Editorial Board and Message from Editor-in-Chief

List of Papers

**Microwave Related:**
- Switchable UltraWideband Antenna for Indoor and Automotive Vehicular Band Application.
  - D. Thiripurasundari and D. S. Emmanuel
  - pp: 1-5
- A 2.1 to 4.6 GHz WideBand Low Noise Amplifier Using ATF10136
  - M. Meloui, I. Akhchaf, M. Nabil Srifi, and M. Essaaidi
  - pp: 6-10
- Design of Triple Band Slot Antenna for 802.11a/b WLAN and Upper UWB Application Using Pentagonal Tuning Stub
  - Pratap N. Shinde and B. K. Mishra
  - pp: 11-17
- Propagation of Modes in the Corrugated Waveguide Made of the Silicon Carbide Material
  - T. Gric and L. Nickelson
  - pp: 18-24
- Geo-textile Based Metamaterial Loaded Wearable Microstrip Patch Antenna
  - J. G. Joshi, Shyam S. Pattnaik, and S. Devi
  - pp: 25-33

**Optics Related**
- All Optical Solution for Free Space Optics Point to Point Links
  - Daigo Hirayama and Banmali Rawat
  - pp: 34-44

**Call for Papers:**
14th International Symposium on Microwave and Optical Technology (ISMOT 2013).
Message from the Editor-in-Chief
Banmali S. Rawat

First of all I would like to wish “Very Happy and Prosperous New Year-2013” to all our authors, subscribers, Editorial Board Members and reviewers. It gives me great pleasure to bring out the 1st issue of the International Journal of Microwave and Optical Technology (IJMOT) for the year 2013. Once again I would like to apologize to you all for publishing this issue late due to some serious web related problems at our end. This issue contains good mixture of papers in the areas of: various types of microstrip antennas, LNA, slot antenna, ring resonator for wireless applications, multiband antenna, corrugated waveguides and all optical solutions for free space point to point link.

If your current research paper submitted to IJMOT or any other journal is in similar area as published previously in IJMOT, please make sure to cite the reference of IJMOT. This would help in improving the impact factor of IJMOT.

As you all know that now we collect flat fee $100 per published paper up to 8 pages. For additional pages beyond 8 pages we would charge $30 per page. Once the paper is accepted for publication, the authors would be asked to pay publication/page charges as per invoice before the paper is published. If the page charges are not paid until the date of next issue, the paper would be removed from IJMOT data base. However, in order to help the authors, we have also decided that if the authors’ organization/university/institution is an annual subscriber of IJMOT during that period, the publication fee up to 8 pages would be waived off. It means, if the authors would like to have publication fee waived off (up to 8 pages), they should request their organization/university/institution to subscribe IJMOT without any delay.

You can also present your papers in the ISMOT conferences and get published in IJMOT as a full length papers if selected by the technical program committee. All these papers also go through normal review process before being finally accepted for publication. The 14th ISMOT is going to be organized in Kuala Lumpur, Malaysia from October 28-31, 2013 under the leadership of Profs. Le-Wei Li of Monash University and Chuah Hean Teik of Universiti Tunku Abdul Rahman, Malaysia. For details please visit the ISMOT-2013 website at: www.utar.edu.my/ismot2013. Other future conferences are tentatively scheduled as: ISMOT-2015- Guadalajara, Mexico and ISMOT-2017 – Dresden, Germany. We hope to see many of you participating ISMOT-2013 and future ISMOT conferences.

I am very pleased to inform our authors/subscribers that IJMOT is now indexed by SCOPUS, Google, EI-Compendex, EBSCO, ISI and Media Finder. We are contacting other indexing agencies also in this regard.

I would like to thank all the editorial board members and reviewers for their continued help and support for IJMOT. My special thanks to our web manager Mr. Syam Challa for improving the IJMOT website and helping in the publication of every issue since 2007.

Banmali S. Rawat
Dated: February 27, 2013
Switchable Ultra Wideband Antenna for Indoor and Automotive Vehicular Band Applications

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Abstract - An ultra-wideband antenna that can effectively switch between communication band (3.1-10.6 GHz) and automotive vehicular band (22-29 GHz) and vice versa is presented. A rectangular monopole antenna is initially designed to provide 3.01-10.4 GHz band. The antenna is further modified by introducing a slit across the patch to form a parasitic patch to cover the 22-29 GHz band. Reconfigurable capability is achieved by introducing switch across the slit. Bandwidth of 110% is obtained in the communication band and 27.45% in the vehicular band respectively. The radiation pattern is consistent throughout the band with average gain of 3.26 dBi in the communication band and 4 dBi in the vehicular band respectively.

Index Terms - monopole antenna, UWB, PIN diode, reconfigurable

I. INTRODUCTION

UWB technology is the emerging and fast growing technology since 2002 after the allocation of 3.1-10.6 GHz by FCC for indoor communication application [1]. This technology promises short range communication at very high data rate (500 Mbps for distance less than 2m and 110 Mbps for distance up to 10m) for wideband wireless communication. FCC subsequently allocated 22-29 GHz band for automotive short range radar application. There are two types of automotive radar, long range radar at 77 GHz with a range capability up to 200m for automatic cruise control (ACC) and short range radar at 24/26 GHz and 79 GHz up to 30m for anti-collision detection. Long range radar with narrow radiation beam enables automobile to maintain a cruising distance, while short range radar has recently attracted attention because of its wide applications such as pre-crush warning, stop and go operation and lane change assistance.

In recent years automotive wireless market has expanded greatly where the antenna is the key element in determining the performance and size of the communication systems. Modern cars contain multiple antennas that capture AM/FM broadcasts, Satellite Digital Audio Radio Services (SDARS) signals, GPS data and Electronic Toll Collection [2].

The requirement for increased functionality such as direction finding radar, control and command within a confined volume places a greater burden. Solution to the aforementioned problem is the reconfigurable antennas. These antennas have become more attractive with increased demand for number of multiband antennas. Multiband antennas provide high level of functionality to a system by eliminating the need for complicated wideband antenna solution [3]. A number of reconfigurable antennas have been presented [3-6]. In [3] a reconfigurable planar dipole antenna, a cactus antenna and a fractal antenna are presented. In [4] a UWB antenna which can electronically reconfigure between single/dual band notch by placing 3
PIN diode switches integrated within the antenna is presented. A multiband Sierpenski fractal antenna with three sets of RF MEMS switches with different actuation voltages used to sequentially activate and deactivate parts of antenna is presented [5]. A reconfigurable antenna operating between two complementary sub-bands is investigated in [6]. Report is not available on reconfigurable antenna with two different widely separated switching bands of 3.1-10.6 GHz and 22-29 GHz.

In this paper a rectangular monopole antenna that can be configured to operate in two frequency bands, 3.1-10.6 GHz (Band-I) and 22-29 GHz (Band-II) is presented. The reconfigurable capability can be realized by switching a PIN diode ON and OFF on the patch. The proposed reconfigurable antenna design enables the use of single antenna in indoor as well as vehicular application simultaneously. The design has been successfully implemented and the experimental result for the reflection coefficient and radiation characteristic is presented.

II. ANTENNA DESIGN

The geometry of the proposed antenna is shown in Fig.1. The antenna is fabricated using FR4 substrate with dielectric constant 4.4, substrate height of 1.6 mm and loss tangent 0.022. A single antenna element can be modified to operate at two different frequencies by changing the length of the antenna element. The initial design process started with the design of rectangular patch antenna to operate for band-I with the center frequency of 6.85 GHz using the design equations [7]. The quarter wave length of the antenna is 11 mm at the center frequency of 6.85 GHz. The dimension of the antenna is optimized to 13.3 mm to provide wide bandwidth. The radiating length required for the antenna to operate in band-II at the center frequency of 25 GHz is 6 mm. The antenna structure is modified by introducing a slit at a distance of 6 mm which corresponds to half wavelength. The width of the slit is chosen to be 0.3 mm.

The antenna can operate with two different states the ON state and OFF state. When the diode is ON, it connects the two parts of the patch to form a large rectangular patch and operate at band-I. Alternatively, when diode is turned OFF, the length of the radiating element decreases and operates in band-II and the top portion of the
The dimension of the radiating patch can be varied by employing a PIN diode between the radiating elements. The HPND PIN diode of dimension $0.3 \times 0.9 \text{ mm}^2$ is modeled as metal pad in the simulation and experiment. The validity of this simplification has been demonstrated in [8-11].

**III. RESULT AND DISCUSSION**

The reflection coefficient characteristic of the antenna with and without metal pad is shown in Fig.2 and Fig.3 respectively. With the pad the antenna operates at communication band and without pad, where the physical length of the antenna is less, the antenna operates between 21-30 GHz. The position of the switch is optimized through simulation using CST Microwave studio. When the switches are at edge of the patch, current concentrate near the switch and are opposite in phase in slit and cause a notch band of 5.5-7.5 GHz. When switches are away from the edge, the concentration is less due to which the bandwidth widens. The above said action can also be investigated from the current distribution illustrated in Fig.4. Hence the optimum position of switch is chosen to be at 0.3 mm from the edges.

Reflection coefficient characteristic of the antenna with and without metal pad is shown in Fig.5. During the ON state the antenna operates in band–I and a fractional bandwidth of 110% is achieved. During OFF state the antenna operates in band–II and a bandwidth of 27.45% is achieved as illustrated in Fig.5. The group delay of the antenna is consistent throughout the band with less than 1.5ns variation as illustrated in Fig.6.
The reflection coefficient characteristic of the antenna is measured using N5230A Agilent network analyzer. The Fabricated antenna and their corresponding reflection coefficient are shown in Fig. 7, 8 and 9 respectively. The comparison of simulated and measured reflection coefficient characteristic of the antenna with metal pad is shown in Fig.8. It can be noted that the antenna exhibits ultra wide bandwidth of 3.1-10.45 GHz. Though the antenna without metal pad operates between 21-30 GHz the reflection coefficient of the antenna is verified up to 20 GHz due to the limitation in the testing facility. The antenna is non-responsive up to 20 GHz as expected except for 5.13 - 6.97 dB as illustrated in Fig.9. The smaller band which is appearing may be due to the SMA connector.

During OFF state (without metal pad), the antenna shows non-directional pattern in azimuth plane and degraded pattern in elevation plane at 25 GHz as shown in Fig. 10. The measured radiation characteristics for the ON state (with metal pad) at 3.5 GHz, 6.5 GHz and 8.5 GHz respectively are shown in Fig. 11. The proposed antenna shows good omni directional radiation pattern in H-plane and monopole like in the E-plane.
A reconfigurable rectangular monopole antenna with wide operational bandwidth is designed. The simulated and measured results demonstrate that the antenna’s operational bandwidth shifts from 3.1-10.4 GHz to 22-29 GHz. This UWB reconfigurable single antenna can be used for short range data transfer as well as vehicular application.

VI. CONCLUSION

REFERENCES

A 2.1 to 4.6 GHz Wideband Low Noise Amplifier Using ATF10136

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Abstract — This paper presents a novel low-noise amplifier (LNA) for the 2 to 4.6 GHz applications. The LNA uses a circuit topology consisting of two gain stages with two feedback loops using ATF technology. The noise figure of this LNA varies from 1.5 to 5.4 dB, and its gain varies from 5 to 13 dB throughout the operating frequency band. The S11 and S21 simulated and measured results are presented, and show good agreement.

Index Terms — low noise amplifier, ATF-technology, feedback amplifier.

I. INTRODUCTION

The Federal Communications Commission approved the use of UWB technology for commercial applications in the unlicensed 3.1–10.6 GHz frequency range. The UWB technology, which is an ideal candidate for many applications due to the transmission and reception of the trains of short pulses having very low power, can achieve very good time and spatial resolutions [1]–[3]. For this reason, the use of the low noise amplifiers (LNAs) in UWB communication systems must balance the specification of bandwidth, linear high gain and low noise figure at low DC consumption [4]. A low noise amplifier (LNA) forms the first block that amplifies the desired band of signals without adding significant noise to the signal. The primary role of the LNA is to achieve large gain and low noise figure.

