

STUDY OF ULTRA WIDE BAND ANTENNAS FOR FUTURE WIRELESS COMMUNICATION DEVICES

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CERTIFICATE

I, **Deepa Negi** hereby declare that the work contained in the thesis entitled "**STUDY OF ULTRA WIDE BAND ANTENNAS FOR FUTURE WIRELESS COMMUNICATION DEVICES**" being submitted by me to the **Department of Electronics and Communication Engineering, Thapar Institute of Engineering and Technology, Patiala** in fulfillment of the award of the degree of "**Doctor of Philosophy**" is a record of authentic research work carried out under the supervision of **Dr. Rajesh Khanna** and **Dr. Jaswinder Kaur**. The matter presented in this thesis does not incorporate any material previously published or written by any other person except where due references are made in the text. The results obtained in this thesis have not been submitted in part or full to any other institute or university for the award of degree or diploma.

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ABSTRACT

The current work provides a detailed analysis of ultrawideband antennas, exploring their various crucial features such as broad band impedance matching, rejection for interfering wireless bands and optimum gain and bandwidth values utilizing novel metamaterial structures. It has been observed from the literature study that as a UWB antenna covers large unlicensed band from 3.1 to 10.6 GHz, therefore it needs to have consistent or stable performance in the entire working bandwidth. It led to the utilization of a broadband impedance matching technique called distributed impedance matching method to obtain satisfactory matching properties in the present work. Along with proper impedance matching, the second important requirement of an UWB antenna is to have good notching characteristics. The method of attaching tuning metallic stubs/strips to antenna structure is presented in this thesis. It is one of the simplest rejection techniques which mainly alters the surface current distribution and provides high impedance at notched frequencies. Now another crucial need of UWB antenna is to have optimal gain and bandwidth values. For this purpose, we designed various novel shaped metamaterial unit cells and then utilized them as a reflector plane along with the designed UWB antennas. As the study of metamaterials have gained huge attention now a days, hence detailed analysis of these unit cells including their complete structural progression, electric and magnetic field distribution, bending analysis, extraction of material properties along with circuital analysis is performed. Further the double negative metamaterials are considered one of the most important members amongst the metamaterial family due to its vast applicability in various fields, hence it is focused more in this work. Further two flexible metamaterial antennas designed gave satisfactory outcomes in terms of their gain, bandwidth and forward radiation characteristics. All the proposed antennas and metamaterial cells have novel structures and gave better outcomes when compared with the latest existing designs.

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LIST OF ABBREVIATIONS

AMC	Artificial Magnetic Conductor
BCWC	Body Centric Wireless Communication
CP	Circular Polarization
CPW	Co-planar Waveguide
CSRR	Complimentary Split Ring Resonator
CST	Computer Simulation Technology
DGS	Defected Ground Structure
DIM	Distributed Impedance Matching
DNG	Double Negative
EBG	Electromagnetic Band-gap
EM	Electromagnetic
EPS	Expanded Polystyrene
FCC	Federal Communications Commission
FF	Fidelity Factor
FFS	Fidelity Factor of the System
FR4	Flame Retardant 4
IFS	Iterated Function System
IR	Impulse Radio
ISM	Industrial Scientific and Medical
LCP	Liquid Crystal Polymer
LHM	Left-Handed Media
MIMO	Multiple Input Multiple Output
MLM	Main Lobe Magnitude
MMW	Millimeter Wave
MNG	Mu-Negative
NRI	Negative Refractive Index
NZIMs	Near-Zero Refractive Index Metamaterials
OFDM	Orthogonal Frequency Division Multiplexing

P	Parallel
PCBs	Printed Circuit Boards
PDMS	Polydimethylsiloxane
PE	Perfect Electric
PEC	Perfect Electric Conductor
PIFA	Planar Inverted F Antenna
PIN	Positive-Intrinsic-Negative
PM	Perfect Magnetic
QWT	Quarter Wave Transformer
RF	Radio Frequency
RFIDs	Radio Frequency Identification
S	Series
SAR	Specific Absorption Rate
SNG	Single Negative
SRRs	Split Ring Resonators
SSF	System Fidelity Factor
UWB	Ultrawide Band
VNA	Vector Network Analyzer
VSWR	Voltage Standing Wave Ratio
WBANs	Wireless Body Area Networks
WIMAX	Worldwide Interoperability for Microwave Access
WLAN	Wireless Local Area Network
ZOR	Zeroth Order Resonance

GLOSSARY OF SYMBOLS

R	Patch Radius
f_r	Resonant Frequency
ϵ_r	Substrates Dielectric Constant
ω	Angular Frequency
BW	Bandwidth
Q	Quality Factor
L	Equivalent Inductance
R	Equivalent Resistance
C	Equivalent Capacitance
W_b	Ground Layer Width
W_n	Ground Step Width
L_b	Ground Layer Length
L_t	Length of Top U-Shaped Structure
S_{11}	Reflection Coefficient
L_{stub}	Initial Lengths of Stubs
λ_0	Free Space Wavelength
f_n	Notching Frequency
C	Velocity of Light
ϵ_{reff}	Effective Dielectric Constant
W_t	Width of Topmost U shape Slot
Z	Total Impedance of Band Stop Structure
Z_A	Input Impedance of Proposed Structure
Z_n	Impedance of Notch
BW_n	3 dB Bandwidth of notch
ω_n	Angular Frequency of Notching Structure
Q_n	Quality Factor of Notching Structure
S_{21}	Transmission Coefficient

ε	Effective Permittivity
μ	Effective Permeability
N	Refractive Index
ε'	Imaginary Effective Permittivity
μ'	Imaginary Effective Permeability
z_m	Wave Impedance
K	Wave Vector
Q	Substrate Height
D	Air Gap
δ	Loss Tangent
H	Gap between Antenna and Human Phantom
X	Air Space within Antenna and Metal Sheet
σ	Tissue Conductivity
E	Electric Field's Root mean square Value
ρ	Density of Human Tissue
Z	Phantom Depth
S	Length of Arc
R	Cylinder Radius
θ	Bending Angle
F	Correlation Factor
τ	Delay Factor
θ^0	Rotation Angle

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CHAPTER 1

INTRODUCTION

1.1 UWB TECHNOLOGY

As the present-day wireless system is expanding, the ultra-wide band (UWB) technology unlocks the new opportunities for wireless communication system. It is a technology which uses extremely small power level digital signal spread over a large frequency band resulting in huge data rates, short range and large channel capacity. Federal Communications Commission (FCC) allowed unlicensed usage of 7.5 GHz of frequency band (3.1 to 10.6 GHz) for various commercial communication systems. As a result, in 2002, the UWB technology obtained a major breakthrough. Apart from this the technology due to its large spectrum interferes with different wireless bands (WiMAX, WLAN) as well [1].

The UWB technology has a peculiar property of using very low energy consumption. As in this the average energy of a UWB signal gets distributed over a wide frequency range resulting in a signal with low power level (lower than noise floor). Therefore, according to FCC spectral mask in a UWB system an antenna utilizes the highest power level of the order of 0.5 mW only. Due to its lower power levels, which is comparable to the noise floor, it is quite difficult to spot and intercept by the intruders. In the present-day scenarios, for various applications including existing as well as upcoming, the UWB technology is preferred [2]. Previously, the technology was mostly utilized for indoor, small distance, handheld and indoor applications. At present UWB antennas have acquired huge recognition not only in academia but also in industry for various wireless transmission system applications [3]. Presently its various applications include wireless communication systems [4], Sensing and Monitoring Systems (RFIDs and rescue operations) [5], Positioning and Tracking systems (Military and Sports) and finally in Radars and Bio-Imaging Systems [6]. Further there are two competing technologies used for the UWB wireless communication; Impulse Radio (IR) UWB technology and Multiband Orthogonal Frequency Division Multiplexing (OFDM) technology.

IR-UWB systems uses very small usually 2 to 3 ns duration pulses and large bandwidth for transmitting and receiving the data. Moreover, apart from having very small spectral density, these systems need to have very small usually in sub milli watt of effective radiated power as well. These

features make the IR-UWB technology more preferable as compared to other signaling techniques. The IR-UWB systems with large bandwidth helps to obtain accurate distance estimation usually in centimeter-level. Moreover, the distance estimations are immune to multipath channel interferences and now a days an IR-UWB transceivers consuming lesser power with less complexity are easy to develop. Today IR-UWB is the foundation of one of physical layers stated by IEEE 802.15.4a standard and is used in cognitive radio applications. Apart from communication applications, these systems are also used in ground penetrating radars as well as the imaging systems.

The multi-band OFDM (MB-OFDM) system uses a time frequency code for transmitting data across 528 MHz large bands which are generated across the entire UWB spectrum. Each sub band utilizes an OFDM modulation technique for transmitting data across UWB spectrum. The integration of multi-banding with OFDM systems have reduced the multipath interference issues, further the benefits of pulsed multi-band systems can also be relished. The multi band OFDM system possess various important features including the potential to capture multi-path energy using a single RF chain, immune to variations due to group delays, and the potential to maintain good data rates even with high small band interferers.

Latest advances in wearable technologies, combined with ever-growing miniaturization of electronic devices, have resulted in the development of a large number of devices which can either be placed on human body or carried in the pockets. It can be seen as a succession of the trend initiated by mobile phones, which during the previous year's turn out to be more appropriate and compact for customized operations. Parallel to this trend, research is going on for designing various body-centric communication systems including dedicated jobs like military personnel, firefighters and paramedics. With increasing presence of wireless communication systems in our day to day lives, the body-centric wireless communication (BCWC) systems will be focused more in the evolution of the fifth generation (5G) mobile communication systems or even further development.

For a wireless communication system, the antennas play the crucial role. Similarly, in a UWB system, UWB antenna is the most essential part. These antennas have acquired notable research as well as recognition in the recent times. Vast improvement is seen in the UWB antenna designing because of growing acceptance for UWB systems. A single UWB antenna can be utilized in place

of multiple narrow-band antennas, as a result the total quantity of antennas is minimized in a system. Oliver Lodge [7] was the first person to bring the idea of UWB antenna design in 1898 and from that day on, a large number of UWB antennas were designed. Unfortunately, these designs were not very useful and suffered from limitations like interference, high cost, large size, complex structure and were found inconvenient for applications including higher frequencies [8].

Features including production price along with complications in designing techniques are very crucial aspects while designing these UWB antennas. Latest UWB antenna configurations concentrate on compact printed antennas due to their potential to integrate with different components on similar printed circuit boards (PCBs) and easy fabrication technique. The UWB systems co-exists with many other narrow bands therefore to reduce the interference from these unwanted bands, notches are introduced.

Ever since the first research article on wearable antennas was published in 1999, huge number of wearable antennas with different arrangements working in the UWB range have been designed during recent years. The utilization of conformal substrates is the most apparent development trend amongst all studied wearable antennas. As compared to the standard metallic shapes like solid and flat copper plates on rigid substrates, conformal antennas are preferred more as these can be easily worn on human body for BCWC systems. Further the wearable antennas aside from being conformal, it also needs to be robust, small size and maintains optimum bandwidth, radiation efficiency, gain values when placed close to human body.

UWB textile and conformal antennas are becoming attractive due to their light weight, conformal, comfortable to wear and easily combined with clothes nature. Textile antennas are amongst the most appealing and progressive research topic of present times as it creates a wearable interface between the machine and human. Because of present day demand for miniaturized wireless devices, the use of wearable textiles in the antenna designing is on high demand. As a result, for body-centric communication systems the study of wearable textile antennas has turned out to be the most leading research areas. In future UWB antennas will be used in wearable and millimeter wave applications.

Although numerous UWB utilizations are developed and proposed in the past years, still more research needs to be done for the complete and proper utilization of UWB band. UWB antenna which is compact, free of interference and highly flexible which can be mounted on any shape of

surface is required to meet the demands of future wireless communication systems. Various challenges faced while designing UWB antennas are interference occurring from other unwanted bands, as it co-exists with other narrow bands. As UWB antennas are extensively used for the wearable applications, therefore back radiations need to be reduced as it lowers the overall efficiency of the antenna. Many studies have been dedicated to assess the performance of UWB antennas. The UWB system must meet the requirements listed below.

1.2 SPECIFICATIONS OF UWB ANTENNA

A UWB antenna also plays a significant part in a UWB system as in case of standard wireless communication systems as well. Moreover, to design a UWB antenna is quite difficult when compared to the narrow band antenna design [8, 9].

Broad operating bandwidths: According to FCC, a UWB antenna is the one which transmit at a large operating bandwidth of equal to/greater than 500 MHz or having fractional bandwidth of 20%. So, it's the large bandwidth of 7.5 GHz which differentiates a UWB antenna from other antennas.

Stable Radiation patterns and gains: A designed UWB antennas performance needs to be uniform and stable in the complete working frequency range. Preferably, the antenna impedance matching along with radiation characteristics and gain values must be consistent over the whole operational band. Meantime, the UWB antenna should also deliver band rejection features so that it can exist together with another narrow-band systems [10, 11].

Large gain transmissions in preferred direction: The properties of the radiation pattern whether it should be omni-directional or directional, it is totally determined by the application scenario. Mostly in case of hand-held and mobile systems, the omni-directional radiation patterns are generally preferred and further the directional radiation patterns are mostly beneficial for radar systems in which large gain values are required.

Low Profile: An appropriate antenna requires to be quite small in size, so as to be compatible with the UWB unit particularly for portable devices and mobiles. An antenna with compact size and easily integration with PCB is mostly preferred.

High transmission efficiency: For the good performance of overall system, a UWB antenna should have optimum outcomes. For instance, the antenna must be constructed in such a way that the entire device including the antenna and radio frequency (RF) front end should agree with the compulsory power emission limits specified by the FCC or other regulative agencies.

Consistent group delays: A UWB antenna is desired to show acceptable time domain features. For transmitting digital signal, a UWB system uses short pulses having small pulse duration along with low duty cycle. In frequency domain the signal with small pulse width is spread over large band, therefore occupying the extensive bandwidth. Therefore, at present a UWB antenna needs to be viewed as a “band-pass filter” instead of a “spot filter”. The antenna imposes a considerable influence on the input signal in such scenarios [12]. Consequently, the main focus of an acceptable UWB antenna is to show adequate time domain characteristics, which means to have a received signal with least pulse contortion as the received signal contains useful information.

Fidelity factor: The actual fidelity factor (FF) definition considers the transmitting antenna effect only and further the correlation factor between the input excitation signal and the radiated signal from the antenna under test are considered [13]. It is the term used to measure the antenna capability to maintain its actual signal shape, which is further utilized for excitation of test antenna.

System Fidelity Factor: The system fidelity factor (SFF), includes a transmitting and receiving antenna connected through a radio channel in between them. Further the correlation factor between the input excitation signals and the received signals by the receiver antenna are studied for calculating the system fidelity factor.

Fidelity Factor of the System: The “fidelity factor of the system” (FFS), considers the antenna spatial radiation characteristics. The correlation factor within the radiation signal received in arbitrary direction and the signal is main lobe direction for two antenna system is studied here. Unlike above two factors it does not measure the amount of signal distortion due to radio channel in between. For calculating it, we need to rotate the receiver antenna to achieve the transmitted signal whose direction is different in comparison to the actual reference direction.

So, from above study it can be concluded that a UWB antenna which is flexible, cost effective and free of interference along with stable/consistent bandwidth, gain along with good forward radiation features in complete operating band is desirable for future wireless communication systems.

1.3 PROBLEMS FACED WHILE DESIGNING UWB ANTENNAS

Large and complex structures: Most of the researchers have focused mainly on the antenna parameters like bandwidth, gain, directivity and large complex structure is not taken into consideration [14-16]. To achieve high bandwidth stacking [17], aperture coupling and fractal geometry are extensively used by the researchers. So, antenna miniaturization has to be considered as these antennas have to be mounted on wireless devices or human body. For miniaturization, we can use DGS, shorting pins and metamaterials etc.

Interference: Due to various narrow band technologies which operates below 10.6 GHz give rise to interference in UWB systems as they work in similar UWB frequency region. A band stop filter which shows notching function at particular stopband frequency is required which will reduce the interference due to various existing systems. Conventionally different lumped elements including capacitors and inductors are used as filters, however these increases the overall size of the systems along with increased system complexity and cost. Different notching methods are used by the researchers and among them insertion of slots [18] and introducing parasitic strips is comparatively easier and further helps in size reduction.

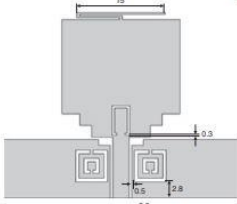
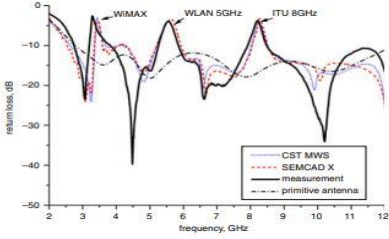
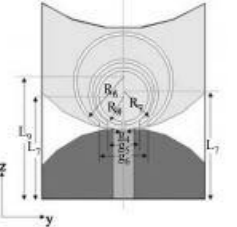
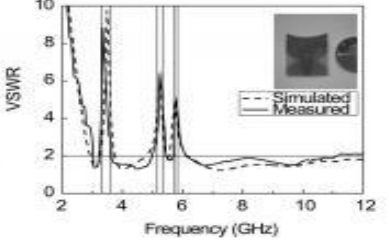
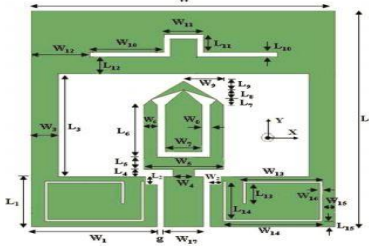
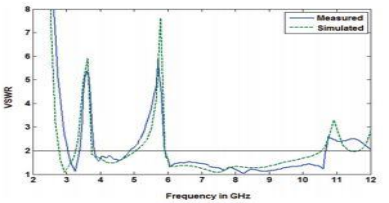
Different notching techniques used by the research workers are categorized on the basis of slot's position whether it is located on feed line, ground plane, radiating patch or in the region nearer to the radiation element [19]. In [20] a tunable band notch features rejecting WLAN interference is acquired by inserting a pair of L shape slots in the corner of an opened radiation slots. Band notch antenna has impedance bandwidth from 2.6 GHz to 13.6 GHz except the notched band from 4.9 GHz to 5.85 GHz. In [21] a dual notch band features rejecting bands from 3.40-3.48 GHz and 5.40-5.98 GHz are generated by using a single complementary split-ring resonator (CSRR). Moreover, the notching element is placed over radiation patch of a monopole antenna operating from 2.90 to 12 GHz. In [22] two band rejected antennas notching both lower and upper narrowband WLAN bands and operating in the range from 3 to 12 GHz are proposed. First antenna uses a single separated strip to reject band from 5.15-5.825 GHz. Second antenna uses two separated strips to reject bands from 5.15 GHz to 5.35 GHz and 5.725 GHz to 5.825 GHz.

