \( P_{in,0} \) represent harmonic power supplied by the device, the maximum theoretical conversion efficiency is given as \( \eta_{th} = \frac{P_{in,0}}{P_{in,0} + P_{out}} \). When \( m \) is the multiplication order, this can be shown to be 100% equal conversion loss at sine wave as a input with minimum power 3 dBs. The triple circuit satisfies the power level at 0 dBm. The quantum efficiencies are greater than 90% for fundamental, second and fourth harmonic. Regarding phase noise to noise power in the baseband, the noise levels are increased by the multiplication factor as frequency multiplication is effectively a phase multiplication process. The increase in noise level is given by 20log\( n \), where \( n \) is the multiplication factor. Thus a frequency tripler will lead to an increase of at least 9.54dB to noise level. In our circuit we achieved excellent phase noise performance and increase in phase noise was observed only 9.05dB.

CONCLUSION

Diode-based frequency multiplier (9520/65GHz) has been designed, fabricated and tested at Ka band frequency. Its major advantages are effective suppression of the unwanted fundamental, even order harmonics and minimum conversion loss to the desired third harmonic. This has been demonstrated by computer simulation and further verified by measured results for this triple circuit. The absence of biasing, tuning greatly simplifies the use and measuring of schottky barrier diode in tunable line make it also easier to fabricate. This frequency tripler has been used in development of Ka-band phase locked oscillator.

ACKNOWLEDGEMENT

The authors express their sincere gratitude and thanks to Mr. Anuk Siv, Director IITAL for his kind permission to publish this work. The authors are also grateful to M.I.C and CAMI group for fabrication of the hardware. The authors wish to thank M.E.K. Dulka and Mr. Ashok Kumar for their initiatives and technical assistance.

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A 18-40 GHz COPLANAR WAVEGUIDE MEMS SHUNT SWITCH

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ABSTRACT: In this paper, we discuss about the design of high isolation, low insertion loss shunt switch using MEMS. As a first step single shunt switch has been designed and was analyzed using EMDOCS. Using the simulated results, model was extracted for single shunt switch. In the second step, shunt switch for high isolation and low insertion loss was designed by using two single shunt switches connected by a high impedance line to reduce the insertion loss and inductive loading was also incorporated to achieve high isolation.

INTRODUCTION

Silicon-electro mechanical actuators (MEMS) have several advantages such as low DC power consumption, large down-to-up state compliance ratios, low intermodulation products and can be fabricated on any substrate. The monolithic actuator switching speeds and high actuation voltages (15-80 V). However, these shortcomings can be traded in a low-loss high isolation telecommunication switches and radars with low scanning rates.

Switches with high isolation and low loss at high frequencies are required in several applications. This requires an extremely careful analysis and design.

MEMS SHUNT SWITCH FABRICATION

The single MEMS switch is implemented on a 400 μm thick high resistivity Si substrate (ρ = 11.9) covered with 600μ A-SiO2. The CPW layers (Cu/W, 600 100-60 μm) are defined by the layers of 300° A-Ho/SiO2/A-Ti/Au layer. Plasma enhanced chemical vapor deposition of SiNx layer is carried out for a thickness of 0.15μm. A 1.35μm thick sacrificial layer of photomask is deposited and patterned. The thickness of this layer determines the nominal gap height go of the membrane. The sacrificial layers are removed using TMAH, as the element. An Au layer of 0.2μm is used to form the bridge layer. Typical pull down voltage for these switches is 12 ~ 35 V.

Fig.1 shows the up state and top view of the shunt switch. Fig.2 shows the equivalent model of the shunt switch.
MEMS SHUNT SWITCH MODEL

The parallel plate capacitance of the MEMS shunt switch is

\[ C_{\text{par}} = \frac{\varepsilon_0 \alpha W W}{S_{\text{eff}}} + \frac{L \varepsilon_0 f}{d} \]
The second term in the denominator is due to the thin dielectric layer of SiC. For a dielectric thickness of 0.05mm and a dielectric constant of 3.6 and a bridge height of 1.5mm, the parallel plate capacitance is 517 pF. The Down-State capacitance is given by  
\[
\frac{C_{DL}}{C_{UP}} = \frac{1}{1 + \frac{d_{SiC}}{C_{UP}}} 
\]
where  
\[
d_{SiC} = 0.05 \text{mm} \times 0.5 \text{mm} 
\]
for a capacitance area of 80 mm x 820 mm and a bridge height of 1.5 mm.

**SIMULATION RESULTS**

Simulation of shunt switch has been carried out using 3D full wave solver. Fig.3 shows the insertion loss and return loss of the switch in the up state. The insertion loss is about 6.1 dB at 40 GHz. Fig.3 shows the insertion and return loss of switch in down state. The isolation is about 34.0 dB at 40 GHz.

![Return Loss](image1.png)

![Insertion Loss](image2.png)

Fig.3. Return Loss and Insertion Loss of switch in up state.

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From Fig 3 we see that with a single shunt switch the isolation at low frequencies is very poor. This is due to finite value of capacitance which results in high impedance at low frequencies and a short at high frequencies. Also the insertion loss is about -0.5 dB at 40 GHz and the Return Loss is about -7.5 dB at 40 GHz. To improve the performance, band pass filter characteristics have to be incorporated. This can be implemented by using two single shunt switches separated by a high impedance line and by incorporating inductive tuning. Fig 5 shows MEMS shunt switch for 1- 40 GHz.

![Figure 4: Isolation and Return Loss of a switch in the off state.](image)

**Fig. 4.** Isolation and Return Loss of a switch in the off state.

**Fig. 5.** Proposed implementation of two MEMS bridge switches.
Simulation of Two MEMS shunt switch has been carried out using 3D full wave solver.

Fig. 6 shows the insertion loss and return loss of the switch in the up state. The insertion loss is about - 0.42 dB and return loss is about -14 dB at 40 GHz. Fig. 7 shows the isolation and return Loss of the switch in down state.
CONCLUSION

Design procedure of high isolation and low insertion loss switch using two MEMS bridges separated by high impedance line and by inductive tuning has been discussed. The performance results show that isolation has improved and the insertion loss could be reduced to a considerable extent.

REFERENCES


COMPUTER-AIDED DESIGN PROCEDURE OF MICROWAVE NON-LINEAR CIRCUITS
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ABSTRACT
This paper presents a detailed CAD procedure as applied to non-linear microwave circuits with a
emphasis on the use of simulation engines employed in the evaluation of performance parameters. It
also provides a general idea of the CAD procedure and brings out the scope of the differences in
the design steps of some of the nonlinear circuits.

INTRODUCTION
All Electronic circuits are nonlinear; this is a fundamental truth of cimputer
engineering. The linear assumption that underlies most modern circuit theory is in practice
only an approximation. Some circuits, such as analog amplifiers, are only very weakly
nonlinear, however, and are used in systems as if they were linear. In these circuits, nonlinearities are minimized. Other circuits such as mixers, frequency multiplier, oscillators
exploit the nonlinearities in their circuit elements; these circuits would not be possible if nonlinearities did not exist. In these, it is often desirable to maximize (to some zero) the

Effect of the nonlinearities, and even to minimize the effects of any linear phenomena. The
problem of analyzing and designing such circuits is usually more complicated than for
linear circuits. This article deals with the computer-aided design procedure of microwave
nonlinear circuits like mixers and oscillators.

NONLINEAR CAD
In order to be useful, CAD techniques must obviously keep pace with technological
realities, which means that conventional circuit-oriented CAD must evolve into modern
system-oriented CAD. This involves the need for nonlinear capabilities, since system
performance always requires nonlinear functions, and the ability to deal with very large size
problems. From this viewpoint, nonlinear circuit CAD marks an essential step toward the
technological update of computer-aided techniques. This clearly establishes the present trends in
nonlinear microwave CAD. On the one hand, we have the circuit design problem, an
intriguing one with several interrelated aspects, a tentative list of which is given below:

- Analytical Simulation of a known circuit;
- Optimization of a nonlinear circuit;
- Multiple-frequency excitation (limit cycle analysis);
- Frequency conversion (mixing);
- Stability analysis;
- Noise analysis;

State-of-the-art computer-aided design (CAD) methods for active microwave circuits rely heavily on models of real devices. The equivalent circuit device models must be based
upon accurate parameter extractions from experimental measured data. The specification of design goals marks one of the essential differences between linear and nonlinear circuits. The
following rows come of the performance indices of non-linear circuits such as power amplifier,
mixer, oscillator and frequency multiplier respectively. They are:

- The output power from a given port at any harmonic;
- The spectral purity of the output signal at a specified harmonic, from a given
The return loss at any port connected with a free RF source.

- The power transfer efficiency from the dc bias sources to the output signal at a given port and harmonic.
- The transducer gain between given input and output ports at specified harmonics.
- The power-added efficiency between given input and output ports.

Many integrated CAD packages contain additional tools to aid in oscillator design. Since an oscillator is a continuous feedback loop, special non-linear signal injection ports are provided for the purpose of calculating loop gain. Table 1.0 gives an overview of some of these CAD packages for use in the design of oscillators.

<table>
<thead>
<tr>
<th>S No</th>
<th>Simulator Engine</th>
<th>Domain</th>
<th>Stimuli</th>
<th>Applications</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>DC</td>
<td>Frequency</td>
<td>Multiple DC well levels</td>
<td>Establish operating curves, trace DC bias curves</td>
</tr>
<tr>
<td>2</td>
<td>Linear</td>
<td>Frequency</td>
<td>Single Small-signal sinusoid</td>
<td>Small-Signal steady-state analysis; Can calculate network parameters; Linear amplifiers</td>
</tr>
<tr>
<td>3</td>
<td>Harmonics</td>
<td>Frequency &amp; Time</td>
<td>Multiple Large-signal sinusoids</td>
<td>Large-Signal steady-state behavior of PA's, Mixers, Oscillators</td>
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<tr>
<td>4</td>
<td>Voitser Series</td>
<td>Time</td>
<td>Multiple Large-signal sinusoids</td>
<td>Account Large-Signal steady-state behavior of PA's, Mixers, Oscillators</td>
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<td>5</td>
<td>Transient</td>
<td>Time</td>
<td>Multiple arbitrary time-varying signals</td>
<td>Oscillator Start-up</td>
</tr>
<tr>
<td>6</td>
<td>Convolution</td>
<td>Frequency &amp; Time</td>
<td>Multiple arbitrary time-varying signals</td>
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<td>Frequency</td>
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<td>Noise Analysis of non-circuit components</td>
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<tr>
<td>8</td>
<td>Envelope</td>
<td>Frequency &amp; Time</td>
<td>Multiple arbitrarily modulated large-signal sinusoids</td>
<td>Non-linear distortion effects</td>
</tr>
</tbody>
</table>
CAD Procedure of Mixer Design

Following is the CAD Procedure that can be implemented in the design of a Mixer. Before that we must consider the following choices:

Choose a mixer type (Active or Passive). An active mixer is one that can provide power (or conversion) gain from RF to IF. The widely used double-balanced (Gilbert or Jones) style of mixer is an example of an active mixer that is provided in most of the CAD tools. A passive mixer is one that can only provide conversion loss. For example, the FET double-balanced ring or diode double-balanced ring mixers are popular examples that are available in the CAD package. It is often easier to obtain large gain compression threshold and intercept point with passive mixers.

Choose an IF configuration (Single-ended or Differential). This is an important design decision. Differential is generally preferred for RFIC applications, as it lends itself to double balancing without the need for passive baluns or transformers. There is also an advantage in rejection of noise coupled through the substrate due to common-mode rejection. However, if the differential interface on-chip is provided by active baluns, the gain compression and intermodulation performance are often limited by the single-ended to differential conversion stage at the input to the mixer. The input balun is often implemented in the output of a low noise amplifier.

Choose an application (Down conversion or up conversion). This choice is clearly dictated by our requirements for IF, LO, and IF frequencies.

Once the schematic design of the mixer is initiated the following templates assist in analyzing the performance parameters of the circuit. They are:

- DC Biasing
- Output Spectrum, Conversion gain and InterationSimulations
- Gain Computation
- Intermodulation Distortion and Intercepts
- Noise Figure
- Dynamic Range
- Adjacent channel power with digital modulation

Secondly, the design process depends on whether the mixer input and output must be matched to some on-chip impedance source or load, or if the mixer is interconnected on-chip. When needed, matching networks will provide some passive gain and so should be designed before going further with evaluation and optimization of the large-signal performance or noise figure of the mixer. On-chip interconnections should be modeled with appropriate source and load impedances. These may be complex impedances at a single frequency. In some cases or if a wide bandwidth design is required, the source and load could be modeled with equivalent circuits. Once the correct LO power is determined and matching networks are designed, then you should proceed with both a noise figure and a large signal evaluation. Which one you do first depends on which one of these is most critical for your application. The mixer design approaches for low noise and for large signal handling often are in conflict, so tradeoffs must be made to favor whichever is most critical.

Finally to evaluate the large-signal capability of a mixer to apply two or more signals to the input, these dual or multiple signals (tones) mix together and form intermodulation products, which one you do first depends on which one of these is most critical for your application. These results expressed in terms of the carrier-to-MID power ratio and calculated second and third-order (TOI) intercepts. These results are representative of how well the mixer will perform in a true multi carrier environment. In some cases you will want to evaluate the
Similarly in the design of an oscillator the large signal characterization cannot of the following simulations.

- Single-frequency oscillation.
- Phase noise.
- Tuned oscillations.
- Frequency Pulling.
- Frequency Pushing.

Harmonic balance simulation is an excellent way to analyze many oscillators in the frequency domain. Occasionally you might have an oscillator that converges in a time-domain simulation, but the harmonic balance oscillator algorithm is unable to find the solution. In such a case certain CAD tools provide certain techniques for solving those oscillators:

- Analyzing the large-signal loop gain in harmonic balance to find the point of oscillation and using that as an initial guess for the full harmonic balance oscillator analysis.
- Using a transient analysis to produce an initial guess for the harmonic balance oscillator analysis.

CONCLUSIONS

A general CAD procedure for the design of microwave nonlinear circuits such as mixer and oscillator has been presented which gives an amount of confidence in the performance prediction of these circuits with higher degree of accuracy which results in minimum deviation from the measured results.

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DESIGN AND ANALYSIS OF COMPACT CIRCULAR TO RECTANGULAR WAVEGUIDE TRANSITION AT 94 GHz

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Abstract: Design concept for a compact rectangular to circular waveguide transition is presented. The transition can be integrated with the flange at the apex of any conical feed horn. Conversion of TE₀₂ to TE₀₅ can be achieved using modal matching techniques. Instead of using conventional taper or abrupt stepped transitions, a cone has been used in rectangular waveguide. The thickness of transition is such that its can be integrated with the flange. Return loss better than 17.13 dB and insertion loss of 0.15 dB has been achieved over the bandwidth of 4 GHz with center freq of 94 GHz. The design methodology is verified by excellent agreement between theoretical analysis using HFSS software and practical measurements.

INTRODUCTION

Parabolic reflectors and dielectric lens are mostly used to collimate the output horns as primary feed. The circular to rectangular transition is required to have interface with standard rectangular waveguide type components, commonly used for commercial telecommunication applications. Direct integration of a rectangular to circular transition within the conical feed system, flange, provides compactness, handling and therefore low overall feed system cost. Such a concept requires design capabilities for short and compact transition as well as incorporation of ailment tail in the cross section of the flange for an easy manufacturing process. General stepped or stepped circular to rectangular transitions, commercially available are very large in length[1]. This paper describes a new compact circular to rectangular transition at 94 GHz with a cone in rectangular waveguide section. A trade off between length and return loss has been achieved over the bandwidth of 4 GHz (92 GHz - 96 GHz).

THEORETICAL DESIGN CONSIDERATIONS

The basic rectangular to circular waveguide transition considered has WR-40 rectangular port fit one side and circular cross section at the other side. The TE₀₅ mode has been considered for rectangular port and TE₀₂ mode has been taken for circular port [2]. If a single mode enters the transition from one port it will travel along the transmission line and emerge at the other end as a single mode. The electromagnetic field in a waveguide of varying cross section can be described by varying eigen function [3].

\[\vec{\psi} = g₁ \psi₁ + g₂ \psi₂\]

where \(\psi₁\) and \(\psi₂\) are monotonically differentiable function of \(x\) with
$g(0) = 0$

$g(L) = 0$

where $L$ is the length of transition if input and output modes are $TE_0$ modes, the cross sectional boundaries of the transition as determined by the conditions that the normal derivative of $g$ vanishes at the boundary. The eigen function of the rectangular $TE_0$ mode can be written as $g(r = k_1 sin(x))$. Where $k_1$ is the coordinate parallel to the long sides of the rectangular waveguide. The eigen function of the circular section and for the $TM_11$ mode is $g(r) = e^{j2\pi r_0 \sqrt{2\pi k_1\nu}}$, where $j$ is the first order Bessel function of $k_1$, $k_1$ scaling factor, $r_0$ Radial coordinate in $\mu$m radius off the circle.

The dimension of the circular waveguide has been chosen to have same cut-off number as that for the input rectangular waveguide. To keep for $k_1/k_2$ constant throughout the transition it is reasonable approximation to determine the required angle such that maximum magnitude of $E$ vectors in the rectangular and circular cross section are equal. If the losses are neglected, cone angle can be determined by equating the power flow through input and output cross section of the transition.

**ELECTRICAL PERFORMANCE**

The designed compact transition has been manufactured using FDM machine. The fabricated transition is shown in figure 1. The length of the designed transition is 3.7 mm with one port in WR-10 (2.54mm x 1.27mm) and the other port has a circular diameter of 2.1 mm.

![Fabricated rectangular to circular waveguide transition](image)

Figure 1. Fabricated rectangular to circular waveguide transition.

Return loss for the compact transition has been calculated over the band of 4 GHz from 92 GHz to 96 GHz. The transition was modeled in HFSS and the simulated return loss obtained is better than 18 dB for the band of 4 GHz with centre frequency of 94 GHz. The performance of fabricated transition has been evaluated using vector network analyzer. The measured return loss at rectangular port is better than 17 dB, which is very close to the value of the simulated result at that port. The simulated and measured return losses at the rectangular port are shown in the figure 3. The simulated return loss at the circular port is approximately same as that at rectangular port. The measured insertion loss is 0.15 dB, as compared to the simulated insertion loss of 0.06 dB at 94 GHz.

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CONCLUSION

A rectangular to circular transition has been designed and developed using FMM process at 94 GHz with 4 dB bandwidth. The measured and the simulated results are in conjunction with each other. Return loss better than 17 dB and insertion loss of 0.85 dB has been achieved practically. This compact flange integrated circular to rectangular transition can be integrated with flange at apex of any conical horn feed.