In this paper the feedback topology and Agilent Technologies which are candidates to degrade the noise figure and reduce the maximum power gain available from an active device. ATF-10136 is a high performance GaAs FET, and it is appropriate for use in the first stage of low noise amplifiers operating in the 0.5–12 GHz frequency range. The GaAs ATF-10136 MESFET technology is chosen because it has an advantage in terms of the gain and the noise figure [4].

According to these requirements, this paper report the design of a two stage broadband low noise amplifier using GaAs MESFET ATF-10136 technology. To achieve wideband matching and load design, a few different topologies have been reported: a multi-section LC matching network LNA [5], a distributed LNA [6], and a resistive feedback or common-gate LNA [7], [8]. The simulated results are done by using Agilent ADS commercial simulator.

II. WIDEBAND LNA CRITERIA DESIGN

The critical LNA parameters design are High gain, low noise figure and very wide band (2 GHz to 4.6 GHz). The proposed low noise tow stages gain amplifier is illustrated in fig.1. It consists on tow stages feedback amplifiers on ATF technologies and matching networks. The designed LNA is based on two-gain stages approach, where the first wideband gain stage is connected to the second wideband second stage via an inter-stage matching as shown in fig.1. In the first stage, the shunt feedback technique formed by the resistors R2, R3 and a capacitor C3, placed between the gate and the drain of the transistor Q1. The resistance value of
R2 is maximized to reduce its noise contribution for an acceptable input return loss. The resistor R1 affect largely the bandwidth, where: \( R2 > R1 \). The first stage is optimized for acceptable low noise and gain. For this, an input matching network, formed by inductors L1, L2 and a capacitor C2, is introduced.

Fig. 2 ADS schematic of the stages feedback LNA

Another important component is the capacitor C4, which contribute on the extension of the bandwidth and make flat gain. The capacitor C8 ensure the same impact in our design. Cascading a second stage that is formed with the same topology such the first stage. A shunt feedback resistors R6, R7 and capacitor C7 are connected between the drain and gate of transistor Q2. The resistor R6 has an impact on extension of bandwidth where: \( R6 > R5 \). The first and second stages are matched via an inter-matching which is formed by an inductors L4, L5 and a capacitor C6 in \( \pi \)-shaped structure. This allows good compensation for the noise figure (NF) and degradation of gain. The tow stage cascade architecture can be used at 1V.

III. LC-SECTION MATCHING NETWORK AND FEEDBACK TOPOLOGY

A. \( \pi \)-Shaped LC-Matching

In the general case, the essential requirement to achieve maximum gain is a good matching network between the load and the source. The quality factor Q is the most important factor to construct the matching network [1]. It is defined as:

\[
Q = \omega \frac{\text{Energy stored}}{\text{Average power dissipated}} \quad (1)
\]

In our LNA design, to increase the bandwidth and maintain a flat gain as much as possible, three \( \pi \)-shaped LC-matching are used, i.e. input matching, inter matching and output matching networks. The \( \pi \)- shaped wide band matching topology is shown in fig.2.

Fig. 3 \( \pi \)-shaped matching network

B. Feedback Topology

Table I: Components design identification

<table>
<thead>
<tr>
<th>C1</th>
<th>C2</th>
<th>C3</th>
<th>C4</th>
<th>C5</th>
</tr>
</thead>
<tbody>
<tr>
<td>50 pF</td>
<td>0.13 pF</td>
<td>25 pF</td>
<td>0.145 pF</td>
<td>0.1 pF</td>
</tr>
<tr>
<td>C6</td>
<td>C7</td>
<td>C8</td>
<td>C9</td>
<td>C10</td>
</tr>
<tr>
<td>0.155 pF</td>
<td>0.25 pF</td>
<td>0.21 pF</td>
<td>0.3 pF</td>
<td>0.01 fF</td>
</tr>
<tr>
<td>C11</td>
<td>L1</td>
<td>L2</td>
<td>L3</td>
<td>L4</td>
</tr>
<tr>
<td>5 pF</td>
<td>0.335 nH</td>
<td>55 nH</td>
<td>4 H</td>
<td>55 nH</td>
</tr>
<tr>
<td>L5</td>
<td>L6</td>
<td>L7</td>
<td>R1</td>
<td>R2</td>
</tr>
<tr>
<td>0.155 pF</td>
<td>0.1 H</td>
<td>4.2 nH</td>
<td>10 kΩ</td>
<td>100 kΩ</td>
</tr>
<tr>
<td>R3</td>
<td>R4</td>
<td>R5</td>
<td>R6</td>
<td>R7</td>
</tr>
<tr>
<td>50 Ω</td>
<td>1 kΩ</td>
<td>10 kΩ</td>
<td>100 kΩ</td>
<td>50 Ω</td>
</tr>
<tr>
<td>R8</td>
<td>R9</td>
<td>Transistors: Q1 and Q2</td>
<td></td>
<td></td>
</tr>
<tr>
<td>1 kΩ</td>
<td>75 kΩ</td>
<td>ATF-10136</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Wideband LNAs have been frequently implemented in feedback topologies [7], [9]. The requirement criteria LNA design, an ultra wideband width and a flat gain, can be met using a feedback topology. This technique have been largely used for amplifiers design aver an octave bandwidth. On the other hand, the feedback technique seems to be the best solution due to their advantages, especially it can be used in the broadband amplifiers to control gain flatness and reduce the input and output VSWR at the same time, and to make the circuit more robust, and to de grade the noise figure and control the extension bandwidth of the amplifier [3]-[5].
In general case of the shunt feedback topology shows in fig. 3, the input impedance is given by:

$$Z_{in} = \frac{R_f}{1 + G + \frac{R_f}{Z_Q}}$$

(2)

And if we ignore the input impedance of the basic amplifier when $|Z_Q| >> |R_f|$, the input impedance is given as:

$$Z_{in} = \frac{R_f}{1 + G}$$

(3)

And the gain $G$ will be chosen when:

$$Z_m = R_s$$

(4)

In our design, the first stage is formed by the transistor $Q_1$, and based on the shunt feedback topology shown in fig. 4.a) via the resistors $R_{f1}$ and $R_1$, where:

$$R_{f1} = R_2 + \left\{ R_3 // \frac{1}{j\omega C_3} \right\}$$

(5)

The second stage consists on a transistor $Q_2$ and the shunt feedback loops via resistors $R_{f2}$ and $R_5$, as shown in fig. 4.b):

$$R_{f2} = R_6 + \left( R_7 // \frac{1}{j\omega C_7} \right)$$

III. EXPERIMENTAL AND SIMULATED RESULTS

The designed LNA was fabricated and mounted on FR4 substrate ($\epsilon_r = 4.4$). The back of the substrate was covered with copper. Fig. 7 shows the photograph of the prototype of the proposed LNA. Fig. 5 shows the simulated scattering (S) parameters ($S_{21}$ and $S_{11}$). The poor match between simulated and measured $S_{11}$ parameters is due to the power of the amplifier, which has been polarized 5V, whereas ADS simulator is not based on the polarization in $S_{11}$ and $S_{21}$ calculation. The measured $S_{11}$ parameters, plotted in fig. 6, are less than -10 dB from 2 to 4.6GHz. It is minimal at 4.5 GHz with -40 dB. The measured gain of the proposed LNA is maximal around nominal 13dB at 2.5 GHz and rolls off to 7 dB at 4.6 GHz. The noise figure of the amplifier is less than 3.6 from 2 to 4.6 GHz, and less than 4 dB from 2 GHz to 6 GHz as shown in Fig. 9.
Fig. 6: Simulated S21 and S11 of the LNA

Fig. 7: Measured gain and S11 of the proposed LNA

Fig. 8: The photograph of the proposed LNA.

Fig. 9: Simulated noise figure (NF) of the proposed LNA.
Table II: Comparison the proposed amplifier to some previous LNA design works

<table>
<thead>
<tr>
<th>Works</th>
<th>Topology</th>
<th>Technology</th>
<th>Bandwidth (GHz)</th>
<th>S21 (dB)</th>
<th>NF (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[5]</td>
<td>Band-pass Chybychev</td>
<td>0.18-µm CMOS</td>
<td>2.3-9.2</td>
<td>9.3</td>
<td>4.2-9.2</td>
</tr>
<tr>
<td>[10]</td>
<td>Distributed</td>
<td>0.18-µm CMOS</td>
<td>DC-14</td>
<td>10.6</td>
<td>3.4-5.2</td>
</tr>
<tr>
<td>[10]</td>
<td>Distributed</td>
<td>0.18-µm CMOS</td>
<td>DC-22</td>
<td>7.3</td>
<td>4.3-6.1</td>
</tr>
<tr>
<td>[11]</td>
<td>Feedback</td>
<td>0.18-µm CMOS</td>
<td>2.0-4.6</td>
<td>9.8</td>
<td>2.3-4.0</td>
</tr>
<tr>
<td>[1]</td>
<td>Distributed</td>
<td>0.18-µm CMOS</td>
<td>0.03-6.3</td>
<td>8.6</td>
<td>4.2-6.2</td>
</tr>
<tr>
<td>[2]</td>
<td>Feedback</td>
<td>0.18-µm CMOS</td>
<td>3.6</td>
<td>16</td>
<td>4.7-6.7</td>
</tr>
<tr>
<td>[3]</td>
<td>Distributed</td>
<td>0.18-µm CMOS</td>
<td>2.7-9.1</td>
<td>10</td>
<td>3.8-6.9</td>
</tr>
<tr>
<td>[4]</td>
<td>Two stages Feedback</td>
<td>ATF-10136 ATF-34143</td>
<td>3.6-4.2</td>
<td>21</td>
<td>1.1</td>
</tr>
<tr>
<td>This Work</td>
<td>Two stages Feedback</td>
<td>ATF-10136</td>
<td>2.1-4.6</td>
<td>7-13.6</td>
<td>&lt;3.6</td>
</tr>
</tbody>
</table>

The performances of the proposed LNA are summarized and compared with previous works in Table II. The proposed LNA is characterized with its flat gain, its wide bandwidth, and low cost of manufacture.

IV. CONCLUSION

A novel wideband low noise amplifier (LNA) for 2.1 GHz to 4.6 GHz applications was designed. The LNA utilizes an improved shunt feedback topology exploited in a tow stage amplifier and a three π-shaped broadband LC-matching networks. The proposed LNA was implemented on a FR4 substrate using ATF-10136. It achieves NF lower than 3.6 dB over the entire desired band, and S11 less than -12 dB, and S21 larger than 7dB from 2.1 to 4.6 GHz. This LNA presents a good candidate for mobile and wireless communication applications.

REFERENCES

Design of Triple Band Slot Antenna for 802.11a/b WLAN and Upper UWB Application Using Pentagon Tuning Stub

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Abstract- A planar antenna with pentagon shaped tuning stub and asymmetric rectangular ground plane width along with wide rectangular aperture is presented for triple band operation. A novel concept of use of feed coupling slot as inherent band-notch at 4 GHz and design of wide rectangular aperture with length and width at λ/4 and λ of WLAN 802.11b and 802.11a resonating frequency respectively with variation in rectangular ground plane width for multiband operation is discussed. Antenna geometry is very simple and have compact dimension 32 mm × 42 mm. The operation is stable over all the operating bands centered at 2.66 GHz, 5.27 GHz and 8.78 GHz with average gain of 1.86 dBi, 0.54 dBi and 5.28 dBi respectively. The bandwidth VSWR ≤ 2 of the 802.11b, 802.11a and upper ultrawideband is 620 MHz, 860 MHz, and 3.67 GHz respectively, which corresponds to fractional bandwidth of 24.43%, 15.37%, and 32.23 %.

Index Terms- rectangular aperture, pentagon shaped tuning stub, segmentation, CPW feed, WLAN, UWB.