In [23] a notched band antenna rejecting three bands from 3.3 GHz to 3.9 GHz, 5.2 GHz to 5.35 GHz and 5.8 GHz to 6.0 GHz is proposed. The notching features are acquired by inserting Omega shape slot on patch along with DGS on ground with a semi-octagon shaped ring behind the

designed antenna. In [24] triple band rejection is achieved at 3.5, 5.5 and 8.2 GHz of notch frequencies by inserting three notch elements on the antenna. In both [25] and [26] the WiMAX and WLAN bands are discarded by using different resonating slot structures. The papers discussed above from [20-26] are given below in Table 1.1 in the tabular form to better follow the text.

Table 1.1 Pictures and obtained S parameters of band notched antennas

Ref.	Photo	Band rejected/Size	S-parameters/VSWR
[20]		4.9-5.85 GHz/ 40×40×1	
[21]		3.40-3.48, 5.40-5.98 GHz/ 40.9×34×1	
[22]		5.15-5.825, 5.15-5.35, 5.725-5.825 GHz/ 24×36×1.524	
[23]		3.3-3.9, 5.2-5.35, 5.8-6 GHz/36×34×1	

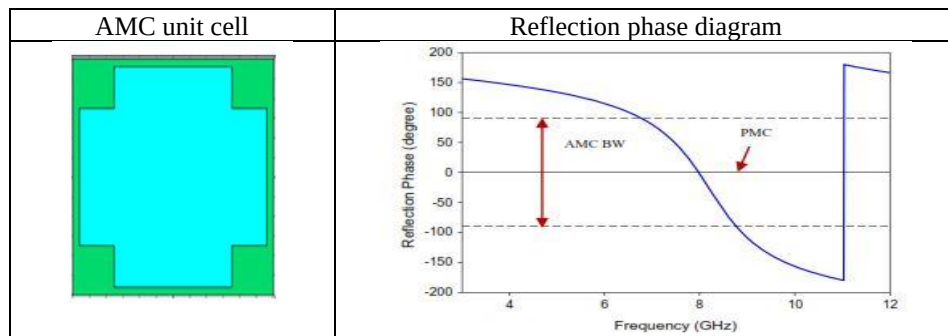
[24]		3.5, 5.5, 8.2 GHz/40×40×0.812	
[25]		3.4, 5.25, 5.78 GHz/30×25×1.5	
[26]		3.5, 5.7 GHz/28×23×0.8	

Back radiations: Metamaterials helps to design compact profile antennas and also offers benefits of better radiation features and lower price. As antennas have to be mounted on different devices, rigid surfaces and human bodies, so our biggest concern is to make a unidirectional antenna with least back radiations [27]. As one lowers the back radiations, lesser will be losses and energy absorbed by the surface. Recently, artificial magnetic conductors (AMCs), electromagnetic band gaps (EBGs), double negative (DNG) and negative refractive index (NRI) metamaterials are used by the researchers to reduce the back radiations.

Starting from the beginning and then with the evolution of theory of “metamaterial” has brought revolution and has also gained enormous recognition from the research workers over the past years [28]. In generally these can be described as the artificially designed electromagnetic materials showing unique features which are not found in nature [29]. Further on the basis of special characteristics shown by these artificial structures, they are further classified into various categories in literature [30].

AMCs: These are the band gap structures which uses the surface behaving as magnetic conductor in a particular frequency region. These are mainly acquired by using different metallization designs and repeated dielectric substrates. The AMC has an unusual boundary condition. Hence, these are capable of utilizing a novel kind of ground plane. The usage of unique AMC ground planes rather than perfect electric conductors (PECs) for various microwave and antenna applications has gained attention due to its advantages like enhanced gain and back radiation reduction. The most peculiar feature of AMC materials includes its continuously varying reflection phase curve from 180° to -180° with linearly decreasing frequency. Further about $\lambda/10$ reduction in antenna size is seen by using AMC reflectors with no impact on its performance when contrasted to PEC reflectors along dipole antennas. The AMC unit cell along with its reflection phase diagram is shown in the Table 1.2 given below [31].

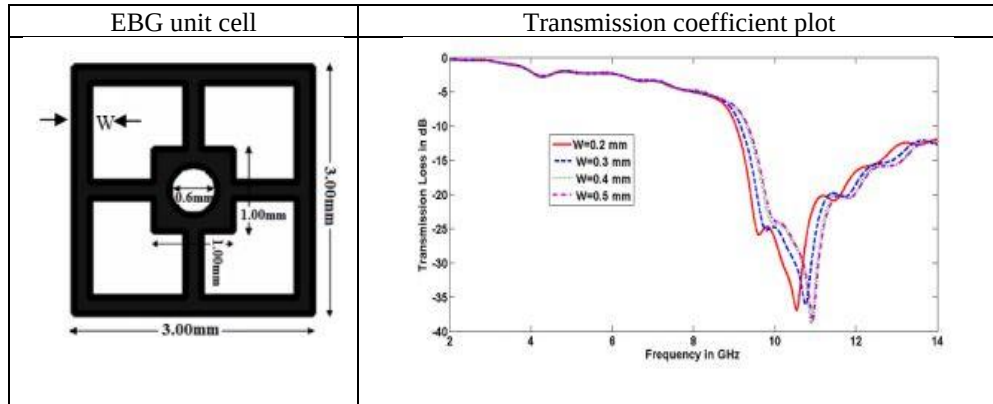
Table 1.2 AMC unit cell with its reflection phase diagram



EBGs: These consists of a distinct group of newly found electromagnetic materials showing unique electromagnetic behavior [32], therefore giving favorable solutions to various microwave complications, particularly in the field of communication [33]. These materials are mainly determined as the artificially designed repetitive or non-repeated structures which obstructs the electromagnetic (EM) wave propagation for different types of polarizations and angle of incidence in a particular frequency range. These also lowers electromagnetic coupling amongst antenna elements in an antenna array and the EM waves get diffracted before arriving to the boundaries of configuration hence resulting in the subduing of surface waves [34]. Due to this lesser back radiation are observed in the antenna which further enables smaller interference among adjoining microwave elements with minimum EM wave absorption towards the user of the device. The EBG

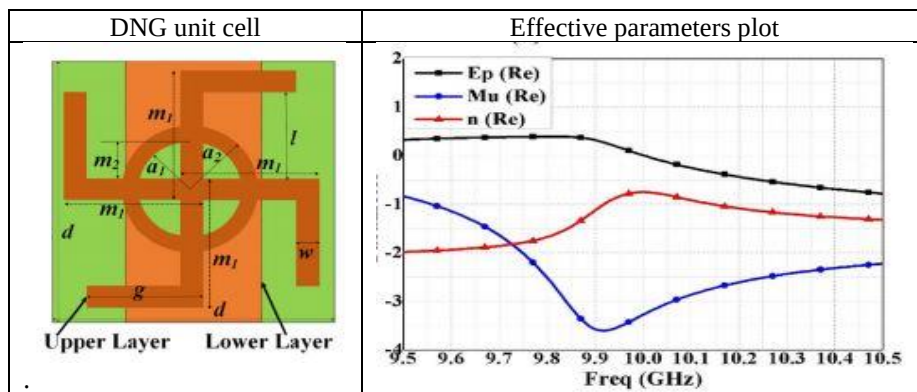
unit cell photo along with its transmission coefficient curve is shown in the Table 1.3 shown below [35].

Table 1.3 EBG unit cell with its transmission coefficient curve



DNG Metamaterials: These are the artificially structured materials which gives the frequency region of both permittivity and permeability negative therefore resulting in frequency range with negative refractive index values. These structures are also determined as Left-handed materials (LHMs). The Left-handed media contributes to the back propagation of EM wave, further it also displays antiparallel group and phase velocities with reverse Snell's law and doppler effect. DNG unit cell along with its effective parameter curve is shown in the Table 1.4 below [36].

Table 1.4 DNG unit cell with its effective parameters plot



Need for flexible antennas: Flexible antennas are designed using the substrates which can be effortlessly unified with non-uniform surfaces without ruining its working properties [37, 38]. Emergence of these unique materials having quite distinct chemical properties, with latest

fabrication techniques, further results in the evolution of dielectric substrates that provide satisfactory outcomes along with high efficiency, conformal nature, light weight and reproducible behavior. Following are the different types of flexible antennas.

Using Textile: These use textile fabrics [39] including cotton, zari and nylon etc. These are mostly preferred for applications where the antennas need to be unified with clothes and diverse wearable devices.

Using Polymer: In this the conformal and mechanically robust antenna are manufactured using the substrates made of polymers and their derivatives. These are mostly preferred substrates in the generation of conformal electronic devices. Further the liquid Crystal Polymer (LCP) and Polydimethylsiloxane (PDMS) are one of the repeatedly utilized flexible substrates amongst the different types of polymers accessible in the market.

Using Paper: The usage of paper (organic based material) as flexible substrate material has gained huge attention during the recent years. The popularity of these conformal paper-based substrates is mainly due to their productiveness, low cost and easy availability in the market. Based on their coating, texture, thickness and density, there are different varieties of paper available. Further the manufacturing technique used is also fast (printing processes) [40, 41], as compared to the conventional etching process used in the fabrication of rigid substrates. Moreover, these conformal paper-based antennas also encourage the concept of “green” electronic devices apart from being climate friendly.

1.4 SOLUTIONS FOUND IN THE PROPOSED THESIS FOR THE PROBLEMS FACED BY UWB ANTENNAS

The first problem faced is the large and complex geometries of the UWB antennas as to achieve large bandwidth and high gain, most of the researchers have used fractal and multilayered geometries. In the proposed work (chapter 3) a UWB antenna with optimum gain and bandwidth performance is designed using simple circular patch antenna as the base structure. To achieve large bandwidth the concept of “combining resonant structures” is used along with DGS structure.

The second problem faced is the interference from unwanted narrow microwave bands. To solve the problem of interference in the proposed work (chapter 4) an easy technique of attaching tuning

stubs to the antenna structure is used. The notching stubs are tuned in such a way that it creates a wide notch band rejecting both WiMAX and WLAN bands.

The third problem faced is the problem of back radiations drastically affecting the complete UWB antenna performance. Now for this problem in the proposed work, we used the array of metamaterial unit cells as a reflector plane below the UWB antenna. First of all, we designed three flexible metamaterial unit cells (chapter 5 and 6). The final metamaterial antennas gave optimum gain and bandwidth enhancement along with good forward radiation characteristics with reduced back lobes.

Fourth problem was the need of flexible/ conformal antennas as conventional flat and rigid substrates are difficult to integrate in human clothing and moreover there is a problem of frequency shift when a rigid conventional antenna is bent, rolled or placed over a curved surface. Now to solve this problem in the proposed work (chapter 5 and 6) we used flexible FR4 substrates for designing various unit cells as well as metamaterial antennas.

1.5 ORGANIZATION OF THE THESIS

The thesis is categorized in seven chapters and its chapter-wise summary is given below:

Chapter 1: INTRODUCTION

The chapter includes UWB technology along with its various advantages, characteristics and applications. Thereafter the various requirements and problems faced by the UWB antennas are introduced. Further the possible solutions of the problems faced by UWB antennas along with the organization of thesis have been introduced.

Chapter 2: LITERATURE REVIEW

The current chapter, a comprehensive review of related literature is presented. It includes various research articles categorized into four parts i.e. (a) UWB antenna design (b) UWB antenna showing interference rejection (c) UWB antenna including reduced back radiations (d) UWB antenna on flexible substrate along with the motivation and objectives of the thesis. Finally, based on the literature gaps and objectives, the methodology for the current work has been decided.

Chapter 3: DESIGN OF MULTI RESONATING UWB ANTENNA

This chapter presents the designing and fabrication of multi-resonating UWB antenna using distributed impedance matching (DIM) technique to achieve wideband matching characteristics across the whole working frequency range covering 3.1 to 10.6 GHz. A comparative analysis between proposed and existing matching methods is also presented.

Chapter 4: GUITAR SHAPED BAND NOTCHED ANTENNA

This chapter presents the designing and fabrication of a guitar shaped band notched antenna with broad band stop characteristics (rejecting both WiMAX and WLAN). The broadband design in the previous chapter is utilized as the base antenna structure in which resonating stubs are used as the notching elements.

Chapter 5: METAMATERIAL CELL FOR VARIOUS MICROWAVE BANDS

In this chapter a broadband DNG conformal dual sided metamaterial unit cell is designed for various microwave frequency bands such as S, C, X and Ku. The cell is observed to have largest DNG and NRI bandwidth as compared to some recent metamaterial unit cells.

Chapter 6: FLEXIBLE BROADBAND ANTENNAS USING METASURFACES

This chapter presents the design and testing of two conformal broadband structures using two distinct type of metamaterial layers namely mu negative (MNG) and NRI metamaterials. Further in both the structures the metamaterial is utilized in the form of array configuration behaving as a reflector layer to enhance various antenna properties.

Chapter 7: CONCLUSIONS AND FUTURE WORK

Finally, the conclusions of the proposed work and its possible future scope are documented in this last chapter of the thesis.

CHAPTER 2

LITERATURE REVIEW

2.1 INTRODUCTION

Literature review is divided in four different sections. Section 2.2.1 discusses the various types of wideband antennas designed for different applications. In section 2.2.2 various research papers including wideband antennas with different band rejection techniques to remove interference from unwanted bands are discussed. Section 2.2.3 contains wideband antennas using different types of metamaterials to enhance forward radiation characteristics, therefore resulting in reduced back lobes. Section 2.2.4 includes various conformal antennas using different types of flexible materials for various applications.

2.2 LITERATURE REVIEW

2.2.1 UWB antenna design

In this article [42], various fractal designs using hexagonal shaped structures along with different orientation of patch are designed and studied. Further all these antennas are contrasted in terms of their gain values, impedance bandwidth, matching features and amount of frequency bands. Moreover, the antenna having negative orientation gave better results, further these fractals antennas having various orientation of patch has been proposed and manufactured in the manuscript.

A tapered slot Vivaldi antenna using resonant cavities to extend the lower frequency band is designed [43]. Further to study the effect of dielectric substrate on antenna performance a full-wave time-domain method is used. Apart from dielectric constant, the permittivity was observed to play a crucial role in case of wideband antennas with similar geometries. It is observed that as we increase the permittivity, bandwidth gets improved and as we add the resonant cavity the lower end frequency response is extended even lower than 500 MHz.

A wideband miniaturized antenna with slotted ground and reduced patch with the elevated coplanar waveguide feeding is discussed [44]. Proposed structure gave broad -10 dB operating bandwidth of 126% (2.8-12.4 GHz). Further the proposed design shows steady and almost

omnidirectional radiation features across the entire working frequency range. Further the impact on the antenna performance due to variations in different parameters is also analyzed.

The $50 \times 50 \text{ mm}^2$ sized antenna using FR-4 substrate, consisting of three semi-circular structures with distinct lengths which are attached to a round disk and an altered ground plane having half circular shape. Proposed antenna gave impedance bandwidth of 210 dB (1.5-11) GHz however various discontinuities in the internal working bands are observed. Further, two pairs of split ring resonators (SRRs) are inserted close to the antenna feed to enhance the performance of designed antenna and also to acquire good impedance matching across entire working frequency range [45].

This paper presents a unique, implantable and moon-shaped conformal antenna covering both industrial, scientific and medical (ISM) band (2.45 GHz) and UWB band (3.1-10.6 GHz) for neural recording systems along with head implants [46]. Further to increase the antenna design procedure the designed structure is studied by using a multilayered box shaped model of human head. Further the simulation results of the proposed antenna gave good impedance matching, far field characteristics and directivity over the entire working frequency range.

A simple structured low profile planar inverted F antenna (PIFA) antenna using a wide feed line, a top loading element and a shorting element is designed for broadband applications. No parasitic elements are used in the designed antenna and it further provides the large impedance bandwidth value of 142% (3.1-18.5 GHz). The technique of bringing resonances to proximity is used to achieve broadband features and further by using the knowledge of the volume ratio of proposed antenna, few portions of antenna ground are eliminated which results in reduced antenna size [47].

In this article a unique and diminished half elliptical planar dipole antenna is designed in which 45% of reduction in its electrical size is acquired by simply placing a concaved arm closer to the original structure. In the entire operating range (2-9.9) GHz, the proposed antenna gave wide measured impedance bandwidth along with satisfactory omnidirectional radiation pattern [48].

In this article a compact $24 \times 32 \text{ mm}^2$ sized novel self-similar fractal UWB (3.1-12.5 GHz) multiple input and multiple output (MIMO) antenna using iterated function system (IFS) is designed, which further uses self-identical fractal structure to acquire compact size and broadband features [49]. Further the self-identical parameter of designed structure is 1.79, that is a fractional type of parameter. At second iteration level the metallic area of both the unique self-identical fractal

antennas are reduced by 29.68%. Further a T shaped strip and a vertical slot inserted in ground enhances the impedance bandwidth and isolation of designed antenna.

A $40 \times 40 \text{ mm}^2$ sized half ellipse UWB antenna with slots in the ground plane and giving fractional bandwidth of 83% from 4.40 to 10.60 GHz is designed for radar systems [50]. In this initially a perpendicular half elliptical slit is placed in the bottom plane further the axis of half ellipse is diagonally rotated in substrate direction. Also, the half ellipse axes are further uniformly improved across the patch to function as a reflector plane. The half elliptical slit is again horizontally revolved to boost the bandwidth of antenna. The designed structure gave 1-3 dBi improvement in gain along with 5 dB reduction in the half power beam width, when compared with the previously designed structures in the literature.