ACKNOWLEDGEMENT

We are thankful to Dr. R. Sivasankar for granting permission to publish this paper. We are grateful to Mr. K. Sivas Kumar for group Director for his continuous encouragement and support. We thank CAME and CAD group for manufacturing the compact rectangular to circular transition.

REFERENCES


A COMPACT OMT DESIGN FOR A Ka, BAND BEACON ANTENNA

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Abstract
The design and realization of a relatively compact orthogonal transducer (OMT) is reported in this paper. The OMT is intended for use onboard ISRO's experimental spacecraft, GSAT-4, in conjunction with a dual-polarized Ka-band beacon horn antenna. The design was made especially challenging by the high frequency of operation and the frequency reuse requirement (i.e. identical spot frequencies in both polarizations.) The proposed OMT eliminates transistors and mixer transformers at both spurious bands and is consequently compact (<1.7Q long.) The simplified geometry assists the fabrication complexities due to the stringent tolerances of these frequencies and minimizes the chances of errors. An FEM-based solver was used to analyze the OMT geometry (Ansoft-HFSS.) The realized OMT was tested to achieve a return loss of <20 dB at both ports with an isolation of > 40 dB.

INTRODUCTION
Orthomode Transducers (OMTs) are antenna feed chain components sequentially encountered when a single antenna is required to operate two polarizations simultaneously. Typical applications are TV/IF antenna for terrestrial or satellite communications where orthogonal polarizations are used for the two respective functions. In this case, the OMT allows the high-power signal to be routed to the radiation elements. The received low-power signal is routed to the receiver equipment maintaining adequate isolation between the two signals. The two polarizations handle different bands e.g. to a C-band satellite communications system, the band 1.7-2.3 GHz is used for downlink and the band 5.80-7.02 GHz is used for uplink purposes. Further class of applications is referred to frequency reuse systems. In these, overlapping or coincident frequency bands are implemented in a single antenna for a TV/IF or TV/RF function. From a design point-of-view, OMTs for the latter class of applications are more complex as will be described. Due to a common frequency band, the choice of waveguide dimensions becomes restricted. The design configuration must be selected carefully and close tolerances maintained to achieve good isolation between orthogonal ports. This becomes even more stringent at Ka-band (20-30 GHz.)

This paper reports a relatively compact OMT designed for dual-polarized Ka-band TV/IF beacon antenna meant to operate onboard the GSAT-4 experimental spacecraft. The functional requirements of the OMT are given. The next section describes typical OMT configurations and that selected for the present OMT. The design aspects and the analysis procedure are next presented. The computed results using an FEM solver and the experimental results are given in the end.

FUNCTIONAL REQUIREMENTS
ISRO's experimental spacecraft, GSAT-4 is carrying a Ka-band payload for the first time. It is expected to fulfill several experimental requirements including that of propagation studies.
for fixed or mobile earth stations. This will be supported by the provision of two dual-polarized beams at 70.7 and 50.5 GHz respectively. At these frequencies, each beam will radiate both polarizations on a continuously swept basis. The boresight footprints are required to cover the entire Indian subcontinent with a medium gain but very high polarization purity to allow access to all parts of the subcontinent with a sufficiently clear path. Each beam will require an OMT to combine the two polarizations into the common antenna input. The derived specifications of the 26-GHz beacon OMT being reported here are:

- Frequency: 20.2 GHz (L-Band signal)
- 20.2 GHz (L-Band signal)
- Interface: V-Port and WR51 waveguide
- 16 Port WR51 waveguide Common port 1 13.22 mm 129ω
- Return Loss: 17 dB, min (L-Band Lna-H port)
- Isolation: 35 dB, min (L-Band to Lna-H port)
- Power handling: 2W CW per port (simultaneously)

The critical parameter is isolation, as the beacon requires a high order of polarization purity.

OMT CONFIGURATIONS

The configuration of an OMT is based on a common section that interfaces to the feed / antenna element. With a central horn, a usual requirement is a circular waveguide. In most OMTs, the common section is usually continuous with one of the output ports (polarizations) while referred to as the through port. The other output port (polarization) is derived as a branching section from this axial geometry. This is usually called the coupled port. The choice of the common section, transition to the required through/port and the method of branching are the three key structural elements of an OMT and are determined by the operating frequencies, return-loss and isolation requirements. A typical configuration is a circular-to-rectangular waveguide transition at the common port followed by a tapered / stepped transition to the through port and a narrow axial iris coupling to the branch section. The branch guide may also have a stepped or taper transition or a standard section to match the WR impendence. With two different frequency bands in the through and coupled ports, it is possible to achieve filtering by choosing appropriate quasi-rectangular dimensions for the two waveguides. These can be selected non-stigmatic to either suppress the cross coupling or to reduce the abrupt change in impedance at any step e.g. after the iris. With a frequency reuse OMT; this freedom is not available as both polarizations carry the same frequency. In this case, the symmetry and precise orthogonality of the two signal paths becomes very important. The proposed configuration takes advantage of the 3.40 frequency requirement to realize an extremely compact and straightforward design. The usual transition to a square section is eliminated. As a symmetrical linear circular-to-rectangular taper is used to match the 111.3mm circular section to a standard WR51 waveguide. An appropriately narrow longitudinal iris is used to couple the orthogonal polarization to another WR51 waveguide. The symmetrical taper ensures that the coupling iris is uniformly thick. Further, no tees are used to match the iris to the succeeding waveguide section. The taper section and the WR51 waveguide provide a virtual shorting plane at the operating frequency by forcing the r.m.s. wave to be canceled. The position of the iris is optimized with respect to the shorting plane at the operating frequency. A short length of WR51 waveguide, which later forms part of the signal plumbing, ensures the continuity of the short plane (see Figs. 1 & b). In this way, an extremely compact OMT is realized by eliminating transitions and step transitions. The
DESIGN AND ANALYSIS CONSIDERATIONS

Unless fixed from the third antenna side, the common section is selected to satisfy the following requirements:

a) adequate separation between waveguide cut-off and the lowest operating band-edge; and
b) prevention of higher-order mode generation or propagation in the common section.

Placement of tuning elements of discontinuities in the common section is

on the symmetry. Elements symmetrical to both X- and Y-axes exist and TM_0, modes with

odd m. and even n. indices. With Y-symmetry, only the n indices are even, and with X-symmetry, the m

indices are odd [1]. Presence of these modes affects both impedance match and

isolation. The choice of asymmetry or discontinuities in the OMT structure must account for

these modes and their possible impact on the performance.

This kind of structure is unsuitable for method-of-moments or modal processing due to a

continuously varying geometry along the device axis. Also, the coupling iris would span several

such modules. Consequently, a solver based on the finite-element method was selected for

analyzing the proposed geometry (Ansoft-HFSS). The solver uses adaptive mesh refinement and

is initialized by a manually seeded mesh in the coupling iris region. A mesh facility in the

code was used to generate repeated models of the geometry to optimize the iris dimensions

and the placement of the virtual short behind the iris location.

SIMULATION & MEASURED RESULTS

Simulations of the final geometry included a Three-Port Return-Loss of 25.8 dB and a Coupled-

Port Return-Loss of 33.4 dB at 20.2 GHz. The coupled response was relatively narrow as

was expected due to the simplified design geometry. The isolation was 40.7-dB.

Measurements indicate that an adequate impedance match is achieved at both the ports and an

isolation value close to the predicted value is attained (see Fig. 2.).

FABRICATION ASPECTS

The critical areas for fabrication are the transition and the iris. Wire-cut EDM was used for the

asymmetrical taper portion. The slot was also realized using conventional EDM and the

specified tolerance of ±0.001 in. was just met. The proximity of the coupling iris to the WR31

thru-putt flange made the inner taper design quite complex. A non-standard denting with central ridge

to one side was devised to overcome this problem.

ACKNOWLEDGEMENTS

The authors express their gratitude to Dr. S. B. Sharma, Group Director-A&G for his inspiration

and guidance. We acknowledge Mr. Rupa Bandyas, a colleague, provided useful suggestions for

a mesh design and other aspects from his own OMT design experience. Finally, we owe a
debt of gratitude to Sh. N. N. Pandya, who provided mechanical design expertise and Sh.

Rajpati Gobir who modeled the OMT geometry.

REFERENCE

Fig. 1: Conceptual drawing of the proposed CBLT in an airport environment.

Fig. 2: Measured performance of the 28-GHz Hi-Rand beacon with CBLT for case 1-4.
MINIATURIZED DESIGN OF A MICROMACHINED PHASE SHIFTER USING MODIFIED GROUND COPLANAR WAVE GUIDE

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Abstract: In this paper we study the effect of modifying the ground trace of a CPW in miniaturizing the design of a micromachined phase shifter. Simulated results based on the proposed design show a 30% improvement in the figure of merit of the phase shifter in addition to a reduction of the device length by an equal fraction. A micromachined phase shifter designed for 500GHz with this approach has an overall area of 6.5 mm² and an Insertion loss less than 2dB.

INTRODUCTION

A phase shifter is one of the main components in phased array antennas used for communications and radar systems. Phase shifter causes a change in the phase of output signal w.r.t. the input signal. The attenuation and phase shift can be varied continuously by varying the length (mechanical or dielectric). Without the phase shift, the signals are summed to achieve a certain phase difference. This is achieved by using a number of sections of the same device. In radio frequency systems, these devices are often used in arrays of hundreds or thousands of elements.

However, these have several disadvantages. For example, ferroelectric phase shifters are complex in nature and have a high fabrication cost, whereas the semiconductor devices (e.g., PIN diode-based phase shifter) have high insertion loss. Hence, the trend is moving towards the application of micromachined devices for phase shifter applications. Micromachined phase shifters have low insertion loss, low DC power requirement, more compact and can be fabricated using mature silicon manufacturing processes.[2,3]

The micromachined phase shifter presented here works as a distributed system with variable capacitances. Here the tunable diode has a coplanar waveguide, periodically loaded by micromachined phase interdigital capacitors used to perform the phase shifting. The capacitance is determined by the length, spacing and number of fingers in the interdigital structure. A thin film of alternating dielectric layers is created on this to incorporate the electronic tunable diode. BST is a ferroelectric material belonging to a class of nonlinear dielectrics that exhibit an electric field-dependent dielectric constant [4-6]. By changing the bias voltage between these fingers, the dielectric properties of the thin film can be made to change, thereby resulting in a change in capacitance.

In this paper, we strive to reduce the overall size of a micromachined phase shifter by incorporating defects detected ground CPW [2]. Here spiral defects are introduced into the ground traces of the CPW on which the phase shifter is constructed to alter the propagation characteristics. The following study has demonstrated that such spiral defects can introduce an additional capacitance into the equivalent circuit of the line. We have utilized this in reducing the overall size as well as in improving the performance characteristics of the phase.
The basic configuration of the phase shifter is shown in Figure 1. This design consists of a number of parallel tunable capacitors arranged periodically along a CPW. The equivalent circuit for each section is also shown there. The equivalent LC model for a single section of the phase shifter is similar to that of single transmission lines, with only a variable capacitance varied in parallel with the transmission line capacitance. Hence, using the equivalent circuits as shown in Figure 1(a) we can calculate the characteristic impedance and propagation constant of the section of the line. Due to the periodic nature of the device, the Bragg diffraction frequency should also be considered for determining the upper cut-off frequency of operation.

\[ \text{Figure 1 (a) Basic configuration of a microstrip line phase shifter using plunger tunable inter-digital capacitors with BST thin film. (b) A section of the device. (c) Equivalent circuit of a section.} \]

The characteristic impedance, phase velocity and Bragg frequency of the "loaded" line are:

\[ Z_0 = \frac{Z_{00}}{\sqrt{1 + \frac{I_C}{e}}} \]

\[ v = \frac{c}{\sqrt{1 + \frac{I_C}{e}}} \]

\[ f_{\text{Bragg}} = \frac{1}{2\pi p_0 \sqrt{1 + \frac{I_C}{e}}} \]

where \( C_0 \) and \( I_C \) are the capacitance and inductance of the lossless "unloaded" CPW line. The length of each section is \( s \) and the total capacitance of the phase interdigital sections is \( C_{DI} \).

The phase shift, \( \Delta \phi \), provided by a single section at any given frequency, \( f \), can be obtained from:

\[ \Delta \phi = -\pi f f_{\text{Bragg}} \left( \frac{1}{C_{DI}} \right) \left( \frac{1}{\varepsilon_{\text{BST}}} \right) \times 10^9 \text{ Degree/Unit} \]

where \( Z_{00} \) and \( \varepsilon_{\text{BST}} \) are the impedances of the line corresponding to the extreme values of capacitance realizable by changing the dielectric properties of BST thin film. Using the above equations, an approximate design of the phase shifter can be done. The accuracy analysis for the evaluation of this analytic approach can be done using full-wave EM simulator such as HFSS. These simulations require substitution of dielectric properties of BST (for applied bias voltage). We have obtained these using analysis of dielectric properties.
In order to realize phase shifter designs having shorter section length along with an overall improved phase characteristics, we investigated means of varying the inductance of the line without significantly altering the capacitance. The simplest method to vary the inductance is to change the section spacing but this changes the CPW capacitance as well. Alternatively, we explored the possibility of having the ground plane of the CPW modified (e) using spiral inductors (Figure 2), where sections of ground coplanar with the signal plane alone are modified to introduce a suitable inductance. Although there has been research reported on implementing compact phase shifters using lumped spiral inductors, ours is the first attempt towards using modified ground (instead of lumped inductors) to optimize the phase shifter. The structure in Figure 2, shows two inductors per section in each ground plane.

To model the proposed modification, we have used an extra inductor layout in series with the distributed inductance composing of the modified CPW. EM simulations showed that the modified-ground has a very little impact on the total capacitance. The equivalent circuit based on Telegrapher’s equation is shown in Figure 2(b).

Periodic structures such as the phase shifters we consider here, have an upper limit of operational frequency introduced by Bragg diffraction. In our study we have enforced the condition that the Bragg frequency remains unchanged after reducing the section spacing. This can be done to retain the frequency range of operation.

Table 1 Design parameters for the phase shifter

<table>
<thead>
<tr>
<th>CPW</th>
<th>CPW dimension (G/WG) μm</th>
<th>90/70/90</th>
</tr>
</thead>
<tbody>
<tr>
<td>Width</td>
<td></td>
<td>1.2</td>
</tr>
<tr>
<td>Height</td>
<td></td>
<td>1.2</td>
</tr>
<tr>
<td>Interdigital Capacitor</td>
<td>Number of fingers</td>
<td>12</td>
</tr>
<tr>
<td></td>
<td>Overlap length μm</td>
<td>32</td>
</tr>
<tr>
<td></td>
<td>Finger width μm</td>
<td>5</td>
</tr>
<tr>
<td></td>
<td>Finger gap μm</td>
<td>2</td>
</tr>
</tbody>
</table>

RESULTS AFTER OPTIMIZATION

A set of mixed electromagnetic and network simulations were performed using
different reductions in section length along with modified (spiral) ground.
The full-wave simulation of each section is done using IE3D. The results from this simulation are used to extract the parameters in the equivalent circuit. Table 1 summarizes the extracted parameters used to calculate the Bragg frequency. The extracted inductance value for the modified ground shows good correspondence with the analytically determined inductance of the spiral structure.

The modified ground phase shifter gives about 30% improvement in the phase shift per unit length with much variation in the insertion loss. However, since there is a reduction in line length (by as much as 60%), the insertion loss is reduced. However, section spacing cannot be reduced any further in this design would require a larger spiral inductance and hence a larger area for the spirals in the ground. Hence 50% appears to be the best for the current configuration. Figure 1 shows the variation of phase shift and insertion loss for different reductions in section lengths at 20 and 40 GHz. The overall size area for the device is described. Although the final design is not validated against experimental ones, we have found the results to be almost equal to the analytical predictions at this stage (e.g., prediction of material properties of BST for various voltages, prediction of capacitance of parallel variable capacitors, effect of spiral defects on ground of CPW) match experimental results reported in the literature.

SUMMARY

We have demonstrated a CPW with its ground traces modified with spiral slots can effectively be used to improve the device performance. This approach has been implemented to optimize a micromachined phase shifter. Based on full-wave electromagnetic simulations using IE3D and a modified topology of network modeling we have determined that a complete 360° phase shift (at 30 GHz) is possible using 1.2 cm long CPW, with a minimum insertion loss of 2 dB. The modified device has a figure of merit of 30% (decrease) compared to 15% in the original design. Hence the performance improvement can be put at 30%. To apply similar principles to miniaturize the design of other RF/microwave circuit components currently underway. We are also trying to fabricate such micromachined phase shifters to validate the novel miniaturization approach presented here.

REFERENCES

<table>
<thead>
<tr>
<th>Reduction in section length</th>
<th>0 (original)</th>
<th>20</th>
<th>40</th>
<th>50</th>
<th>60</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{op} (\text{Before})$</td>
<td>103.3</td>
<td>105.5</td>
<td>107.3</td>
<td>109.5</td>
<td>111.5</td>
</tr>
<tr>
<td>$C_{op} (\text{Before})$</td>
<td>36.2</td>
<td>37.1</td>
<td>38.0</td>
<td>38.9</td>
<td>39.8</td>
</tr>
<tr>
<td>$L_{op} (\text{After})$</td>
<td>180.8</td>
<td>190.77</td>
<td>199.57</td>
<td>209.04</td>
<td>219.7</td>
</tr>
<tr>
<td>$C_{op} (\text{After})$</td>
<td>35.5</td>
<td>36.2</td>
<td>37.1</td>
<td>38.0</td>
<td>38.9</td>
</tr>
<tr>
<td>$I_{op} (\text{Before})$</td>
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<td>20.2</td>
<td>20.2</td>
<td>20.2</td>
<td>20.2</td>
</tr>
<tr>
<td>$I_{op} (\text{After})$</td>
<td>12.3</td>
<td>22.8</td>
<td>28.2</td>
<td>33.8</td>
<td>39.5</td>
</tr>
</tbody>
</table>

Figure 3: Performance of the optimized design of phase shifter containing 43 periodic units shown in Figure 2. $\Delta l$ is the reduction in section length.
RESEARCH SESSION 7
Dielectric behaviour of some mesogenic, non-mesogenic and organic molecules

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Technology for Land mine detection - a structured review

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An introduction to cellular neural network and its application

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The study of the microwave Dielectric and absorption behaviour of Brz-Zr-Co

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The Graphical User Interface (GUI): Adaptive Array

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The effect of dopants on the microwave dielectric properties of SrLa2TiO6

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2.7 Preparation and Dielectric Properties of Composite Substrates With Varying Filler Concentration
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2.8 \((1-x)\text{MnO}_2-x\text{TiO}_2\) (Mn-Zn, Mg) dielectrics for microwave substrate applications
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2.9 Cu-x Mg x Ni, TiO\(_2\), Ceramic Dielectric Resonators for Microwave Telecommunication Applications
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2.10 Application of metamaterials for defence and communication
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2.11 Application of metamaterials for defence and communication
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2018
Dielectric Behavior of Some Mesogenic, Non-Mesogenic and Organic Molecules

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Abstract: Dielectric behavior of some mesogenic and non-mesogenic compounds are presented and discussed from the data obtained using a microwave frequency 9.98 GHz and 8.74 GHz by applying Higashi Method. The obtained results are compared with earlier results obtained by different methods, wherever available.