I. INTRODUCTION

Considering the drastic and rapid development of wireless industry in the recent years, there is a need of small, compact, low profile microstrip antenna with high gain and bandwidth. The wireless industries are engaged in development of UWB antennas with characteristics of high data rate transfer in interference free environment, to satisfy the need of consumer electronics. As a key component of UWB communication system, the study of UWB antennas electrical characteristics for indoor and outdoor applications is reported by many researchers. Federal Communications Commission (FCC) has approved spectrum from 3.1-10.6 GHz of bandwidth of 7.5 GHz for UWB communication system [1]. Planar antennas are studied for wide band operation with various shaped tuning stub like ellipse, cone, semi-circular, U and rectangular round cornered [2-5]. All these articles have well reported the advantages of planar geometry such as wide impedance bandwidth and omnidirectional radiation pattern.

Today’s wireless gadgets need narrow operating bandwidth at various resonating frequencies in UWB spectrum. Therefore an antenna design with dedicated UWB bandwidth for single wireless application will degrade the channel efficiency. Multiband operation has demonstrated by segmenting the UWB spectrum. The segmentation is achieved by embedding parasitic patch in antenna geometry and various shaped slots like rectangular, C and many more in radiating patch as well as in ground plane [6-8]. These additional efforts taken for generation of multiband in interference free environment will result antennas structure more complex. Also the size of the existing antennas are large and most of them need a special substrate material such as ROO3003, RT / Duroid 5870 etc. An attempt is made in this paper to achieve a multiband operation with simple and compact antenna geometry with a rarely used pentagon shaped radiating patch.
This paper discusses a design of multiband antenna, which can be used for existing wireless applications like WLAN and upper ultra-wideband application without the use of any embedded frequency notch element in antenna geometry. This triple band operation uses a simple wide rectangular aperture with pentagon shaped radiating patch fed by a coplanar waveguide (CPW) feed. A concept of radiation from ground plane along with radiating patch and use of transmission line coupling slot with coplanar ground plane as frequency notch is discussed to have triple band operation. A segmentation of UWB band by controlling ground plane parameters at respective center frequencies of resonating bands with wide rectangular aperture is presented. The bandwidth of each segmented band is more than 500 MHz for high speed WLAN and upper UWB application.

II. ANTENNA DESIGN AND ANALYSIS

The proposed rectangular slot antenna with CPW feed structure printed on FR4 substrate of dimension 32 mm × 42 mm is shown in Fig. 1. The antenna comprises pentagon shaped tuning stub and rectangular ground with rectangular aperture of area 16.6 mm × 33 mm². The thickness of substrate is 1.6 mm having relative dielectric constant \( \varepsilon_r = 4.4 \) with tangent loss \( \tan\delta = 0.02 \). The parameter \( r_1 \) controls the side length of the pentagon, which in turn controls the coupling gap \( s \) between tuning stub and the ground plane. The aperture area as well as coupling gap \( s \) is controlled with variation in bottom ground height \( h_1 \). The rectangular aperture area, side length of pentagon, ground plane height \( h_1 \), and \( h_2 \) are optimized to have 50Ω impedance matching for dual as well as triple band operation. The optimized parameters of the proposed antenna are listed in Table 1.

A rectangular aperture of vertical length 16.6 mm have been designed at quarter wavelength of center frequency 2.4 GHz of first resonating band. The bottom ground height \( h_1 \) is adjusted to quarter wavelength of rejecting band which corresponds to wide notch band centered at 4 GHz. The coupling gap \( g \), which couples bottom ground with feed line act as an open ended slot. The effective length of slot equal to \( h_2 \) attenuates the band centered at 4 GHz. The horizontal length of aperture \( 2w \) is adjusted to 63 mm, which is close to a wavelength of second resonating band centered at 5.1 GHz. The pentagon side length is designed at 7.40 mm, which is closer to the quarter wavelength of third resonating band resonating in the vicinity of 9 GHz. From the preceding analysis it is clear that by exciting ground plane as well as patch for radiation with appropriate dimension of rectangular aperture and feed slot as discussed above, the
Table 1: Parameters of the proposed antenna

<table>
<thead>
<tr>
<th>Parameter</th>
<th>( W )</th>
<th>( L )</th>
<th>( w )</th>
<th>( r_1 )</th>
<th>( h_1 )</th>
<th>( h_2 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Unit (mm)</td>
<td>42</td>
<td>32</td>
<td>33</td>
<td>6.8</td>
<td>10.4</td>
<td>5</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameter</th>
<th>( w_f )</th>
<th>( S )</th>
<th>( g )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Unit (mm)</td>
<td>3</td>
<td>1</td>
<td>0.3</td>
</tr>
</tbody>
</table>

A triple band operation for wireless communication will be possible without use of any band-notch element.

### III. SIMULATION AND MEASURED PERFORMANCE

The simulation of the antenna is carried out with electromagnetic software HFSS V.11, which is based on the finite element method. The reflection coefficient of the antenna for variation in the side length of pentagon shaped tuning stub in relation with \( r_1 \) is shown in Fig. 2. The bands generated at 2.42 GHz and 9.82 GHz shows, there is a good coupling between feed, wide rectangular aperture and tuning stub, at radius \( r_1 = 6.8 \) mm and \( s = 1 \) mm. The shift in the first resonating band towards higher side in the vicinity of 2.5 GHz as well as decrease in resonant frequency and lower edge frequency of ultrawideband (UWB) can be achieved by lowering the side length of the tuning stub. The impedance matching performance of these two bands can be verified by controlling the coupling gap \( s \) with variation in ground plane parameter \( h_1 \) instead \( r_1 \). The optimized values \( h_1 = 10.4 \) mm and \( s = 1 \) mm give rise dual band of 10 dB impedance bandwidth 550 MHz and 5 GHz respectively as shown in Fig.3. The resonating mode in the vicinity of 5 GHz is made dominant with variation in ground plane height \( h_2 \) as shown in Fig.4. The multiple resonating bands can be increased.

**Fig. 2.** Simulated reflection coefficient \( S_{11} \) curves for variation in \( r_1 \) and \( s \) at \( W = 42 \) mm, \( L = 32 \) mm, \( w = 33 \) mm, \( h_1 = 10.4 \) mm, \( h_2 = 7.5 \) mm, \( w_f = 3.16 \) mm, \( g = 0.22 \) mm.

**Fig. 3.** Simulated reflection coefficient \( S_{11} \) curves for variation in ground plane height \( h_1 \) at \( W = 42 \) mm, \( L = 32 \) mm, \( w = 33 \) mm, \( r_1 = 6.8 \) mm, \( h_2 = 7.5 \) mm, \( w_f = 3.16 \) mm, \( g = 0.22 \) mm.

**Fig. 4.** Simulated reflection coefficient \( S_{11} \) curves for variation in ground plane height \( h_2 \) at \( W = 42 \) mm, \( L = 32 \) mm, \( w = 33 \) mm, \( r_1 = 6.8 \) mm, \( h_1 = 10.4 \) mm, \( w_f = 3.16 \) mm, \( s = 1 \) mm, \( g = 0.22 \) mm.
Fig. 5. Simulated reflection coefficient $S_{11}$ curves for variation in feed width at $W = 42$ mm, $L = 32$ mm, $w = 33$ mm, $r_f = 6.8$ mm, $h_1 = 10.4$ mm, $h_2 = 5$ mm, $w_f = 3$ mm, $s = 1$ mm $g = 0.3$ mm.

Fig. 6. Measured and Simulated reflection coefficient $S_{11}$ curves of proposed antenna.

Fig. 7. Simulated VSWR curves of proposed antenna.

or decreased by tuning inductance and capacitance value of aperture with variation in $h_2$. Fig. 4 shows that, the first band is drastically shifted towards higher frequency with a resonance at 2.62 GHz as well as lower edge of UWB is shifted considerably towards lower side generating second band at 5.1 GHz. Hence by controlling the ground plane height $h_2$ the segmentation of UWB band as well shift in first band at 2.62 GHz is achieved for WLAN 802.11 a, 802.11 b and up link X- band satellite communication application. The bandwidth of dual band and triple band operation is summarized in Table 2. Fig.5. indicate the bandwidth of all the operational bands have been increased with fine tuning of feed width $w_f$. A good agreement between measured and simulated reflection coefficient is shown in Fig.6. A mismatch at higher frequencies occurred because of soldering effect and variation in substrate dielectric constant.

Table 2: Performance of dual and triple band antenna

<table>
<thead>
<tr>
<th>Band</th>
<th>Dual Band</th>
<th>Triple Band</th>
</tr>
</thead>
<tbody>
<tr>
<td>Parameter</td>
<td>WLAN 802.11 b</td>
<td>UWB</td>
</tr>
<tr>
<td>Lower Cut-off, $f_c$</td>
<td>2.16</td>
<td>5.19</td>
</tr>
<tr>
<td>Upper Cut-off, $f_c$, GHz</td>
<td>2.71</td>
<td>10.20</td>
</tr>
<tr>
<td>10 dB Impe. BW, GHz</td>
<td>0.55</td>
<td>5.01</td>
</tr>
<tr>
<td>Center freq. $f_c$, GHz</td>
<td>2.435</td>
<td>7.69</td>
</tr>
<tr>
<td>FBW (% )</td>
<td>22.58</td>
<td>65.15</td>
</tr>
<tr>
<td>Return loss (dB)</td>
<td>22.33</td>
<td>41.86</td>
</tr>
<tr>
<td>Reso. freq. (GHz)</td>
<td>2.41</td>
<td>9.82</td>
</tr>
</tbody>
</table>
Fig. 7. shows simulated VSWR of proposed triple band antenna, which could provide sufficiently wide impedance operating bandwidth (VSWR ≤ 2) of 620 MHz, 860 MHz, and 3.67 GHz for WLAN 802.11b, 802.11a and upper ultrawideband application respectively. The wide notch band of bandwidth 1.93 GHz covers WiMAX (3.3-3.6 GHz) interference band successfully. This inherent notch band property and radiation performance by ground as well as patch is studied by surface current distribution taken at center frequency of notch band 4 GHz and operating frequencies 2.6 GHz, 5.1 GHz and 8 GHz as shown in Fig. 8. When the antenna is excited at 4 GHz, the maximum concentration of current is around the feed coupling slot instead of feed line, which implies that the less energy radiated at 4GHz, resulting into wide band notch. When antenna is invoked at frequencies 2.6, 5.1, and 8 GHz the current accumulates maximum on vertical length, horizontal length of rectangular aperture and slant side length of pentagon respectively, which implies antenna radiates effectively to these resonating frequencies.

Fig. 9. Measured gain of proposed antenna.

Fig. 9. shows the gain of the antenna is very less in the vicinity of 3-4 GHz, which
implies less radiation due to inherent notch-band characteristics. The gain at upper UWB is almost constant in the range of 5.2 dBi. Fig. 10. show the radiation pattern in E-plane and H-plane taken at 2.6 GHz, 5.1 GHz and 8 GHz. It indicates that an omni-directional characteristic is altered at higher frequencies because of manufacturing imperfection and soldering effect at feed. The radiation patterns on both sides of the antenna are similar. The E-plane pattern possesses an eight figure indicating bidirectional antenna feature, which is similar to conventional dipole antenna. At higher frequency nulls are present in both directions. The cross polarization in H-plane at 5.1 GHz is stronger than pattern at 2.6 GHz and 8GHz which indicates that the band resonating at centre frequency 5.1 GHz of bandwidth 810 MHz is likely to be circularly polarized.

III. CONCLUSION

In this study a triple band antenna with pentagon shaped tuning stub is proposed and investigated for wireless applications like WLAN and upper UWB application. A radiation characteristic of ground plane as well as radiating patch and inherent band-notch characteristics is studied to achieve ultra-wide triple band operation with wide aperture area. The feed coupling slot is demonstrated as frequency notch element. The entire antenna structure is less complex and operates with steady gain. The three operating bands, WLAN 802.11a, 802.11b and X-band uplink satellite communication have fractional bandwidth (FBW) 24.43%, 15.37%, and 32.23% with average gain of 1.86 dBi, 0.54 dBi and 5.28 dBi respectively.