A broad axial ratio wideband (2.96-12 GHz), circularly polarized (CP) antenna consisting of altered ring-shaped patch, tapered microstrip feedline, rectangular shaped partial ground plane is proposed for UWB applications. By using a rectangular shaped slot and a stub in the ground layer a broad axial ratio of 5.52 GHz is attained [51].

An ultrawide band antenna consisting of a pair of consistent staircase slits, decreased ground layer, and, a cut in ground layer for broadening the operating bandwidth, is proposed for wearable and IoT applications [52]. Antennas are proposed on two dissimilar substrates i.e. denim and FR4 and a satisfactory match between measured and simulated outcomes is found. The measured impedance bandwidth for both substrates is 7.71 GHz and 7.95 GHz, further the achieved gain values are 5.12 dBi and 3.57 dBi.

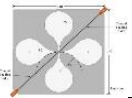
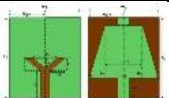
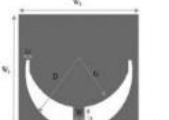
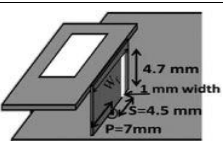
A $30 \times 25 \times 0.8 \text{ mm}^3$ sized compact UWB (3.1-10.6 GHz) antenna using semi conformal Duroid substrate along with a novel patch shape that look likes the geometry of Ashoka chakra is proposed for body wearable applications. Proposed antenna gave fractional bandwidth of 114%, gain value of 5.15 dB, peak radiation efficiency 91% and specific absorption rate (SAR) values, obtained with small value of input power (100mw), in accordance with the standards of IEEE [53].

This paper presents an elliptical shaped ($20 \text{ mm} \times 20 \text{ mm}$) UWB antenna consisting of an ellipse patch combined with a stub to cover low frequency regions, a strip loaded at rear side, and modified ground plane [54]. Designed antenna gives good wideband features covering bands from 2.68-16 GHz, 18.8–20 GHz, 0.89–0.92 and 1.54–1.64 GHz. Further it also provides maximum gain and

directivity values of 5.48 dB and 6.9 dBi. A good match between simulated and measured results is acquired and the practical measurements are carried out even in the presence of dense wood, plywood and air.

In this a fractal UWB antenna consisting of a flower shaped radiating element, CPW feed line with an uneven stub, and ellipse shaped ground structure is designed [55]. The proposed 15.7 mm × 11.4 mm antenna gave a wide -10 dB impedance bandwidth of 12.12 GHz (4.58-16.7 GHz), gain of 5 dB, fidelity factor higher than 75%. It also gave desirable omni- directional radiation characteristics and also showed satisfactory match between the measured and simulated outcomes. Table 2.1 shows the pictures and obtained results of various antennas discussed above from [42-55].

Table 2.1 Antenna pictures with their relevant results

Ref.	Photo	Size (mm ³)	Bandwidth (GHz)	Radiation pattern	Peak gain/efficiency	Transient performance analysis
[42]		140×140×0.508	(2-10 GHz)	Yes	4 dB/ (90-98%)	No
[43]		30×30×1.6	6.10 GHz	Yes	7.14 dB	No
[44]		254×150	1-6 GHz	Yes	NA	No
[45]		25×25×1.6	2.8-12.4 GHz	Yes	4.5 dBi/76%	No
[46]		50×50×1.6	1.5-11 GHz	Yes	25 dBi/90%	No
[47]		8×9×1.6	(2-11GHz)	Yes	12 dBi/NA	No
[48]		20×12×6	(3.1-18.5 GHz)	Yes	8 dB/greater than 85%	No

it. As a result, the size of final altered EBG cell is decreased by 70%. The proposed antenna gives good radiation characteristics along with average gain of 4 dB in the passband region, however good band rejection of 0 dB is obtained at notching frequencies [57].

This article presents the features of a composite right/left-handed resonator along with its usability for various UWB notched band antennas. It is noticed that the numerical simulations of asymmetrical metamaterial are able to optimize both rejected frequencies as well as zeroth order resonance (ZOR) frequency. Further periodicity assumption allows to forecast the rejected frequencies and with the help of theoretical dispersion diagram the ZOR frequency can be easily figured out [58].

A ,12 mm ×19 mm, sized triple notch band UWB (2.9-12 GHz) antenna contains an ellipse shaped patch with two inserted L shape slits along with a split ring complementary resonator slot for WBANs applications [59]. It provides triple band notch features by rejecting interferences from WiMAX, WLAN and X bands (3.3 GHz to 3.8 GHz, 5.1 GHz to 5.825 GHz and 7.25 GHz to 7.75 GHz). Moreover, proposed structure is designed using FR4 substrate and as compared to latest research the designed structure gives 42% more reduction in size.

The manuscript discusses a wideband antenna having reconfigurable notched band features. It comprises of a rectangular shaped modified patch with defected ground layer [60]. Further an open ended and inverted U-shaped slot are introduced on the top and bottom layers of antenna for rejecting both WiMAX and WLAN bands. Further to acquire reconfigurability, on/off states of two PIN diodes placed inside the slots is changed allowing the switching between the two notch bands.

A printed antenna (2.87-13.3 GHz) using CPW feeding along with rejecting five notched bands is proposed here. The five stop bands are created by using a split ring circular resonator, two symmetrical and circular shaped split ring resonators (SRRs) and a circular inverted arc creating three stop-bands. Further two symmetrical slits behaving as DGS structures is created in the ground to achieve one more band notch feature. The designed antenna shows band rejection at five narrowband wireless bands with center notch band frequencies at 3.96 GHz, 4.35 GHz, 5.7 GHz, 8.54 GHz and 9.95 GHz [61].

In this work a reconfigurable UWB triple notch band antenna using three PIN diodes is designed. The triple band rejection is achieved by insertion of a single slit in between the feedline and the radiator and two strips at the topmost position of the radiator. The proposed antenna generates easily tunable notch bands in the frequency range from 3.3 GHz to 3.6 GHz, 3.7 GHz to 4.2 GHz and 5.725 GHz to 5.825 GHz as a result eliminating disturbances from C band and upper/lower WLAN bands. Antenna shows consistent radiation features with gain value of 3 dBi, further lumped elements are used for providing correct biasing of these diodes [62].

This paper presents a CPW (2.5-20.1 GHz) fed UWB monopole on-chip antenna in which a U shape slit is inserted in the microstrip feed to create single band-notch characteristic from 7.9 GHz -8.4 GHz, therefore rejecting disturbance from microwave bands used for up linking in satellite communications (X band). Designed antenna shows satisfactory standing wave ratio values lesser than 2 in the entire operating range of frequency except rejected frequencies. Antenna obtains fractional bandwidth of 156 % and achieves bandwidth ratio of 8.04:1 [63].

In this article, a UWB antenna with 7.6 GHz of operating bandwidth (2.9-10.5 GHz) is generated with penta-notch band characteristics [64]. The five notching regions cover ISM, radar surveillance, WiMAX, C and WLAN bands with notching frequencies at 2.33 GHz, 2.81 GHz, 3.38 GHz, 3.87 GHz and 5.75 GHz. Five notches are generated by combining three altered U shape slots in the top layer and two meandered shape bandgap structures. Out of these meandered shaped bandgap structures, the first is inserted on the top layer and second is positioned adjacent to the feed. The EBG cells are mainly responsible for showing notching characteristics at both the lower and higher resonance frequencies.

In this paper a 31.2 mm $\times$ 38.4 mm sized UWB (2.9-11 GHz) fractal reconfigurable notch antenna using 50 Ω strip line feeding on FR4 substrate is designed [65]. The radiating element of designed antenna comprises of two altered C shape slots along with ten switches which contribute to the reconfigurable notch band characteristics at center notching frequencies of 5.4 GHz, 5.8 GHz, 6.1 GHz, 7 GHz, 7.9 GHz and 8.4 GHz.

This work presents a UWB (2.2-10.8GHz) fractal antenna showing dual rejection characteristics from 2500 MHz to 2690 MHz and 3300 MHz to 4200 MHz, hence removing interference from LTE 2500, WiMAX and some regions of C band. The antenna contains a nested trapezoidal ring-shaped structure as radiating patch which is excited by using a strip line fed together with a

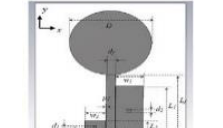
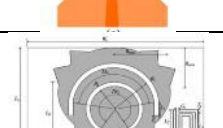
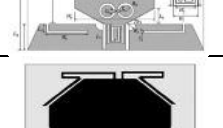
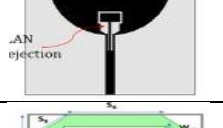
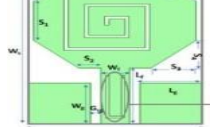
defected ground plane. Further designed structure achieves radiation efficiency value above 80 % and realized gain values in the range of 1.68 dBi to 3.1 dBi in the passband region [66].

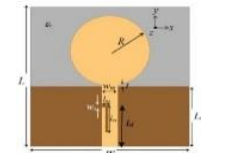
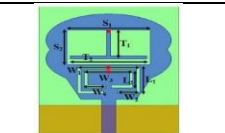
In this article a $36 \times 34 \times 1 \text{ mm}^3$ sized compact semi-circular radiator based ultra-wideband antenna with band rejection features and using co planar wave guide feed is proposed. To acquire double band rejection features, an ellipse shaped slot is created in the radiating patch to reject interference from 3.97 GHz to 4.48 GHz. Further a circular ring-shaped metamaterial structure is produced at the back side of the strip line feed to suppress frequency regions from 5.79 GHz to 6.57 GHz and 7.30 GHz to 7.60 GHz. By properly tuning the capacitance of varactor diode, the lower rejected region generated because of ellipse shaped slot is further moved to the lower notching frequencies [67].

In this article, very simple technique to create desired notch frequencies is used. In this paper $\lambda/4$ single and multiple spur lines are created on the strip line feed, which generates spur's resonance and further introduces notching frequencies in wideband antenna [68]. The electromagnetic interactions within the spur lines and strip line feed results in the generation of desired notching frequencies. Further a circular wideband microstrip antenna is used as a reference to show the frequency notching features. Moreover, the notching structure as well as the radiating element are totally independent to each other as a result the given technique may be easily implemented to any UWB antenna with strip line feed.

In this article, a $20 \times 22 \text{ mm}^2$ sized reconfigurable triple band notch UWB (2.28-19.35 GHz) antenna with fractional bandwidth of 157.75% rejecting WiMAX, WLAN and downlink satellite bands is presented [69]. To create triple band notch, a T shape inverted strip and two C shape slots are etched on the antenna radiator. Further the proposed antenna gave large standing wave ratio values of 14.59, 39.40 and 6.43 at three notching frequencies of 3.69 GHz, 5.42 GHz, 7.57 GHz. The three rejected bands are controlled and made reconfigurable by utilizing the PIN diodes. Table 2.2 discusses the various relevant results of antennas from [56-69].

Table 2.2 Antenna photos with their relevant results

Ref.	Photo	Notching technique	Size (mm ²)	Notched band/frequency (GHz)	Radiation pattern	Gain/efficiency in notched region	Transient response
[56]		Slots, SRR	41.5×32	3.3-3.7, 5.15-5.35, 5.725-5.825, 7.25-7.75	Yes	Less than -3 dB at all 4 bands	Yes
[57]		EBG cells	35×39	5.2 and 5.8	Yes	Drops to 0 dB	No
[58]		CRLH unit cell	60×50	2.45, 3.6, 8.3, 10.6	yes	NA	NO
[59]		CSRR, slots	12×19	3.3-3.8, 5.1-5.825, 7.25-7.75	Yes	Below 0 dB/below 60%	No
[60]		Slots with pin diodes	31×18	3.32-3.83, 5.15-6.94	Yes	-4 dB, -6.1dB	No
[61]		SRR, slots	29×24	3.96, 4.35, 5.7, 8.54 and 9.95	Yes	Below -4dB/Below 40%	No
[62]		Slot with pin diodes	0.22×0.28×0.005 (λ)	3.3-3.6, 3.7-4.2, 5.725-5.825	Yes	Below -3 dB/Below 35	No
[63]		slots	8.5×11.5	7.9-8.4	Yes	NA/below 60%	Yes
[64]		EBG cell, slots	34.9×31.3	2.81, 3.38, 3.87	Yes	Below 0 dB/NA	No
[65]		Slot with MEMS switches	31.2×38.4	5.4, 5.8, 6.1, 7, 7.9, 8.4 GHz	Yes	NA	No

[66]		Nested trapezoidal rings	36×36	2500-2690 MHz, 3300-4200 MHz,	yes	Below 1.5 dB/Below 20%	Yes
[67]		Elliptical and circular ring resonators	36×34	3.97-4.48, 5.79-6.57, 7.30-7.60 GHz	yes	Below -5 dB/below 50%	No
[68]		Spur lines on feed	50×50	4.62, 5.70, 7.73	yes	NA	No
[69]		Slots, Stubs	20×22	3.69, 5.42, 7.57	Yes	Below -15 dB/below 30	yes

2.2.3 UWB antenna with reduced back radiations

A CPW fed UWB (2.51-14 GHz) metamaterial antenna with fractional bandwidth, peak gain and peak radiation efficiency value of 139.19%, 6.6 dBi and 86.84% is discussed for wideband applications [70]. Proposed metamaterial antenna consists of a SRR and one cell of composite right/left hand transmission line. Broadband characteristics are acquired by merging the resonances created due to the unit cell, ring resonator and the ground layer. Further, to enhance the gain values without influencing the wideband characteristics, a moderately reactive surface is inserted to the rear side of proposed structure. Moreover, across entire impedance bandwidth a quite consistent group delay having alterations within 1 ns is achieved.

In this a CPW fed metamaterial antenna using AMC surface is presented. The basic broadband antenna using CPW feed antenna uses the altered oval shaped radiating element along with a semicircular ground plane. Further double layers of proposed AMC surface consisting of 5×5 elements of altered square ring resonators is used with initial antenna to achieve required outcomes for breast imaging applications. Final designed antenna enhances the overall gain values to 5 dB and also generates directional radiational patterns [71].

In this article a novel fractional antenna using metamaterial is used to obtain multi resonating features. Initial design contains a Sierpinski fractal triangle, L shape slot as the main radiation

elements in which further a circular ring resonator is introduced in the bottom layer. Multiband antenna operates at resonance frequencies of 3.3 GHz, 5.5 GHz, 7.3 GHz and 9.9 GHz, and giving fractional bandwidth values of 3.5%, 5.01%, 13.2% and 5.77%, covering WiMAX, satellite television and X bands. Further the resonance frequency of 3.3 GHz is achieved by introducing the circular ring resonator unit cell in the ground layer [72].

In this article three distinctive antennas like dipole, loop, and monopole antenna operating in the microwave regions from 5 GHz to 15 GHz are presented. Further 8×8 array of a mushroom type band gap structure is proposed to enhance the antenna properties such as reflection coefficient impedance bandwidth, directivity, realized gain, front back ratio and the level of side lobes [73].

A compact metamaterial inspired wideband antenna is discussed in this manuscript. Designed antenna uses an arrangement of metamaterial unit cells which help in the excitement of various working modes in the smaller frequencies [74]. It provides measured bandwidth of 25% with 5.1 GHz of center resonance frequency and 6.6 dB of peak gain. The electrical size lesser than 40% as compared to other reviewed designs makes it a good contender for various narrowband wireless communication applications as well apart from WLAN and WiMAX bands.

In this paper a simple patch antenna array using a layer of metamaterial loaded as superstrate for WiMAX wireless band at 5.8 GHz is presented [75]. Initially in unloaded case, the antenna array has low gain value of 4.31 dBi and 425 MHz of impedance bandwidth. However, when both the patches of microstrip patch array antenna are combined with the metamaterial layer as superstrate consisting of circular ring resonator and wires, the gain value rises to 11.9 dBi and the bandwidth of antenna reaches to 685 MHz, that shows 7.6 dBi enhancement in the gain values and 260 MHz increment in bandwidth values. Further approximated circuital model of proposed antenna is designed and studied.

A compact CPW fed broadband metamaterial antenna analyzed using characteristic modal analysis is proposed for body implant utilizations [76]. Broadband characteristics is achieved by exciting various modes of an unsymmetrical complementing ring resonators. Further the miniaturization and good matching features are acquired by using a pair of unsymmetrical half curve shape ring resonator in the major radiating patch. As gold and silicon shows good biostability with human body structure as a result these two materials are utilized as radiating element and substrate material.

Article presents a unique antenna showing resonance at five frequencies of 2.58 GHz, 3.17 GHz, 3.42 GHz, 4 GHz and 5.3 GHz with diminished frequency ratio values of 1.2286, 1.078, 1.169, and 1.325 within two successive resonance frequencies. Both the simulated and measured outcomes of designed structure demonstrate satisfactory match. Designed structure also demonstrates stable radiation features and therefore can be applicable for various applications including different wireless system applications along with the radar applications [77].

In this paper, a metamaterial dual band antenna using co planar waveguide feed is presented for narrowband WLAN applications [78]. It is analyzed that by using metamaterial cell, 2.5 times reduction in antenna profile is observed as contrasted to conventional triangle shaped antennas. To achieve the dual band (2.40 to 2.48 GHz, 4.7 to 6.04 GHz) features two triangular split ring resonators are used and further an open-ended stub is introduced to provide the impedance matching. The proposed antenna achieves 160 MHz and 1.97 GHz of measured bandwidth values at two of its working bands.

In this article, a circular shaped antenna along with a near-zero refractive index metamaterials (NZIMs) is designed [79]. A novel analytical methodology is used for designing the NZIMs. The proposed metamaterial is designed by using triple piled up layers of holed metal surfaces, and every layer consists of a repetitive pattern of square holes (fishnet pattern). The effective parameters extracted using inversion method are used to study the low refractive index feature of the designed unit cell. The unit cell helps in increasing the gain values of the proposed circular shaped antenna. Further the designed metamaterial shows 3 dB enhancement in the antenna directivity and also gives low reflection coefficient values below -20 dB.