INTRODUCTION

Dielectric property forms one of the powerful tools for the determination of molecular structure (1). For example, in the case of some derivatives of Syphonaria rather high values of electric dipole moments could be explained in a satisfactory way by introducing cyclic meso-stereic structures to share compensate rather than improbable bicyclic structures (2-5). The dielectric relaxation time and electric dipole moment are the important molecular parameters that are helpful in understanding the structure, size, shape of the molecules; the ion-, and neutral-, solvate, solvate forms etc. In particular the relaxation time can be evaluated by measuring the high-frequency dielectric permittivity, $\varepsilon'$ and $\varepsilon''$ at either 1) different frequencies, 2) different temperatures, of pure liquids or 3) different concentrations of a polar molecule in a non-polar solvent. However, such studies have no advantages over pure liquids that the strong dipole-dipole interactions in dilute solution phase are greatly reduced and also that they permit to study the effect of viscosity on the molecular parameters, in addition to the effect of temperature only in the case of pure liquids.

By assuming that the behavior of the dilute solutions conform closely to that predicted by the Debye theory, the two molecular parameters $\varepsilon'$ and $\varepsilon''$ can be evaluated by the concentration variation method due to Gouy (6), hypothesis methods due to Higashi (7), Higashi et al. (8), dielectric conductivity methods due to Kudrjawzew and Chintjav (9), Murphy et al. (10), and also a method due to I. Franks (11) based on Field's equation. However, instead of carrying out dielectric measurements on a set of fixed concentrations of a polar-solute molecule in a non-polar solvent, it is possible to carry out similar measurements on a single appropriate concentration at several frequencies (12, 13), or at only two frequencies (14, 15) around the frequency corresponding to maximum absorption and to determine the loss tangent (tan $\delta$) from which both $\varepsilon'$ and $\varepsilon''$ can be evaluated. In particular, the later method seems to be convenient in a situation where the polar substance under study is in solid phase at the temperature of measurement or the quantity is not enough, which is generally the case for liquid-crystal substances, to prepare a set of varying dilute solutions of it in a non-polar solvent. However, this procedure for the measurement of $\varepsilon'$ and $\varepsilon''$ and evaluation of $\varepsilon'$ and $\varepsilon''$ from their losses, is not being widely used by the investigators.

Therefore, the author feels, it is worthwhile to check the validity of the method by carrying out dielectric measurements on some polar molecules at a single frequency in the microwave region following the concentration variation method and also on a

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single arbitrary concentration of it that gives measurable maximum dielectric absorption at two different frequencies not well separated by each other and compare the results and draw correlations on the basis of comparison of the results.

In view of the considerations in the above paragraphs, dielectric measurements in benzene at room temperature on the pure samples of 2-chloro-diketonephene, 2-chloro-6-fluoro-, benzaldehyde, p-fluoro-phenyl acetate, 2-methyl benzonitrile, p-bromo nitrobenzene, m. bromo nitrobenzene and diphenyl sulphone are carried out at a frequency of 9.08 GHz employing concentration variation method. Similar measurements, on a single weight fraction of each of them at 9.08 GHz and also at 8.74 GHz, both at the room temperature are carried out. Because of the want of enough quantity of the substance dielectric measurements on a single weight fraction in benzene of each of the liquid crystal samples, namely, HoOAB (4, 4'-Bip (biphenyloxy)-oxybenzene), HoOCD (4, 4'-Bis (biphenyloxy)-oxybenzene), PoOAB (4, 4'-Bis (pentyl)oxybenzene) and OOBa (4-phenyloxy-4-trifluoromethyl-4-butyloxylene) are carried out at the said two frequencies and temperature.

MEASUREMENTS AND RESULTS:

The real |

-imaginary — parts of the complex dielectric constant at 4" and 6" were measured by employing standing wave techniques as described earlier [10], respectively. The static and optical permittivity of the solvents and solutions, were obtained in a routine way by using Franklin oscillator set-up and Abbe refractometer, respectively. The quantities X and Y required in the Gopalakrishna's method, the slopes m, n, a, b, c, and y, in the Higasi and Higasi et al. method, the slope of "K" and "K" in the dielectric conductivity method —, to evaluate the molecular parameters and were determined by the appropriate linear plots of the dielectric parameters. For the determination of these two molecular parameters in the case of single concentration — frequency measurements, the data are analyzed in the light of Whiffen and Thompson approach (14).

In table 1 a comparison of the evaluated dielectric parameters, p and τ is made from different methods is present. The measured dielectric permittivities along with the determined quantities p and τ by subjecting these data to the method of Whiffen and Thompson are presented in table 2.

DISCUSSION

It is seen from table 1 that there is a good agreement between the values calculated from conductivity and Gopalakrishna's method and of the Higasi et al. method except in the case of p-bromo-nitrobenzene. Where the agreement between η and τ obtained from conductivity, and Gopalakrishna's method is rather poor. Further, the τ values obtained from Higasi method differs from other values, the difference being well outside the experimental error. Such a trend of variation is not unexpected because the former methods do not account for the distribution of the present time whereas the latter do. It is supported by the finite non-zero values of the distribution parameter η in all the cases as can be seen from this table.

As regards the dipole moment values of these molecules for agreement between the values obtained from Gopalakrishna and Higasi methods seems to be good if not better, except in the case of p-bromo-nitrobenzene molecule. A look at the last column of the table 2 which contains p values of these molecules obtained by two-frequency method tend Support that the p values obtained are of right order.

It is observed from table 2 that the agreement between the τ values obtained

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From two-frequency absorbance and other methods Table 1 is reasonably good. Regarding the \( \varepsilon \) values of liquid crystalline substances, in values from 3 p.s to 25 p.s. via 11 p.s for three homologues of POAB, XHOAB, and HPOAB and thus varies according to their molecular weights. In view of the general observation (6) that a two-ring benzene system in benzene would have \( \varepsilon \) in the range of 16.30 p.s., the observed values in the present investigation seem to be shorter. But such a situation is not tenable in the literature. Since the \( \varepsilon \) values of these liquid crystalline materials in their isotropic phase differ widely, they would not have been reported in the literature as far as the author is aware. Therefore, it may be said that the values are of right value. As regards the dipole moment values of these materials, except for the length of hydrocarbon chain, which may contribute small moments due to inductive effect, the three molecules POAB, HPOAB and XHOAB contain the same strong anony dipole and the total electric dipole moment arises from this dipole. But the observed values do not support this point and indicate that this method gives physically meaningless results. But, the ground state dipole moment values for these molecules determined by an independent method (modified Coggeshall) reported elsewhere [17] do not seem to contradict these values at not supporting them. Thus, it may be said that the single concentration - two frequency method can serve at least as giving some qualitative trends but not otherwise.

<table>
<thead>
<tr>
<th>Molecular</th>
<th>Microscopic</th>
<th>G.K. Method</th>
<th>Nigami Method</th>
</tr>
</thead>
<tbody>
<tr>
<td>Compound</td>
<td>( \varepsilon/pS )</td>
<td>( \varepsilon/pS )</td>
<td>( \varepsilon/pS )</td>
</tr>
<tr>
<td>2-Chloro-Phenol</td>
<td>1.5</td>
<td>1.5</td>
<td>1.5</td>
</tr>
<tr>
<td>2-Chloro-4-Nitro-</td>
<td>1.0</td>
<td>1.0</td>
<td>1.0</td>
</tr>
<tr>
<td>Toluene</td>
<td>1.5</td>
<td>1.5</td>
<td>1.5</td>
</tr>
<tr>
<td>2-Methyl-Benzene</td>
<td>3.2</td>
<td>3.2</td>
<td>3.2</td>
</tr>
<tr>
<td>p-Xylene</td>
<td>5.6</td>
<td>5.6</td>
<td>5.6</td>
</tr>
<tr>
<td>Weight</td>
<td>to</td>
<td>to</td>
<td>to</td>
</tr>
<tr>
<td>Distribution</td>
<td>0.00091</td>
<td>0.00091</td>
<td>0.00091</td>
</tr>
</tbody>
</table>

Table 2

<table>
<thead>
<tr>
<th>Molecular</th>
<th>Weight</th>
<th>( \varepsilon' ) at 5.8 ( \times ) 10^{-3}</th>
<th>( \varepsilon' ) at 5.8 ( \times ) 10^{-3}</th>
<th>( \varepsilon' ) at 5.8 ( \times ) 10^{-3}</th>
<th>( \varepsilon' ) at 5.8 ( \times ) 10^{-3}</th>
<th>( \varepsilon' ) at 5.8 ( \times ) 10^{-3}</th>
</tr>
</thead>
<tbody>
<tr>
<td>SodiumSalicylate</td>
<td>0.0128</td>
<td>0.0128</td>
<td>0.0128</td>
<td>0.0128</td>
<td>0.0128</td>
<td>0.0128</td>
</tr>
<tr>
<td>4-Chloro-anisole</td>
<td>0.0303</td>
<td>0.0303</td>
<td>0.0303</td>
<td>0.0303</td>
<td>0.0303</td>
<td>0.0303</td>
</tr>
</tbody>
</table>

2.1
REFERENCES:

TECHNOLOGY FOR LAND MINE DETECTION: A STRUCTURED REVIEW

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Abstract: Land mines are a serious threat to the progress of advancing troops and to render terrain impassable. Several promising technologies such as active and passive infrared, microwave and MMW, acoustic, Neutron activation, magnetic field sensing and probing etc. for detection of land mines are being used currently on global basis, however each have their own strengths and weaknesses. This paper reviews various technologies for land mine detection and presents the results of initial measurement carried out at 400 MHz. It has been shown that 400 MHz signal yields better target to clutter ratio and land mines could be identified from a safe distance.

INTRODUCTION:

Mines are lethal weapons that have been used since World War-I and American civil war when the networks of tunnels, where dug under enemy positions and filled with explosives to destroy or disable enemy troops. Mine clearance efforts at operational level is known as mine warfare. The cost of each land mine, comes to few hundred rupees i.e., the cost of mine detection, identification and destruction is exorbitantly high nearly 1000 times. Further the injuries due to land mine minor to fatal have disabled many and their misfire are unacceptable. Several promising technologies for detection of mines such as Infrared, microwave, mm-wave, acoustic wave, neutron activation, probing and magnetic field sensing are being used currently on global basis. Each technology has its own strength and weaknesses.

As an ideal land mine detector should be able to indicate accurately the position including depth. These mines are made of various materials, the only certain means of identification is detection of explosive fillings. Some of the explosives like TNT have high resistance to shock or friction and can be handled with ease while others like nitroglycerine are sensitive they are mixed with inert desensitizer for practical use.

Land mines are concealed either in brush or garbage or buried at least 12 inches deep in the earth. In all the cases the scattering, emission or reflection of EM energy is complicated by the presence of the earth surfaces. Several authors (3-5) have investigated the EM scattering by buried objects. When the air-interface is perfectly flat, Korn, 1981, However gave mathematical foundation for the propagation and scattering of plane wave in layered media (6). This method combines the antenna transmit and receive pattern with multiple bounce of EM wave with buried target. The wave length in the medium is $k = \frac{1}{\sqrt{\epsilon r}}$ .

Which at 300 MHz is about 50 cm for sandy soil with dielectric constant approx. 3. This paper presents in the subsequent section the brief of different methods of land mine detection and results of initial measurement carried out at 400 MHz. It has been shown that 400 MHz signal have better target to clutter ratio and land mines particularly abandoned could be identified from a safe distance.
INFRARED SENSOR:

Load mine return or release here as a rate rule than their surroundings thus during natural temperature variations, infrared sensors monitor the thermal contrast between the surrounding i.e. loosely packed soil with buried mine. Explosives are generally packed in a solid, plastic, wooden, or dielectric casing which are sensitive to temperature variation. To the ideal case the intensity of black body radiation can be written as $I = KT^4$ (watt meter$^{-2}$ steradian$^{-1}$ Hz$^{-1}$). Where $K$ is the Boltzmann constant $= 1.38 	imes 10^{-23}$ joule*K$^{-1}$, $\lambda$ is free space wavelength in meters $T$ is physical temp in $K$

Thus it is clear that intensity of black body radiation and physical temperatures are directly related with each other. No metallic body can exist an intensity greater than that of black body at the same physical temperature. Even a small temperature contrast of 0.2 °C between buried mine and surrounding disturbed soil can be noticed by infrared theories and trusted mines are identified.

MM WAVE SENSOR

Modern mines are created in plastic mold. The electrical properties of plastic, explosive and surrounding soil are more and less identical and thus the identifiability is a tough task. However each mine contains a metallic pin of the size 1-3 centimeter to trigger the explosive. Wavelength here greater than 3 centimeters often fails to detect the smaller pin but it is easily detected by mm-wave active and passive sensor.

In mm-wave band soil has high permittivity and low dielectricity on the other hand metal has low emissivity and higher reflectivity.

As an ideal case the emissivity of a black body in thermal equilibrium, ε equals unity and reflectivity $\rho$ equals zero. Conversely the reflectivity $\rho = 1 - \varepsilon$. Where $0 < \varepsilon < 1$

In such bright conditions temperature $T$ of the targets is related in physical temperature of the medium and can be written as

$T = \frac{1}{\varepsilon}$

Where $0 < \varepsilon < 1$

Therefore, mm-wave passive & active sensor can help identify the metallic pin of plastic mines concealed in earth, bushes by using submicron signal processing.

NEUTRON ACTIVATION

Most explosive used in load mines have abundant nitrogen nuclei. Neutrons emitted by an accelerator or radio isotopic source when subjected on nitrogen nuclei, specific gamma rays are emitted and detected. A spectrometer analysis of different eigenvalues are stored and used for conformity test. Customs departments uses neutron activation techniques at air ports to identify narcotics, drugs etc.
ACOUSTICS

Acoustic waves with frequency greater than 20 KHz can penetrate very wet and heavy ground such as clay. The reflected sound wave from varying targets or boundaries between materials help ascertain the location and depth of buried mine, but it needs a complete signature analysis using advance signal processing techniques so that false alarm rate from shrapnel could be minimized.

CONVENTIONAL METAL DETECTOR

In this method, the sensor do not radiate any energy, but only measure the disturbance of the earth's or targets natural electromagnatic field (7). It is very difficult to differentiate a mine like object from metallic debris. In most battle field the soil is contaminated by large quantities of shrapnel, metal scrapes, cartridge cases etc leading to innumerable false alarms. Thus the deminer uses a metal detector followed by probing of the ground. This process of demining is slow and warrants a new methodology which must enhance the state of mine detection and automate these tasks whenever possible to prevent mine clearing personnel to meet an accident.

400 MHZ RADIO WAVE SENSORS

The civil engineering applications of GPR (Ground Penetrating Radar) to locate pipes, reinforcing rods, imperfections in structural concrete is well known and very well documented elsewhere. (8). In some cases subsurface targets are resonant scatterer sin they may be simple one dimensional reflectors while in other cases they may be simple non resonant scatterer. Due to propagation through the varying soil condition difficulties are often faced when there is low contrast between buried object and surrounding soil parameters.

PRINCIPLE OF SIGNAL INTERACTION AND FEATURE EXTRACTION

Soil is opaque by nature and one can thick soil can conceal the object from naked eye. Further, the earth being leaxy and inhomogeneous due to varying, consistent with varying degrees of electrical parameters its dependency on weather makes the identification of buried object difficult. The task becomes truly formidable due to the presence of the ground interface. The reflected signal from the buried object depends on the that targets shape, its electrical parameters (σ, ε, μ) and the orientation of the target.

The conceptual design of 400 MHz radio wave sensor is dictated by the electrical properties of the ground, the simulated depth of the target and the characteristics of the target. The electric field amplitude is freq dependent for a lossy media. The medium act as a km pass filter. It only allows a frequency to penetrate which is below the cut off frequency given by the relation:

\[ f_{\text{min}} = 3.76 \times 10^{6} \text{Hz} \]  

Where \( f_{\text{min}} \) = frequency at which electric field amplitude is 3dB less than the initial freq. (Hz)

\[ R = \text{distance to observation point in meter} \]
When an electric field impinges on the target the induced current acts on the scattering and radiation fields of its own. The induced current is distributed on the targets in amplitude and phase as function of incidence angle, polarization and frequency, thus the scattered energy also depends on angle, polarization and freq. In terms of wavelength λ/4 resonance dimension, 3, these regions are defined.

Rayleigh : Wave length (λ) is large compared to the target dimension Resonance : Wave length (λ) and target dimension are of the same order.

Optical : Wave length (λ) is much smaller than the target dimension.

To identify an object like land mine, particularly and with ease of finite dimension, application of Rayleigh regime is preferred so that there is no variation in phase of the incident wave over the scattering body i.e., all portions are exposed simultaneously.

** CONFIGURATION OF 400 MHz SENSOR**

As shown in fig 1, 20 watt RF power transmitted from signal source through a λ/4 size dipole antenna. The angle of incidence was adjusted 35°, 45° & 55° so that the reflected and forward scattered energy are received roughly at the same angle following optical theory of reflection. A metallic cylinder roughly of 12 inch diameter was buried inside the soil at the depth of 30 cm, the normal location of buried and tank mine. The experiment was repeated with varying soil conditions like tightly packed earth, loosely packed earth, and by increasing the moisture content of the target area. Fig 2 shows the results of signal received and its oscillation from targets without cylindrical structure, and with cylindrical structure.

With no buried target a very small oscillation was observed. It may be due to the component of direct field coupling between trans. A receive antenna.

A very strong oscillation was observed in received signal with buried metallic cylinder. Loop packing and light packing of soil has minor effect on received signal strength but a significant decrease in reflected signal strength was observed in moisten in the soil.

The initial measurement provides basis for back scattered measurements and ascertain that 400 MHz can be used over a suitable mobile platform to locate the buried abandoned IED or tank mine from a safe distance.