ACKNOWLEDGMENT

The authors are thankful to Management of Sinhgad Academy of Engineering, Pune, TCET.
Mumbai, and Dean of NMIMS University, Mumbai for constant encouragement and providing valuable research facilities.

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Propagation of Modes in the Corrugated Waveguide Made of the Silicon Carbide Material

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Abstract—The object under investigation is the corrugated SiC (Silicon carbide) waveguide in the operating frequency range 5–70 GHz at temperatures $T = 1500^\circ\text{C}$, $T = 1000^\circ\text{C}$ and $T = 20^\circ\text{C}$. The permittivity of this material is $8 - 2\iota$, when $T = 1500^\circ\text{C}$, $7 - \iota$, when $T = 1000^\circ\text{C}$ and $6 - 0.5\iota$, when $T = 20^\circ\text{C}$ at the cutoff frequencies of the main modes. We have investigated the complex dispersion characteristics of the main modes propagated in these corrugated waveguides. It is the complicated electrodynamical problem. The analysis is made by the Singular Integral Equations’ method. All the boundary conditions at the waveguide walls are satisfied including the edge conditions. The imaginary part of the permittivity could be very large. We discovered the particularity of main mode losses of corrugated waveguide.

Index Terms—corrugated waveguide, Singular Integral Equations’ method (SIE), absorptive material, silicon carbide, main mode.

I. INTRODUCTION

The corrugated structures have been used for more than 20 years in radiating systems, telecommunications, and radio astronomy. The corrugated structures are used for polarization-insensitive shields, reflector panels, stealth structures in RCS-related problem, and horn antennas. A 2.50-GHz corrugated transmission line with a directional coupler for forward and backward power monitoring has been constructed and tested for use with a 25-W continuous-wave gyrotron for dynamic nuclear polarization experiments in [1]. Several new synthesis techniques are described for the design of tapered corrugated waveguide low-pass filters in [2]. A new, approximate, corrugated waveguide filter design method is developed for thin-film optical waveguides in [3]. The method determines both the corrugation period and depth measured along the guide's surface, given a specification of the filter’s reflection coefficient. The design technique is based on a combined effective index approach and Fourier transform, inverse-scattering theory for one-dimensional, dispersionless, dielectric media. Use of the general technique is illustrated by the design of two corrugated waveguide filters. The design results are compared with those obtained using the first Born approximation, nonlinear renormalization, and the exact Gel'fand-Levitan-Marchenko method for two component inverse-scattering systems.

As we see the corrugated waveguide are used in a wide area of applications. Thus it is important to investigate such waveguides and to have some possibility to control the waveguide phase constants and losses, for example, by temperature.

The analysis of modes propagating in the corrugated waveguides attracts much attention today. The group-index engineering in a one-dimensional periodic silicon structure consisting of a deep-etched laterally corrugated waveguide with circular holes patterned onto its wide section is demonstrated in [4]. The theoretical analysis, supported by experimental results, shows that the first-order optical mode can propagate inside the Brillouin zone with a relatively high group index over a wide frequency range. A general formulation for the characterization of corrugated waveguides is presented in [5]. The formulation is based on modal expansion in the different smooth-walled waveguides which constitute the corrugated structure and on the use of mode
matching at discontinuities. The use of an admittance matrix formulation and a suitable root-finding algorithm leads to a rigorous and efficient technique. Dispersion curves are presented for corrugated waveguides of circular and rectangular cross-sections. The effect of the finite thickness and width of teeth and slots on the dispersion behavior is shown.

A linearly polarized mode basis set for oversized, corrugated, metallic waveguides is derived for the special case of quarter-wavelength-depth circumferential corrugations. The relationship between the linearly polarized modes and the conventional modes of the corrugated guide is shown in [6].

The paper [7] presents the design of square-to-circular waveguide transitions. Design and analysis procedure is performed through the commercially available field solver HFSS and CST. The field matching technique was used for the modal analysis of a circular waveguide [8], which is corrugated to form discs between corrugations. The novelty of our work is the analysis of the corrugated waveguide made of radiation absorptive material by means of our software.

Here we review the analysis of the SiC corrugated waveguide presented in Fig. 1 by the Singular Integral Equations' method. By the means of this method and our Matlab computer program we could analyze the corrugated waveguides of different cross-sections shapes and investigate all the waveguide eigenmodes. The dimensions, shapes of tooth and the distances between them could be arbitrary. In this work we review the dependencies of the main mode upon the temperature. We have already made analysis of some circular and rectangular SiC waveguides [9].

II. INVESTIGATIONS OF THE CORRUGATED WAVEGUIDE BY THE SIE METHOD

Here the corrugated waveguide (see Fig. 1) made of silicon carbide (SiC) material is investigated by the SIE method. It should be mentioned that the analyzed waveguide structure has a lot of rectangular corners. The conditions at the waveguide edges are satisfied automatically using the SIE method. The integral solutions (1) of the Maxwell’s equations are transformed to the Cauchy’s integrals at small distances from the waveguide edges. The Cauchy’s integral is the exact solution of the Laplace’s equation. The SIE method allows taking into account all boundary conditions which are in the full agreement with the configuration of the waveguide structure. We discovered the particularity of the dispersion characteristics of such waveguide structure. The computer programs have been created in Matlab. The software was approbated and results of the simulations of the conventional waveguides have been compared with the results of the other authors [10].

Fig. 1. The cross-section of the investigated corrugated waveguide. \( a = 7 \) mm, \( b = 7 \) mm, \( t_1 = t_2 = t_3 = t_4 = 1 \) mm

A. The integral representation of solution to Maxwell’s equations

The Maxwell’s equations for this boundary problem have been solved by the electrodynamically rigorous SIE method. This method attracts much attention due to its positive features. The shape of the cross-section of the investigated waveguide structure may be arbitrary (Fig. 2).
In Fig. 1 the contour $L$ separates materials with different permittivities and permeabilities. We assume that the area $S$ is the waveguide surrounding media. All the boundary conditions are satisfied. We present here our researches of the square waveguide which boundary surfaces which are parallel to the $x$- and $y$- axes. At the boundary which is parallel to the $x$-axis we assume that $E_x^+|_{Lx} = E_x^-|_{Ly}$, where contour $L$ is composed from segments $Lx$ and $Ly$. At all the boundaries we assume that $E_z^+|_L = E_z^-|_L$. The analogical boundary conditions are satisfied for the magnetic field. The longitudinal components of the electric $E_z(r)$ and magnetic $H_z(r)$ fields at the boundary $L$ have the integral representation:

$$E_z(r) = \left( \int_L r s (H_0^{(2)}(r^*) k_z r^*) ds \right)$$

$$H_z(r) = \left( \int_L h s (H_0^{(2)}(r^*) k_z r^*) ds \right)$$

(1)

where $H_0^{(2)}$ is the Hankel function of the zeroth order and the second kind.

![Fig. 2. Geometry of an arbitrary cross-section rod waveguide and the SIE method notations](image)

Here the magnitude $s$ is the arc abscissa and $ds$ is an element of contour $L$. The magnitudes $\mu_h(\vec{r}_s)$ and $\mu_e(\vec{r}_s)$ are the unknown functions satisfying the Hölder condition [10]. We apply the Krylov–Bogoliubov method whereby the contour is divided into $n$ segments and the integration along a contour $L$ is replaced by a sum of integrals over the segments. The expressions of all the electric field components are presented below.

The longitudinal components of the electric field for the area $S^+$ and $S^-$ are:

$$E_z^+ = \sum_{j=1}^{n} \mu_e^+(s_j) \int_{\Delta L} H_0^{(2)}(k^+_{L} r^*) ds$$

$$E_z^- = \sum_{j=1}^{n} \mu_e(s_j) \int_{\Delta L} H_0^{(2)}(k^-_{L} r^*) ds$$

(2)

After substitution of the longitudinal field components (2) in the transverse component formulae [11] we obtain the expressions at the contour points:

$$\left( E_x^+ \right) = -\frac{2\mu_0 \mu_e^m \omega \cos \theta}{k^+_{L}^2} \mu_h^+(s_j) -$$

$$- V^+ \left[ k^+_{L} \sum_{j=1}^{n} \left( \mu_h^+(s_j) \right) \int_{\Delta L} H_1^{(2)}(k^+_{L} r^*) \frac{y_S-y_0}{r^*} ds \right] +$$

$$+ Q^+ \left[ k^+_{L} \sum_{j=1}^{n} \left( \mu_e^+(s_j) \right) \int_{\Delta L} H_1^{(2)}(k^+_{L} r^*) \frac{x_S-x_0}{r^*} ds \right]$$

$$\left( E_x^- \right) = -\frac{2\mu_0 \mu_e^m \omega \cos \theta}{k^-_{L}^2} \mu_h^-(s_j) -$$

$$- V^- \left[ k^-_{L} \sum_{j=1}^{n} \left( \mu_h^-(s_j) \right) \int_{\Delta L} H_1^{(2)}(k^-_{L} r^*) \frac{y_S-y_0}{r^*} ds \right] +$$

$$+ Q^- \left[ k^-_{L} \sum_{j=1}^{n} \left( \mu_e^-(s_j) \right) \int_{\Delta L} H_1^{(2)}(k^-_{L} r^*) \frac{x_S-x_0}{r^*} ds \right]$$

(3)

(4)

(5)
\[ \left( E_y \right)_j = \frac{2\mu_r^a\mu_t^a\omega\cos\theta}{ \left( k_\perp \right)^2} \mu_h(s_j) - \] (6)

\[ -Q \left[ \sum_{j=1}^{n} (\mu_c(s_j)) \int_{\Delta L} H_1^{(2)}(k_\perp^r) \frac{V_s - V_0}{r'} dr \right] \]

\[ -V \left[ \sum_{j=1}^{n} (\mu_h(s_j)) \int_{\Delta L} H_1^{(2)}(k_\perp^r) \frac{x_s - x_0}{r'} dr \right], \]

where \( \chi_+ = i\varepsilon_\sigma \varepsilon_\sigma^m \omega / \left( k_\perp^+ \right)^2 \), \( \chi_- = i\varepsilon_\sigma \varepsilon_\sigma^m \omega / \left( k_\perp^- \right)^2 \),

\( V_+ = i\mu_\sigma \mu_\sigma^m \omega / \left( k_\perp^+ \right)^2 \), \( V_- = i\mu_\sigma \mu_\sigma^m \omega / \left( k_\perp^- \right)^2 \),

\( Q_+ = ih / \left( k_\perp^+ \right)^2 \), \( Q_- = ih / \left( k_\perp^- \right)^2 \), \( h = h' - ih'' \) is the complex propagation constant where \( h' \) is the real part (phase constant) and \( h'' \) is the imaginary part (losses) of the value \( h \). The field components and the values of the unknown functions \( \mu_c(s_j) \) and \( \mu_h(s_j) \) are noted in the upper–right corner with the sign corresponding to different area \( S' \) or \( S \), i.e., the unknowns functions inside of waveguide are \( \mu_c^+(s_j) \), \( \mu_h^+(s_j) \) and outside of waveguide are \( \mu_c^-(s_j) \), \( \mu_h^-(s_j) \).