In this paper, a metamaterial-based MIMO antenna with 23.5% of operating bandwidth, 9.2 dBi of gain value, 18 dB of isolation value and 73% of radiation efficiency value is designed for WLAN applications (at 5.8 GHz) [80]. The proposed antenna consists of two pairs of 2×2 array elements as the radiating patch along with a unique unit cell design in the bottom layer. The bottom layer including a 2×5 array elements of unit cells is used to enhance the antenna performance.

This article presents a novel large gain metamaterial antenna working at resonance frequency of 9.8 GHz [81]. The unit cell has star shaped structure and display small impedance features at its resonant frequency. A slot antenna behaving as magnetic dipole along with 3 dBi gain value is used as the reference antenna. However, with the introduction of unit cell the gain enhancement

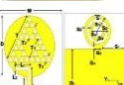
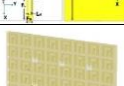
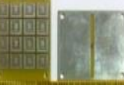
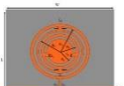
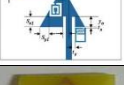
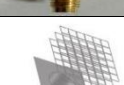
value higher than 5 dB is observed in the designed antenna. The final antenna shows compact size, 8.24 dBi of high gain and applicable for both line of sight and wireless communications.

Article presents a cavity antenna with co planar wave guide feed and large gain values [82]. Further it uses a flat edge arrow shaped radiating patch that is mostly dependable for achieving wideband characteristics. The patch antenna is then combined with square shaped patches that behave as partly reflective structures and further results in its gain improvement. The square shaped parasitic patches are manufactured at the bottom side of substrate and then placed at the high point of 1.51 above dielectric substrate by using the rods.

In this article a novel wideband metamaterial loaded antenna is studied and fabricated [83]. The dispersion features of unit cells are also analyzed graphically and numerically. Further the zero order resonances are utilized for acquiring metamaterial miniaturization. The large fractional bandwidth of 109% and 125% is attained by combining the zero order resonances with other modes and optimizing the metamaterial parameters accordingly.

Article presents a 25 mm $\times$ 22 mm $\times$ 1.6 mm sized CPW fed UWB (3.1 to 13.5 GHz) metamaterial inspired antenna for wideband applications. It provides 10.4 GHz of impedance bandwidth and 4 dBi of gain. Main radiating element of proposed structure contains three ring resonators kept at one axis. Further to boost the matching characteristics of proposed antenna, an altered ground layer together with co planar wave guide feeding is utilized. Moreover, the proposed antenna also provides omni-directional radiation patterns [84]. Table 2.3 shows different antenna parameters along with their pictures from [70-84].

Table 2.3 Antenna photos together with their relevant results

Ref.	Photo	Material /type	Size (mm ³)	Bandwidth (GHz)	Radiation pattern	Peak Gain /efficiency	Transient behavior
[70]		SRR, CRLH-TL/loaded	28.8×32×1.6	2.51-14 GHz	Yes	6.6 dBi, 86.84%	yes
[71]		AMC/layer	76×44×1.6	3.1-7.6 GHz	yes	10 dBi/90%	Yes
[72]		SRR/loaded	30×24.8×1.6	3.24-3.44, 5.31-5.62, 6.99-7.52, 9.78-10.08	yes	1.75 dB/NA	No
[73]		EBG/layer	80×80×1.6	10.756 GHz	Yes	9.9dBi/NA	No
[74]		CRLH metamaterial/loaded	0.75×0.60×0.07 (λ^3)	4.45-5.71 GHz	yes	6.6 dB/NA	No
[75]		DNG metamaterial/loaded	52.5×67.10×1.48	685 MHz (5.4-6.1 GHz)	yes	11.9 dBi/NA	No
[76]		Asymmetric CSSR/loaded	16×16×0.275	1.35-3.5 GHz	yes	-16.5 dBi/NA	No
[77]		Rectangular SRR and IDCLLR/loaded	35×35×1.5748	2.58, 3.17, 3.42, 4.0 and 5.3 GHz	yes	4.01 dBi/89.35	No
[78]		Triangular SRR/loaded	20×24×1.6	2.40-2.48, 4.7-6.04,	yes	2.277/93.25%	No
[79]		NZIM metamaterial/loaded	NA	NA	NA	NA	NA
[80]		Metamaterial cell/loaded	137×77×3.048	5.8 GHz	Yes	9.3 dBi/73%	NA
[81]		Metamaterial cell/layer	89×50×1.6	9.8 GHz	Yes	8.24 dBi/NA	NA
[82]		Square parasitic patches/layer	169×169×1.524	3.8-8.8 GHz	Yes	8 dB/NA	NA

[83]		CRLH unit cells/loaded	25×15×1.6	4-17.2 GHz	yes	4.8 dBi/89%	NA
[84]		SRR/loaded	25×22×1.6	3.1-13.5 GHz	Yes	4dBi/80%	NA

2.2.4 UWB antenna on flexible substrate

In this article, a CPW fed flexible composite UWB (3.3-12 GHz) antenna consists of a conformal substrate made of epoxy resin and E glass fiber in which a very slim and conducting sheet of fabric is inserted to produce a flexible antenna is proposed [85]. The flexible antenna components are manufactured accurately by using laser printing technique. Proposed composite antenna gives measured radiation efficiency above 70% along with satisfactory radiation pattern, gain and bending performance.

A compact conformal textile antenna using C and H shape resonators along with co planar wave guide feeding and EBG structure is designed [86]. The H shaped resonator stimulates various frequency bands, the pair of C shaped resonators improves impedance bandwidth and the band gap unit cell is used to acquire lower SAR values. Proposed final design gave measured impedance bandwidths of 1.5 GHz to 3.0 GHz for first band and 4.5 GHz to 6.5 GHz for the second band.

A four layered embroidered textile antenna using zari and cotton cloth is proposed in this article for ISM band (frequency of 2.45 GHz) [87]. To enhance the antenna efficiency and reflection coefficient characteristics an end to end slot structure is presented. The antenna having four layers is fabricated utilizing the traditional embroidery technique and then all the layers are sewn jointly to generate the final antenna design. Proposed structure has been studied under different bent, crumpled, and wet conditions. Moreover, the structure is studied for body wearable applications and in combination with Arduino Uno board.

Article presents a dipole antenna with slightly shifted feed line and uses an array size 2, 4 and 8 elements for millimeter wave utilizations [88]. Array antenna using 8 elements having 43.6×25.1

$\times 0.25 \text{ mm}^3$ sized dimension is manufactured using liquid polymer substance and achieves 15.7 GHz (24.7 GHz-40.4 GHz) of wide operating bandwidth. The mutual interactions amongst the array size of 8 elements is lower than -35 dB and also provides a gain of 12.62 to 13.1 dB. Further the antenna shows better performance regarding its matching characteristics, radiation characteristics, group delays along with polarization patterns.

In this paper a vinyl polymer-based flexible antenna is proposed for on-body WBAN utilizations. The antenna uses triangular shaped patch, fractal design (Koch) along with meandered slots and defected ground structures to acquire miniaturization of size and bandwidth enhancement [89]. The proposed antenna provides the impedance bandwidth of 7.75% (2.36-2.55 GHz), maximum gain value of 2.06 dBi and 75% of antenna radiation efficiency value. Further antenna has been studied for structural deformation for various bending angles which results in only 0.6% detuning in antenna frequency with consistent bandwidth features, all these features confirms the high performance of proposed structure for conformal utilizations.

A $60 \times 60 \times 2.4 \text{ mm}^3$ sized single resonance metamaterial loaded fabric antenna is produced for wearable utilizations with metamaterial cell acting either as band gap structure or as a magnetic conductor [90]. The main advantages of using metamaterials in flexible wearable antennas is to reduce the effects of antenna detuning and also to lower the amount of the backward radiations towards the human body. Furthermore, the metamaterial antenna shows the ability for managing the lower absorption rate values and 14.5%, improvement in impedance bandwidth, 13.7 dB front-to-back ratio and 7.5 dBi gain values. Designed antenna is also studied for different bending conditions on actual and simulated human body.

The article presents a conformal wearable and tunable slot antenna using artificial magnetic structures [91]. The tunability is produced by continuously changing the radiation characteristics of antenna with the help of a PIN diode utilized in between the stub and the folded slot. Further it also utilizes a double band polarization sensitive AMC structure to enhance the radiation pattern and decrease the SAR values. Both the stub and the slot show distinct polarization states and working frequency range and further the antenna properties are studied for both smooth and bent surfaces.

Article presents a broadband large gain 4×4 patch antenna using circular polarization along with a conformal polyethylene terephthalate material as substrate [92]. Comparative study between the

smooth and bent surfaces is also shown to demonstrate the array bending effect on the functioning of the designed antenna. Further to enhance the impedance matching and axial ratio bandwidth, the two levels of consecutive rotations are utilized in the antenna designing process. Good consistency is observed in both the measured and simulated outcomes.

Article presents a broadband double ring implantable slot antenna using a slim and biostable metamaterial layer as superstrate with CPW feeding to provide flexibleness and human body isolation for biomedical applications [93]. The proposed antenna gave very low realized gain value of -12 dB. A metamaterial layer with very high permittivity values is used as a superstrate layer resulting in 3 dB gain enhancement in the proposed antenna.

In this article a dual band design using split triangle shaped radiating patch along with ellipse shaped ring structure and CPW feed is proposed using polyimide material [94]. The designed antenna covers dual bands with two resonance frequencies at 2.60 GHz and 3.48 GHz. Further it shows reflection coefficient values of -21.92 dB and -32.14 dB, 1.75 dBi and 2.39 dBi of gain values along with bandwidth values of 100 MHz and 330 MHz at both of its resonance frequencies.

In this paper a nanocomposite material-based double band antenna is designed for different wireless systems, body-centric and Wi Fi applications [95]. The nanocomposite material with conductivity of 7500 S/m is produced by combining a conducting polymer substance with nanostructures. The designed antenna parameters including reflection coefficient, SAR, radiation characteristics and gain values are measured over the human body as well as in vacuum. It shows 1.22 dB of measured gain on body.

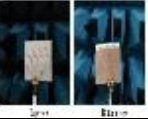
This article presents a conformal antenna using polydimethylsiloxane as substrate material and has a leaf like structure for WLAN ISM band. Further to decrease the SAR values, the layer of meta material cells is introduced. Antenna utilizes a copper layer with 3 mm of height as the conductive layer. The metamaterial layer in the bottom side of proposed antenna is used to cover only the area with maximum back radiations apart from complete antenna and also helps in reducing the SAR value. The proposed metamaterial antenna gave low absorption rate value of 1.53 W/kg over 1 g of body tissue at 2.4 GHz of resonance frequency [96].

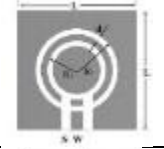
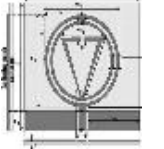
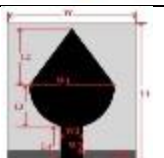
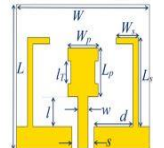
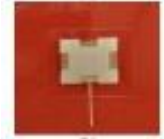
This paper presents a liquid crystal polymer (LCP) based flexible millimeter wave (MMW) antenna, two element antenna arrays for wireless systems working at K microwave band [97]. The

proposed antenna consists of rectangle shaped microstrip antenna which is tapered using two perpendicular slots and also uses CPW feeding. In the bottom layer L shaped stubs are used for converging the scattered radiation characteristics hence enhancing the gain and directivity values. A unique and fast technique post-printing sintering is used in this. The designed array antenna displays 14 GHz of impedance bandwidth, maximum gain value of 11.35 dBi along with stable high gain value above 9 dBi in the entire K microwave band.

A $22 \times 31 \text{ mm}^2$ sized conformal multilayered fractal antenna using inkjet printing technique operating in the frequency range from 4.79 GHz-5.04 GHz is designed [98]. Designed antenna uses 0.125 mm thick Kapton polyimide as the substrate material. It contains two layers of radiating patches; first layer is the 0.12 mm thick SU-8 polymer which is directly inkjet-printed on the substrate. Further to acquire compact size and desired impedance matching the second layer including Minkowski fractal geometry patch is designed and is directly inkjet-printed above the first layer of SU-8 polymer. Table 2.4 shows the antenna photos along with their relevant results from [85-98].

Table 2.4 Antenna photos along with various important results

Ref.	Photo	Material	Size (mm ²)	Bandwidth	Gain/efficiency	Transient analysis	Radiation pattern
[85]		Conductive fabric/E-glass	40×40	3.3 to 12 GHz	3.1 dB/94%	No	yes
[86]		Textile	70×70	1.5-3.0 GHz, 4.5-6.5 GHz	7.7 dBi/NA	No	Yes
[87]		Textile	58×63	2.45 GHz	8dBi /67.03%	No	Yes
[88]		Liquid Crystal polymer	43.6×25.1	24.7-40.4 GHz	13.1 dB/96.5%	No	Yes
[89]		Vinyl polymer	39×39	2.36-2.55	2.06 dBi/75%	No	Yes
[90]		Fabric material	60×60	2.4GHz	7.5 dBi/NA	No	Yes

[91]		Rogers RO3003	83×89	2.45	NA	No	Yes
[92]		PET film	14.85×14.85	7.7-8.3 GHz	16.90/87%	No	Yes
[93]		Kapton polyimide	57% (1.5-3.5 GHz)	2.45	-9dB/NA	No	Yes
[94]		Polyimide material	33×30	2.60, 3.48	2.39 dBi/88.37	No	Yes
[95]		Polymer based	NA	1.8-2.45, 5.15-5.825	3.12 dB/NA	No	Yes
[96]		Polydimethylsiloxane substrate	50×45	2.4 GHz	2.88 dB/68.06%	No	Yes
[97]		Liquid crystal polymer	11×12	26.5-40 GHz	11.35 dBi/	No	Yes
[98]		Kapton polyimide	22×31	4.79-5.04 GHz	4.5 dBi/NA	No	Yes

2.3 MOTIVATION

After the comprehensive study of the literature, it has been concluded that extensive research is already carried out in the area of UWB antennas but still there is a scope for an improvement as most of the papers reported a tradeoff between the one or the other parameters of the UWB antenna. To design a UWB antenna which is flexible having optimum size along with good bandwidth, gain and radiation characteristics is still not reported by many researchers. Also, the various metamaterial unit cells reported in the literature are taken from earlier published articles and in most of the papers the detailed analyses of unit cells are unreported. Also, the flexible antennas

designed mostly used expensive and rare materials with complicated fabrication techniques. All these drawbacks motivated us to design a UWB antenna with optimum parameters without compromising any of its important properties. Subsequently, an interest arises to implement new, novel and flexible metamaterial unit cells which can be used for various applications and also provide large DNG and NRI bandwidths.

Based on the relevant literature study and facts, an attempt is made to outline the gaps from the study and to suggest the statement of the problem for the proposed work.

2.4 GAPS IN THE STUDY

After doing the detailed literature survey on UWB antennas as in the above section 2.2, some gaps are found and presented here as the motivating factor for doing further research.

1. Researchers have used stacking and fractal geometry to have large bandwidth but it further has increased the complexity [99] and size of antenna, therefore size reduction and simple structure should be introduced to develop a compact and mechanically robust antenna.
2. As we are using UWB technology which co-exists with many other narrow bands, therefore designing of a notch filter is required to reduce the interference occurring from unwanted bands.
3. Most of the researchers have focused on bandwidth and gain enhancement only and back radiations are not taken into consideration. As back radiations can drastically affect the overall antenna performance. So, back radiations have to be reduced by placing a reflector plane (AMCs and EBGs) between the antenna and surface.
4. As conventional flat and rigid substrates like FR-4 and Duroid are not easy to integrate in human clothing, there is a need of highly conformal and flexible antennas, therefore to solve this problem textile antennas can be designed which easily adapts to the surface on which it is mounted.
5. There is a frequency shift when antenna is bent, rolled or placed over a high dielectric constant material. So, this issue has to be solved by designing a broadband antenna with uniform bandwidth and return loss in the entire frequency band.

2.5 OBJECTIVES

Based on the research gaps and problem statements identified above, the main objectives of the thesis are as follow:

1. Design and simulation of a compact planar UWB antenna for future wireless devices.
 - Parametric studies of the proposed antenna
 - Fabrication and Testing of proposed antenna
2. Design of a notch in proposed UWB antenna to reject unwanted band.
 - Parametric studies of the proposed antenna
 - Fabrication and Testing of proposed antenna
3. Design of UWB antenna with suppressed back radiations using metamaterials.
 - Fabrication and Testing of proposed antenna
4. Design of UWB antenna on flexible substrate for future wireless devices.
 - Fabrication and Testing of proposed antenna

CHAPTER 3

DESIGN OF MULTI RESONATING UWB ANTENNA

The designing of ultrawideband antenna is considered among the most encouraging area of research for upcoming wireless devices. This chapter discusses the designing, fabrication and testing of a multiband broadband antenna. It is found through literature survey that a UWB antenna needs to have a consistent and steady performance across the whole working frequency region covering 3.1 to 10.6 GHz and hence it is quite crucial for these antennas to have good impedance matching features across its broad bandwidth. The method of “Distributed Impedance Matching” is utilized here, where we alter the structure geometry as such rather than utilizing passive networks to achieve matching characteristics.