**CONCLUSION**

An effort has been made in this paper to review the various mode of land mine detection and result of 400 MHz in situ measurement has been presented. The trade off in ground penetration low antenna gain for the different freq, band is a key consideration in the design of a system used to detect buried objects. Lower freq, band 300-400 MHz may yield better target to clutter ratio with specifically 300-300 MHz may have less transmission loss. It is concluded that there is no full proof mechanism available till date to identify the abandoned and land mine in bulk. The existing method of land mine detection is very slow and therefore warrants investigation on a new mode of land mine detector, so that location of land mine could be identified from a safe distance and finally the mine could be cleared without causing any
damage to life and property of people residing near border areas particularly in the western sector

Fig 1 SCHEMATIC OF 2MHz SENSOR

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AN INTRODUCTION TO CELLULAR NEURAL NETWORK AND ITS APPLICATION

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The Cellular Neural Networks have been an area of an increasing development since the paradigm was presented in 1988 [1]. One of the areas involved in this effort is the core that research some of the roles that describe the patterns that we found in nature. Cellular Neural Networks (CNN) constitute a class of nonlinear, recurrent and locally coupled arrays of identical dynamical cells that operate in parallel. ANALOG chips are being developed for use in applications where sophisticated signal processing at low-power consumption is required. A CNN is very well suited for high-speed parallel signal processing. Its local interconnection features makes it tailor-made for VLSI implementation. Potential applications for Cellular Neural Networks are signal processing, pattern recognition and image processing. In this paper an overview is given regarding CNN and its applications.

INTRODUCTION

A cellular neural network (CNN) is an artificial neural network [2,3] which features a multi-dimensional array of neurons and local interconnections among the cells. The basic CNN is in two well-known techniques: the artificial neural networks and the cellular automata. A neural network passes through weights (synapses), summations (body of the neuron), multiplications (denominators) and a special function called activation function, which controls the values so that they remain between certain boundaries. The cellular automaton (CA) theory [4] treats the universe as a collection of identical components, called cells.

ARCHITECTURE OF CNN

The cells are distributed over an n-dimensional space with every cell related with (exactly k cells (a k-regular grid). Refer Figure 1, in a 2- dimensional 8-regular networks the cell in the i-th row and the j-th will be referenced as C(i,j). The neighborhood of a cell is commonly defined with help of a parameter r, which is the maximum distance between a pair of cells C(i,j), C(i',j'). But that they are still related. There is a single rule that acts on every cell. The neighbourhood is defined as follows:

\[ N(i,j) = \{k : \max(|i-i'|,|j-j'|) \leq r \} \]  

An example of a neighborhood for a bidimensional 8-regular CA with r=1 and r=2 is shown in Figure 2. The role over a multidimensional 1-regular cellular automaton is defined by the relation article [3] uses the following equations:

\[ c_{i,j}^{n+1} = f(c_{i-1,j}^{n}, c_{i+1,j}^{n}, c_{i,j-1}^{n}, c_{i,j+1}^{n}) \]
So this relation can be stated by the use of a table with all the combinations and the new value of the cell. The use of this table in many cases is impossible. But it is practical due to the number of combinations; there are many ways to simplify this, which include an arbitrary rule, symmetrical rules and the use of a kernel. A kernel is an index similar to the weights on an ANN, only that every cell will have the same set of weights not one for every cell -something like in an ANN. The kernel is used in combination with the convolution operator.

This operator is defined for a 4-regular bidimensional CA as follows:

\[ f_{i,j}(u) = \sum_{c} k_{c} u_{i-1, j-1} \]

A Cellular Neural Network later uses this operator, together with the characteristics of a regular space, and the neighborhood.

The CNN model as stated in the original article [1] uses the following equations:

\[
\frac{\partial y_{i,j}}{\partial t} + x_{i,j} = y_{i,j+1} + y_{i,j-1} + y_{i+1,j} + y_{i-1,j} + \sum_{c} k_{c} y_{i-1, j-1} \equiv 0
\]

In a cellular neural network a cell CA, i.e., contains three different value one for an input, one for the state value, and one to represent the output. For a CNN the input is assumed to be a constant, \( A \) and \( B \) represent a pair of matrices with the same dimension of the one defined by the neighborhood and together with the scalar \( k \) they are called the closing term (these are the equivalent to the weights on an ANN or to the kernel for a CA).

APPLICATION OF CNN

A CNN is a nonlinear analog circuit, which processes signals in real time. It is made of a massive aggregate of regularly spaced closed circuit called cells, which communicate with each other directly only through nearest neighbors. Cellular Neural Networks (CNNs) constitute a class of nonlinear, recurrent and locally coupled arrays of identical dynamical cells that operate in parallel. Analog chips are being developed for use in applications where sophisticated signal processing at low power consumption is required. Signal processing via CNNs only becomes efficient if the network is implemented in analog hardware. In view of the physical limitations that analog implementations entail, robust
operation of a CNN chip with respect to parameter variations has to be


REFERENCES


CONCLUSION

In this paper the basic of CNN is discussed with some of the applications. Simple examples are discussed. It is possible to extend high-speed scientific applications of CNN in the field of medical, DSP, Antenna, and Communication etc.

REFERENCES


STUDY OF THE MICROWAVE DIELECTRIC AND ABSORPTION BEHAVIOUR OF MIL-ZN-CO FERRITES CERAMICS AT X-BAND

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Abstract: Low fines high permittivity dielectric materials have revolutionised the microwave technology. Microwave dielectric constant and absorption coefficient characteristics play a significant role in cellular mobile handsets. Therefore in the present paper an attempt has been made to measure the complex dielectric constant and absorption coefficient using Mn-Zn-Fe₀.₆₃-Co₀.₃₇ powders at X-band frequency. The measured value of the absorption coefficient was found to be matched with calculated one.

INTRODUCTION

MICROWAVE dielectric in ceramic form have been exploited in many applications such as filters, resonators, etc. During the past decade materials compositions have been developed to suit the single crystal material requirement. In the HTCC thin film applications in microwave. Microwave dielectric materials are widely used for the applications in the communication market. The advantages of microwave ceramics include high temperature stability, the good power handling and low insertion loss at high frequencies. These features in combination with new ideal designs will be built for continuing success of microwave ceramic in 3rd generation of mobile communication. The introduction and use of microwave materials is essential for further miniaturization of discrete microwave ceramic components. A material having dielectric properties to reduce the component size by a factor of at least to presented along with possible applications for such a ceramic.

In the present paper an attempt has been made to study the microwave dielectric and absorption behavior Mn₆.6 Zn₀.₃₇ Co₀.₆₃ Fe₂0₄ electronic ceramics in composition as well as frequencies. The absorption characteristics of Mn₆.6 Zn₀.₃₇ Co₀.₆₃ Fe₂0₄ ferrite ceramic predicts its application in cellular mobile phone.

(A) COMPOSITION OF THE SAMPLE

The properties of a ferrite ceramics material depend upon its composition as well as the microstructure. Mn₆.6 Zn₀.₃₇ Co₀.₆₃ Fe₂0₄ powder with a slight excess of electrical and magnetic properties which take them useful material for a number of ceramics particularly at high frequencies. A ferrite ceramics materials series of the composition Mn₆.6 Zn₀.₃₇ Co₀.₆₃ Fe₂0₄ where in x varies from zero to 0.45 in steps of 0.05, has been prepared by the normal ceramic and hot pressing techniques. The AR grade Mn₆.6 Zn₀.₃₇ Co₀.₆₃ Fe₂0₄ powders where balanced with defatted winder is an agate mortar and pestle. The starting was dried and calcinated at 1000°C for 3hrs, followed by cooling in N₂ atmosphere. It was washed again and then granulated using a small quantity of polyvinyl alcohol as a binder. A part of this material was moulded in

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to pellets and toroids which were finally sintered at 1300°C for 3hr to prepare ferite samples by the wet method.

The resulting powder was used to prepare ferrites by traditional hot pressing technique in which the material was sintered at 1250°C for 3hrs under the maximum pressure of 35MPa using the slurry mixing reported earlier. The coding of both the samples was taken in N2 atmosphere.

**MEASUREMENT OF DIELECTRIC CONSTANT AND ABSORPTION COEFFICIENTS**

The technique used in this measurement programme was the infinite sample method described by the N.M. Abrahams [8] 1. A X-band microwave bench with a slotted section and crystal detector was used for the measurement of VSWR and silt of mini coax. In this technique, the complex dielectric constant \(\varepsilon\) was determined using the relation

\[
\varepsilon_r - \varepsilon_i = \frac{\varepsilon_r + \varepsilon_i}{2} + \left(\frac{\varepsilon_r - \varepsilon_i}{2}\right)^2
\]

Where \(\varepsilon_r\) and \(\varepsilon_i\) are the real and imaginary parts of the dielectric constant respectively. \(\varepsilon_r\) is voltage standing wave ratio (VSWR) and \(D\) and \(\theta\) are the position of first minima with and without the sample connected. The sample was filled and pressed manually in a designed cavity of required dimensions and it was terminated in a matched load. The wave of \(L\), \(D\) and \(\theta\) were determined using a dial indicator on the slotted line section (least count 0.001 cm). VSWR values were determined using double minimum power method.

A very efficient technique of measuring microwave absorption in ceramic materials at X-band frequency consisted of resonant cavity type frequency meter, which is commonly referred to as an absorption wave meter. In this experiment the absorption coefficient \(a\) is a X-band signal will be determined by adjusting the frequency meter until a sharp drop in power is registered at the detector. A precision attenuator will be used to set a required operating level for the detector and SWR meter in addition to insulating the signal sources from the cavity of the frequency meter. The Q factor of absorption wave meter are determined as

\[
Q = \frac{f}{f - f_0}
\]

Where \(f_0\) is the operating frequency and \(f\) and \(f_0\) are the adjacent power points which are \(\pm 0.5\) below.

**CALCULATION OF MICROWAVE ABSORPTION COEFFICIENT**

The microwave absorption coefficient is determined using microwave absorption model and given by

\[
K_{abs} = \frac{K_{abs}(\varepsilon_r + \varepsilon_i)\varepsilon_r}{\varepsilon_r + \varepsilon_i}
\]

Where \(K_{abs}\) is wave impedance, \(\varepsilon\) is permittivity of frequency \(f\) in the dielectric constant of the ferite ceramic sample. \(f\) is frequency 8-12 GHz and \(\varepsilon\) is the size of the pellet size sample. In above formula, the high power of a neglected therefore absorbance coefficient becomes

\[
K_{abs} = \frac{20K_{abs}(\varepsilon_r + \varepsilon_i)\varepsilon_r}{\varepsilon_r + \varepsilon_i}
\]

**RESULTS AND DISCUSSION**

A complete programme has been written in basic language for calculation of dielectric constant and absorption coefficient using Eq.(4). The variation of dielectric
The dielectric constant decreases gradually with rise of frequency from 8 to 12 GHz for all compositions. The decrease is to dielectric constant with frequency may be explained on the basis of heterogeneous structure of ferrites compressing low continuity grain separated by high relativity grain boundaries as proposed by Koss. Such as heterogeneous lead to interfacial polarization which results in decrease in dielectric constant with frequency in microwave region.

The variation of measured and calculated value of absorption coefficient with frequency are shown in Fig. 1. It is found that absorption coefficient is increases with frequencies. The composition x = 0 have a small value of absorption coefficient, that 40%. It shows that absorption coefficient are also increases with composition of the sample. The absorption phenomena is occurred due to photon localization. The measured value of absorption is matched with calculated values using Eq. (3) and (4). The behavior is shown in Fig. 3.

ACKNOWLEDGEMENTS

Author are gratefulful to professor Devendra Kumar and professor Ope Nikash professor of Electrical and Electronic ceramics Dept. Of Ceramic Engineering IIT. J.H.U. for useful discussion during laboratory visit.

REFERENCES

GRAPHICAL USER INTERFACE (GUI): ADAPTIVE ARRAYS

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ABSTRACT: This paper presents Graphical User Interface (GUI) for Adaptive Arrays. The GUI is a simulation tool enabling the user to study, understand and experiment on the adaptive array antenna principles, through visualization. It is used to facilitate and complement various adaptive array algorithms like LMS, Apfelbaum, Conant-Modular Algorithm under various conditions of interference and desired signal directions.

INTRODUCTION

An adaptive array is an antenna array that contains its own adaptive means to feed back control, while assuming operation. It consists of a set of spatially disposed sensor elements connected to a single or multiple-channel adaptive signal processors. Processes in order to detect/maintain a signal arriving from a particular direction, the phases and amplitudes of the currents on the antenna array elements can be electronically adjusted such that received signals from that direction add in phase, and minimum gain is achieved in that direction. Using different kinds of algorithms, adaptive receiving arrays can steer them actively automatically to pick up the signal without knowledge beforehand the direction of arrival of the signal in the presence of interference. Also they have ability to form a null in the direction of interference. This forms the basis of the Smart Antennas. Adaptive arrays are a key technology expected to dramatically improve the performance of future wireless communication systems because they have potential to expand coverage, increase capacity, and improve signal quality.

Looking at the rapid growth and fast evolution of adaptive arrays and smart antennas into the modern wireless communication, it is felt that there should be a simulation tool to equip ment on these principles. The GUI presented in this paper graphically demonstrates principles of adaptive arrays using different algorithms like Least Mean Square (LMS), Apfelbaum Algorithm and Constant Modulus Algorithm (CMA) under stationary and non-stationary environments.

ADAPTIVE ARRAY FORMULATION

An adaptive array for processing narrowband signals is shown in Fig. 1. Each individual element array is driven connected to a variable weight. Each individual array element output is multiplied by a variable complex weight and then all the output signals of the array elements are summed. These weights are adjusted according to a performance index such as mean square error or input signal to noise plus interference ratio (SNIR) by adaptive processor. The first of these leads to LMS array and the second Apfelbaum...
array. The optimum weights computed by applying adaptive techniques, steer the main beam towards the desired signal direction and the null towards the interference signal direction. The Constant Modulus Algorithm adjusts the weight vector of the adaptive array so as to minimize the variation of the envelope of the desired signal at the array output.
algorithm techniques developed in recent times. Following adaptive array algorithms are used for the simulation in this paper.

1. MS Algorithm: It tries to minimize the square of the error between the steering vector and the actual received output signal. The ideally sized weight vector for the stationary environment is given by [2]

\[ W_{opt} = H^H R^{-1} \]

\[ \text{where } H = E[\phi(a,s)/s] \text{ and } R = E[\phi(a,s)\phi(a,s)^H] \]

2. RLS (Recursive Least Squares) Algorithm: This algorithm tries to minimize the error in the output of the system [2]. The optimum weight vector is given by

\[ W = H^H R^{-1} \]

\[ \text{where } H = E[\phi(a,s)/s] \text{ and } R = E[\phi(a,s)\phi(a,s)^H] \]

3. Graphical User Interface: All the algorithms given by equations 6 - 11 are simulated in the program, integrated in MATLAB. User can enter the number of elements, desired signal direction, number of interferencers and their direction, power of desired signal, interference power etc. and observe the radiation pattern. User can choose between classic and polar

\[ J(x) = \frac{1}{2} [w^H(x) \phi(k) - y(k)]^2 \]

\[ \text{where } y(k) = \sum w^H(x) \phi(k) \]

\[ \text{and weight vector is updated by:} \]

\[ w(k+1) = w(k) + \mu e(k) \phi(x) \]

\[ e(k) = \frac{y(k) - y(x)}{\phi(x)} \]

Fig. 2 Snap Shot of Graphical User Interface

[Diagram showing the radiating pattern and graphical interface]
In this paper, we investigate the adaptive arrays by visualization. It can be observed through this GUI that the LMS array adapts its desired signal direction even if the interference signals are stronger than it requires the reference signal [2]. The OAM array adapts to desired signal direction, but it requires the desired signal direction to be known [3]. The CMA array adapts to any stronger signal, even if it is interference [3].

REFERENCES
EFFECT OF DOPANTS ON THE MICROWAVE DIELECTRIC PROPERTIES OF SRLa4Ti4O15

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Abstract: AB6BO415 type cation deficient hexagonal perovskite SRLa4Ti4O15 was prepared through solid state reaction route. Dopants with different valences (like WO3, CoO2, MoO3, Nb2O5, Bi2O3, SnO2, and Al2O3) were added to the octahedral SRLa4Ti4O15 powder and sintered. Microwave dielectric properties of the ceramics were studied by resonance method. Pure SRLa4Ti4O15 has a dielectric constant (e) of 47.6, quality factor (Q x f) of 38,000 GHz (at 4.7 GHz) and T = 9 ppm/°C. The doped SRLa4Ti4O15 show microwave dielectric properties, e in the range 47-85, quality factor (Q x f) in the range 37000-43500 liters and T in the range -7 to -10 ppm/°C. With the addition of dopants there is an enhancement in the quality factor and the dielectric constant with marginal improvements in T.

INTRODUCTION

Most microwave systems are located in the 300MHz to 30 GHz range. An essential component of any telecommunication device is a filter by which a particular frequency range can be selected and all others blocked. The introduction of ceramic dielectric resonators (DRs) into this field made possible the miniaturization of communication equipment due to their relatively high permittivities. A good microwave dielectric resonator must therefore display a high dielectric constant (e) for miniaturization, high unloaded quality factor (Q) for selectivity and small temperature variation of resonant frequency (e) for thermal stability of microwave devices. Typically, ceramics with 20 < e < 100, Q > 2000 and e < 20 ppm/°C are useful for various kinds of applications. Cation deficient hexagonal perovskites are useful in this regard. Recently, the microwave dielectric properties of some AB6BO415 type cation deficient hexagonal perovskites such as BaSrNb2O6, Sr2Sn2CuO5, Sr2Y2O415, Ba2LaTi4O15 and AB6BO415 (M = Sr, Na and Ca) have been reported [1, 2]. They are characterized by high dielectric constant up to 54, high quality factor with Q x f of up to 350,000 GHz and T in the range -25 to 478 ppm/°C. In the present paper the effect of dopants on the microwave dielectric properties of SrLa4Ti4O15 is investigated.