These functions at the same contour point are different for the field components in the regions \( S' \) and \( S \), i.e., \( \mu_c^+(s_j) \neq \mu_c^-(s_j) \). The magnitude

\[ k_\perp^+ = \sqrt{k^2 - e_c^2 \mu_t^m} \] is the transversal propagation constant for the SiC medium in the area \( S' \) and

\[ k_\perp^- = \sqrt{h^2 - k^2 e_c^2 \mu_t^m} \] is the transverse wave numbers in the area \( S \) (Fig. 2). Magnitudes

\[ \varepsilon_r^m = \text{Re}(\varepsilon_r^m) - i\text{Im}(\varepsilon_r^m) \]

and

\[ \mu_r^m = \text{Re}(\mu_r^m) - i\text{Im}(\mu_r^m) \] are the complex permittivity and the complex permeability of the SiC media. Magnitudes \( \varepsilon_r^a \) and \( \mu_r^a \) are the permittivity and the permeability of air around the waveguide. The value \( k = \omega / c \) is the wavenumber in air, \( \omega = 2\pi f \), where \( f \) is an operating frequency. The contour \( L \) is divided into \( n \) segments and the length of a segment is \( \Delta L = L/n \) where the limits of integration in the formulas (2–6) are the ends of each segment \( \Delta L \). \( H_1^{(2)} \) is the Hankel function of the first order and the second kind. The angle \( \theta \) is equal to \( 90^\circ \cdot g \) from 1 till 4 for rectangular contours in the formulae (3–6). The expressions for the magnetic field components are similar. The system of the algebraic equations obtained from the boundary conditions is homogeneous. The condition of solvability is obtained equalizing the determinant of the system to zero. The determinant is our dispersion equation. We have used the Müller’s method to find the complex roots. The roots of the dispersion equation give the propagation constants of waveguide modes. After obtaining the propagation constants of desired modes we can determine the EM field of these modes (see, formulae 2–6). For the correct formulated problem the solution is one–valued and stable with respect to small changes of the coefficients and the contour form [10].

III. RESULTS AND DISCUSSIONS

The object under investigation is the corrugated SiC waveguide with teeth (see Fig. 1) in the operating frequency range 5–70 GHz at temperatures \( T = 1500^\circ\text{C} \), \( T = 1000^\circ\text{C} \) and \( T = 20^\circ\text{C} \). Here we used the experimental data of SiC permittivities dependant upon temperature [12]. The permittivity of this material was taken 8 for the main modes. In our calculations the imaginary part of the permittivity is dependent upon frequency like:

\[ \text{Im}(\varepsilon_r^m) = \frac{1}{\omega \varepsilon_0 \rho}, \] (7)

where the value \( \rho \) is the semiconductor SiC material specific resistivity, \( \omega \) is the cyclic operating frequency, \( \varepsilon_0 \) is the dielectric constant.

The dispersion characteristics of such waveguide structures are presented in Fig. 3. The main mode propagated in the corrugated waveguide when \( T = 1500^\circ\text{C} \) is denoted the solid line, the main mode at \( T = 1000^\circ\text{C} \) – by the dashed line, the
main mode at $T = 20^\circ C$ – by the dotted line. In Fig. 3 (a) we present the dependence of the real part of longitudinal propagation constant upon frequency, in Fig. 3 (b) – the dependence of the imaginary part of longitudinal propagation constant upon frequency.

In Fig. 3 we see that there are complicated dependencies with peaks of $h'/k$ and $h''/k$ values on the frequency. We see that real and imaginary parts of dispersion characteristics have unusual protrusions with minimal and maximal values in the narrow frequency range. The last essential attribute can be used for creation of sensitive modulators, filters, absorbers and so on. The dependency of position on the frequency axis of protrusion peaks is an intricate one and should be considered for each specific parameter, i.e. waveguide dimensions and shapes of corrugated waveguide tooth as well as SiC permittivity.

In Fig. 3 (a) we see that when the temperature decreases the cutoff frequency of the main mode moves to the higher frequency range. In Fig. 3 (a) we see that when the temperature increases, the value of the phase constant peak increases. The wavelength of the main mode decreases when the temperature increases. Analyzing the characteristics depicted in Fig. 3 (b), we see that dependence of the waveguide losses upon temperature is quite complicated. We should notice on the base of our additional investigations that there are some frequencies where the values of the losses are smaller for corrugated waveguide than for the rectangular one with the same sizes.

The appearance of the points of minimum and maximum in Fig. 3 could be explained as follows. The electromagnetic field concentrates near the waveguide boundary, i.e. at the interface of the air and SiC environments. The interaction between the SiC and air acts as the concentrator of electromagnetic field of the waveguide main mode. Changing the frequency and the temperature we could achieve the case when the field is drawn into or pushed out the waveguide.

In Fig. 4 we present the dependencies of the transverse propagation constants $k_{\perp}$ outside of SiC waveguide upon frequency. The distribution of the value $k_{\perp}$ is an interesting one for the analysis because this value is a part of Hankel function argument. The Hankel function describes the behavior of the electric and magnetic fields (see formula (4) and (6)). In Fig. 4(a) we present the dependence of the real part of transverse propagation constants $k_{\perp}$ and in Fig. 4(b) – of the imaginary its part on frequency at different temperatures.
IV. CONCLUSIONS

- We have created the new algorithm for investigation of the corrugated waveguides made of absorptive materials.
- We have discovered the anomalous behavior of the dispersion characteristics of chosen corrugated waveguide at different temperatures in the wide temperature range.
- The real and imaginary parts of dispersion characteristics have unusual protrusions with minimal and maximal values in the narrow frequency range. This essential attribute can be used for creation of sensitive modulators, filters, absorbers and so on.
- It should be mentioned that the losses of the main mode at the temperature $T = 1000^\circ$C are smaller than the losses of the main mode at the temperature $T = 20^\circ$C.

REFERENCES


Geo-textile Based Metamaterial Loaded Wearable Microstrip Patch Antenna

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Abstract- In this paper, a geo-textile material that is polypropylene based metamaterial loaded wearable T-shaped microstrip patch antenna for public safety band applications is presented. Under unloaded condition of T-shaped microstrip patch antenna, poor matching is observed in the public safety band. Further, this antenna is loaded with metamaterial split ring resonators (SRRs). This loading provides better matched condition in the public safety band. In loading condition, the antenna resonates at 4.97 GHz with the bandwidth and gain of 50 MHz and 6.40 dBi respectively. The electrical dimension of the proposed wearable antenna is $0.37\lambda \times 0.37\lambda$. The antenna is fabricated, tested, and the measured results are presented in the paper. An equivalent circuit of the designed antenna under loading condition is also prepared and analyzed.

Index Terms- Public safety band, geo-textile, polypropylene, metamaterial, split ring resonator, negative permeability, equivalent circuit.

I. INTRODUCTION

The Federal Communication Commission (FCC) allotted a separate frequency spectrum of 4.94 GHz to 4.99 GHz that is 50 MHz band for public safety applications dedicated to the protection of human life, health, and property where point-to-point or point-to-multipoint connectivity is necessary [1-2]. It covers the applications such as fire fighter, police vehicles, offsite workers and rescue teams, private ambulance services, military services, airport and seaport surveillances so that the interior and sensitive locations can be monitored round the clock for the protection of human life and property. The compact, light weight, efficient, and easily installable antennas are essential in the public safety band. Looking into the safety of human life it is essential to develop the wearable antenna that can be easily integrated as a part of military uniforms, fire fighting and police garments, military tent clothing, seat belts and covers of military and police vehicles. The limitations of conventional microstrip patch antennas are more deposition of electromagnetic signals in the human body that is high specific absorption rate (SAR) though their physical size is large. Secondly, due to size it is difficult to integrate and make them hidden inside the clothing of wearer [3-9]. Hence, a wearable antenna that is the textile based antenna is one of the better alternatives for such type of applications. The wearable antenna should be light weight, flexible, compact, hidden and should be easily integrated within the clothing and it should not affect the health of wearer.

In practice, different natural as well as synthetic textile materials such as nylon, cotton, Jean, polyester, Teflon, Nomex, liquid crystal polymer (LCP), fleece fabric etc. are used as a substrate to manufacture the wearable antennas for industrial scientific and medical (ISM) band applications [3-9]. In the literature, different wearable antennas are fabricated on various textile substrates for body centric communication systems are reported that covers Wi-Fi, Wi-Max, WLAN, HYPER LAN, body area networks (BAN), Bluetooth applications. Hall P. S. and Hao Y. presented a study on the necessity of wearable antennas for personal area networks (PAN), body area network (BAN) and ISM band [3]. Recently, the authors reported a metamaterial embedded wearable rectangular microstrip patch antenna for IEEE...
802.11a WLAN applications [4]. C. Hertleer et al. presented ISM band microstrip patch antenna on flexible pad foam substrate for protective clothing of fire fighter [5].

Metamaterial inclusions are directly used as loading element for size reduction, enhancement of gain, bandwidth, directivity and efficiency of microstrip patch antennas. In the literature, different metamaterial loaded microstrip patch antennas are reported [2], [4], [11-18]. These inclusions match the impedance at frequency which is lower than the initial resonant frequency of the unloaded microstrip patch antenna. Under loading condition, the microstrip patch antenna generates sub-wavelength resonances due to modifications of the resonant modes. In 1968, Veselago theoretically predicted that metamaterial possesses negative values of magnetic permeability (µ) and electric permittivity (ε) [19]. Some metamaterial structure consists of SRRs to produce negative permeability and thin wire elements to generate negative permittivity [11-23]. Authors have reported the effect of mutual inductance on the resonant frequency, bandwidth, gain, and size of metamaterial loaded electrically small microstrip patch antenna when the loading distance between the metamaterial element and the antenna gets varied [14]. This concept is used by the authors to design and fabricate the polypropylene based metamaterial loaded microstrip patch antenna for public safety band applications.

In this work, the authors proposed a geo-textile material that is polypropylene substrate based wearable antenna for public safety band applications. The polypropylene is a non-woven type of geo-textile which is used as a substrate because of its features such as light weight, the polypropylene sheets are available in different thickness which avoids the processes like sewing to obtain the substrate of desired thickness. The objective of this paper is to design, fabricate and test the metamaterial loaded polypropylene based wearable antenna for public safety band applications. In this antenna, a T-shaped microstrip patch is loaded with four metamaterial square SRRs of equal dimensions by placing them around the patch. The proposed antenna is fabricated, tested, and the measured results are presented in the paper. An equivalent circuit model of the T-shaped microstrip patch antenna under loading condition is also prepared and analyzed in this paper.

The paper is organized into following sections. In Section II, the geometrical sketch and design of the proposed wearable antenna is presented. Section III presents the results of T-shaped microstrip patch antenna under unloaded and loaded conditions. The metamaterial characteristics of the square SRR are also studied and presented in this section. In Section IV, the equivalent circuit analysis of the designed antenna is presented. The paper is concluded in Section V.

II. ANTENNA DESIGN

Fig.1 shows the geometrical structure of metamaterial loaded wearable T-shaped microstrip patch antenna designed and fabricated on the geo-textile polypropylene substrate.