3.1 GUITAR SHAPED WIDEBAND ANTENNA

The final configuration of designed broadband structure is given in the Fig. 3.1 and its improved measurements are specified in the Table 3.1. Originally circle shape microstrip antenna using incomplete ground plane is used as initial design structure, further (3.1) [100] is utilized for calculating the circle shaped patch radius.

$$R = \frac{87.94}{f_r \sqrt{\epsilon_r}} \quad (3.1)$$

Here R denotes patch radius, f_r denotes resonant frequency and ϵ_r represents the substrate's dielectric constant. Consequently, various resonance structures are established in the initial design to enhance its matching characteristics and reflection coefficient at various resonance frequencies. Foremost, a slot including a plus and ring shape structures is inserted in the circle shape antenna patch. It enhances the matching characteristics at five distinct resonance frequencies including S, C, X microwave bands. Next the defected ground structure (DGS) is inserted in half ground plane that promotes more enhancement in matching characteristics at its resonance frequencies. In the end, U shape slot along with a staircase shaped structure is introduced in the top layer of antenna, further two slits placed at the corners of partial ground plane along with parasitic plane at topmost right corner are placed in the bottom layer of the initial antenna for enhanced matching features at various resonance frequencies. The ultimate altered circle shape patch antenna displayed in Fig. 3.1(a) encloses the entire unauthorized broad frequency region covering 3.1 to 10.6 GHz of

microwave band. Structural progression of designed structure to boost the matching features along with reflection coefficient values is examined in the following section.

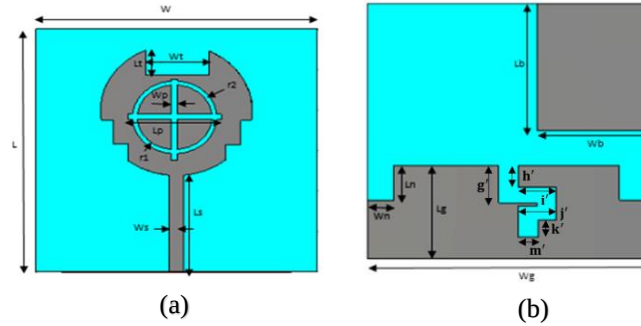


Fig. 3.1 Geometry of proposed structure (a) Front layer (b) Back layer

Table 3.1 Improved dimensions of designed structure (mm)

Parameter	L	W	Lt	Wt	r1	r2	Wp
Dimension	30	30	4	6.8	4	4.5	0.65
Parameter	Lp	Ws	Ls	Lb	Wb	Ln	Wn
Dimension	10	1.58	11.6	15	12	4	3
Parameters	g'	h'	i'	j'	k'	m'	
Dimensions	4.5	2.5	4.5	4	2	2.15	

3.1.1 Structural progression of the ultimate design

The step by step structural progression of the ultimate structure of designed antenna is presented in Fig. 3.2. The correspondent reflection coefficient value for each structural alteration is displayed in Fig. 3.3. In the beginning in Fig. 3.2 (a) a circle shaped patch along with incomplete ground plane is utilized for constructing a broadband antenna structure. The impedance characteristic of design S1 is shown in first column of the Table 3.2 at various resonance frequencies. It's quite evident that at different resonance frequencies the input impedance feature of the designed antenna is not close to 50Ω . Further by inserting a DGS along with a plus and ring shape slot, as displayed in S2 design, the S1 input impedance features are enhanced as presented in second column of the Table 3.2. Later a staircase shape is carved from S2 design as displayed in S3 design that further enhances antennas impedance characteristics (column 3 of Table 3.2). Finally, as a U shape slot, parasitic ground along with corner slits are inserted, as displayed in S4, input impedance of

proposed antenna eventually come closer to the 50Ω value at five resonance frequencies (column 4 of Table 3.2). Enhancement in reflection coefficient values along with input impedance features of proposed antenna with every structural alteration at various resonance frequencies is displayed in the Fig. 3.4 and Table 3.3.

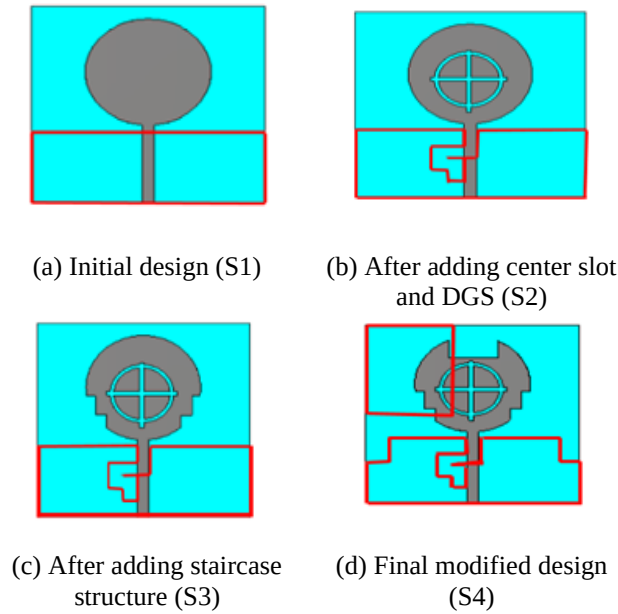


Fig. 3.2 Structural progression of proposed structure

Table 3.2 Input impedance (Reactance and Resistance) values (Ω) after every structural alteration at various resonance frequencies together with resonance type indicated by S/P (S: series resonance P: parallel resonance)

Frequency (GHz)	Input impedance (Ω) after each geometrical variation of antenna			
	S1, S/P	S2, S/P	S3, S/P	S4, S/P
3.4	(89, 4.12), S	(75, -11.56), S	(72, -15.7), S	(52, -0.70), S
5	(75, -6.00), S	(60, -16.53), S	(60, -9.97), S	(52, -0.07), S
6.5	(79, -1.64), P	(68, 2.28), P	(78, -4.85), P	(52, -0.39), S
7.7	(75, -12.04), P	(64, 11.3), P	(56, 5.96), S	(54, 0.98), S
9.3	(40, 9.1), S	(53, 15.89), S	(56, 22.62), S	(58, 2.6), S

Table 3.3 Reflection coefficient (dB) after every structural alteration at five resonance frequencies

	Return loss (dB) after each geometrical variation of antenna			
Frequency (GHz)	S1	S2	S3	S4
3.4	-10	-13	-13	-28
5	-13	-15	-17	-30
6.5	-12	-15	-13	-30
7.7	-12	-15	-22	-26
9.3	-16	-16	-13	-21

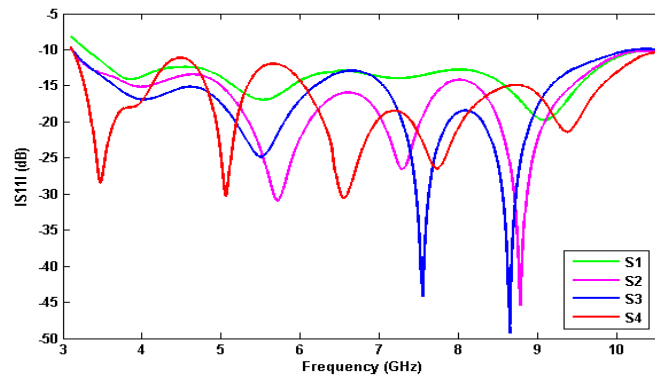


Fig. 3.3 Reflection coefficient curve displaying the results of several structural alterations done to the proposed antenna

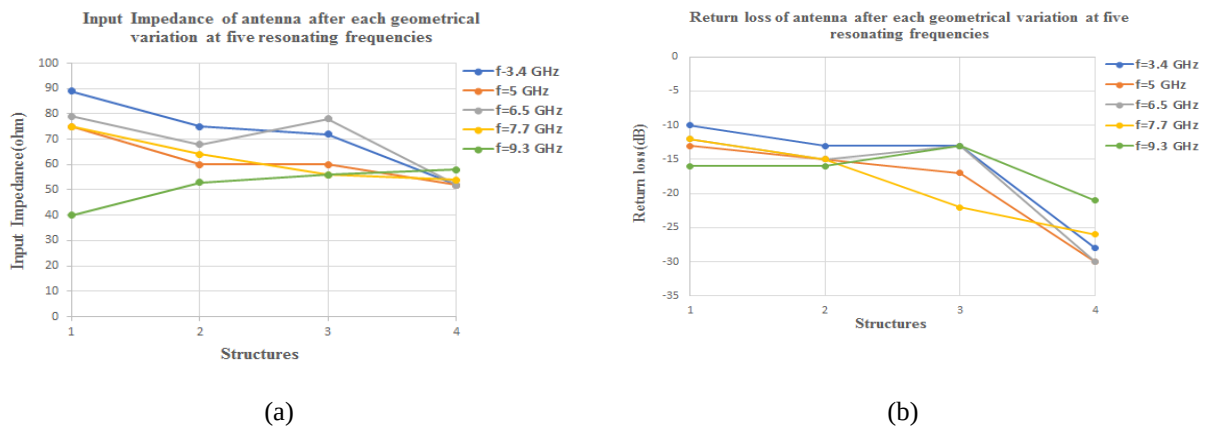


Fig. 3.4 (a) Input impedance (Ω) plot at various resonance frequencies (b) Reflection coefficient (dB) plot at various resonance frequencies

3.1.2 Analytic investigation of designed antenna

In this segment the logical analysis of the designed antenna on the bases of input impedance and smith chart curves are considered for deeper awareness of the introduced methodology, that further verifies the simulated outcomes acquired in the section 3.1.1.

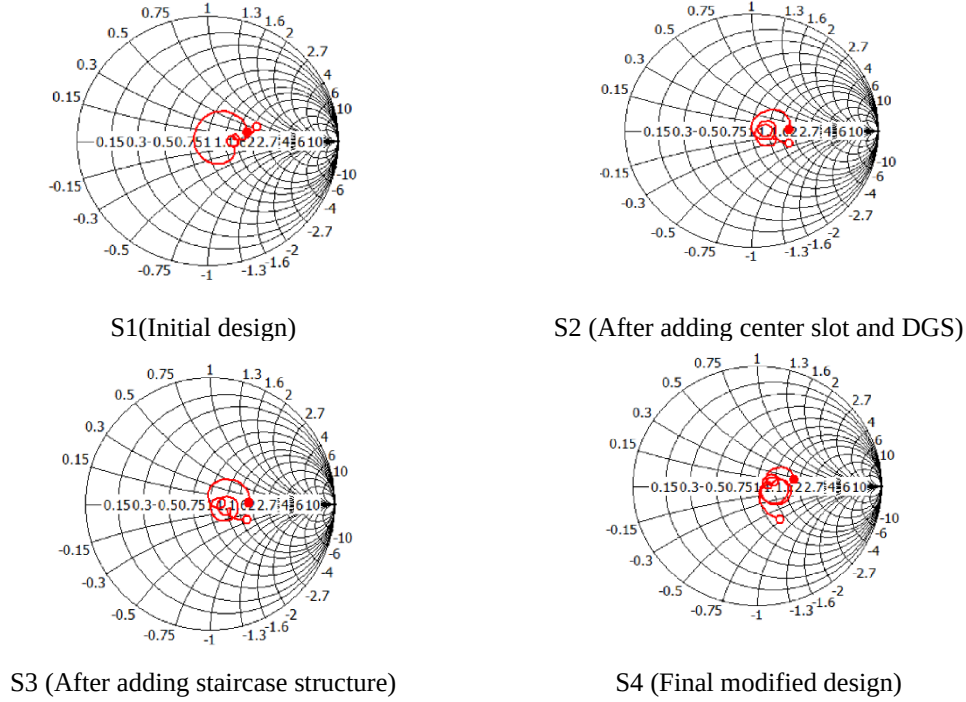


Fig. 3.5 Normalized Smith chart plots with matching characteristics at four distinct designs

Figure 3.5 displays the 50Ω normalized and clockwise input impedance plots of given structure at four distinct design stages (S1, S2, S3 and S4). By observing the smith chart plot of S1 design, it's analysed that input impedance plot displays inductive type of reactance in the frequency bands covering 3.1-4.68 GHz, 5.53-6.47 GHz and 9.13-10.6 GHz along with capacitive type of reactance in the region covering 4.69-5.52 GHz and 6.48-9.12 GHz, further it gives 3.99 GHz of entire inductive along with 3.47 GHz of entire capacitive frequency regions. Broad input impedance curve too far from middle 50Ω resistive line evidently shows impedance mismatch features in original structure (S1).

From smith chart at S2, a total of 3.95 GHz of capacitive reactance covering the bands from 3.1-5.73 GHz, 6.63-7.25 GHz and 8.07-8.77 GHz along with 3.5 GHz of entire inductive reactance in the frequency regions from 5.74-6.62 GHz, 7.26-8.06 GHz and 8.78-10.6 GHz is observed. It is

quite evident that the input impedance plot has slightly moved towards the right direction of the smith chart, therefore giving enhanced resistive values with higher quantity of resonance frequencies moving towards the middle of normalized smith chart. Hence it is clear that the two resonance structures (DGS and circle shape slot with ring) brings the capacitive effect to the basic antenna design. As these structures have shifted the input impedance plot along the bottom right side resulting in the enhancement in capacitive and resistive values of the antenna input impedance.

At S3 design, with the addition of a staircase shaped structure, the capacitive type of reactance values are observed covering 3.1-5.51 GHz, 6.50-7.4 GHz and 8.2-8.5 GHz (3.61GHz) and inductive values are found from 5.52-6.49 GHz, 7.5-8.1 GHz, and 8.6-10.6 GHz (3.57 GHz). Input impedance plot achieved at this stage is quite identical to the one obtained at S2 along with minute shift in upwards indicating a rise in the inductive effect of the antenna input impedance.

At S4 design, frequency bands covering 3.1-3.51 GHz, 4.38-5.06 GHz, 5.55-6.60 GHz and 7.25-7.65 GHz (2.54 GHz) shows the capacitive type of reactance values and the frequency bands from 3.52-4.37 GHz, 5.07-5.54 GHz, 6.61-7.24 GHz and 7.67-10.6 GHz (4.88 GHz) displays inductive type of reactance values. Hence the rise in inductive reactance values can be observed in the concluding antenna design. Further at S4 design the input impedance plot meets the middle line of smith chart many times referring to the reactive component moving towards zero and the resistive part of 50Ω is only observed that could be seen clearly in the column 4 of the Table 3.2 as well.

All the smith chart outcomes acquired show good harmony with reactance values observed at different distinct resonance frequencies as shown in the Table 3.2. Further multiple resonance along with multiband characteristic of proposed antenna as found in the Fig. 3.3 is also justified with the help of smith charts, as the total amount of impedance loops have enhanced by introducing various resonance structures in the base antenna design.

It is also noticed that the above resonance structures act as LC resonance circuits hence producing capacitive and inductance effects, where dielectric gaps produce capacitive and metallic strip produce inductive effect as noticed in the Fig. 3.5 as well. The dependency of designed antenna bandwidth on L , R and C values produced by various resonance components can be observed from (3.3) and (3.4) given below.

$$\omega = 2\pi f = \frac{1}{\sqrt{LC}} \quad , \quad (3.2)$$

$$BW = 1/RC \quad , \quad (3.3)$$

$$Q = \frac{\omega}{BW} = R\sqrt{C/L} \quad , \quad (3.4)$$

Here ω , BW and Q represents the angular frequency, bandwidth, and quality factor of antenna. L , R and C indicates the equivalent inductance, resistance and capacitance values generated by resonance circuit. It's quite evident from above analysis that the antenna impedance bandwidth relies on equivalent capacitance, inductance and resistance values generated by the various resonance components.

3.1.3 Parametric investigation of antenna

In this the different measurements of designed antenna are altered separately to realize their correspondent consequence on the antenna reflection coefficient plot. By utilizing this parametric analysis, we can select the suitable measurements of proposed antenna in agreement with the required outcomes.

Impact of alterations in the partial ground step width 'Wn' on the antenna reflection coefficient

The value of W_n is altered from 1 to 5 mm. 3 mm is observed to be the optimal W_n value. Hence as we reduce the W_n value beneath 3 mm of optimal value, the antenna reflection coefficient rises at 4.3 GHz of resonance frequency from -28 to -17 dB and at 6.5 GHz of frequency from -30 to -19 dB and little enhancement in antenna reflection coefficient is found at left resonance frequencies as shown in Fig. 3.6. As we go above 3 mm of W_n value, the reflection coefficient value at 9.3 GHz of frequency rises strongly from -21 to -10 dB and at resonance frequency of 3.4 GHz, it decreases from -27 to -46 dB, moreover the resonance at 6.5 GHz combines with 7.7 GHz of frequency, as a result at $W_n = 5$ mm at most three resonant frequencies are left.

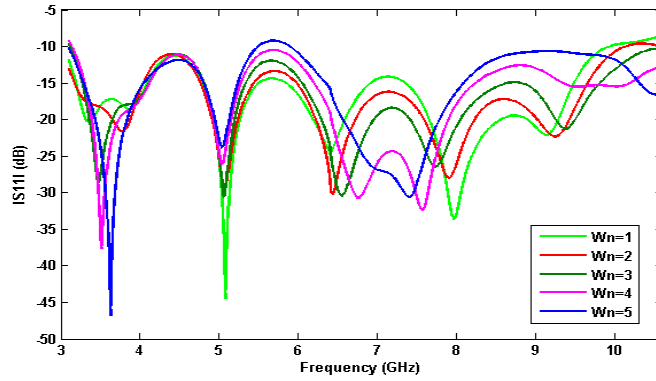


Fig. 3.6 Impact of alterations in 'Wn' on the antenna reflection coefficient

Impact of alterations in parasitic ground layer length 'Lb' on the antenna reflection coefficient

The value of Lb is altered from 13 to 17 mm. 15 mm is observed to be the optimal Lb value. As we enhance the Lb value more than its optimal value of 15 mm, the resonances observed at 7.7 and 9.3 GHz frequencies stay almost unchanged, however the reflection coefficient value decreases at 6.5 GHz of resonance frequency from -29 to -44 dB and at both the frequencies of 3.4 and 5 GHz, the reflection coefficient changes from -28 to -16 dB and from -30 to -11 dB as displayed in the Fig. 3.7. As we reduce Lb beneath 15 mm, reflection coefficient value at resonance frequencies of 6.5, 7.7 and 9.3 GHz rises and at Wb=13 mm strong rise is found at 6.5 GHz of frequency from -30 to -15 dB, further with decrement in Lb value, the enhancement in reflection coefficient values at larger frequencies is observed.