The ceramics were prepared through the solid state ceramics route. High purity SrCO3 (99.99%), TiO2 (99.9%) [Aldrich Chemicals] and La2O3 (99.99%) [HILL LTD.] were used as the starting oxide powders. 150°C was heated at 1000 °C for 3 hours before molding in a high vacuum. The powders were mixed according to the stoichiometry and ball milled in distilled water medium for 24 hours in a plastic bottle using zirconia balls. The wet mixed powder was dried and calcined at 1250 °C for 4 hours, ground and again calcined at 1400 °C for 4 hours. A small amount (0.2 wt %) of
differently doped samples (such as WO3, CuO, MoO3, NiO, CoO, H2O2, SnO2, and Al2O3) were added to the calcined mixture and were ground well. The powder was mixed with Poly Vinyl Alcohol (PVA) as binder, mixed, dried, and again ground. The resultant fine powder was unidirectionally pressed in a tungsten carbide die under a pressure of 100 MPa such that the pressed disks have 6-8 mm in length and 14 mm in diameter. Siccative acid was used as a lubricant. The green pellets were sintered at 1150 °C for 2 hours. The sintered pellets were polished well. The bulk densities were measured using Archimedes method. The phase purity of the sintered samples were checked by the X-ray diffractometry (XRD) technique using Phillips X-ray diffractometer. The microscopic dielectric properties of the samples were measured using an Agilent 8757 ET Network Analyzer using cavity method and 17600R made resonant frequency [6]. The temperature coefficient of resonant frequency of was measured in the range 200°C to 700°C.

RESULTS AND DISCUSSION

The powder X-ray diffraction pattern of pure SrLa4Ti4O15 was taken and is in well agreement with the reported data [3, 4]. SrLa4Ti4O15 crystallized into ABO4O5 type cubic deficient hexagonal perovskite with Sr and La in the A site with co-ordination number 12 and Ti in the B site with co-ordination number 6. The addition of 0.2 wt% of dopants did not affect the phase purity of SrLa4Ti4O15.

The sintering temperature of pure SrLa4Ti4O15 was optimized in the temperature range of 1550°C to 1650°C. From Fig. 1 it is evident that the grain growth increases with temperature and attains maximum value at 1625°C and then decreases for higher temperatures. There is a slight decrease in the value of as sintering proceeds. The addition of SrLa4Ti4O15 was selected for maximum unloaded quality factor. WO3, CuO, MnO, NiO, CoO, H2O2, SnO2 and Al2O3 (0.2 wt% each) were added as dopants. For the pure SrLa4Ti4O15 the percentage of density was 96%. The addition of R2O3 increased the sintered density to 97.7%. With the addition of dopants, microscopic dielectric properties were found to be improved with a reduction in processing temperature of 1640°C to 1200°C. The microscopic dielectric properties of doped SrLa4Ti4O15 are summarized in Table 1. The dielectric constant of all the ceramics were corrected for porosity applying the following formula

\[ K = \frac{1 - \frac{\varepsilon_d - 1}{2\varepsilon_r + 1}}{1 - \frac{\varepsilon_d - 1}{2\varepsilon_r + 1}} \]

where \( \varepsilon_d \) is the dielectric constant corrected for porosity, \( \varepsilon_r \) is the experimentally determined constant, and \( \varepsilon_d \) is the fractional porosity [5].

The addition of dopants increased the dielectric constant slightly. The dopants such as WO3, CuO, MnO, NiO, CoO, H2O2, SnO2, and Al2O3 all decreased the experimentally measured density but increased the unloaded quality factor. On adding Al2O3 the quality factor is reduced. It is evident from Table 1 that the addition of all dopants decreased the if
Fig. 1. Variation of dielectric constant and unloaded quality factor of undoped SrLa4TiO15 as a function of sintering temperature.

Table 1: Microwave Dielectric Properties of pure and doped SrLa4TiO15

<table>
<thead>
<tr>
<th>Temperature (°C)</th>
<th>ϵr (corrected)</th>
<th>ϵr (corrected)</th>
<th>f0 (GHz)</th>
<th>Q×f0 (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>500°C</td>
<td>4.12</td>
<td>4.45</td>
<td>45.2</td>
<td>91.1</td>
</tr>
<tr>
<td>900°C</td>
<td>4.31</td>
<td>4.45</td>
<td>45.2</td>
<td>91.1</td>
</tr>
<tr>
<td>1100°C</td>
<td>4.43</td>
<td>4.45</td>
<td>45.2</td>
<td>91.1</td>
</tr>
<tr>
<td>1300°C</td>
<td>4.43</td>
<td>4.45</td>
<td>45.2</td>
<td>91.1</td>
</tr>
<tr>
<td>1500°C</td>
<td>4.43</td>
<td>4.45</td>
<td>45.2</td>
<td>91.1</td>
</tr>
</tbody>
</table>

CONCLUSION

The SrLa4Ti4O15 was prepared through solid state ceramics route. Effect of dopants (like WO3, CeO2, MgO, Nd2O3, Bi2O3, SrO, and Y2O3) with different weight percentages on the densification and microwave dielectric properties of SrLa4Ti4O15 was studied. Addition of dopants decreased the sintering temperature from 1625 to 1600 °C. Pure SrLa4Ti4O15 has a dielectric constant (ϵr) of 47.6, quality factor
(Q x f) = 37,300 GHz and \( f' = 91 \) ppm mol. The doped SrLa4Ti4O15 show in the range 46.6 - 48.1 quality factor (Q) in the range 36000-42000 kHz and \( f' \) in the range -6.6 to -9.8 ppm mol. The quality factor was increased for samples such as W03, CoO2, MoO3, Ni2O3, Bi2O3, Sb2O3. The dielectric constant is increased for doped SrLa4Ti4O15 except for MoO3 and Bi2O3. In general the \( f' \) is decreased with the addition of dopants.

ACKNOWLEDGEMENT

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REFERENCES

PREPARATION AND DIELECTRIC PROPERTIES OF COMPOSITE SUBSTANCES WITH VARYING FILLER CONCENTRATION


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Abstract: The preparation, microstructural evaluation and dielectric properties of ceramic filled PTFE substrates have been studied. Five powders of rutile are used as particular filler materials. Sintering soaking has been employed using polystyrene foam preforms (PTMS) to preclude the hydrophilic nature of the filler material. Void content in the thermostabilized composite substrates has been evaluated using X-ray diffraction measurements. A co-axial reflection method approach has been employed to theoretically derive the effective dielectric constant of the composite substrates and validated using experimental results.

INTRODUCTION

New materials with high dielectric constant and low-loss tangent are needed in electronic industry for transmission of electrical signals at microwave frequencies as a means to enable the miniaturization of the electronics. Many polymeric composite materials are presently available for the aforementioned applications. Present among these materials are composite systems based on polytetrafluoroethylene (PTFE) and other fluoropolymers such as polytetrafluoroethylene/kieselguhr/porcelain etc. Composites of the above (CMF) filled polytetrafluoroethylene (PTFE) chat with copper are attractive candidate materials for printed circuit boards (PCB) and integrated circuits (IC) chip carriers primarily due to their unique combination of mechanical, chemical and dielectric properties [1-3].

The PTFE/Cu based composite substrates show high dielectric constant, low-loss tangent, low costliness of thermal expansion, as well as chemical inertness. However, if untreated ceramic is used as filler, the resultant composites have unacceptable high porosity, high water uptake in humid environments, and poor mechanical integrity. It is reported that the treatment of glass, ceramic or silicon organic functional coupling agents on ceramic filler materials improves the mechanical and electrical properties and prescribes the filler from moisture absorption [4-6]. In the present study, PTFE/Cu based composite substrates have been prepared and dielectric properties are correlated with the particular filler concentration.

EXPERIMENTAL

An optimum dispersion of polytetrafluoroethylene with 60 weight percent solid content is used as the polymer matrix. The amount of particular filler material, polymer matrix and microfiber glass is selected based on the theoretical modeling. Appropriate quantity of aqueous PTFE dispersion is mixed thoroughly with particular ceramic filler and microfiber glass. Porous case rutile with 3 µgr/cubic cm size are used as the filler material. The ceramic filler material is coated with a silicon coupling agent, which reduces the surface of the ceramic.
hydrophobic and provides improved resistance to moisture absorption, tensile strength, peel strength and dimensional stability. Sol-gel process is employed for silane coating of the rutile particles using Phenyl trimethoxysilane (PTMS) as the precursor. The coupling agent (1 wt%) phenyltrimethoxysilane was hydrolyzed in formic acid and lown free water with a pH of 3. The TiO2 particles were added into the hydrolyzed coupling agent solution and the particles are well dispersed in the solution ultrasonically. The coated powders were further dried in an oven at 150°C. The slurry made out of colloidal PTFE, silane coated rutile and microfibre glass are mixed thoroughly using heavy duty stirrer for 30 minutes. In order to agglomerate the mixture of polymer, ceramic and fiber, polyethyleneimine is added as a flocculating agent. The liquid is removed from the agglomerated materials by filtering and the filtrate is dried at 100°C. The dough thus obtained is ram extruded to form preforms of 4mm thickness. Calendering of the lubricated preforms has been employed to obtain thin uniaxial tape (≤100 µ) with fine filler distribution. Thermalisation of the calendered green tapes has been done at a temperature of 350°C with an initial pressure of 120 Kg/cm².

Low frequency dielectric properties of the thermally laminated substrates have been measured up to 13 MHz region using an HP 4192 Impedance Analyzer. Poreosity measurements of the composite substrates have been done using CE Instruments make Hg pycnometer and Philips make Scanning Electron Microscope is used for the microstructural analysis.

RESULTS AND DISCUSSION

MICROSTRUCTURAL STUDIES

The dielectric properties of composites in general depend on the structure, crystallinity, morphology, and the presence of fillers or additives. The PTFE composites with different filler concentrations (up to 60 vol%) have been prepared using silane coated rutile particles. The scanning electron microscopic picture of a typical PTFE/TiO2 substrates is given Fig. 4.

![Fig. 1 Typical SEM pictures of a) silane coated rutile particle b) PTFE/TiO2 composite substrate](image)

It can be seen from the SEM picture of the composite substrate (Fig 1(b)) that the TiO2 particles are uniformly distributed out through the PTFE matrix. The PTFE/TiO2 interfacial strength and better dispersion of the filler are due to the coupling agent treatment.

THEORETICAL MODELLING

Precise prediction of the effective dielectric constant is a very important step for the design of composite materials. In mixture models the dielectric permittivity of a composite
material is expressed in terms of the dielectric permittivity of every constituent material and their volume fraction. There are several dielectric mixture models [7,8] and these formulae are derived on the basis of various theoretical assumptions and experimental data. Among these, the Lichtenecker-Kitto-Quekida or the 'logarithmic law of mixing' gained wide attention. The general formula for calculating permittivity of a two component system is as follows:

\[
\log \varepsilon = \log \varepsilon_1 + \frac{\varepsilon_2 - \varepsilon_1}{\log \varepsilon_2 - \log \varepsilon_1}
\]

or in general form for a mixture of m components,

\[
\log \varepsilon = \sum_{i} \frac{\varepsilon_i - \varepsilon}{\log \varepsilon_i - \log \varepsilon}
\]

Equation (2) in the Lichtenecker-Kitto-Quekida equation or logarithmic equation.

In the present case, theoretical modeling is employed using Lichtenecker-Kitto approach to arrive at suitable composite ratios, and theoretically predicted values are validated with experimental results. It is reported that a filler concentration of 50 to 60 vol.% is ideal of dimensional stability [9]. Hence, permittivities of the composite system in this range are theoretically calculated and results are compiled in Table 1. It can be seen from Table 1 and Figure 2 that there is a slight difference in the calculated and observed permittivities.

**Table 1. Calculated & observed dielectric constant of PTFE/TiO\textsubscript{2} composites**

<table>
<thead>
<tr>
<th>wt% of</th>
<th>Calculated (\varepsilon) (Lichtenecker-Kitto)</th>
<th>Experimental (\varepsilon)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>6.47</td>
<td>6.31</td>
</tr>
<tr>
<td>20</td>
<td>9.30</td>
<td>9.33</td>
</tr>
<tr>
<td>40</td>
<td>19.34</td>
<td>19.08</td>
</tr>
</tbody>
</table>

In order to understand the difference in the theoretically calculated and experimentally observed permittivities, the porosity of the thermolaminated substrates were measured using \(\varepsilon\) porometry. The pore size distribution of the thermolaminated PTFE/TiO\textsubscript{2} composites are given in Fig. 3. It is observed that the samples under study have a total porosity of about 3%. The slight difference in the calculated and observed permittivities of the PTFE/TiO\textsubscript{2} composites can be attributed to the presence of open porosity (3%) in the thermolaminated substrates. The particle size and morphology also affect the dielectric properties of composite systems.

**Fig. 2 PAO concentration Vs dielectric constant of composite substrates**

**Fig. 3. Pore size distribution of PTFE/TiO\textsubscript{2} with 3% porosity.**
The dielectric anisotropy is a very crucial property in composite substrates, which mostly depends on the distribution of the filler in the polymer matrix. In the present study, it is observed that the anisotropy in the dielectric properties are well within the allowed tolerance ($\varepsilon_1$, $\varepsilon_2$, $\Delta$). It further confirms that the processes methodology employed in the present case is adequate for the fabrication of PTFE/TOGS composite substrates with tight dielectric tolerance.

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(1-XMALA04 - XTiO2 (M=ZN, MG) DIELECTRICS FOR MICROWAVE SUBSTRATE APPLICATIONS

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Abstract: The dielectric properties of ZnAO4 and MgAO4 spinels were studied by preparing monocrystals mixtures with TiO2 in an effort to tune their resistivity constant. It was found that this low dielectric constant materials 0.83ZnAO4-0.17TiO2 and 0.75MgAO4-0.25TiO2 have excellent dielectric properties for possible applications in microelectronic packaging. The substrate characteristics of the new temperature stable low loss materials 0.83ZnAO4-0.17TiO2 which is even advantageous over alumina have been investigated.

INTRODUCTION

Modern electronics are based on integrated circuits and these circuits are built on insulating materials called substrates. The electrical isolating properties of dielectric ceramics have found extensive application as advanced ceramic materials for substrates and packages. The typical characteristics needed for a high density package substrate are: (a) low dielectric constant, (b) low dielectric loss and (c) matching coefficient of thermal expansion to that of the material attached to it. These substrate materials must exhibit high Q factors in order to maintain overall high Q levels by lowering power dissipation. Typical dielectric properties of some commonly used low-permittivity ceramic substrates such as alumina, show less reliable properties for the MIC applications.

A recent study by Nurendana et al., reported that ZnAO4 and MgAO4 are low loss dielectrics with high Q and negative temperature coefficient of resonant frequency. It is well known that TiO2 is a low loss material with positive ε and high dielectric constant. Hence it is possible to gather the microwave dielectric properties of the materials by making solid solution or mixture of positive ε material with a negative ε material. In this investigation we carried out a detailed study on the mechanical characteristics of ZnAO4 and MgAO4 with TiO2.

EXPERIMENTAL

The (1-X)MgAO4-xTiO2 ceramics were prepared by the conventional mixed oxide route. High purity ZnAO4 and TiO2 (Johnson Matthey Chemical Co.) were used as the starting materials for the synthesis of spinels. The calcination temperature was 1100°C for 4 hours. It was then ball milled with Anatase TiO2 (Ahlrich 99.9% pure) according to the formula (1-X)ZnAO4-xTiO2 (x=0.0, 0.1, 0.12, 0.14, 0.15, 0.17, 0.18, 0.19, 0.20, 0.25, 0.3, 0.4, 0.5, 0.6, 0.7, 0.9 and 1.0) for 24 hours using deionized water as the milling medium. The homogenized mixture was then pressed into cylindrical disks of about 14 mm diameter and 6.8 mm thickness. The sintering temperature is in the range 1375-1425°C for 4 hours. The well polished pellets were used for microwave measurements. The dielectric properties ε’ and ε” of the materials were measured as the microwave frequency range using a network analyzer HP 8510 C (Hewlett-Packard, Palo
RESULTS AND DISCUSSION

Fig. 1 Variation of ε of ZnAl2O4 (left) and MgAl2O4 (right) with mole fraction of TEG2 addition.

The dielectric constant of ZnAl2O4 is 8.5 while that of MgAl2O4 is measured to be 3.75. The dielectric constant of TEG2 is 105. The variations of the dielectric constant of ZnAl2O4 and MgAl2O4 with TEG2 addition is plotted in Fig. 1. The dielectric constant of a mixture can be calculated using the general Maxwell–Wagner form factor:

\[ \varepsilon = \varepsilon_1 + \varepsilon_2 \]

where \( \varepsilon_1 \) and \( \varepsilon_2 \) are the volume fraction and relative dielectric constant of the two materials.

Fig. 2 Variation of (Q) ε and dielectric frequency of (1-x)ZnAl2O4+ xTEG2 with (x) for x = 0.1, 0.3, 0.5, and 0.7 of MgAl2O4 with mole fraction of TEG2 addition (8H2O).

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Footer text:

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As shown in Fig. 2, the uncoated quality factor of pure ZnMgO is MgZnO and MgGaZnO is reasonably good (Qi is 56,000 and 68,000 GHz) which is well above that of many of the conventional low dielectric constant materials. The product, Qi x f is maximum (~1,40,260 GHz) around 0.17 in (x=ZnMgAlO=xT102) but close to 2 (d) where it assumes a minimum value (two). It is noteworthy to note that the microstructure for both shows rather better quality factors compared to the Nd phase components which is unusual in its dielectric mixtures. As the TiO2 concentration in MgZnO increases (see Fig. 2, right), the microstructure quality factor also increases reaching a maximum value of Qmax=105,400 GHz for x=0.25 in (x=ZnMgAlO=xT102). For higher TiO2 contents, the quality factor decreases.

Table 1. Comparison of properties of 0.83ZnMgO:0.17TiO2 with alumina and aluminium nitride nitride substrates

<table>
<thead>
<tr>
<th>Material</th>
<th>aluminium</th>
<th>aluminium nitride</th>
<th>ZnMgAlO:TiO2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Density</td>
<td>3.8</td>
<td>3.20</td>
<td>4.05</td>
</tr>
<tr>
<td>Dielectric Loss</td>
<td>0.004</td>
<td>0.001</td>
<td>0.004</td>
</tr>
<tr>
<td>Dielectric Constant</td>
<td>9.8</td>
<td>8.7</td>
<td>5.8</td>
</tr>
<tr>
<td>Electrical Conductivity (ohm cm)</td>
<td>210</td>
<td>213</td>
<td>59</td>
</tr>
<tr>
<td>Thermal Conductivity (W/m·K)</td>
<td>24</td>
<td>26</td>
<td>90</td>
</tr>
<tr>
<td>Thermal Expansion (ppm/°C)</td>
<td>6.0</td>
<td>7.6</td>
<td>3.2</td>
</tr>
</tbody>
</table>

The above discussion revealed that the compositions 0.83 ZnMgAlO:0.17TiO2 and 0.75MgZn:0.25TiO2 are ideal temperature stable, high Q dielectric resonators, which can be compared to many of the conventional low dielectric constant microwave dielectrics like Al2O3. In spite of the fact that alumina has high quality factor at microwave frequency ranges. Its ε' is 10 ppm/°C and being a well-known refractory, its processing temperature is high (>1700°C). The new substrates are intrinsically in all these aspects. Their substrate characteristics such as thermal conductivity, thermal expansion coefficient, electrical resistivity, temperature coefficient of the dielectric constant, mechanical strength and reactivity of the material with metallic silicon etc. of one of the material 0.83ZnMgAlO:0.17TiO2 are investigated and are given in Table 1.