![Fig. 1. Geometrical sketch of metamaterial loaded wearable T-shaped microstrip patch antenna on geo-textile polypropylene substrate](image-url)
Fig. 2 (a) and 2 (b) respectively depicts the photographs of the radiating patch and ground plane of the fabricated antenna. The antenna consists of two rectangular microstrips which are overlapped on each other to form a T-shaped microstrip patch. The purpose to select the T-shape is to increase the resonant length of the antenna so as to reduce the size of antenna and to accommodate the metamaterial SRRs to achieve the further size reduction. The dimensions of the T-shaped microstrip patch are; length of horizontal and vertical microstrips \( L_h = 22.5 \, \text{mm} \) and \( L_v = 10 \, \text{mm} \) respectively. The width of both microstrips is \( W_r = 3 \, \text{mm} \). The geometrical dimensions of a square SSR are; length of outer split ring \( L_s = 9 \, \text{mm} \), gap at the split of both rings \( g \), width of the rings \( w \) are set to; \( g = w = 1 \, \text{mm} \), and separation between the inner and outer split rings \( s \) is 0.5 mm. The electrical size of the square SRR is 0.253 \( \lambda \times 0.253 \, \lambda \) (\( \lambda \) is the free space wavelength at resonance frequency of square SRR 8.45 GHz). The T-shaped microstrip patch antenna is loaded with such a four square SRRs that are placed at the distances \( g_1 = 1 \, \text{mm} \), \( g_2 = 0.75 \, \text{mm} \), \( g_3 = 4.5 \, \text{mm} \), and \( g_4 = 0.6 \, \text{mm} \) to obtain the resonance frequency of public safety band. The aspect ratio of horizontal rectangular microstrip that is length \( (L_w) \) to width \( (W_r) \) is fixed to 7.5 similarly, the ratio of the gaps between upper square SRRs \( (g_3) \) to \( (g_1) \) is also set to 7.5. The ratio of the gaps of lower square SRRs \( (g_4) \) to \( (g_2) \) is set to 1.33. The length of vertical rectangular microstrip \( (L_v) \) is fixed to 1.33 times the aspect ratio of horizontal microstrip. According to the designed dimensions and shapes the radiating patch, square SRR, and ground plane of the antenna are cut from the self adhesive copper tape of thickness 0.1 mm and tightly adhered on the polypropylene substrate. The size of this antenna at resonance frequency 4.97 GHz is 0.369 \( \lambda \times 0.369 \, \lambda \). The finely cut microstrip patch and the SRRs are tightly adhered on the polypropylene substrate. This antenna is designed and simulated on polypropylene (PR 30) substrate of thickness \( h = 1.9 \, \text{mm} \) and dielectric constant \( \varepsilon_r = 2.2 \) supplied by TECHFAB India, Mumbai, India. The antenna is co-axially fed at \( x = -7.2 \, \text{mm} \) and \( y = 0 \, \text{mm} \). The proposed antenna is entirely handmade and high degree of accuracy is maintained in the entire fabrication processes. Method of moment based IE3D electromagnetic simulator is used to simulate this antenna.

III. RESULTS & DISCUSSION

Fig. 3. Simulated return loss \( (S_{11}) \) characteristics of unloaded wearable T-shaped microstrip patch antenna on polypropylene substrate
Fig. 3 shows the return loss ($S_{11}$) characteristics of unloaded T-shaped microstrip patch antenna. In this configuration, poor impedance matching is observed at 4.95 GHz that is in the public safety band. However, the feed point location is rigorously determined on the entire patch to obtain the good matched condition in the proposed frequency band. Hence, to obtain the good impedance matching of the antenna at the public safety band the proposed T-shaped patch is loaded with metamaterial square SRR inclusions as shown in Fig. 1. The metamaterial characteristics of the square SRR are verified and presented before analyzing the loading effect on microstrip patch.

Fig. 4 shows the reflection ($S_{11}$) and transmission ($S_{21}$) coefficient characteristics of square SRR that resonates at 8.48 GHz. The effective medium theory is used to verify the permeability ($\mu_r$) and permittivity ($\varepsilon_r$) from the reflection and transmission coefficients ($S$-parameters). The Nicolson-Ross-Weir (NRW) approach is used to obtain the effective medium parameters. The expressions of equations (1) and (2) are used to determine these effective parameters. The metamaterial characteristics of the SRR are verified using the $S$-parameters obtained from IE3D electromagnetic simulator and MATLAB code with mathematical equations (1) and (2) [13-18],[21-23].

$$\mu_r = \frac{2}{jk_0 h} \frac{1 - V_2}{1 + V_2}$$  \hspace{1cm} (1)

$$\varepsilon_r = \frac{2}{jk_0 h} \frac{1 - V_1}{1 + V_1}$$  \hspace{1cm} (2)

where $k_0$ is wave number, $h$ is substrate thickness; $V_1$ and $V_2$ are composite terms to represent addition and subtraction of $S$-parameters. The factor $k_0 h = 0.336$ which is $<<1$ [21-22]. The values of $V_1$ and $V_2$ are calculated using equations (3) and (4) [13-18], [21-23].

$$V_1 = S_{21} + S_{11}$$  \hspace{1cm} (3)

$$V_2 = S_{21} - S_{11}$$  \hspace{1cm} (4)

From Fig. 4 good matching is observed near the resonant frequency of SRR that is at 8.48 GHz in the range of 8.35 GHz to 8.7 GHz.

Fig. 5 depicts the relative permeability ($\mu_r$) characteristics of the SRR which indicates that the SRR structure is single negative (MNG) metamaterial. The value of permeability ($\mu_r$) is negative in the frequency range of 8.35 GHz to 8.7 GHz. Such four square shaped SRRs are used to load the T-shaped microstrip patch as shown in Fig. 1. In loading condition, a good matching is obtained in the public safety band.
Fig. 6 depicts the simulated return loss ($S_{11}$) characteristics of the metamaterial loaded polypropylene based wearable microstrip patch antenna. This antenna resonates at 4.97 GHz (in the frequency band of 4.94 GHz to 4.99 GHz) with the bandwidth and gain of 50 MHz and 6.38 dBi respectively. The directivity of the proposed antenna is 7.56 dBi. Fig. 7 shows the photograph of the experimental set up of return loss measurement of the fabricated antenna using Bird site analyzer® (Model No. SA-6000 EX, Frequency range 25 MHz to 6 GHz). Fig. 8 shows the measured return loss ($S_{11}$) characteristics of the T-shaped microstrip patch antenna loaded with SRRs. This antenna will find its applications in different public safety band applications.

IV. EQUIVALENT CIRCUIT ANALYSIS & THEORETICAL DISCUSSION

Fig. 6. Simulated return loss ($S_{11}$) characteristics of metamaterial loaded wearable T-shaped microstrip patch antenna on polypropylene substrate

Fig. 7. Photograph of experimental set up to test metamaterial loaded polypropylene based wearable T-shaped microstrip patch antenna

Fig. 8. Measured return loss ($S_{11}$) characteristics of wearable T-shaped microstrip patch antenna on polypropylene substrate.

Fig. 9 shows the equivalent circuit diagram of metamaterial square SRR loaded polypropylene based wearable T-shaped microstrip patch antenna.
In loading condition, the four SRR inclusions are SRRs and calculated to 0.819 nH by using equation (7) [14], [17-18], [24].

\[ M_1 = \frac{\mu_0 L_s}{2\pi} \left[ 0.467 + \frac{0.059w^2}{L_s^2} \right] \]  

(7)

Consider, \( M_2 \) is mutual inductance between the upper SRRs and the horizontal microstrip of the T-shaped patch. Let \( M_3 \) is mutual inductance between the two lower SRRs and horizontal microstrip of the patch. \( M_4 \) is mutual inductance between the two lower SRRs and vertical microstrip of the patch. The mutual inductance \( M_2 \) to \( M_4 \) are respectively calculated to 0.861 nH using equation (8) [14], [17-18], [24].

\[ M_2 = M_3 = M_4 = \frac{\mu_0 L_s}{2\pi} \left[ 0.467 + \frac{0.059W_r^2}{L_s^2} \right] \]  

(8)

The inductance of horizontal \( (L_{hr}) \) and vertical \( (L_{vr}) \) rectangular microstrips of the T-shaped patch is calculated to 14.6 nH and 5 nH respectively using equation (9) [14], [17-18], [24].

\[ L_{hr} = L_{vr} = \frac{\mu_0 L_s}{2\pi} \left[ \ln \left( \frac{2L_s}{W_r} \right) + 0.5 + \left( \frac{W_r}{3L_s} \right) \right] \left[ \frac{W_r^2}{2L_s^2} \right] \]  

(9)

The values of \( L_{hr} \) and \( L_{vr} \) are calculated at the lengths \( L_r = L_h = 22.5 \text{ mm} \) and \( L_r = L_v = 10 \text{ mm} \) respectively. In loading condition, the SRRs are positioned proximity to the horizontal and vertical microstrips of the T-shaped patch. The microstrip patch is coaxially excited hence, due to electromagnetic induction the time varying flux induces the current on each of the square SRR used for loading the patch. Thus, the electric field is induced across the gap capacitance of the splits and mutual capacitance (capacitance per unit length) between the inner and outer splits rings of the SRRs. The inductance of rectangular microstrip patch antenna with the capacitance of SRRs and the mutual inductances forms the LC resonant circuit.
of the loaded antenna. This capacitance compensates inductance of T-shaped microstrip patch and the good matching is obtained at the lower resonant frequency 4.97 GHz. Thus, the capacitance of SRRs and the mutual inductance are sufficiently large to match with inductance of the T-shaped microstrip patch. Therefore, the negative permeability SRRs act as matching elements at the lower resonant frequency 4.97 GHz. In loading condition, the gain of antenna is enhanced because the SRRs acts as matching elements and accepts maximum power from the source for the radiation. Thus, the SRR loading reduces the resonant frequency of the proposed antenna by better impedance matching at 4.97 GHz by reducing the antenna size.

Fig. 10. Simulated current distribution of metamaterial loaded wearable T-shaped microstrip patch antenna on polypropylene substrate (a) Surface (b) Vector

Fig. 10 (a) and Fig. 10 (b) respectively depict the simulated surface and vector current distribution along the designed metamaterial loaded wearable T-shaped microstrip patch antenna. The current is uniformly distributed along the antenna structure as shown in Fig. 10 (a). The arrow shows current flow along the T-shaped microstrip patch and the square SRRs as depicted in Fig. 10 (b). The current is induced in the SRRs because of mutual coupling. Fig. 11 (a) and 11 (b) respectively depicts the azimuth and elevation radiation patterns of the polypropylene based metamaterial loaded T-shaped microstrip patch antenna indicating the gain of 6.38 dBi.

Fig. 11. Radiation patterns of metamaterial loaded wearable T-shaped microstrip patch antenna on polypropylene substrate (a) Azimuth (b) Elevation
Fig. 12. Positioning of the metamaterial loaded polypropylene based wearable T-shaped microstrip patch antenna

Fig. 12 (a) shows the photograph of on body positioning of the fabricated wearable antenna on the clothing of security personnel. Fig.12 (b) depicts the integration of fabricated antenna in the seat cover of a vehicle.

V. CONCLUSION

In this paper, a geo-textile material polypropylene based metamaterial SRR loaded T-shaped microstrip patch wearable antenna for public safety band applications is presented. Under loading condition, the T-shaped microstrip patch resonates in the public safety band. Thus, metamaterial loading is an advantageous approach for size reduction with considerable gain and bandwidth. The SRR loading introduces the inductance, capacitance and mutual inductance to match the impedance at desired resonance frequency. The advantages of proposed antenna are small size, inexpensive, light weight, and easy integration within the clothing. This antenna will find its application in the clothing and helmets of rescue teams, military, fire fighters, security and police personnel, as well as garments of the military tents.

ACKNOWLEDGEMENT

The support of Director, National Institute of Technical Teachers Training and Research (NITTTR), Chandigarh, India is thankfully acknowledged. J. G. Joshi is highly indebted to Director, Directorate of Technical Education, Mumbai (M.S.), India and Principal, Government Polytechnic, Pune, India for sponsoring him to pursue full time Ph.D. under AICTE sponsored Ph.D. QIP (POLY) scheme.

REFERENCES


All Optical Solution for Free Space Optics
Point to Point Links

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Abstract – The focus of this paper is to eliminate the electrical devices for FSO point to point links by replacing them with optical devices. The concept is similar to an extended beam connector. The aim is to achieve a detectable signal of 1nW at a distance of 500 meters at a wavelength of 1500-1600nm. This leads to application in building to building links and mobile networks. The research examines the design of the system in terms of generating the wave, the properties of the fiber feeding the wave, and the power necessary to achieve a usable distance. The simulation is executed in Code V by Synopsys, which is an industry standard to analyze optical systems.

Index Terms: Fiber Optics, Communications, Axicons, Free Space Optics, Bessel and Code V.