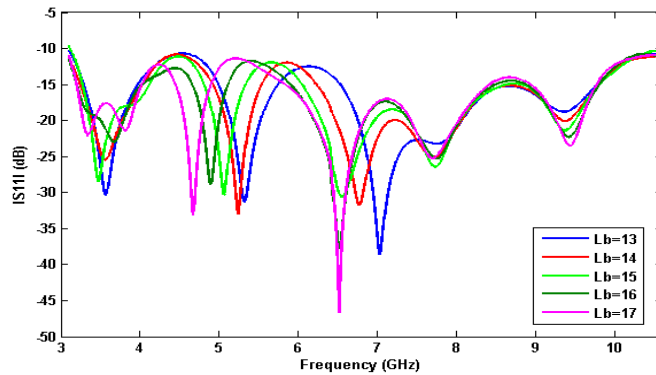


Fig. 3.7 Impact of alterations in 'Lb' on the antenna reflection coefficient

3.1.4 Outcomes and discussions

Designed broadband structure is manufactured and experimentally tested utilizing KEYSIGHT E5063A (100 KHz-18 GHz) vector network analyzer (VNA) and electromagnetic chamber.

Picture of the manufactured structure is displayed in the Fig. 3.8.

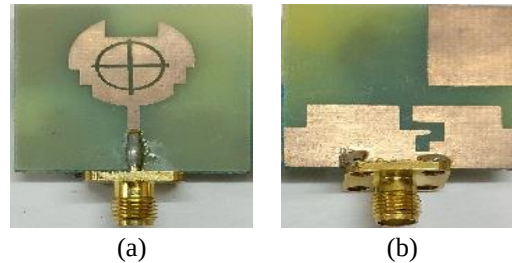


Fig. 3.8 Picture of manufactured structure (a) Upper layer (b) Back layer

Figure 3.9 displays measured and simulated reflection coefficient plots of designed antenna. From simulated plot it's evident that broadband antenna works in unauthorized frequency region covering 3.1 to 10.6 GHz of microwave region together with steady reflection coefficient values underneath -20 dB at five distinct resonance frequencies. Moreover, in measured plot a persistent 0.2 GHz of right shifting is found at first two resonance frequencies along with enhanced reflection coefficient values at other three resonance frequencies. The working bandwidth values obtained in both the measured and simulated plots look identical with small discrepancies. The disparity in experimental outcomes is observed due to faults generated when manufacturing the designed antenna using wet itching method and also because of misalignment problems produced while connecting the designed antenna to the vector network analyzer.

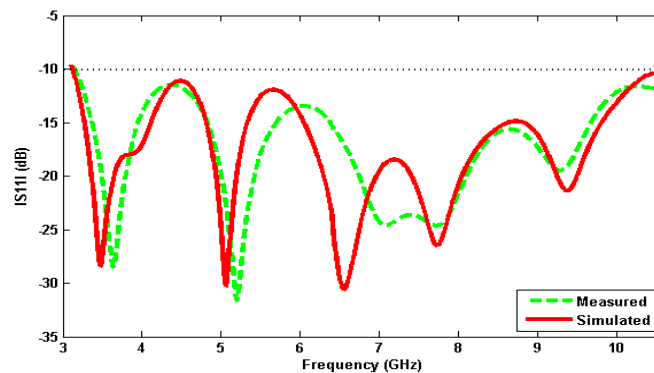


Fig. 3.9 Measured and simulated reflection coefficient plots of designed structure

Figure 3.10 displays measured and simulated standing wave ratio curves of designed structure. It's quite evident from two curves that proposed wide band structure shows wave ratio values below two in complete wideband working frequency region from 3.1 GHz to 10.6 GHz. Still minute alterations can be seen in the measured outcomes because of faults produced during the manufacturing and testing process.

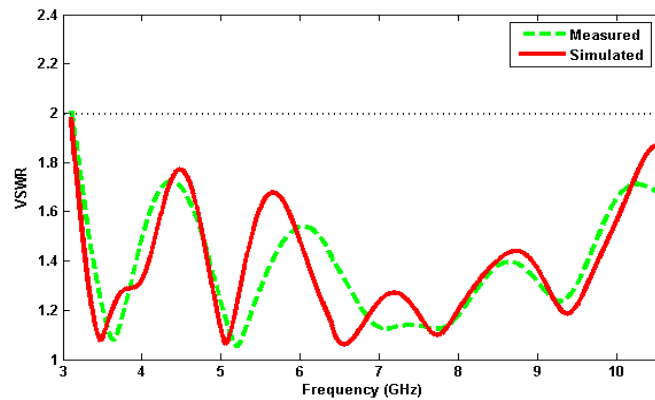


Fig. 3.10 Measured and the simulated VSWR curves of designed structure

Figure 3.11 displays measured and the simulated gain plots of designed structure. It is quite evident from simulation curve that proposed wide band antenna maintains steady gain values across the complete UWB band with 5.8 dB of highest gain value at 9.4 GHz of frequency and 1.9 dB of lowest gain value at 4.2 GHz of resonance frequency. Further from the measured plot, it can be seen that the enhancement in gain values is found in the frequency bands covering 3.1 to 3.6 GHz as well as 6 to 10.6 GHz, with maximum gain at 9.4 GHz frequency where 0.3 dB of increment in gain value is achieved from 5.8 dB to 6.1 dB. Moreover, in remaining frequency region both measured and simulated gain plots show little discrepancies because of testing and manufacturing errors.

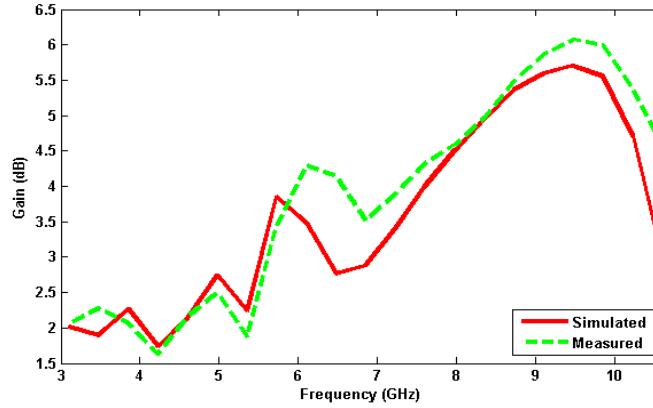


Fig. 3.11 Measured and simulated gain plots of designed structure

Figure 3.12 displays the simulated and measured radiation efficiency plots of designed structure. It is evident through simulation curve that the proposed wide band antenna generates steady radiation characteristics across the whole UWB band with 94% of highest radiation efficiency value at 3.4 GHz and 68% of lowest radiation efficiency value at 10.6 GHz frequency. Moreover, little alterations are found across the whole UWB band in case of measured plot showing smallest radiation efficiency value at 10.6 GHz, where 4% enhancement in efficiency value is found from 68% to 72% along with similar highest radiation efficiency value at 3.4 GHz of resonance frequency.

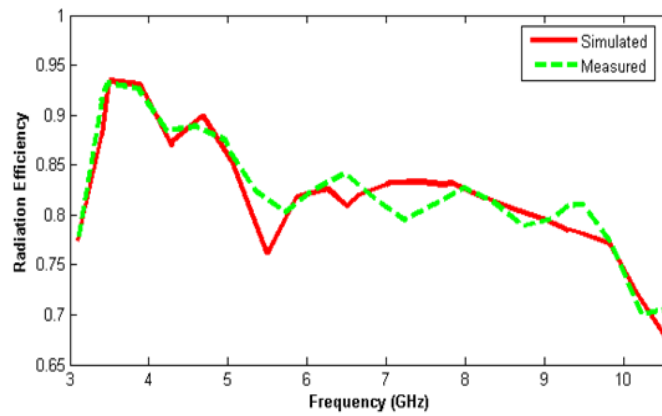


Fig. 3.12 Measured and simulated Efficiency plots of designed structure

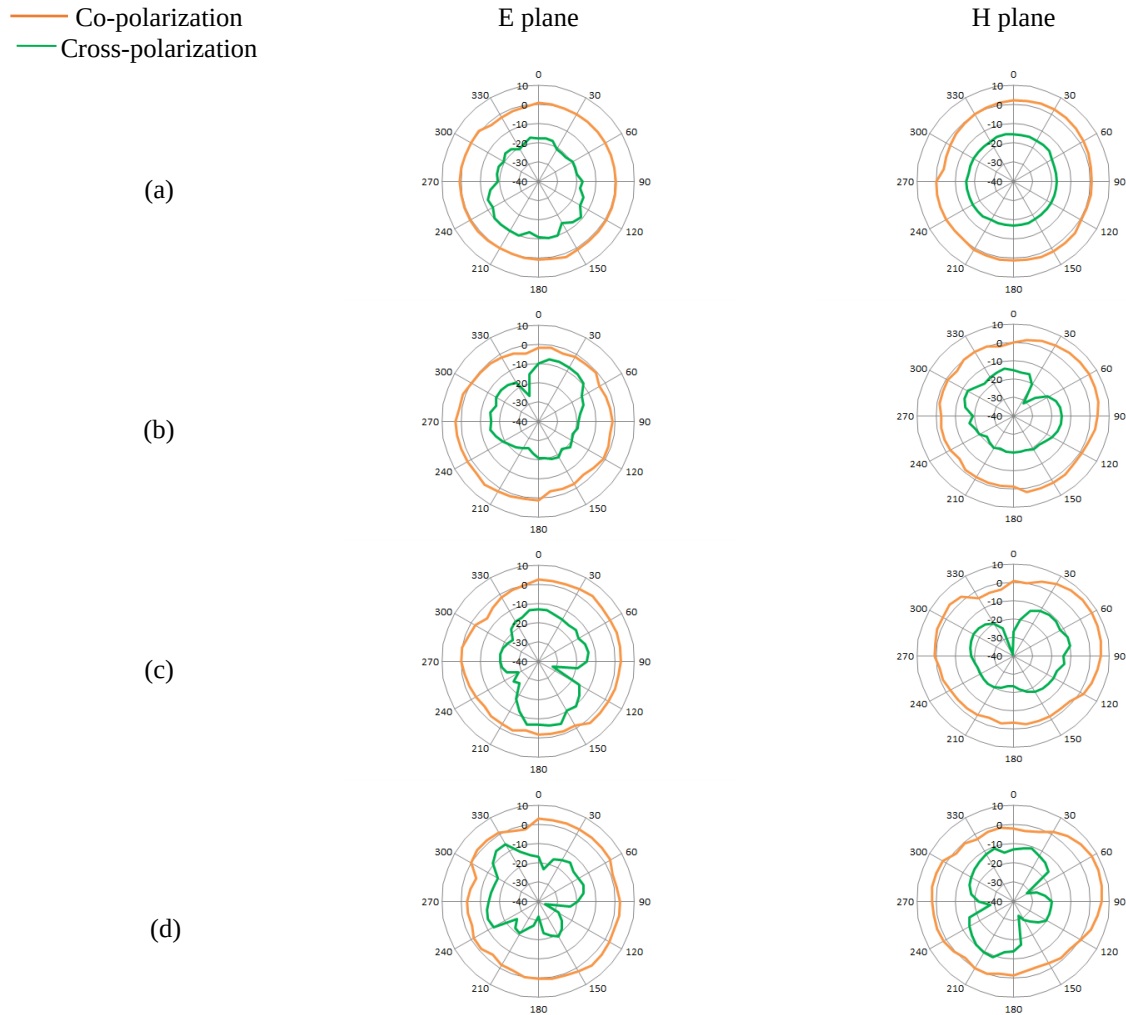


Fig. 3.13 Measured radiation characteristics of wideband antenna at four resonance frequencies (a) 3.4 GHz (b) 5 GHz (c) 7.7 GHz (d) 9.3 GHz

E and H plane with respect to antenna geometry mainly depends on the placement of the antenna in the coordinate system or the antenna orientation with respect to the axes. In the present work yz plane is the H plane ($\Phi=90^\circ$) while xz plane represents the E plane ($\Phi=0^\circ$). Figure 3.13 displays normalized radiation characteristics of designed wideband antenna along with co and cross polarization amplitudes at four distinct resonance frequencies of 3.4, 5, 7.7 and 9.3 GHz. It is found that the E as well as the H plane co polarization radiation patterns have omni-directional nature at smaller frequencies, however at higher frequencies due to high order modes the radiation patterns become directional. Moreover, the cross-polarization amplitudes achieved in both the field planes are quite smaller mainly at smaller frequencies as compared to the corresponding co polarization

amplitudes. The co polarization amplitudes larger than 1.25 dBi along with cross polarization amplitudes smaller than -5.73 dBi are achieved in the two principle field planes as displayed in the Table 3.4.

Table 3.4 Co polarization and cross polarization amplitudes at four resonance frequencies

Frequency (GHz)	Main lobe magnitude (MLM) of UWB antenna			
	E plane		H plane	
	Co-polarization (dBi)	Cross-polarization (dBi)	Co-polarization (dBi)	Cross-polarization (dBi)
3.4	1.94	-10.02	2.95	-16.9
5	3	-7.89	5.46	-15.4
7.7	2.86	-6.92	7.13	-10.7
9.3	1.25	-5.73	7.05	-9.19

3.1.5 Comparison with current impedance matching methods

In the past years, there are numerous impedance matching methods utilized by the scientists. In general, these techniques are widely split into two groups namely broad band and narrow band impedance matching methods. The narrow band matching is achieved by utilizing lumped components (inductor and capacitor), transformers, tuning stubs with altered feedlines, quarter wave transformers (QWT) and impedance matching circuitry. As the work presented mainly focusses on achieving the broadband impedance matching in antenna, hence some of these methods are considered below [101].

1. Multiple section transformers. The proposed method uses numerous $\lambda/4$ segments of the transmission line behaving as the transformers that are linked with each other, further this technique can be viewed as the modern form of quarter wave transformer (QWT) method.
2. Tapered lines. The method utilizes an exponentially expanding and reducing microstrip feed lines to acquire broadband impedance matching characteristics.
3. Distributed Impedance Matching (DIM). The method uses geometrical alterations in the antenna structure to acquire impedance matching characteristics. These alterations modify the concentration of surface current and further establishes reactive loading effect in the antenna structure.

The above mentioned two methods i.e. the multi section and tapered lines are extensively utilized now a days to obtain the broadband impedance matching characteristics. Still these methods suffer

some drawbacks as these methods utilize either numerous segments of transformer or extensive tapered lines to achieve impedance matching at different resonance frequencies. As a result, the overall antenna profile increases. However, the method utilized in present work only uses geometrical alterations without enhancing the overall antenna size.

Merits of DIM method:

Amongst all the methods analyzed before, the DIM designing approach utilized in the present work seems to be quite workable, cheap, user friendly, methodical and organized to achieve impedance matching characteristics. In this method just by modifying the geometry and parameters of resonance structures (LC resonance circuits), the antenna input impedance features at different resonance frequencies can be controlled with ease as seen in Table 3.2. DIM predominantly manages the reactive loading in the designed antenna through geometrical alterations.

The present technique produced better radiation features, radiation efficiency and gain values as alterations in antenna design changes the surface currents pathway that further modifies the radiation characteristics of designed antenna. Moreover, different slots inserted in antenna design which behaves as resonance structures permits the surface currents to follow lengthy and curved routes at larger frequencies that further decreases the proposed antenna's electrical length, hence providing a small or diminished structure. Further the effective area of designed antenna is enhanced by utilizing meandered edges and parasitic ground layer which further causes gain enhancement in proposed structure. Moreover, proposed technique utilizes neither lumped components nor added circuitry that can cause antenna losses and unwanted radiations; hence it shows smaller conductor losses along with satisfactory radiation efficiency when contrasted with other impedance matching methods. Further the proposed method also provides larger degree of freedom while modelling an antenna as in this case numerous antenna performance parameters are easily modified by varying different resonance structures.

Table 3.5 Comparison of presented matching method with various current methods on basis of acquired outcomes

Ref.	Impedance matching method used	Size (mm ²)	Substrate material used (high cost/low cost)	Broadband/narrowband impedance matching	Bandwidth (GHz)	Gain (dB)	Radiation efficiency (%)
[102]	Open-ended microstrip lines	30×20	RT duroid 6002 (high cost)	narrowband	3.92-4.21 (0.29 GHz)	4.5	NA
[103]	Using slot and stub	40×70	FR4 (low cost)	narrowband	2.4-2.484, 5.15-5.35 (0.284 GHz)	Slightly lower than 6	NA
[104]	Using hybrid feeding technique	NA	Glass epoxy (low cost)	narrowband	4.51-5.4 (0.89 GHz)	7.52	NA
[105]	Using slots	86×46	NA	narrowband	2.4-2.52 (0.12 GHz)	NA	NA
[106]	Loading antenna with dielectric material	NA	ROTM10i substrate (high cost)	narrowband	3.1-4.52 (1.42 GHz)	NA	NA
[107]	Modifying periphery of patch and ground	70×90	Teflon glass fiber (high cost)	broadband	3.1-10.6 (7.5 GHz)	5.5	NA
[108]	Multilayered antenna giving coupled resonances	70×70	Duroid and FR4 (high cost)	broadband	3.35-6.95 (3.6 GHz)	8.8	93
[109]	Using slots	50×50	FR4(low cost)	narrowband	905-925 (0.02 GHz)	3	NA
[110]	Using matching circuit	NA	FR4(low cost)	broadband	0.5-3.5 (3 GHz)	NA	NA
[111]	Using matching circuit	50×13.5	FR4(low cost)	narrowband	0.60-1.25, 1.60-2.53 (2.23 GHz)	4.8	90
[112]	Loading metamaterial unit cell	31×24.57	FR4 (low cost)	broadband	3.87-7.63 (3.87 GHz)	2.51	93
[113]	Using modified patch	30×30	FR4 (low cost)	broadband	3.5-10.6 (7.1 GHz)	5.72	NA
[114]	Structural modification	120×120	FR4 (low cost)	narrowband	1.03-2.31 (1.28 GHz)	2.3	NA
[115]	Using matching network	60×40	TLYA-5CH200 (high cost)	narrowband	(1.728 GHz)	NA	NA

[116]	$\lambda/4$ transmission line (QWT)	58.35 × 38.5	FR4 (low cost)	narrowband	850-963, 1450-1650, 2175-2425 (0.563 GHz)	1.3	75
Proposed work	Using multi resonating structures	30×30	FR4(low cost)	broadband	3.1-10.6 (7.5 GHz)	6.1	94

It is quite evident from Table 3.5, that the suggested matching method provides 94 % of largest radiation efficiency as contrasted to remaining impedance matching methods, showing least losses. It is found in the Table 3.4 too, as co polarization amplitude more than 1.25 dBi along with cross polarization amplitude smaller than -5.73 dBi is acquired in the antenna's principle plane patterns (E and H plane). The proposed work provides 6.1 dB of large gain value except [99, 103], however when comparing the remaining outcomes with [99, 103] the designed antenna shows better performance in respect of its complete profile, impedance bandwidth and acquired radiation efficiency. Moreover, it provides 7.5 GHz of large impedance bandwidth in comparison with all the outcomes discussed above. Through the outcomes observed above it is quite evident that the suggested design technique provides comparatively better outcomes than the available design methods.