The 0.83ZnMgAlO:0.17TiO2 substrate has a temperature coefficient of resistivity (TCR) of ~15 ppm/°C. The electrical insulating properties of this material is good: t<30 MΩ which provide better isolation between microelectronic components in integrated Chip Circuits. The TCR of TCO in the range 0.0-150°C, established the thermal chemical stability of this material. The thermal expansion coefficient of the composition 0.83ZnMgAlO:0.17TiO2 was measured in the range 530-197°C at 6.3 ppm/°C which is comparable to that of metallic silicon used in microelectronic circuitry. The thermal conductivity of this ceramic was measured at 59 W/m·K which is almost double the value of alumina substrate. A mixture of the newly developed substrate material with metallic silicon when heated at various temperatures did not form any additional phase with it apart from solid silicon formed due to the oxidation of silicon. These results indicate that the mixture ceramic 0.83ZnMgAlO:0.17TiO2 proves most of the qualification to be projected as an ideal substitute for alumina substrate which is commonly used in microelectronic packaging.

CONCLUSION

Two dielectric materials 0.83ZnMgAlO:0.17TiO2 and 0.75MgZn:0.25TiO2 were developed for microelectronic substrate applications. These temperature stable materials have high quality factor (Qmax = 100,000 GHz) and low dielectric constants. The 0.83ZnMgAlO:0.17TiO2 substrate has a comparable thermal expansion coefficient (6.3 ppm/°C) as that of

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alumina, but the thermal conductivity (59 Wm⁻¹K⁻¹) is more than twice of it. The 0.83%Al₂O₃+0.17TiO₂ substrate has a small value of ε (4.15 ppm/K) which illustrates its suitability in microwave integrated and microelectronics packaging circuits.

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Abstract: Ca$_5$Mg$_3$Nb$_2$Ti$_2$O$_{12}$ ceramics have been prepared through conventional solid-state ceramic route for $0 \leq x \leq 1$. The crystal structures of the ceramics were studied by X-Ray diffraction techniques and dielectric properties were measured at microwave frequencies (3 - 5 GHz). In Ca$_5$Mg$_3$Nb$_2$Ti$_2$O$_{12}$ system as $x$ increases from 0 to 1, or decreases from 48 to 38, Qx f increases from 28,000 GHz to 39,000 GHz and if decreased from 48 to 38 by a ppm of Ca. Ca$_4$Mg$_3$Nb$_2$Ti$_2$O$_{12}$ dielectric ceramics were found to have stable resonant frequency with temperature and are potential candidates for applications in personal and satellite communication systems in the S and C bands.

INTRODUCTION

Dielectric materials are continuing to play a vital role in the microwave telecommunication technology. These materials are key components in the realization of low-loss, temperature-stable resonators, filters, circulators and multi-layer circuit modules for satellite and broadcasting equipments and in many other microwave devices.1-4 Dielectric resonator (DR) materials are selected for miniaturization and better performance of wireless systems, as well as for the terminals and base-stations as well as for handsets.5 Complex perovskite (ADDS) type materials are an indispensable material base for microwave dielectrics with important properties such as high dielectric constant (εr), high unloaded quality factor (Qux) and near-zero temperature coefficient of resonant frequency (Δfr/ΔT). Recently Bijumon et al.3 reported the microwave dielectric properties of CaSrNb2TiO12 (CaSrCa$_4$Sr$_2$SrTiO$_{12}$) in complex perovskite form (CPP) ceramics with ε = 48, $Q$ x f > 28,000 at 45.8 GHz and Δfr/ΔT = 740 ppm/°C, when the samples were sintered at 1550°C/hr. More recently, this material is reported to be used for the bandwidth enhancement of DM loaded microstrip patch antennas and for the fabrication of wide band dielectric resonator antennas.6,7 Efforts have also been made to tailor the microwave dielectric properties of these materials by doping and solid solution formation.8,9 High dielectric constant along with relatively high quality factor makes these ceramics compact enough to use as DRs. But the relatively high if value precludes their use in practical circuits. Solid solution phase formation between compounds with positive and negative if values is reported as an effective method to tailor the dielectric properties of...
polycrystalline ceramics. With this point of view, effort has been made to synthesize and characterize Ca$_x$Mg$_{2-x}$Ni$_2$TiO$_4$ (0 ≤ x ≤ 1) with improved microwave dielectric properties. In this study the dielectric properties of Ca$_x$Mg$_{2-x}$Ni$_2$TiO$_4$ ceramics at microwave frequencies were investigated as a function of Mg$_2$O content for x varying from 0 to 1, to see the effect values near to zero.

EXPERIMENTAL

The starting materials for the synthesis of polycrystalline ceramics employed in this study were high purity CaCO$_3$, (MgCO$_3$)$_2$, Mg(OH)$_2$, MgO, TiO$_2$ (Alrich electronic) and NiSO$_4$ (NPC, India) powders. The powders were strategically weighed as per the formula Ca$_x$Mg$_{2-x}$Ni$_2$TiO$_4$ for x = 0, 0.2, 0.4, 0.6, 0.8 and 1. The mixtures were ball milled in polyethylene bottles using 0.08 lb in de-ionized water for 24h. The slurry was dried and grounded well. The powders were fired in air in platinum crucibles at 1300°C for 4h. The calcined powders were well ground in an agate mortar and pestle for 2h. It is then mixed with 4 weight % of poly vinyl butyral as binder, dried and again ground into fine powder. Polycrystalline pellets of 18mm diameter and 7mm height were pressed under a pressure of 150Mg/cm$^2$. The ceramic disks were sintered in air at 1550°C for 4h on platinum plates.

The phase purity of the samples was examined by X-ray diffraction (XRD) method using CuK$_a$ radiation. Well polished ceramic samples were used for microwave dielectric measurements. A transmission/reflection network analyzer (Agilent E5071B) was used for microwave characterization of the materials. The dielectric constant was measured by Hake-Coleman post resonator method by exciting the TE$_{011}$ resonant mode of the DR using the electric dipole of an antenna as suggested by Courtyard. The quality factor was measured by using the TE$_{011}$ mode in a well polished copper cavity 8. The value of dielectric constant obtained by careful absorbed method is verified using same method. The temperature variation of TE$_{011}$ mode is noted at every 5°C interval to calculate the temperature coefficient of resonant frequency.

RESULTS AND DISCUSSION

Careful X-ray diffraction (XRD) study of Ca$_x$Mg$_{2-x}$Ni$_2$TiO$_4$ ceramics revealed that the unit cell has an orthorhombic symmetry and belongs to Prima space group 31. The XRD pattern of Ca$_x$Mg$_{2-x}$Ni$_2$TiO$_4$ specimen for x = 0 to 1 exhibited the same structural that of the parent mineral, throughout the entire compositional range except a slight shift in the position of diffraction peaks. It is evident from this fact that, a solid solution is formed for all ratios of Mg$_2$+ fraction in the Ca$_x$Mg$_{2-x}$Ni$_2$TiO$_4$ system. The slight shift of diffraction peaks to the higher angle region with mole fraction of Mg$_2$+ is an indication of decrease in lattice parameters and a resultant enhancement in the theoretical density of the ceramics. The variation of experimental density with mole fraction of Mg$_2$+ ions in Ca$_x$Mg$_{2-x}$Ni$_2$TiO$_4$ (0 ≤ x ≤ 1) is shown in Fig. 1. Pure Ca$_x$Mg$_{2-x}$Ni$_2$TiO$_4$ has an experimental density of 4.06g/cm$^3$ (69%) when sintered at 1550°C. It can be seen from the figure that the concentration of Mg$_2$+ ions in polycrystalline pellets of 18mm diameter and 7mm height were pressed under a pressure of 150Mg/cm$^3$ when x = 1. This increase in density is due to the substitution of Mg$_2$+ ions in place of Ca$_2+$ in the provokable B-site. Smaller tensile radius of Mg ions as compared to Ca ions lowers the unit cell volume as observed from the
XRD patterns and results in the enhancement of density. Fig. 2 shows the variation of dielectric constant in Ca$_2$Mg$_{1-y}$Ni$_y$TiO$_3$ (0 ≤ y ≤ 1) ceramics with y. When $x$ is varied from 0 to 1, it takes the value from 48 to 58. This decrease in $x$ is due to the lower polarizability of Mg ions as compared to Ca ions, because polarizability plays a major role in controlling the dielectric constant of the ceramics. Moreover most of the Mg-based ceramics possess lower $x$ compared to Ca-based dielectrics and also in agreement with earlier reports.

![Dielectric Constant vs Mol fraction of Mg$^2+$ ion](image1)

Fig. 1 Variation of experimental density of Ca$_2$Mg$_{1-y}$Ni$_y$TiO$_3$ (0 ≤ y ≤ 1) with y.

![Dielectric Constant vs Mol fraction of Mg$^2+$ ion](image2)

Fig. 2 Variation of dielectric constant of Ca$_2$Mg$_{1-y}$Ni$_y$TiO$_3$ (0 ≤ y ≤ 1) with y.

![Dielectric Constant vs Mol fraction of Mg$^2+$ ion](image3)

Fig. 3 Variation of $Q_{\mu}$ with $x$ in Ca$_2$Mg$_2$Ni$_3$TiO$_9$ (0 ≤ x ≤ 1) ceramics.

![Dielectric Constant vs Mol fraction of Mg$^2+$ ion](image4)

Fig. 4 Variation of $Q_{\mu}$ with $x$ in Ca$_2$Mg$_2$Ni$_3$TiO$_9$ (0 ≤ x ≤ 1) ceramics. Bigger Ca ion increased ordering and hence increased the quality factor. Enhanced...

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densification with the substitution of Mg also contributed to the increase in quality factor. From figure 4 it can be seen that Qf decreased from 40 to 33 ppm at x ≈ 0 and varied from 0 to 1. The zero crossing of A occurs at x = x<sub>d</sub>. The Ca-35Mg0.6Ni2Ti0.12 was found to be the temperature compensated composition and has a = 41, Qf x f = 3 300 GHz and γf = 0.

CONCLUSION

Temperature stable complex perovskite Ca5-4Mg3xNb2Ti0.12 (0 ≤ x ≤ 1) ceramics have been prepared through mixed oxide route. The samples were characterized at microwave frequency range. X-ray diffraction analysis revealed no additional phase formation. The εf and γf decreased with increase in molar fraction of Mg substitution whereas Qf was found to be increasing with increase in x. The composition with x = 0.65 in Ca5-4Mg3x.4Ni2Ti0.12 was found to have stable resonant frequency with temperature. Ca5-35Mg0.6 Ni2Ti0.12 has a = 41, Qf x f = 3 300 GHz and γf = 0. Since this low-loss material has the desired characteristics for microwave resonators, it can serve as a potential candidate for use in basic station filtering equipment and dielectric resonator antennas in wireless communications devices.

ACKNOWLEDGEMENT

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BROAD BAND ULTRA HIGH STRENGTH RADOME FOR UNDER WATER APPLICATIONS
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Abstract: A broadband ultra high strength C-sandwich radomes is designed and developed covering multi-octave frequency range of 1-18 GHz for underwater applications. The measured average transmission through the radome is better than 90% over the full frequency range. The radome withstands the water pressure of 60 bar which is crucial for underwater applications. The radome possesses ultra high strength at the same time maintained required electromagnetic transparency. This paper illustrates the design philosophy, analysis and test results obtained on a C-Sandwich radome covering 1-18 GHz.

INTRODUCTION
Electronic Warfare systems require antenna systems operating over wider bandwidth with uniform electrical performance. The performance of an E.W system depends greatly on the performance of its antenna, which in turn depends on the radome housing it. Radomes designed for electronic warfare systems are more than just pieces of pigskin which protect the antennas [1] from environmental hazards, without affecting their electrical performance. To achieve high strength, Radomes with conventional designs are heavier operating over narrow band of frequencies. Radomes for underwater applications require high structural strength which implies a thicker radome. But at the same time it should be electromagnetically transparent with minimum loss and achieve uniform electrical performance operating over wider bandwidth. Hence the design of radome becomes critical after considering structural and the electrical needs of the system. In addition, the radome should be capable of withstanding hostile (severe) environmental conditions of underwater applications. Shape & Size of the radome, material selection and fabrication techniques play a key role in the design of broadband radomes for underwater applications. The radome realizes a wideband antennal covering frequency range of 1-18 GHz and possess high mechanical strength at the same time maintaining low loss and protect the antennas from several environmental conditions.”

DESIGN PHILOSOPHY
Based on the structure, radomes are broadly classified as this wall, solid wall and sandwich radomes. The thickness of the thinnest radome is chosen as 1/20th of λ at it is a highest frequency. Even though thin wall radomes is having low loss operating over broadband of frequencies, it has poor mechanical strength, especially after X-band. In solid wall radomes, to possess high strength, thickness is chosen as solid multiples of 3/2, but it can be operated over narrow bandwidth. In order to have both high strength and ultra broadband characteristics, sandwich type of radomes are employed. These are five types namely “A”, “B”, “C”, and multi-layer type [2]. C Sandwich radome consists of: 3 layers with (low density) low dielectric core material sandwiched between high dielectric skin materials. The skin thickness is generally chosen as 1/20th where λ₃ is at highest frequency. The thickness of core material is normally a quarter wavelength or odd multiples thereof.
THEORETICAL ANALYSIS

The analysis of plane wave propagation through thin, parallel-layered dielectric of infinite breadth is based on the radome design [3,4]. When a plane wave is incident on a multi-layered dielectric interface, a part of it is transmitted and a part of it is reflected as shown in Fig.1. When the position that was transmitted strikes the next surface, the process is repeated. At each interface, the path lengths and transmission angles are calculated using optical ray tracing as shown in Fig.2. The equivalent impedance for each layer is calculated using the path length and angle of transmission. Reflection and transmission coefficients are calculated for parallel and perpendicular components, respectively using the formulae given below:

\[ \rho = \frac{Z_2 \cos \theta - Z_1 \cos \varphi}{Z_2 \cos \theta + Z_1 \cos \varphi} \]
\[ \tau = \frac{Z_2 \sin \varphi - Z_1 \sin \theta}{Z_2 \sin \varphi + Z_1 \sin \theta} \]

Where \( Z_1 \) = Free space impedance, 377 ohms
\( Z_2 \) = Impedance of dielectric layer
\( \theta \) = Angle of incidence
\( \varphi \) = Angle of transmission
\( \rho \), \( \rho_1 \) = Reflection coefficients for parallel & perpendicular polarizations respectively.
\( \tau \), \( \tau_1 \) = Transmission coefficients for parallel & perpendicular polarizations respectively.

A computer program is developed in 'MATLAB' software to determine the transmission and reflection coefficients of multilayered radomes using the equation mentioned above. This program is used for optimizing the acer and skin thickness for the given radome electrical and structural requirements. The inputs to the program are number of layers, material characteristics and thickness of each layer.

SHAPE OF THE RADOME

In general shape of radomes such as hemispherical, Conical, Ogive and Spheroid types can be used. In the present design, Hemispherical shape is chosen because the ray insidnets normally on the surface thereby provides minimum transmission loss. The dimensional details of Hemispherical 'C' shape radome are shown in Fig.3.

MATERIAL SELECTION

In the design of ultra broad band radomes, the material selection plays a key role [5]. The material should have good electromagnetic transparency characteristics and adequate mechanical strength. In the present design of radomes, GFRP is chosen as the skin material and syntactic foam as core material. GFRP is chosen because it has optimum electrical characteristics and moderate strength & stiffness properties. Systactic foam is a combination of 5 to 25 microns dia. hollow micro glass balloons mixed with silica resin. It is the state-of-the-art core material because of its excellent features like low dielectric constant, high contact area, light weight, high strength and low loss. Dielectric constants of GFRP and syntactic foam are 4.2 and 1.6 respectively.
FABRICATION

Using the proposed developed, theoretical studies were carried out for various core thicknesses and for constant skin thickness. From these computed values an optimum skin thickness of 0.4 mm and core thickness of 5.9 mm was chosen. Before fabricating the actual shaped radomes, a C-Sandwich panel of 100 mm x 100 mm is fabricated and transmission and reflection loss was measured over the entire frequency range. There are various methods of radome fabrication such as Hand laying, Mached dye, Vacuum bag, Anto Clave and Filament winding. The selection of a manufacturing method for a given radome design is based on a number of factors including the radome performance requirements and materials and construction. Hemispherical C-sandwich radome is fabricated using machined dye moulding because it gives accurate and repeatable products of high strength and catchless homogeneity.

ELECTRICAL EVALUATION

Reflection loss characteristics of the radome are measured over 1-18 GHz using Network Analyzer. The measured swept frequency band of lens made of hemispherical radome is shown in Fig.4. Radiation pattern measurements are carried out in an Anechoic chamber, by covering the antenna with radome (Ref Fig.5).

RESULTS AND DISCUSSION

The radome skin and core thickness are optimised experimentally using theoretical results. By taking these optimized thickness parameters, radomes were fabricated using machined dye technique. Pressure tests were carried out on the radomes by using pressure caps. The radomes withstand the pressure up to 60 bars. The measured average transmission is better than 98% over the entire frequency band of operation. The beam width variation observed is within ±15%. The main beam ripples, side lobe levels, bore sight errors are within tolerable limits.

CONCLUSIONS

In this paper we describe the analysis and test results obtained on C-Sandwiches are presented. The radome exhibited good electromagnetic transparency at the same time maintained ultra high strength. Development of this radome is very crucial for underwater applications.

ACKNOWLEDGEMENTS

The authors are grateful to Sri C. Komara Swamy Rao, Director, DEER, for his constant support and encouragement. We are grateful to Dr. T. Ramesha Rao & Dr. V.C. More for their technical guidance during the development of this radome. We are thankful to Sri T.N. Yadagiri Rao for giving us the opportunity to work on this development activity. We are thankful to Smt. G. Disha Kiran, N. Siva Kumar & K. Ramesh for carrying out various measurements.