I. INTRODUCTION

Free Space Optics (FSO) provides a hybrid optical solution in point to point Local Area Network (LAN) links. FSO systems are highly preferred when electromagnetic interference is high in a particular region and traditional fiber optics link are too costly. FSO systems may also be useful where the link is temporary or mobile. The amount of conversions needed for a FSO link diminishes its effectiveness. Information travels down the optical LAN backbone to the FSO system. The FSO system’s laser emits an optical beam across the atmosphere to a detector which feeds the signal to the other end user. This system requires four conversions. The optical LAN must convert the signal to an electrical signal to modulate the laser. The laser then converts the electrical signal to an optical signal. On the detector side, the photo detector reverses the process. By removing the conversions and replacing the components with optical devices, system delays are reduced, high fidelity is achieved and overall cost may be reduced through the decrease of power consumption and materials. Furthermore, electrical devices must be phased out of networks to achieve all the benefits of optical devices throughout the entire network. Therefore, this research proposes a new way of establishing point to point FSO links using only optical devices.

This paper proves that replacing all the electrical devices with optical devices is possible and feasible through a lens system. The concept is similar to an extended beam connector. However, where an extended beam connector deals with a gap of a few millimeters, this FSO link covers a distance from 100 meters to one kilometer. This distance covers the normal range needed to establish useful point to point links in a FSO LAN. The aim is to achieve a detectable signal of 1nW at a distance of 500 meters at a wavelength of 1500-1600nm. With 1nW at the edge of the receiving fiber, the signal should be strong enough to flow through the receiving LAN despite any attenuation. Wavelengths of 1500-1600nm are commonly used in fiber optics systems, are subject to less atmospheric noise and due to Erbium Doped Fiber Amplifier (EDFA) being used in optical systems [1].
II. BASIC THEORY AND ANALYSIS

A complete optical FSO point to point link provides an alternative to the conventional FSO point to point link. A complete optical system provides better bandwidth and fidelity to the signal. It reduces the delay caused by the modulation and demodulation of a signal. Furthermore, the system’s dimension and cost makes it ideally suited for mobile networks. The simplicity of the device increases the reliability. There is neither moving pieces to break nor any electronics that would be damaged due to adverse weather conditions. The designed optical FSO point to point link does not have any mechanism to prevent pointing loss due to unwanted movements of the platform. In order to prevent excess pointing loss, the distance between the point to point systems should stay around 100m to 500m. This distance is used commonly in inter-building communication and serves campus locations and mobile networks well.

A. Intensity Analysis

The largest obstacle is in transmitting the beam across the atmosphere. Gaussian beams suffer from dispersion and diffraction, limiting the range of the system. The standard way of overcoming this problem is to boost the power of the signal or to reduce the beam divergence to a few milli-radians. The Bessel beam is able to overcome the problem of diffraction. Dispersion may still exist because the beam generated is not a true Bessel beam, but a Bessel/Gaussian hybrid. As long as R, the radius of the axicon, is larger than the full wave half maximum (FWHM) of the generated pulse, the Bessel beam always outperforms the Gaussian beam. The normalized intensity from the axicon is obtained by taking the square of the Huygens Fresnel equation in cylindrical coordinates as [2]:

\[
I(r, z) = \left(\frac{k}{z}\right)^2 \int_{r_a}^r r' e^{jkr'z^2} \left|J_0(\frac{krr'}{z})\right|^2 dr' \tag{1}
\]

where

\[r' = \text{transverse distance to the optical axis} z \text{ (for on axis intensity, this is equal to} 0)\]
\[r = \text{radius of the feeding beam}\]
\[J_0 = 0^\text{th} \text{ order Bessel Function}\]
\[(r', z) = \text{arbitrary point in the direction of propagation}\]
\[\rho = \text{radius of aperture}\]
\[\rho_a = 0 \text{ (in this case) or inner radius of an annular lens}\]

The system is concerned primarily with the on-axis intensity, as this part carries the signal. When this intensity is evaluated on the axis of propagation the equation is simplified as:

\[
I(0, z) = \left(\frac{k}{z}\right)^2 \int_{\rho_a}^\rho r e^{jkr^2} \left|J_0(\frac{kr}{z})\right|^2 dr \tag{2}
\]

The cross section of the intensity can be calculated by combining the Bessel Equation and the Gaussian equation. This equation is a combination of a pure Bessel function and the equation for a Gaussian pulse. The cross section intensity is given by:

\[
I(r) = \left|J_0(\rho r) e^{\frac{-r^2}{2\sigma^2}}\right|^2 \tag{3}
\]

As the radial wave vector is equal to the radius of the beam formed by the collimating lens, this can be used to relate the NA of the fiber to the collimated beam generating the Bessel beam. The radial wave vector can be related to the numerical aperture by:

\[
\alpha = \frac{f}{\sqrt{\left(\frac{1}{NA \cdot \pi}\right)^2 - 1}} \tag{4}
\]

This equation gives insight to the output and the acceptance angle of the fiber, which defines the initial parameters from which to start designing the axicon.

B. Collimating Lens Analysis.

The collimating lens ensures that the rays hit the axicon parallel to each other. Without this condition, the phase function would not refract
the rays at the proper angle. The collimating lens has a pretty straightforward design. The effective focal length (EFL), can be found from the radius of the axicon and the numerical aperture as [3]:

$$EFL \approx \frac{R}{2 \times NA}$$

(5)

where

R = radius of the axicon
NA = numerical aperture of the fiber

The focal length of the collimating lens defines the distance from the fiber to the lens. The proper placement of the fiber generates a plane wave. Code V uses the following formula to express aspheric lenses [4].

$$z = \frac{cy^2}{1 + \sqrt{1 - (1 - e^2)c^2y^2}} + a_4y^4 + a_6y^4 + \cdots$$

(6)

This formula normally is solved through linear programming. In this research, the optimization program in Code V provides the solution for this equation, adjusting the lens so that the divergence angle from the collimating lens becomes almost zero [4]. Code V is an optical design and analysis program developed by Synopsys Inc [4]. The Optical Telecommunication system industry has been using Code V to design and simulate isolators, couplers, wavelength-division multiplex filters and other optical devices. Code V provides optimization and analysis capabilities for an extremely broad range of lenses and optical surface types. This software includes numerous analysis tools which simulate the propagation of light at a wide spectrum of frequencies and presents the output in graphical format or detailed text. The software is ideal for this research because of its analytical tools and its common use in the industry.

C. Axicon Analysis

As discussed earlier, there are various shapes and designs for axicons. Each has their own advantages and disadvantages. The logarithmic axicon gives a more uniform intensity over a given range. The linear axicon provides four distinct advantages. First, the linear axicon provides more power at the desired distance. The design of the axicon is less complex, which decreases the cost of the axicon. Lastly, the simple design allows multiple ways to design and to simulate the axicon.

In order to extend the beam at large distances, the axicon angle must be very small. To achieve a longer range, it would be desirable for the depth of focus region to start away from the axicon. The traditional cone axicon’s phase function is defined when \(d_1\) equals zero. Thus the traditional axicon does not provide the distance necessary to form the FSO link. This means that an axicon must be approximated using a diffractive or GRIN profile. While the diffractive axicon uses either a conic lens or diffractive gratings to approximate the phase function, the GRIN axicon varies the refractive index of the material to map the phase function. Either design begins by defining the phase function. This phase function dictates the maximum range for the system. Additionally, the angle dictates the location where maximum intensity is obtained. The phase function for a linear axicon is given by [5]:

$$\varphi^+(r) = \frac{[1 + a]r^2 + d_1^2 r^{2/1/2}}{1 + a}$$

(7)

Where

\(r = \) any point \(r\) along the radius of the axicon
\(d_1 = \) the beginning distance of the depth of focus
\(a = \) a variable defined by the following equation as [5]:

$$a \pm \frac{2\pi P_o}{c} = \frac{d_2^2 - d_1^2}{R^2}$$

(8)

Where

\(d_2 = \) the end distance of the depth of focus
\(c = \) estimated slope of the power of the axicon
\(R = \) the radius of the axicon

Keeping the initial power and the average power per unit length as constants, the following relationship is obtained.
This gives a more useful equation for determining the range of the Bessel Beam in terms of the original power, the radius of the axicon and the final power at the end of the depth of focus region as:

$$2\pi P_\sigma R^2 = P_z (d_2 - d_1)$$ \hspace{1cm} (9)

The definition of the depth of focus region provides the basis for the design of the axicon. Theoretically, this distance can extend several kilometers depending on the input power and the detectable power. For practical purposes, this value remains around 100-400 meters. The hyperbolic lens provides a simpler yet just as efficient design and could be readily integrated into Code V. Code V models the hyperbolic axicon with the following equation when $K < -1$:

$$y(x) = \sqrt{2Rx + (K + 1)x^2}$$ \hspace{1cm} (11)

Where

$R =$ Radius of the lens

$K =$ conic constant

The conic equation looks very similar to the phase function given by Equation (7). By setting $R$ equal to 0, the approximation below works well for modeling the axicon:

$$K \propto -\alpha$$ \hspace{1cm} (12)

This formula results in a diffractive axicon with a circular grating. This combination acts similar to the annular lens. In the case of a linear axicon, two terms in the series provide the best match for the phase function.

III. SIMULATION METHOD

Once the lens designs are transformed into the Code V format, the data can be entered into the Lens Data Manager (LDM). This interface allows the user to define the radius and position of the lenses. In this research, the object represents the end of the fiber and the image represents the detector. Code V does not have a way to easily measure the intensity along the axis of propagation, so the image surface must be moved along the axis and the data recorded at each point. The surface properties option in the LDM specifies the diffraction grating or the phase function of the lens. The numerical aperture and pulse type can be set here to best simulate the optical communication system. Code V has numerous optimization tools. These functions create an aspheric lens or other lens shapes that have a near perfect focus at the image distance or other user defined area. These optimization functions could not be used for the axicon, since the lens needs to focus over a distance rather than just at one point. The output of the optical system is simulated by the Code V Beam Synthesis Propagation (BSP) tool. BSP uses a highly accurate beamlet-based diffraction propagation algorithm in order to measure the intensity of a light source at the image location [6]. BSP’s primary output is a color raster chart that displays the image of the beam perpendicular to the axis of propagation over a specified area or optimized area. BSP also produces text results that give more specific information on the beam propagation and power.

IV. RESULTS

Figure 1: The axial intensity vs. the density for different values of the axicon radius.

(a) $I(0,z) \ R = 0.04 \ m$  \hspace{0.5cm} (b) $I_2(0,z) \ R = 0.03 \ m$

(c) $I_3(0,z) \ R = 0.02 \ m$
A. Analytical Results

Based on the analytical procedure previously discussed, the analytical results have been obtained. Figure 1 shows the relationship between axial intensity and distance. It is observed that as the beam propagates along the z axis, different phase components of the beam leaving the axicon interfere with each other constructively and destructively, leading to the oscillation seen as the intensity rises linearly. The design of the axicon phase function dictates where the depth of the focus lies. The system is found to be most effective when the desired distance falls in the middle of the depth of focus range. The steep decrease in intensity is expected for a Bessel beam. The Bessel beam properties only pertain to the depth of focus region, beyond that the different phase components no longer intersect. As seen by the green line on Figure 1(c), a small radius results in a beam that does not deliver sufficient intensity to the target area of 150m to 400m. The depth of focus manifests itself as the max of each plot before the intensity begins to drop rapidly toward zero. The larger the radius the longer the range and the more power arrives at the desired point. This is shown by the red line on Figure 1(a). This is in agreement with the Rayleigh range formula. This increase in intensity occurs without an increase in the input power of the system. Changes in the wavelength require a redesign of the phase function in order to properly align the phase components.

As the numerical aperture becomes larger and travels through the two lenses, the Bessel beam grows narrower. As the numerical aperture shrinks the central beam becomes wider and approaches a Gaussian beam shape due to the collimating lens effects. As the numerical aperture becomes larger the beam gets focused instead of collimated. And as it becomes smaller, both lenses have a lesser effect on the beam causing the Gaussian shape to remain. The rings of the Bessel Beam move closer and closer to each other because the Bessel Beam is more focused. Each ring gets thinner and eventually the system looks like a very sharp, focused beam.