3.1.6 Antenna characteristics in the time domain

The correlation factor between the transmitted and received antenna can be calculated using (3.5) below.

$$F = \max_{\tau} \left| \frac{\int_{-\infty}^{+\infty} S(t) r(t-\tau) dt}{\int_{-\infty}^{+\infty} S(t)^2 dt \cdot \int_{-\infty}^{+\infty} r(t)^2 dt} \right| \quad (3.5)$$

Here $S(t)$ and $r(t)$ denotes the transmitted and received antenna signals and τ denotes the delay factor. As compared to many telecommunication systems, in IR-UWB systems, it is required to achieve good correlation within the transmitting and receiving signals so to have maximum information transfer with minimum distortions due to radio channel. To calculate the fidelity parameter, firstly the input and received signals are studied for two antenna configurations i.e. side by side and face to face. In this study the transmitting and receiving antennas are placed 250 mm apart from each other [117] and simulated results are obtained using CST microwave studio. Now for excitation purpose, the default Gaussian pulse in CST with spectrum ranging from 3.1 to 10.6 GHz is utilized as shown in the Fig. 3.14 (a). As observed below the received signals in Fig. 3.14

(b) and Fig. 3.14 (c) are broadened, still a good correlation is observed within the transmitted and received signals. By utilizing the above equation (3.5), the antenna fidelity factor for two orientations is calculated. The Fig. 3.14 shows the normalized wave signals used for calculating the antenna fidelity factor. Table 3.6 gives fidelity factor values for two antenna orientations.

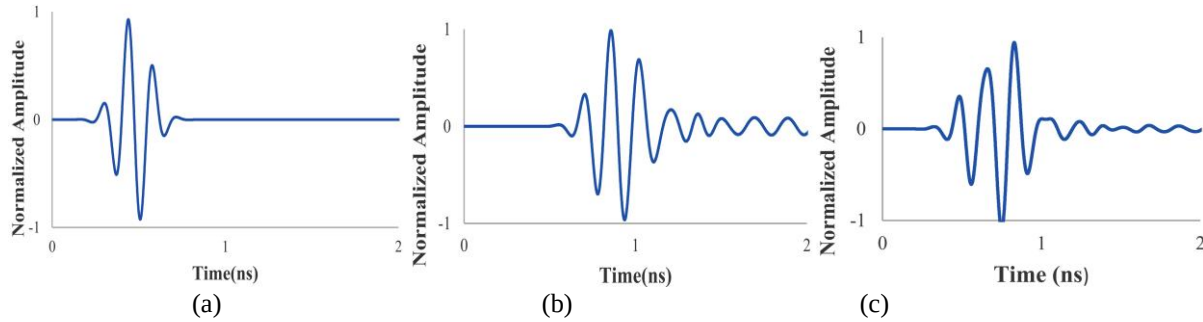


Fig. 3.14 (a) Input excitation signal (b) Received signal (side by side) (c) Received signal (face to face)

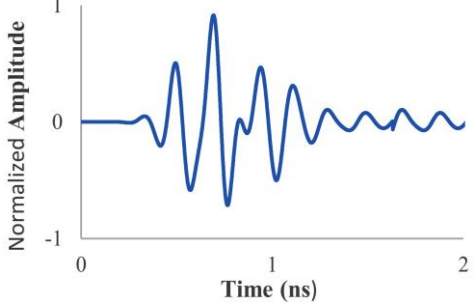
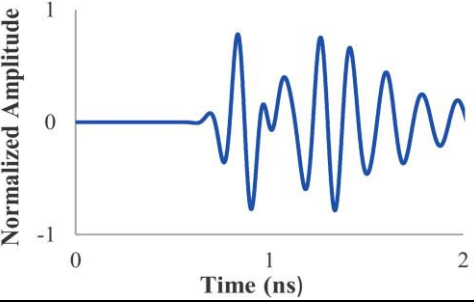
Table 3.6 Fidelity factor for two antenna scenarios

Scenarios	Side by Side	Face to face
Fidelity factor	0.9843	0.9438

For calculating the system fidelity factor, two identical antennas for transmitting and receiving pulses are placed 250 mm apart. The transmitting antenna facing the other antenna is fixed, however the receiving antenna is rotated at different angles. Table 3.7 gives the system fidelity factor values for different receiver antenna rotations.

Table 3.7 System fidelity factor of antenna for different angles of rotation

Rotation angle	Received wave forms	System fidelity factor
$\theta^0=0^\circ$		0.9400

$\theta^0=90^\circ$		0.9267
$\theta^0=180^\circ$		0.9304

3.2 SUMMARY

The main features of this chapter are:

1. The initial geometry comprises of a simple circular patch antenna having 8.1 mm of radius along with $50\ \Omega$ strip line feeding.
2. The antenna shows overall dimension of $30 \times 30\ \text{mm}^2$ and utilizes FR4 substrate having 4.4 of relative permittivity and 1.6 mm of thickness.
3. The multiband antenna envelopes 7.5 GHz of continual bandwidth covering 3.1 to 10.6 GHz of frequency band along with reflection coefficient values under -20 dB at various resonant frequencies.
4. The proposed antenna also displays highest gain and radiation efficiency value of 6.1 dB and 94 %.
5. The matching method utilized apart from giving multi resonant broadband character with better reflection coefficient values also provides increased degree of freedom, lesser conductor losses and antenna size reduction.

In the next chapter various notching stubs will be attached to the wideband antenna discussed in the present chapter. The tuning stubs will provide wide notch for rejecting both WiMAX and WLAN bands.

CHAPTER 4

GUITAR SHAPED BAND NOTCHED ANTENNA

The chapter includes the designing, fabrication and testing of a band notched antenna. As we know the ultrawideband antenna enveloping the unauthorized region from 3.1 to 10.6 GHz exists together with several narrow range wireless systems. Therefore, it is quite essential for any ultrawideband antenna to have good rejection features. The notching method utilized in this chapter is the “technique of inserting strips” which behave as notching structures to achieve band rejection. The proposed method demonstrates two major benefits. In the first place, it shows very minute impact on the design of base antenna, as notching is obtained entirely by inserting strips with no effect on the actual structure. Secondly, it manages to have simpler antenna designing as the notching frequency as well as stop band range is controlled with ease by adjusting the length and gap parameters of notching structures.

4.1 DESIGN OF BAND STOP STRUCTURE

The final configuration of designed guitar shape band stop structure is given in the Fig. 4.1 and corresponding improved measurements are shown in the Table 4.1. The designed structure comprises of an altered circle shape radiating patch along with a ground plane that uses 50 Ω microstrip line feeding. Further various resonance stubs are brought to generate a wide tunable stop band. As shown in Fig. 4.1 (a) a pair of inverted L shaped stubs acting as $\lambda/4$ open ended strips are added at top side of antenna patch to reject frequency from 5 GHz to 5.6 GHz and a pair of parallel I shaped stubs $\lambda/4$ long are attached to the bottom side of the patch adjacent to the feed line to notch out frequency from 3.1 to 5.1 GHz. By inserting these metallic stubs together, a continuous band rejection is achieved from 3.1 GHz to 5.8 GHz. Initial lengths (L_{stub}) of both L and I shaped stubs is $1/4$ of the free space wavelength (λ_0) and is calculated by utilizing (4.1) [118, 119]. However, the initial length dimension of both the strips is further improved to achieve desired band stop characteristics.

$$L_{stub} = \frac{\lambda_0}{4} = \frac{c}{4f_n\sqrt{\epsilon_{reff}}} \quad (4.1)$$

Where $\epsilon_{reff} = \frac{\epsilon_r + 1}{2}$ denotes effective dielectric constant, f_n denotes middle frequency of the frequency region that needs to be rejected, λ_0 signifies the wavelength of wave in empty space, ϵ_r

represents substrate dielectric constant along with ‘ c ’ indicates velocity of light in vacuum with 3×10^8 m/s of value. The geometrical evolution of the final structure is discussed in next section.

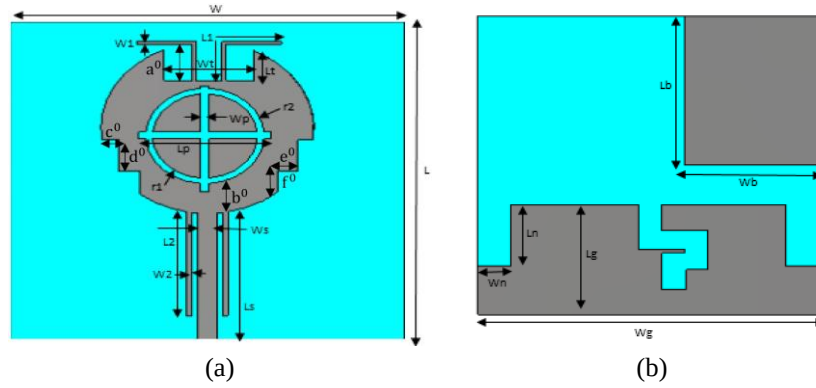


Fig. 4.1 Geometry of proposed band notched wideband structure (a) Upper layer (b) Bottommost layer

Table 4.1 Improved measurements of the designed structure (mm)

Parameters	L	W	L_s	W_s	L_p	W_p	L_t	W_t	$r1$	$r2$
Dimension	30	30	11.6	1.4	10	0.65	4	6.8	4	4.5
Parameters	L_g	W_g	L_n	W_n	L_b	W_b	$L1$	$W1$	$L2$	$W2$
Dimension	11	30	6	3	15	12	8.7	0.3	10	0.4
Parameters	a^0	b^0	c^0	d^0	e^0	f^0				
Dimensions	3.7	2.9	1	3.5	1.5	3				

4.1.1 Structural progression of ultimate structure

The step by step structural progression of ultimate design of given structure and its comparable reflection coefficient curves along with every structural modification are given in the Fig. 4.2 and Fig. 4.3 respectively. Initially in Fig. 4.2 (a) a modified circle shape radiating patch having 8.1 mm of radius with multiple resonating structures like plus and ring shape slot at center along with a defected ground structure (DGS), along with a staircase matched feed line is used to design a basic UWB antenna. Two-inverted parallel L shaped stubs is attached to Antenna 1 as shown in Fig. 4.2 (b) that creates a stop band covering 5 to 5.6 GHz of frequency band. When only two parallel I shaped stubs are attached to the base antenna 1 as displayed in the Fig. 4.2 (c) a stop band from 3.1 to 5.1 GHz is created. Finally, when both L and I shaped stubs is added to the base antenna 1 as shown in Fig. 4.2 (d), the entire region from 3.1 to 5.8 GHz is rejected, thus avoids interference

from WiMAX and WLAN band. Therefore, by combining the two resonant structures (L and I shaped stubs) in a single design a large stop band is achieved. The impact of modifying the various dimensions of proposed structure on its performance is discussed next.

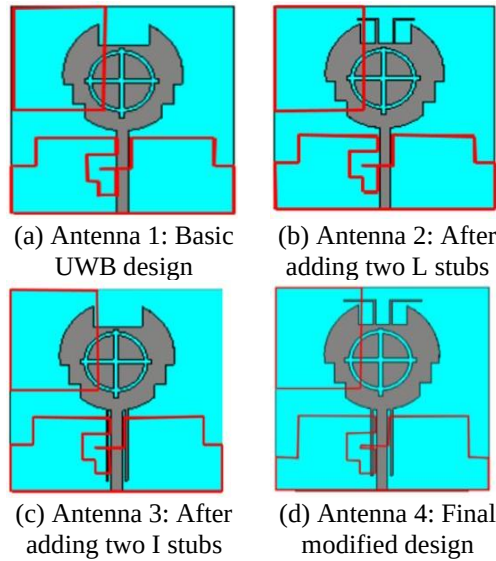


Fig. 4.2 Geometrical variations of the band stop structure

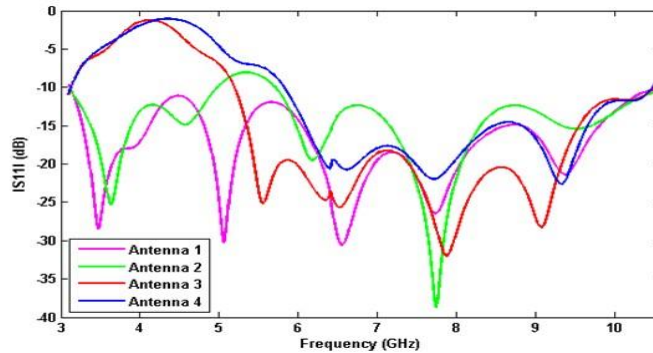


Fig. 4.3 Reflection coefficient plot after every geometrical variation of designed antenna

4.1.2 Parametric alteration of designed structure

Different measurements of proposed structure are separately altered to observe their counterpart impact on the antenna reflection coefficient plots. This analysis assists a researcher to select the optimum measurements of proposed structure for achieving the wanted outcomes.

Geometrical parameters affecting pass band characteristics

Impact of alterations in the length 'Lt' of upmost U shape slot on reflection coefficient. Value of Lt is altered from 1 to 5 mm and 4 mm is observed to be the best Lt value. With the reduction in Lt value under 4 mm to 1 mm, the stop band attenuation level remains same with slight decrement in its bandwidth whereas in pass band region the reflection coefficient value sharply increases in the frequency region from 5.9 to 8 GHz and total bandwidth reduces from 4.9 GHz to 2.8 GHz as given in the Fig. 4.4. Now as we enhance the Lt value more than 4 mm i.e. to 5 mm, slight variations are found in both stop band and pass band regions and by increasing the Lt above 5 mm, upmost U shape slit intersects with plus shape slot in the middle of antenna patch.

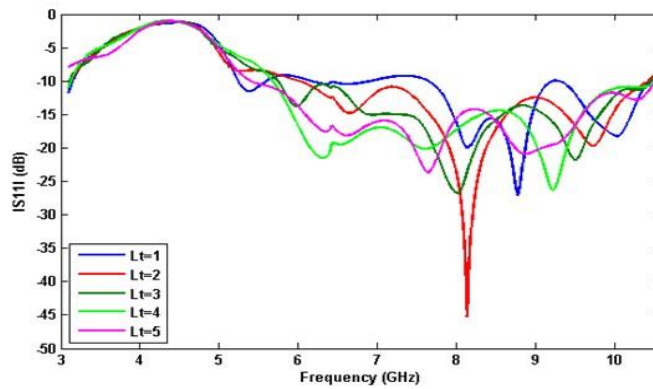


Fig. 4.4 Impact of alterations in 'Lt' on antenna reflection coefficient

Impact of alterations in the half ground layer step width 'Wn' on reflection coefficient. Value of Wn is altered from 1 to 5 mm with a step size of 1mm. 3 mm is observed to be the best improved Wn value. With reducing Wn value, the attenuation level and bandwidth of stopband region remains almost unaffected, whereas in passband region, bandwidth reduces from 4.7 GHz to 3.8 GHz at 1mm as depicted in Fig. 4.5. The reflection coefficient also increases in the entire passband except the region from 7.7 to 9.1 GHz where it drops to a minimum value of -56 dB at 8 GHz. The increase in the value of Wn above 3 mm does not affect the attenuation level of stop band but decreases its bandwidth. In pass band region the reflection coefficient increases in the region from 7.5 to 9.5 GHz and rises to a maximum value of -9 dB when Wn = 5 mm, therefore as we increase the Wn further, the bandwidth of the passband decreases.

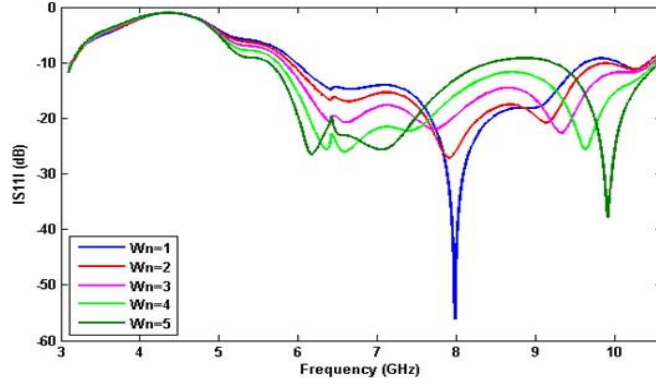


Fig. 4.5 Impact of alterations in 'Wn' on antenna reflection coefficient

Geometrical parameters affecting stop band characteristics

Tuning of the stop band by varying gaps 'G1 and G2', lengths 'L1 and L2' and widths 'W1 and W2'. The gap 'G1' is altered from 0.1 to 2.9 mm with the step size of 0.7 mm, observed in the Fig. 4.6 (a). Below 0.1 mm the two inverted L stubs intersects with each other and above 2.9 mm these stubs intersects with the patch. It is observed that increment in the value of G1 only affects the stopbands higher limiting frequency. Upper limiting frequency increases by only 0.8 GHz as 'G1' is increased from 0.1 mm to 2.9 mm. Further 1.5 mm is observed to be the best improved 'G1' value. Similarly, gap 'G2' is also varied from 1mm to 2.2 mm with a step size of 0.3 as given in the Fig. 4.6 (b). With reducing the 'G2' value under 1 mm, it intersects with microstrip feed line and above 2.2 mm, it gives a huge stop band rejecting the desired bands also. Initially at G2 = 1mm, the bandwidth of the stop band is 0.7 GHz, as we increase the value of 'G2' to 2.2 mm, the bandwidth of the stopband increases to 3.3 GHz. Both the upper and lower cut off frequencies of the stop band is affected by this increase. The optimum value of 'G2' is found to be 1mm, as it gives the desired result. It is observed that closer the gap, smaller will be the obtained stopband bandwidth as specified in (4.6) also.