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Fig. 1. Diagram of mutual boundary

Fig. 2. Generalized Distribution of C"Nannofossil Biozone

Fig. 3. Selected Features of Hbhi - Nannofossil Biozone
APPLICATION OF METAMATERIALS FOR DEFENCE AND COMMUNICATION

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Abstract: Rapidly increasing interest in the left-handed materials (LHM) predicted that certain man-made composite structure would possess, In a given frequency interval, a negative effective magnetic permeability $\mu_e$. Combination of such a structure with a negative effective permittivity medium enabled the construction of metamaterials with both effective permittivity and permeability negative. In this paper the possibilities of Metamaterials application is discussed with special emphasis for the defence and communication.

INTRODUCTION

Most of the laws of physics that the novice designers are taught have been changed for close to 100 years, LHM, first introduced by Veselago (1967) [1], has an unusual EM property the direction of the electric flux is opposite to the wave vector [2]. It follows that Snell's law [3] for reflection at the interface between LHM and regular material has a negative sign [3,4,5]. Metamaterials consist of a lattice of conducting, nonmagnetic, ferromagnetic, dielectric, or metallic nanoscale objects that can be described by an effective magnetic permeability $\mu_e$ and an effective electrical permittivity $\varepsilon_e$, both of which can exhibit values not found in metals or naturally occurring materials. Because the electromagnetic fields in conducting materials can be localized to regions much smaller than the incident wavelength, which can be difficult to perform accurate numerical simulations. The intriguing physical properties of metamaterials are $\mu_e$, $\varepsilon_e$, frequency dispersion [6] and $\mu_e$ and $\varepsilon_e$ are negative over a frequency band. The index of refraction is negative, such as the antiparallelism of the phase velocity and the group velocity [7], can be exploited for several important technological applications.

TIEORY

Model for $\varepsilon_e$:

The dielectric constant $\varepsilon_e$ for metamaterials (in the periodic arrangement of rods) is given by the expression:

$$\varepsilon_e = 1 - \frac{\alpha}{\omega^2 + i\omega \tau}$$

where $\alpha$ is negative over a frequency band, $\tau$ is time delay due to electric fields.

Model for $\mu_e$:

The permeability $\mu_e$ of metamaterials (in the periodic arrangement of rods) is given by the expression.

$$\mu_e = 1 - \frac{\alpha}{\omega^2 + i\omega \tau}$$
\[ \epsilon_r = 1 - \frac{F e_d}{\epsilon_0^2 - \epsilon_m^2} \frac{e_d}{V_e \theta_0} \] (1)

Where \( F \) is the fractional area of the unit cell occupied by the interface of the split ring (F1) and \( \theta_e, \theta_m \) are the losses due to electric and magnetic fields.

In the lossless case we can with \( \theta_e = 0, \theta_m = 0 \) and the models for \( \epsilon_r \) and \( \lambda_0 \) is changed accordingly:

\[ \epsilon_r = \frac{\epsilon_0^2 - \epsilon_m^2}{\epsilon_0^2} \]

\[ \mu_r = \frac{\epsilon_0^2 - \epsilon_m^2}{\epsilon_0^2} F e_d \frac{1 - F}{\epsilon_0} \]

Where \( \epsilon_0 = \epsilon_0 \sqrt{1 - \frac{\epsilon_0}{\epsilon_0}} \)

Therefore

\[ k^2 = \frac{\epsilon_0^2}{\epsilon_m^2} \frac{1}{1 - \frac{\epsilon_0}{\epsilon_0}} \left( \frac{\epsilon_0^2 - \epsilon_0^2 \epsilon_0^2}{\epsilon_0^2 - \epsilon_m^2} \right) \]

After identifying the regions where \( \epsilon_r \) and \( \lambda_0 \) changes sign we can get the relation for \( k \).

**APPLICATION**

Looking at some of the metamaterials concepts as an antenna science, one will start thinking about antennas with very low levels of interference. It will be very useful to have an antenna that received a signal from only one direction and eliminated signals from every other direction. A fundamental, groundbreaking aspect can be that with such an antenna, there is a possibility of overcoming the diffraction limit. This is what limits the resolution of any optical systems. It is not possible to resolve the details of anything smaller than the wavelength of the electromagnetic wave used to examine it. The present-day demand on multi
function monitors requiring high bandwidth. Antennas using metamaterials vary have the potential of multi-directional capability, which need to be studied.

The choice of using meta-material is due to the possibility of realizing broad-band [3] (multi-frequency) antennas. The other advantages of using multi-material are: (a) it has the full design control because one can get the desired or and μ, (b) it reduces the cost and simplicity. The expected application, which may come through, includes possibilities of producing multi-frequency antenna, using metamaterials in place of dielectric. Hence, speed spectrum technology can be easily deployed for jamming and anti-jamming applications. There is a possibility of controlling the directionality of the antenna and hence steering/motion control of the desired direction. This can lead to potential space diversity applications, Bandwidth increase is possible via metamaterial design for fixed size soli antenna, which can lead to minimization of efficient antennas for deployment in missiles, and battlefield vehicles. When used in thin antennas [10], an enhancement is expected to enhance compared to normal dielectric antennas, which can prove in collector strategic data in finer details for analysis. It is expected to project a low Radar Cross Section compared to naturally occurring materials. It has already been established that it shows negative Doppler effect [11], which can have positive application in Radar Technology.

CONCLUSION

In this paper we have studied the basic theory and the possible use of metamaterials for different defense and communication applications. The use of metamaterials may lead to Miniaturization of Antennas, Multi-frequency antenna, Jamming and anti-jamming, learning the RC of antennas etc.

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INVITED TALKS
INVITED TALK 1

USE OF HIGHER ORDER BASIS IN SOLUTION OF ELECTROMAGNETIC FIELD PROBLEMS

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ABSTRACT: The objective of this paper is to illustrate that use of higher order basis functions provide faster convergence. This is also true for finite-element and also for time domain techniques. For integral equations, it guarantees continuity of the charge. Results are presented to illustrate this point.

WHAT IS A HIGHER ORDER BASIS?

A higher order basis function has a higher degree of continuity. For example, the pulse function is piecewise continuous function and is a polynomial of zero degree. The linear triangle function is a first order basis function as it is a polynomial of first degree. A higher order basis in this context will then deal with polynomials of degrees greater than one. We will deal with polynomials up to the ninth degree. Therefore, use of a higher order basis not only guarantees continuity of the function but also a few of its derivatives. However, we have to very careful in dealing with a higher order basis. This is because the charge is discontinuous at the feed point of an antenna and also at the end of the structure where the current with the appropriate orientation either goes to zero or has a singularity. Hence, the charge is discontinuous. We demonstrate in this paper that use of higher order basis over electrically large patch sizes offer a computational advantage as the number of unknowns scales quite moderately with size and frequency [1, 2]. This is true not only for the solution of the integral form of Maxwell’s equations but also for the differential form.

However, in using a higher order basis one has to be very careful as increasing the basis beyond a certain order may deteriorate the condition number of the matrix equation that needs to be solved for. Hence a compromise needs to be made between the choice of the order of the basis and the condition number of the matrix. It has been our experience that if the polynomials beyond the ninth order are not considered in the expansion, then the resulting matrix equations are quite stable and can be solved in an accurate fashion.
APPLICATION TO INTEGRAL EQUATIONS IN FREQUENCY DOMAIN:

Figure 1. Model of a 50 Ω parallel plate transmission line terminated with a 50 Ω load. Length = 17”, height = 1.00”, spacing = 0.16”. (a) Highest polynomial degree $1 \times 1$. (b) Highest polynomial degree $2 \times 2$. (c) Highest polynomial degree $2 \times 2$ for end plates, $2 \times 5$ for middle plates ($5$ along the long dimension).

<table>
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<th>Frequency (MHz)</th>
<th>Impedance (Ohms)</th>
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To illustrate the strength of the entire domain basis consider the analysis of the input impedance of a transmission line as illustrated in Fig. 1. In this problem, there is a strong coupling between the two plates as they are broadside coupled, where the size of the plates are much greater than their separation. To test whether large patches using higher order-basis over the two plates can deliver satisfactory accuracy, a 50 Ω parallel plate transmission line is modeled in Fig. 1. Here, we consider three different discretizations; a finely meshed version serves as a performance baseline (Fig. 1a). The mesh resolution is reduced in Fig. 1b, and finally, the transmission line is reduced to just 6 plates as illustrated in Fig. 1c. The transmission line is fed with a coaxial line source at one end, and terminated in 50 ohms at the other. The input impedance of the transmission line is computed at 7 frequencies between 100 and 700 MHz, and the results are summarized in Table I. The transmission line is 17 inches long overall and is more than one wavelength long at the highest frequency. The width of each transmission line is 1” and the spacing between the plates is 0.16”. For the refined mesh one chooses two terms in the expansion, a constant and a linear expansion on each square patch for each component of the current. For the second mesh, a total of three polynomial terms are used over each patch for each component of the current and for the last large mesh three terms are used as...
before for the transverse variation of the current and a sixth order polynomial is used along the longer direction. Regardless of the mesh size and density, the largest error in the input impedance is less than 2 ohms at any frequency computed using the electromagnetic analysis code WIPL-D [1].

As a second example consider the defect of an Electric Field Integral Equation (EFIE) at an Internal Resonance for closed bodies. It is well known that the Electric Field Integral Equation (EFIE) breaks down at a frequency which corresponds to an internal resonance of the closed structure formed by the closed equivalent surface. The current way to alleviate this internal resonance problem is to either use the combined field or the combined source formulation. These formulations essentially push the complex poles from the imaginary frequency axis into the complex plane. However, both of these formulations essentially double the workload by using both the electric and the magnetic field equations. What we illustrate is that the use of a higher order basis functions to the EFIE only, can practically achieve the same goal. Since, the magnetic field integral equation is not used, it is computationally quite advantageous. The higher order basis localizes the defect and unless one is very close to the pole like up to the second or the third place of decimal of the actual value for the frequency, the formulation does not break down. As a numerical example, let us consider the scattering from a closed conducting cube of side $a = 2\, \text{m}$. It is excited by a normally incident plane wave, whose electric field vector is collinear with one of the edges of the cube. The geometrical model of the cube is comprised of $M = 24$, $M = 54$, and $M = 96$ quadrilateral patches as shown in Fig. 2. When a first-order approximation is used for the currents over each patch, these models require $N = 48$, $N = 108$, and $N = 192$ unknowns, respectively. Figure 3 shows the monostatic RCS, normalized by $\lambda^2$, versus frequency. As expected, a spurious response appears in a frequency range near the resonant frequency. The plots 3A, 3B and 3C are obtained using a first order approximation for the two components of the currents. The responses exhibit spurious responses. By increasing the number of unknowns, the width of the frequency range over which the spurious solution exists decreases and the position of the resonant frequency is moved toward the theoretical value. This frequency range can be further decreased if the second-order approximation is used for the currents to the model with $M = 24$ patches, resulting in $N = 192$ unknowns. When the cube is modeled by $M = 24$ patches and a third-order approximation is used for each component of the currents over each patch, resulting in $N = 192$ unknowns, the frequency range of the defect is very narrow as shown by Fig. 3D. The shape is so narrow that it can hardly be detected, unless the frequency is put in with several significant decimal places.

![Figure 2](image)

(a) $M=24$, $N=48$  
(b) $M=54$, $N=108$  
(c) $M=96$, $N=192$

**Figure 2:** Discretisation of a cube in three different ways using different order of polynomial expansions.
Figure 3. Monostatic RCS of the cubical scatterer in the vicinity of the first internal resonance.

Next, we illustrate that use of a higher order basis can handle analysis of dielectric bodies from very low to very high permittivity. Let us consider a cubical scatterer placed in vacuum with edges collinear with the axes of the Cartesian coordinate system. The cube is excited by an incident plane wave propagating along the minus z-axis, whose electric field vector is collinear with the x-axis. The length of the side of the cube is $a = 0.2\lambda$, where $\lambda$ is the wavelength in vacuum. The cube is made of a dielectric whose relative permittivity is $1 - 100j, 1 - 1000j$ and $1 - 10000j$. A second-order approximation is used for the currents over each patch discretising the cube, resulting in $N = 96$ unknowns. Fig. 4 shows the bistatic RCS in xOz-plane versus the angle. The advantages of using electrically large patches are demonstrated next. Consider the scattering from a flat plate which is normally excited by a plane wave. Consider different sizes for the plate of lengths $\lambda/6, 3\lambda/6, 5\lambda/6, 7\lambda/6, 9\lambda/6$, and $11\lambda/6$. Very accurate results for the monostatic RCS can be obtained by using the following polynomial order of the approximation as 2, 3, 4, 5, 6, and 7. This results in a total number of unknowns which is equal to 4, 12, 24, 40, 60 and 84. The corresponding equivalent number of unknowns per wavelength squared is computed as 144, 48, 35, 30, 27, and 25. The advantages of the choice of a higher order basis now becomes clear. This is because as the patch size increases, the effective number of unknowns used to model the current on the surface actually decreases.

Thus the number of unknowns needed in the analysis depends on the size and complexity of the structure. In the case of electrically long wires this number is 3-4 per wavelength, while for electrically large metallic surfaces this number is 10-20 unknowns per wavelength squared of surface area. As an example let us consider the bistatic RCS of a PEC cube of side $10\lambda$. $\theta$ is measured from xOy-plane. In the analysis two symmetry planes are used. Fig. 5 plots the results obtained by a) $N = 1728$ unknowns (11 unknowns per wavelength squared), and b) $N = 3072$ (20 unknowns per wavelength squared). The corresponding analysis time is approximately $T = 40$ secs and $T = 120$ secs on a laptop.
INSPIRON 5150. The results are almost identical. The results given on Fig. 5b practically do not change by increasing the number of unknowns. When we do not consider any symmetry planes, the number of unknowns becomes 9408, which takes about an hour to solve. Therefore if we consider a cube of $20\lambda$ on the side, the piecewise triangular patch basis functions without any symmetry will require $2\times6\times400\times300 = 1.44\times10^6$ (two components of the current×six faces×surface area per face×approximately 10 unknowns per wavelength) whereas use of an entire domain basis for the same problem like in WIPL-D would take 36,300 unknowns and it can be solved on a 64bit ITANIUM-1 system in approximately less than a day.

**Figure 4.** Bistatic RCS of a cubical scatterer of side $a = 0.2\lambda$, $\sigma/\lambda^2$ versus $\theta$. A third order approximation for the current results in $N=96$.

**Figure 5.** Bistatic RCS of a cube of side $10\lambda$ obtained by a) $N = 1728$ unknowns and b) $N = 3072$ unknowns.

Fig. 6 shows the geometrical models of a full-scale aircraft model made of a) electrically small (relatively irregular) quadrilaterals, and b) of electrically large (relatively
regular) quadrilaterals. The first model requires more than 50% more unknowns than the second one. Fig. 7 shows the bistatic RCS of the aircraft in the xOz-plane (normalized by its maximum value of 27.05 dB), versus angle $\theta$ measured from xOy-plane. For calculations the airplane is placed in the xOz-plane, with its axis oriented along the z-axis. The excitation is a plane wave propagating along the y-axis. One symmetry plane has been used in this analysis. The results are obtained using a) $N = 2711$ unknowns, and b) $N = 4180$, respectively and the field patterns are very similar.

![Geometrical models of a real aircraft](image)

**Figure 6.** Geometrical models of a real aircraft made of a) electrically small (relatively irregular) quadrilaterals, and b) of electrically large (relatively regular) quadrilaterals.

![Bistatic RCS](image)

**Figure 7.** Bistatic RCS in the xOz-plane (normalized by its maximum value of 27.05 dB), versus angle $\theta$ measured from the xOy-plane. For calculations the airplane is placed in the xOz-plane, with the axis oriented along the z-axis. The excitation is a plane wave propagating along the y-axis. The results are obtained using a) $N = 2711$ unknowns, and b) $N = 4180$, respectively.

The next question that arises is, how does this technique scale with frequency when compared to other techniques like the fast multipole method? If we assume that a fast multipole scales as $O(N)$ then as the frequency is doubled, the number of unknowns increase
by a factor of 2 for each side and hence for a surface area, this method will typically scale as \(O(4N)\). To analyze the aircraft model shown in Fig. 6b, the entire domain basis takes 7460 unknowns at 600 MHz. And at 1 GHz, the WIPL-D model requires 9498 unknowns using quadrilateral patches to analyze the structure. Hence, for this problem the fast multipole method will scale as \((1/0.6)^2 = 2.78\), whereas using the higher order basis the problem will scale as \((9498/7460)^3 = 2.05\). Hence the use of entire domain basis is well tailored for the analysis of radiation and antenna problems as a fast multipole method may not be very accurate for handling radiation problems where calculation of near fields and input impedance are of utmost importance.

This methodology has also been applied to the analysis of wire structures in the frequency domain and similar type of conclusions may also be drawn [4]. For analysis of wire antennas use of a polynomial basis over large patch sizes indeed results in a fewer number of unknowns than using the conventional sub-sectional basis and it is computationally more efficient.

**APPLICATION TO FINITE ELEMENT METHOD IN FREQUENCY DOMAIN:**

Use of higher order basis has advantages also with the solution of differential equations. As an example consider the calculation of the capacitance of a square coaxial cylinder as shown in Fig. 8. In Table 2 the computed values of the capacitance are given as a function of the different orders of the basis functions. It is seen that use of a higher order basis indeed does provide a faster convergence in the computation of the value of the charge distribution for the transmission line, where the value of the capacitance is calculated by solving for the static charge distribution using the finite element method.