Figure 2 shows the effect of changing the refractive index of the axicon on the radial wave vector of the Bessel Beam. A larger radial wave vector means the full width half maximum (FWHM) of the Bessel beam becomes more compressed. As calculated from Equation (4), the graph shows that a larger refractive index corresponds to a large radial wave vector and thus a narrower Bessel Beam. However, the greater focus comes at a price. A greater difference in the refractive index increases the reflectivity of the system which results in greater connector return loss. Thus, a trade off must be made between the FWHM of the beam versus the power lost due to reflected beams.

B. Simulation Results.

The system consists of a collimating lens and axicon to produce the Bessel beam. The software Code V 10.4 from Synopsys is selected to simulate the propagation of the system. This simulation uses a fiber with a numerical aperture of 0.37 and a core diameter of 200 µm. These fiber dimensions were chosen because it is commonly used in local area networks. Many local area networks use Cisco network devices to establish the network backbone. These fiber dimensions fit the standard Cisco point to point link between routers and thus fit the optic point to point link well.
Figure 3 shows the Bessel Beam generated by the hyperbolic axicon. It is observed that Figure 3 has significant deformation in the outer rings caused due to hyperbolic lens. Due to the approximation, the lens curves more toward the outside. A smaller axicon radius produces an exaggerated version of this effect. The smaller lens means the aberration occur on the inner rings versus the outer rings. Thus the center airy disk of the Bessel beam forms, but the outer disks compress together creating the larger intensity. A lens with a larger radius also forms a better Bessel beam. However, the aberrations may be used to project more power onto the detector over a short distance, since the system is less stable over a longer range.

In the actual design, a focal length less than the effective focal length prove more useful. This creates a slightly smaller beam that does not go through the edges of the axicon where the shape of the lens deviates from the phase function. As shown in Figure 4, the simulation results follows the curve of the axicon, however, the results do not have a long range and the oscillations are less regular. The simulation also shows that it provides more power for shorter ranges than the theoretically calculated values. The maximum intensity occurs roughly close to theoretical calculations obtained at a distance of about 300m.

A reason for the discrepancy is the difficulty of fitting the phase function to the specifications of the simulation software. The Taylor and conic approximations only work for lens sizes of less than .05m. After that distance, the approximations diverge too far from the phase function to be useful. Even within, the .05m range, the approximations are not perfect fits. This explains the outer rings tendency to blur into each other as shown in Figure 3.

Figure 5 displays the power curve of axicons with different radii. The simulation results do show that the range of the system increases with the radius of the axicon. When the radius doubles, the range also doubles in value. This is in line with the numerical results.
range increases, the input power remains the same. However only increasing the radius is not sufficient in extending the range as shown in Figure 5. It is clear that simply scaling the axicon does not produce the desired results. As the radius increases, the conic constant and the phase function change. The axicon needs to be redesigned for larger radii. However, the range of the axicon is clearly affected by its radius. This is expected since the phase function is dependent on the radius as well as the integral that determines the intensity. Theoretically, as long as the radius increases the range of the device also increases. Practically, this is limited by the ability to create the lens. At very large distances, the phase or conic functions would require unattainable accuracy.

Figure 5 also shows that the power generated from the axicon can be more stable if the power is taken over a larger area rather than at a point. The center airy disk of the axicon maintains a radius of less than 15mm past 500m. With a focusing lens of greater than the airy disk, the oscillations have little effect on the power delivered to the receiver. The intersection of the two curves shows the distance at which the systems need to be replaced. As expected when the input power is raised the power at the detector rises as seen in Figure 6. This causes the increase in the effectiveness of the Bessel’s beam. The measured distance starts at 100m because before that distance, the Bessel Beam is still not fully formed and the width of the beam varies greatly. Throughout the depth of focus region, 150m to 400m, the full width half maximum (FWHM) of the beam stays fairly consistent. This demonstrates that while the central air disk contains less power over a short range, the beam delivers more power over long distances. The drop of in the intensity seen in Figure 4(a) is not only attributed to a drop of power, the FWHM of the beam begins to expand after the distance exceeds the depth of focus region. This causes the intensity at the detector to drop rapidly. The width of the FWHM can be decreased by increasing the refractive index of the material or changing the numerical aperture of the input fiber, but this decreases the power delivered to the target area.

Figure 5: Power curves for axicons with (a) radius = 0.02 m and (b) radius = 0.04 m.

Figure 6: Input power vs. power at detector at peak distance.
Figure 7 shows the numerical aperture for various cross sections of the pulse. It is observed that the pulse cross-section changes with the numerical aperture. Figure 7 (a) shows significant deformation in the Bessel beam. The beam resembles a 1st order Bessel beam, but this is not due to the lens creating a 1st order Bessel beam. The beam seen in Figure 7 (a) is similar to a Gaussian beam. The dip at the maximum occurs because of the center of the axicon. The central spot of the axicon is a source of deformation. The center behaves much like the annular lens center. This means that most of the light is reflected back or refracts away from the axis of propagation. This can be seen clearly when the numerical aperture of the system is less than 0.21. The large dip in the intensity profile was caused by the beam forming around the center of the axicon. As the numerical aperture grows, the beam is wide enough that the outer portions of the beam converge to begin forming the center. As the NA becomes larger and larger, a plane wave is formed before the collimating lens. This causes the collimating lens to do the opposite of its intended purpose and focuses the light instead as shown in Figure 7(b). Recall that a lens that creates an image at infinity generates an image at its focal length when the object is at infinity. Thus a Bessel beam is not formed when the numerical aperture is very large as observed in Figure 7(b).

Figure 8 shows the changes in the Bessel beam profile as a function of the wavelength of the system. It is observed that the wavelength of the design can vary between ±20nm and still retains its ability to project power. The phase function needs to be designed at a certain wavelength. This determines the angle at which the different phase components combine to form the depth of focus region. Figure 8(b) and (c), show the profile of the Bessel beam as the wavelength falls above the tolerance of the axicon design. Both of these figures show a well-defined Bessel beam in the center. However in Figure 8(b), the outer aberrations have become larger and the rings have narrowed. In Figure 8 (a), central airy disk is no longer pronounced and most of the rays are traveling parallel to each other as seen in the large amplitude of the outer rings. The smaller wavelengths are less intense along the axis of propagation and more intense on the outer rings. Figure 8(d) is the most extreme example which demonstrates how the beam would look if the 800nm standard was used on this axicon. Figure 8(c) shows the other side of the spectrum; the larger wavelength generates a wider beam that retains the Bessel properties. The effect of wavelength on the beam can be adjusted by redesigning the lens. The key property here is that the wavelength can vary by 20 nm and the output still remains usable.
V. CONCLUSION

Free space optics is becoming more common in the field of optical communication. FSO systems provide a solution to optical links where fiber installations are too expensive and radio signals encounter too much interference. The bottleneck in the networks due to electrical devices will be eliminated by moving to all optical devices. As the bandwidth and fidelity of optical networks increases, new optical devices are supposed to match these requirements. The results prove that a simple system of lenses can replace the complex electrical components of the traditional FSO point to point systems. The power delivered by the system is enough for an average avalanche photo diode to receive a signal. The system accommodates fibers with a NA between 0.31 and 0.51 without any adjustments.

The system presented here replaces the conventional electrical components with optical components. The design is also very modular. This is the main advantage of an all optical system. While the electrical FSO point to point link requires numerous cables and protection against the elements, this system requires one cable and would not fail due to adverse weather conditions. Cost is another advantage of this design. Due to the casing and electronics in a traditional FSO system, one side of a traditional FSO system can cost at least $6,000 depending on the range and features needed [7]. The raw material cost of the all optical system is around $20 for each element as estimated by Code V cost estimate tool. This brings the estimated total cost of the whole optical system to less than $100. Furthermore, two simple lenses require much less maintenance and care than two electronic devices.

Future research will be adding on to the basic system to provide all the functionality as the electrical point to point links provide while retaining the advantages of the all optical devices. A tracking method to prevent pointing loss needs to be developed, possibly by using higher order Bessel beams. The compatibility
of common coding methods and the Bessel Beam needs to be explored as well as a way of positioning the receiver and transmitter when the beams are beyond line of sight. Lastly, different materials must be explored. Reflective lenses would not experience the reflective loss like the refractive materials. Metamaterials working at the terahertz range could provide many advantages as well.

VI. REFERENCES


CALL FOR PAPERS

The 14th International Symposium on Microwave and Optical Technology (ISMOT-2013) will be held at Kuala Lumpur, Malaysia from 28-31, October 2013. The research fields covered by ISMOT are in the area of microwave components and circuits, optical components, microwave and optical communication systems, electromagnetic theory, antennas, microwave photonics, and all other topics concerning the microwave and optical community. The fundamental goal of ISMOT is to offer an international forum for the exchange of new ideas, thoughts, and realizations on physics, technologies, and applications of microwave, optoelectronics, and related fields. We will be pleased to welcome you to ISMOT-2013 and will be grateful for your contributions at the symposium.

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IMPORTANT DATES & REGISTRATION

Submission of full papers: March 31, 2013
Notification of accepted papers: May 31, 2013
Deadline for pre-registration: August 15, 2013
Symposium registration: October 27, 2013
Symposium dates: October 28 to 31, 2013

All participants to ISMOT 2013 are required to pre-register and pay the non-refundable registration fee. Author registration and payment must be made by September 15, 2013. Papers without a registered author or co-author will not be included in the Conference Proceedings.

Early-bird registration fees (payment by August 15, 2013):
Regular participant: USD 550/-
Student participant: USD 300/-

Pre-registration fees (payment by September 15, 2013)
Regular participant: USD 630/-
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The registration fee includes: the conference kit, attendance in the technical sessions, lunch and refreshments to all the participants. Each presenting author is limited to presenting no more than two papers in oral and/or poster sessions.

Updated and detail information will be posted at http://www.utar.edu.my/ismot2013
The City

Kuala Lumpur (often abbreviated as KL) is the capital and the largest city of Malaysia in terms of population. The city proper, consists of 244 sq kms (94 sq mi), has an estimated population of 1.8 million in 2010. Greater KL, also known as the Klang Valley, is an urban agglomeration of 7.2 million. It is the fastest growing metropolitan region in the country, in terms of population and economy.

Since the 1990s, the city has played host to many international sporting, political and cultural events including the 1998 Commonwealth Games and the Formula One World Championship. In addition, Kuala Lumpur is home to the tallest twin buildings in the world, the Petronas Twin Towers, which have become a symbol of Malaysia’s development.

The Venue

A mere 25 minutes from KL City Centre, this award winning 800 acre "Sunway Integrated Resort City" landmark presents a host of choices of five hotels with over 10,000 square metres of functional space including 55 functional rooms, a convention centre, multiple break-out rooms, the pillar-free Grand Lagoon Ballroom, auditoriums, an amphitheatre, outdoor themed venues, food and beverage outlets, a Balinese inspired spa, London's super dance music club Ministry of Sound-EUPHORIA, a theme park, an upscale shopping and entertainment mall, medical centre, educational facilities and a multitude of distinctive convention, business and leisure facilities.

Travel

In terms of air connectivity, KL is served by KL International Airport (KLIA) at Sepang, Selangor, which is also the aviation hub of Malaysia. KLIA connects the city with direct flights to destinations in six continents around the world, and is the main hub for the national carrier, Malaysia Airlines and low-cost carrier, AirAsia. KLIA can be reached using the KLIA Express high-speed train service from KL Sentral, located at the city center, which takes twenty-eight minutes, while travelling by car via highway will take about an hour. Air Asia flights do not fly out of KLIA main terminal but from the Low Cost Carrier terminal, located adjacent to KLIA, which is served by buses and taxi. It is about an hour’s drive from KL Sentral.

Further details can be found at http://www.utar.edu.my/ismot2013.