**TABLE 2:** Capacitance for different orders of the basis

<table>
<thead>
<tr>
<th>Order</th>
<th>Mesh steps</th>
<th>No. elements</th>
<th>(C_{el} \times 10^{-12})</th>
<th>Relative error (%)</th>
<th>Run-time (sec)</th>
</tr>
</thead>
<tbody>
<tr>
<td>First o.</td>
<td>1=3 e.</td>
<td>5</td>
<td>12</td>
<td>17.25</td>
<td>0.99</td>
</tr>
<tr>
<td></td>
<td>2=12 e.</td>
<td>12</td>
<td>10.844444</td>
<td>5.96</td>
<td>0.99</td>
</tr>
<tr>
<td></td>
<td>3=48 e.</td>
<td>35</td>
<td>10.46224</td>
<td>2.23</td>
<td>1.21</td>
</tr>
<tr>
<td></td>
<td>4=192 e.</td>
<td>117</td>
<td>10.32194</td>
<td>0.86</td>
<td>2.14</td>
</tr>
<tr>
<td></td>
<td>5=768 e.</td>
<td>425</td>
<td>10.20837</td>
<td>0.33</td>
<td>16.25</td>
</tr>
<tr>
<td>Second o.</td>
<td>1=3 e.</td>
<td>12</td>
<td>10.45926</td>
<td>2.2</td>
<td>1.05</td>
</tr>
<tr>
<td></td>
<td>2=12 e.</td>
<td>35</td>
<td>10.32649</td>
<td>0.903</td>
<td>1.14</td>
</tr>
<tr>
<td></td>
<td>3=48 e.</td>
<td>117</td>
<td>10.27083</td>
<td>0.35</td>
<td>1.70</td>
</tr>
<tr>
<td></td>
<td>4=192 e.</td>
<td>425</td>
<td>10.24867</td>
<td>0.14</td>
<td>18.57</td>
</tr>
<tr>
<td>Third o.</td>
<td>1=3 e.</td>
<td>19</td>
<td>10.39838</td>
<td>1.61</td>
<td>1.15</td>
</tr>
<tr>
<td>Serendipity</td>
<td>2=12 e.</td>
<td>58</td>
<td>10.29764</td>
<td>0.62</td>
<td>1.31</td>
</tr>
<tr>
<td></td>
<td>3=48 e.</td>
<td>199</td>
<td>10.25933</td>
<td>0.24</td>
<td>3.52</td>
</tr>
<tr>
<td>Third o.</td>
<td>1=3 e.</td>
<td>22</td>
<td>10.32855</td>
<td>0.92</td>
<td>1.15</td>
</tr>
<tr>
<td>Complete</td>
<td>2=12 e.</td>
<td>70</td>
<td>10.27111</td>
<td>0.36</td>
<td>1.42</td>
</tr>
<tr>
<td></td>
<td>3=48 e.</td>
<td>247</td>
<td>10.24875</td>
<td>0.14</td>
<td>5.00</td>
</tr>
</tbody>
</table>

**APPLICATION TO TIME DOMAIN NUMERICAL TECHNIQUES:**

The time domain methods are derived from the wave equation which a hyperbolic partial differential equation and are conditionally stable. The initial value problems which arise from the solution of hyperbolic partial differential equations, often require the Courant
stability criteria for a stable numerical solution. The Courant stability criteria, relates the spatial discretization with the time discretization through the velocity of propagation. Even though there are implicit formulations for these hyperbolic partial differential equations, they also run into numerical instability problems. Use of an entire domain expansion for the time variable [6, 7] results in a formulation where in the final computations the time variable can be completely eliminated and thereby eliminating the bad properties of a hyperbolic partial differential equation. In addition, since the time variable is eliminated, there is no need for the Courant stability criteria. This is possible through the introduction of the Laguerre polynomials representing the time variable. In a finite-difference time domain (FDTD) formulation, the use of these polynomials can speed the solution process at least by a factor of one hundred [8]. In an integral equation formulation, use of these higher order basis can separate the time and the space variables resulting in a final solution procedure containing the spatial variables only.

**CONCLUSION:**

Use of a higher order basis can provide efficient numerical techniques for the solution of both integral and differential form of Maxwell’s equations. A brief overview has been given to illustrate the claims.

**REFERENCES**

INVITED TALK 2

RADIATION EFFECTS OF PORTABLE COMMUNICATION EQUIPMENT

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Abstract: The biological effects produced by man-made electromagnetic radiation are reviewed, in particular those produced by the now ubiquitous portable phones (cellular) and their associated base stations. Both thermal and non-thermal effects are considered, and safety standards for the maximum permissible exposure (MPE) and specific absorption rate (SAR) are presented.

Introduction In recent years, the number of mobile — or “cellular” — phones in operation has exploded in many countries. The total number of phones sold worldwide during the year 2004 is expected to exceed 620’000’000 (62 crore) units. In many countries there are now more cellular phones than fixed connections. This also means that an increasing number of fixed base stations must be installed, to ensure proper wireless service — and many phone users don’t like to see communications antennas sprout up in their close neighbourhood! This proliferation raises some concerns over potential biological hazards. This presentation aims to put the problem in its proper perspective, by reminding how various applications of electricity and of communications appeared during the 20th century, and at what point potential hazards were detected. The interactions between handset antennas and biological tissues will be evaluated, in terms of the specific absorption rate (SAR). Until now, no convincing evidence was found that cellular phones could be harmful — as long as they are properly operated. However a number of points must be clarified and research in the field goes on actively [1, 2].

Historical background. As soon as Edison’s first light-bulb was switched on, concerned citizens warned of the appalling dangers that the new invention would cause to mankind. When using electrical appliances, some precautions are needed — to avoid electrocution — but these are easily understood, even by young children, and electricity was soon used all over the world. Accidents still do happen, but they are practically always related to carelessness and failure to follow safety rules.

At the beginning of the 20th century, long distance radio started to develop. Huge antennas were erected, first to ensure transatlantic communications, later on regular broadcasting of radio programs, and then of television. At the beginning, the receiver’s sensitivity was very low, so that the transmitters had to “pump” a lot of electromagnetic power (at first mostly noise¹) into the heavens. Some people then absorbed considerable amounts of radiation, but apparently no one did worry about it at the time. Long distance effects of electricity are not

1 I was recently informed that it would practically not be possible to duplicate Marconi’s original experiments, because the resulting noise would now disrupt all communications within a wide area!
easily perceived by the general public, because one does not “see” how a current, carefully contained within an isolated conductor, could affect someone at any distance from it. The concepts of electromagnetic pollution and “electrosmg” only appeared rather recently.

Biological effects of microwave radiation were apparently first studied in relation with radar systems during Word War II. Radars installed on board of ships were found to produce heat and burns when operating personnel came too close. The heating produced by microwaves was then evaluated and safety measures set up — both for operating personnel and for the general public⁴. Theoretical and experimental studies were carried out, covering several frequency bands. They determined that a microwave power density up to 10 mW/cm² (100 W/m²) could be tolerated, even for a long period of time — the thermal (infrared) radiation from the sun can reach about 100 mW/cm² (1 kW/m²), i.e. ten times more.

Many operators, civil and military, worked for many years within this limit, and no adverse effects were noticed. But sometime during the 1970s, safety levels from 100 to 1000 times smaller were found in Russian standards! This caused a considerable commotion, in particular because reasons for this important discrepancy were not readily obvious. It was later on found that US standards referred to maximum radiation, whereas Russians considered an average over the whole body — i.e. the discrepancy was not as great as it did appear at first.

It was then claimed that the difference is due to the existence of “non-thermal” effects, which are not directly linked to the heating of tissues, but to more direct interactions between the electromagnetic field and body cells — such effects being to some extent comparable to those produced by radiation on sensitive electronic equipment. In principle, non-thermal effects could take place at lower power levels, but no conclusive evidence could be found that non-thermal effects can actually be harmful [3]. Non-thermal effects have been investigated over the past three decades, but many of the results obtained, while interesting, proved inconsistent and sometimes could not be replicated. However, making use of the “precautionary principle,” the safety limits for microwave power density were reduced, but not in the same way in different countries. Figure 1 shows the levels for two of the most popular recommendations, the ANSI/IEEE [4] in the United States and the ICNIRP/CENELEC standard of the International Commission on Non-ionizing Radiation Protection [5]. The reasons for the differences are outlined in [6].

It should also be mentioned that considerably higher power levels are commonly applied to patients during therapeutic treatments of a variety of illnesses (diathermy and hyperthermia).

Radiation Basics. Electromagnetic radiation, just like matter, possesses a dual personality: it can be considered either in its granular form, made of photons of energy \( W = hf \) (\( h \) = Planck’s constant, \( f \) = frequency), or in its wavy form (fields, Maxwell’s equations). Quite different interactions with living matter are associated with these two forms For extremely high frequencies, above the visible light range in figure 2, a single photon has enough energy to pull an electron out of an atom, and thus to ionize it, so that such radiation is called ionizing. Chemical changes may then take place, which may induce tan and sunburn (ultraviolet rays) or, for the harder radiation (X and gamma rays), mutations or cancer. At lower frequencies, in visible light, infrared, and all the frequency ranges used in communications, the energy of a single photon is too small to ionize an atom: the radiation is then called non-ionizing.

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² Microwave ovens also appeared around that time, as a byproduct of radar.
Figure 1  Maximum permissible exposure (MPE) for the power density in an uncontrolled (general public) environment.

Figure 2  The Electromagnetic Spectrum ($W$ = photonic energy, $T$ = period, $\lambda$ = wavelength, $f$ = frequency).
It is then the combined effect of many low energy photons that produces either thermal or non-thermal effects. Burns may result, due to excessive heating, when the radiated power absorbed exceeds a safety limit\(^3\).

The two kinds of radiation produce quite different effects — but in common language, radiation primarily means ionizing radiation, and in the past some confusions did occur.

**The specific absorption rate**, or SAR, is the primary parameter used to determine the health risk due to electromagnetic power absorption in the body. It is defined as the ratio of absorbed power per unit mass of tissue. Safety standards, based on a maximum temperature increase of 1°C, require that the SAR averaged over the whole body for 30 minutes or more should not exceed 0.08 mW/g. The ANSI/IEEE standard (USA) limits the SAR averaged over any gram of tissue to 1.6 mW/g, while ICNIRP (International) and CENELEC (European) standards take averages over respectively 100 and 10 grams of tissue, limiting them to 2 mW/kg. Cellular phones must therefore meet different standards in different parts of the world.

When using a cellular phone, the operator places it close to the ear, and then a significant amount of the radiated power is absorbed in the head, producing useless heating of the brain. A large hole appears in the radiation diagram of the antenna so that, as far as the communication is concerned, this power is actually wasted. There are therefore two good reasons to determine the distribution of the absorbed power, and then to design the antenna and the phone casing to reduce the power lost in the operator’s body.

This of course, is by no means straightforward: obviously, one cannot stick thermocouples into the user’s head to determine the increase in temperature. Simulations must be made to evaluate the power distribution and determine the SAR. While several computer approaches have been considered, the Finite Difference Time Domain (FDTD) has emerged as the one best suited to perform this simulation. The three-dimensional domain under study is divided into cubes, within which an approximate formulation of Maxwell’s equations is used. The material properties within each cube can be defined individually, so that properties of skin, bone, and brain can be introduced in the simulation program. Simple models such as homogeneous cubes and spheres were considered first, and increasingly sophisticated models were developed as computer capabilities increased. Figure 3 shows an example of layered human head subdivided into elementary cubes. Experimental simulations are also made, using laboratory head models called *phantoms*. A shell simulating more or less accurately the boundary of the head is filled with liquids having dielectric properties equivalent to those of the human body at the frequencies under consideration. A probe then samples the electric field inside of the phantom, yielding its spatial distribution, from which the SAR can be determined.

The SAR produced by cellular phones is now determined by most phone manufacturers, and values can be found (although with some effort) on the World Wide Web [8].

\(^3\) Cellular phones and microwave ovens both use frequency bands towards the bottom of the microwave range. As a result, similar safety precautions must be taken in apparently very different devices — in ovens, large amounts of power are carefully contained within closed boxes, whereas cellular phones are designed to beam out power, but at much lower power levels.
Non-thermal effects. Contrary to popular belief, non-thermal effects do not remain mysterious due to lack of adequate research in this area: on the contrary, many research projects have been carried out all over the world, and a considerable literature has been devoted to them. The Bioelectromagnetics Society, founded in 1978 [9], presents an international forum for all researchers in the field of electromagnetic interactions with biological bodies. Many of the publications appearing in the magazine “Bioelectromagnetics” present results obtained in biology laboratories, but these results are not so easy to interpret by engineers. More easily understandable are the many publications of Dr. James C. Lin, which regularly appear in the IEEE Microwave Magazine, and in the IEEE Antennas and Propagation Magazine.

It was found in particular that low level modulated microwave fields could interact with the nervous system [10]. More recently, studies were carried out on the “Blood-Brain-Barrier,” or BBB, to determine how nutrients enter the brain from the blood vessels, and how microwave radiation affects the transfer process [11]. The effects of cell-phone radiation on EEG (Electroencephalograms) [2] and on the cognitive function were evaluated [12].

In 1993, a man claimed on a popular TV program that cellular phones did cause cancer, and this created the “cellular phone scare” all over the USA, which resulted in a severe drop in phone sales during the following months [13]. As was pointed by Dr. John Osepchuk, of Raytheon, “all it takes is one plaintiff, a lawyer, and the media to question a huge base of scientific research.” No connections between cellular phone use and cancer were found since then, the scare was probably due to confusion between ionizing and non-ionizing radiation.

Short pulses of modulated microwave radiation can be heard by humans and laboratory animals. However, the “microwave auditory phenomenon,” or microwave hearing effect, only appears at high peak power. A number of projects were initiated to investigate the effect of exposure to wireless communication radiation on the middle and inner ear [14].

Effects on sensitive electronic equipment

Users of cellular phones go everywhere, even in intensive care areas of hospitals. Investigations of instrument sensitivity showed that the operation of most intensive care equipment
can be adversely affected when cellular phones are placed within one meter of these devices [15]. Subsequently, many health facilities instituted cellular phone restrictions, enforced by signs and personnel, but with limited success. Schemes for ensuring electromagnetic compatibility between cell phones and biomedical devices have been proposed [16]. It is highly recommended to users of cellular phones to switch off their devices when visiting hospitals.

Significant compatibility issues were encountered between cellular phones and hearing aids. In late 1995, the cellular phone industry and the hearing instrument industry addressed together the issue of hearing aid interference. Some of the technical challenges and measurement techniques developed to resolve this interference issue are found in [17].

Pacemakers are small implanted electronic instruments that provide electric pulses to sustain deficient hearts. Early pacemakers turned out to be somewhat sensitive to radiation, in the vicinity of radio transmitters, microwave ovens, high voltage lines and storms. More recently, pacemaker manufacturers used electromagnetic compatibility techniques (shielding and filters) to improve the immunity of their products. An extended study showed that recent devices were not affected, even by power levels considerably higher than the thermal safety levels specified in figure 1[18].

**Base Stations.** To set up a communication from a portable phone, a counter station is an absolute must, in order to provide a connection to the fixed phone network. Telephone service providers must therefore install a suitable number of base stations, strategically located to provide a complete coverage over a determined area called cell (hence the name “cellular phone”). The signal beamed by the base station at the center of a cell must reach any cellular phone at the outside boundary of the cell — the power received must be sufficient to carry messages, which must remain understandable. Conversely, the signal beamed by a cellular phone must be able to properly reach the receiver of the base station. The base station must in principle provide wireless connections to all the users within the cell, but with a limited number of channels, which is selected on a statistical basis.

The power levels at both ends of the wireless link are limited by safety requirements. As mentioned previously, the power transmitted by the cell phone must not produce an excessive SAR within the head of the user. Similarly, the power transmitted by the base station must be compatible with the power densities specified in figure 1 — which applies to all persons that may pass by or live in the vicinity of the base station antenna. When a transmitter is located in the open countryside, a suitable exclusion circle can be defined around the antenna. This is not possible when the transmitter is placed in a built up area, and then the antenna must be placed at a sufficient height so that radiation limits are not exceeded.

Considerable research was carried out to determine how signals transmitted from an antenna spread out over the countryside and within built up areas. The problem to be solved is particularly complex because waves get reflected and diffracted, often many times, over a large variety of obstacles, which often exhibit electromagnetic properties that are not well defined [19]. Many approaches were therefore considered, and by now sophisticated computer packages have become available to service providers.

The powers transmitted by the cellular phone and by the base station are continually adjusted to provide a signal level adequate to convey the messages — but there is no point in sending more power that what is strictly necessary. As a result, the power density varies continuously,
depending on how many channels are active at the time and on the relative locations of the active cellular phones. The maximum permissible level defined by safety limitations can thus only be reached when all the channels of a base station are occupied and when all the active cellular phones are at the outer boundary of the cell. This is a case figure that practically never happens in practice, and therefore the power density levels actually encountered remain considerably below the safety limits.

Unfortunately, the setting up of base station antennas regularly causes severe acceptance problems, more particularly when they must be placed within towns or villages [20]. Some people who do not object to using cellular phone themselves sometimes are viscerally opposed to seeing an antenna installed in their neighbourhood… There have been complaints of headaches, nervousness, sleeplessness and other ills, supposedly due to the more or less close presence of base station antennas\(^4\). As soon as plans to install an antenna are published, oppositions flourish, which sometimes slow considerably the installation process. By now, these oppositions are no longer accepted by courts, except when there is a definite proof that the safe power density level would be exceeded, or when the proposed location is somehow unsuitable (unaesthetic, on historical buildings, on roofs of school buildings, etc.).

**Remarks and Conclusion**

Is the use of cellular phones — and that of the other electromagnetic equipment that is regularly appearing on the market — safe? When properly operated, it certainly cannot cause any harmful heating. While some grey areas remain as far as non-thermal effects are concerned (and they will probably remain for a long time, if not for ever), these do not appear to present any significant danger to users. But for many people in the general public, electromagnetic radiation effects remain highly mysterious, and one always tends to fear what one does not understand.

It should be remembered that man-made electromagnetic radiations have been around us for more than a century, so that if even a very small hazard did exist, by now it would have had plenty of time to cause major health problems. The radiation from the many millions of cellular phones now operating all over the world do not appear to cause significant harm on the health of their users.

On the other hand, the use of cellular phones can be lethal, but not because of the radiation that they produce. Carrying on a phone conversation requires a lot of attention, and it is incompatible with many other activities, in particular crossing a street or driving a car. While the latter is strictly forbidden in some countries, a significant number of car accidents (estimated at 6%) were reported as due to the use of cellular phones while driving. And it is hardly worth reminding that the use of private cellular phones and of some other electronic equipment is forbidden in planes during flight.

Some questions should also be raised about the social effects of cellular phones. For some time, it has become customary to hear complete strangers “discussing in public” their family history with some unseen correspondent, while on the street, in trains, etc. While it may be fun to listen to such stories for a while, they soon tend to become rather boring or downright

\(^4\) Service providers mentioned that in some locations, such complaints were voiced when the antenna had been raised, but before the communications equipment was in operation. Evidently, many complaints involve a strong psychological component.
bothersome. In addition, some information picked up while listening might be confidential and could be used improperly. Some need for care and circumspection would be welcome whenever using cellular phones in public.

Another potential cause of disquiet is the fact that, as soon as a cellular phone is switched on, the service providers — and others people too, maybe — know the location of the user. It is thus possible to track people in real time, which may prove quite useful when looking for wanted criminals, but may become objectionable when this information reaches the wrong offices. At a time when people are very sensitive about privacy, rules should be set up to avoid the spurious use of information obtained in this manner.

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[9] www.bioelectromagnetics.org


5 Some parents use cellular phones to check their childrens’ whereabouts. This technique may not be totally reliable however, because it does not take long for kids to find how to switch off their phone.


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