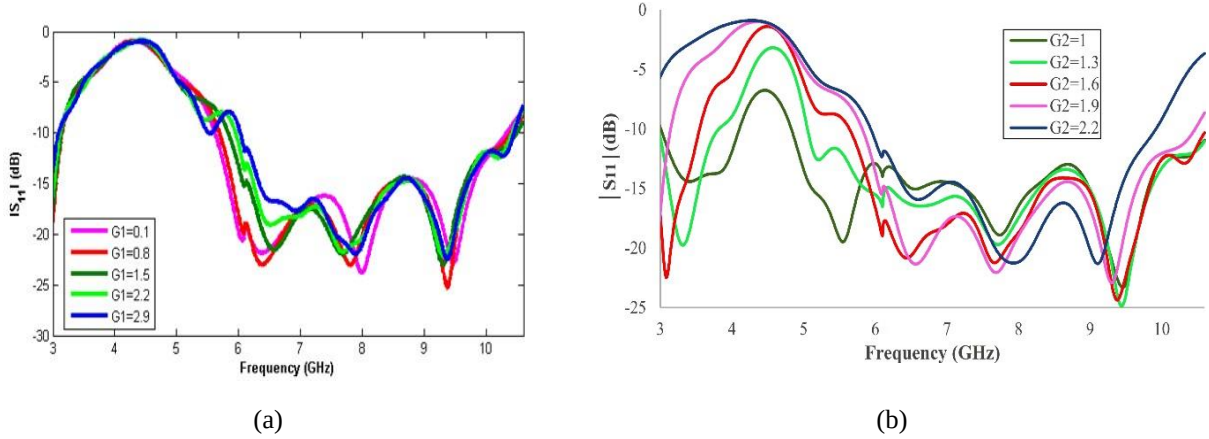


Fig. 4.6 Effect of variation of gaps (a) 'G1' and (b) 'G2' on the S_{11} curve of antenna

The length 'L1' is altered from 0.7 to 9.7 mm as shown in the Fig. 4.7 (a). Till $L1 = 3.7$ mm there is only slight variations in the reflection coefficient curve but above 3.7 mm, the upper stopband bandwidth increases till $L1 = 6.7$ mm. However above $L1 = 6.7$ mm as we further increase the length 'L1', the continuous shift in the higher notching frequencies to the lower frequency band is observed. 8.7 mm is observed to be the ideal 'L1' value. The 'L2' value is altered from 1 to 12 mm as observed in Fig. 4.7 (b). It's found that there is a continuous shift in the lower notching frequencies to the lower frequency side as we increase the value of 'L2', it mainly happens due to inverse relationship within the notched frequency and the length of slot as given in (4.1) also. The reflection coefficient initially increases as we increase the value of 'L2' from 1 to 6 mm. However, above 6 mm, S_{11} values in the passband region starts improving at the cost of reduced passband bandwidth. Therefore, 10 mm is observed to be the ideal 'L2' value, as reflection coefficient under -20 dB is achieved in passband region with desired notch band.

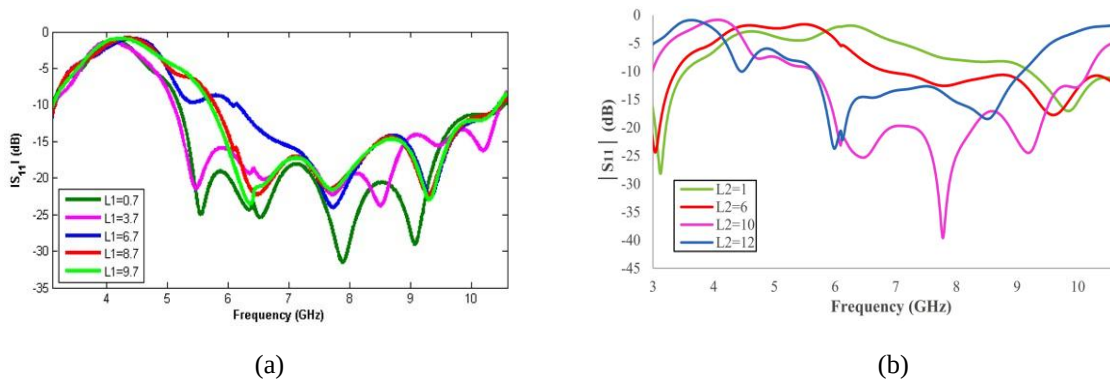


Fig. 4.7 Effect of variation of lengths (a) 'L1' and (b) 'L2' on the S_{11} curve of antenna

The width 'W1' is altered from 0.1 to 0.7 mm with the step size of 0.3 mm as given in the Fig. 4.8 (a) and has almost negligible effect on the S_{11} curve of antenna with increase in the value of 'W1'. The value of 'W2' is altered from 0.1 to 1.3 mm in the step width of 0.3 mm as shown in Fig. 4.8 (b). At $W2 = 0.1$ mm, the obtained stop band bandwidth is 2.5 GHz, as we increase 'W2' value the stopband bandwidth also enhances from both upper and lower cut off frequency and reach to a maximum value of 4 GHz at $W2 = 1.3$ mm. Therefore $W2 = 0.4$ mm is found to be the optimum value. It is observed from the above plots that the variations in the parameters related to I stubs (G2, L2, W2) have more effect on the S_{11} curve as compared to L stubs (G1, L1, W1) which is probably because of higher current density near the feed line, therefore I stubs give comparatively higher notching effect than L stubs.

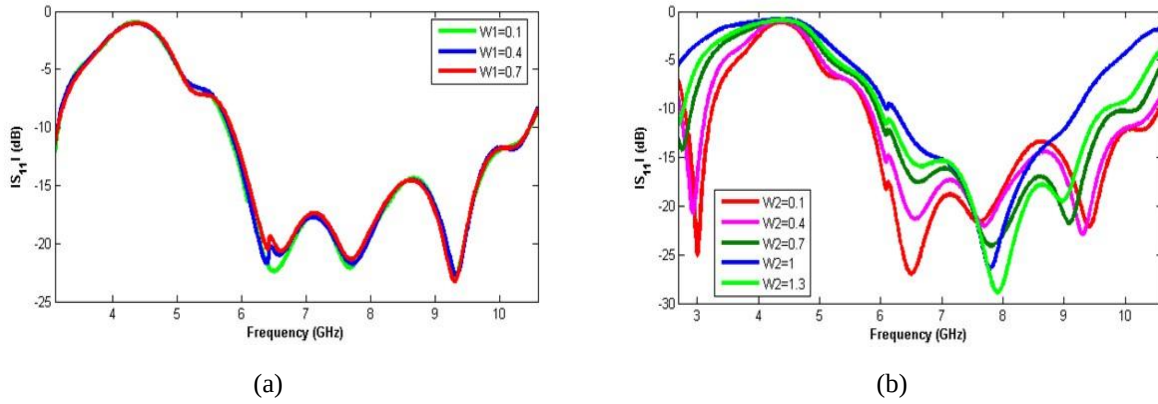


Fig. 4.8 Effect of variation of widths (a) 'W1' and (b) 'W2' on the S_{11} curve of antenna

4.1.3 Equivalent circuit analysis

The notching characteristic of these metallic stubs are easily acknowledged by observing the concentration of surface currents in the proposed structure. Around different notching frequency the metallic stubs attain higher concentration of current to produce the desired attenuation. The length of these stubs shows inductance effect whereas the interactions with in stubs and periphery of the antenna give rise to capacitive effect, which makes these stubs to act as parallel resonance circuit. By inserting these stubs an additional current path is introduced which provides a sufficient quantity of capacitance coupling within stubs and outer borderline of antenna. The final conceptual approximate circuit model of proposed stop band structure is given in the Fig. 4.9 (c) here an RLC parallel resonance circuit analogous to the center frequency (4.3 GHz) of notched band is

connected in series with impedance of broadband structure. The total impedance of designed stop band structure is given as [120].

$$Z = Z_A + Z_n \quad (4.2)$$

Where Z_A denotes input impedance of proposed structure and Z_n denotes impedance of notch. Value of Z_n is given by

$$Z_n = \frac{1}{(1/R) + (1/j\omega L) + j\omega C} \quad (4.3)$$

In (4.3) R , L and C are the equivalent resistance, inductance and capacitance in the final equivalent circuit model as displayed in the Fig. 4.9 (c). According to [121] the resonating frequency (f_n) along with 3 dB bandwidth of notch (BW_n) is calculated by

$$\omega_n = 2\pi f_n = \frac{1}{\sqrt{LC}} \quad (4.4)$$

$$BW_n = \frac{1}{RC} \quad (4.5)$$

$$Q_n = \frac{\omega_n}{BW_n} = R\sqrt{\frac{C}{L}} \quad (4.6)$$

Where ω_n is the angular frequency and Q_n is quality factor of resonant circuit and its also clear from (4.6) that the quality factor of notched region depends on the gaps within the resonators along with the antenna patch. A closer gap produces strong capacitive effect which further increases the stored energy, surface current density, frequency selectivity and hence Q factor of the antenna at the cost of reduced stop band bandwidth.

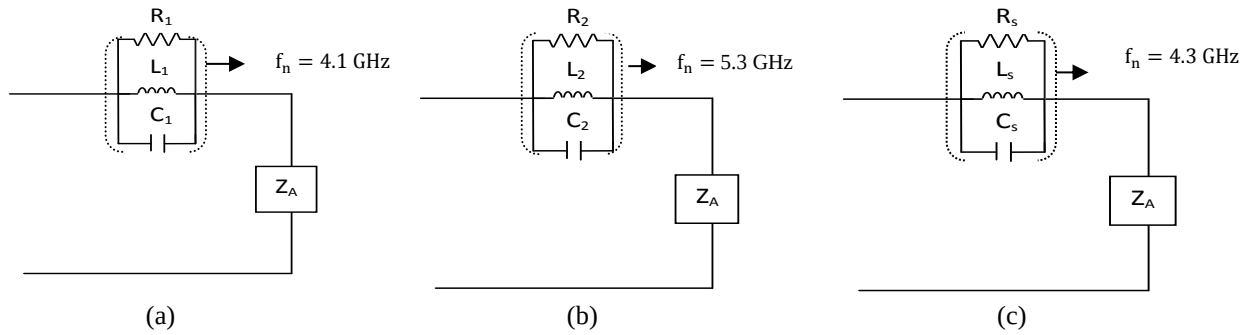


Fig. 4.9 (a) Equivalent circuit model (I strips only) (b) Equivalent circuit model (L strips only)
(c) Circuit model of designed stop band structure (both L and I strips)

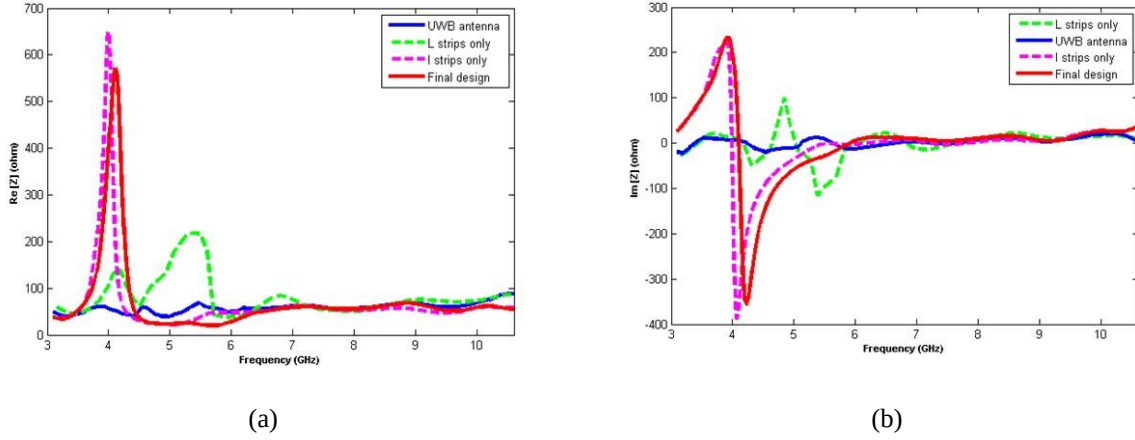


Fig. 4.10 (a) Simulated real input impedance (b) Simulated imaginary input impedance

The simulated imaginary and real components of antenna impedance are shown in the Fig. 4.10. Proper input impedance matching of $Re[Z]$ close to $50\ \Omega$ and $Im[Z]$ around zero ensures the value of S_{11} below -10 dB. A value of input impedance farther from $50\ \Omega$ and zero for real part and imaginary part respectively show the impedance mismatch in the antenna. As in UWB antenna we can observe good impedance matching across entire broadband frequency region from 3.1 to 10.6 GHz. When only L stubs is used, the maximum impedance mismatch can be easily observed at 5.3 GHz with the value of $Re[Z]$ as $218\ \Omega$ and $Im[Z]$ as $-117\ \Omega$, whereas proper matching is observed from 6 to 10.6 GHz. When only I stub is used the maximum impedance mismatch with $Re[Z]$ as $660\ \Omega$ and $Im[Z]$ as $-386\ \Omega$ is observed at 4.1 GHz. Similarly, in the final design plots we observe proper impedance matching in the pass band covering 6 GHz to 10.6 GHz of frequency band and mismatch in rejected region from 3.1 to 5.8 GHz. The maximum impedance mismatch is observed at 4.3 GHz with $Re[Z]$ $585\ \Omega$ and $Im[Z]$ $-357\ \Omega$.

4.2 MEASUREMENT AND SIMULATION OUTCOMES

Designed band notched structure is manufactured and experimentally examined utilizing KEYSIGHT E5063A (100 KHz -18 GHz) model of VNA together with electromagnetic chamber. Further CST Microwave Studio software is used for simulation studies of various antenna parameters. Picture of manufactured antenna is displayed in the Fig. 4.11.

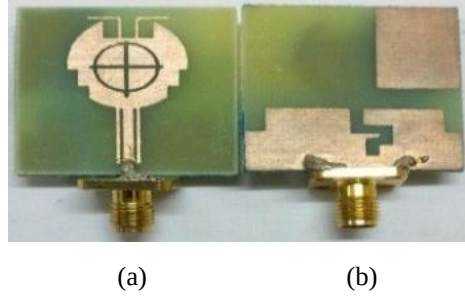


Fig. 4.11 Picture of manufactured band notch antenna showing (a) Top layer (b) Bottom layer

Figure 4.12 shows the magnitude of S_{11} curve of basic UWB antenna and band notched structure. By observing simulation results it is quite evident that the broadband structure works in frequency region covering 3.1 to 10.6 GHz thereby producing 7.5 GHz of enormous and continuous bandwidth. Contrarily, band notched antenna has a wide stopband region from 3.1 GHz to 5.8 GHz with stopband bandwidth of 2.7 GHz and maximum attenuation level of -1.07 dB at 4.3 GHz, whereas the passband region has total bandwidth of 4.8 GHz along with uniform reflection coefficient (under -10 dB) in region from 5.8 to 10.6 GHz. From measurement outcomes it's found that measured bandwidth range of the stopband increases slightly from 2.7 GHz to 2.9 GHz with slight decrease in the attenuation level in frequency band covering 3.1 to 3.4 GHz along with 5.3 to 5.6 GHz whereas in passband slight variations in S_{11} curve is observed. The differences in measurement outcomes is observed due to inaccuracies transported when we manufacture the designed structure using chemical etching method and further because of calibration problems produced while measuring the proposed structure with VNA.

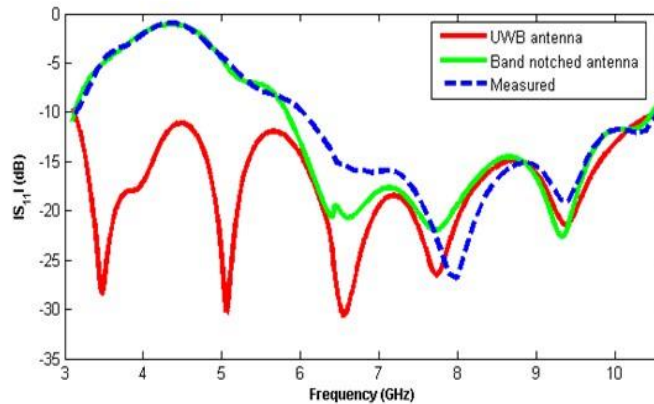


Fig. 4.12 Simulated and measured S_{11} curve for the fabricated antennas

Simulation and measurement wave ratio graphs of designed antennas is shown in Fig. 4.13. Through simulation graph, it is quite evident that proposed structure shows wave ratio value below 2 across the complete wideband range covering 3.1 to 10.6 GHz of frequency and with the introduction of notch the VSWR value has increased from 1.8 to a maximum value of 16.2 at center frequency (4.3 GHz) in the rejected region. In the measured curve, the value of VSWR has increased by 2.5 from 16.2 to 18.7 at center notch frequency, whereas in passband region VSWR below 2 is maintained.

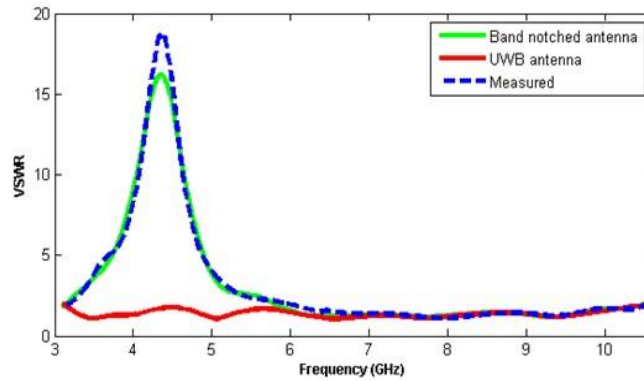


Fig. 4.13 Simulated and the measured VSWR plot of the designed antennas

The simulated as well as measured gain curves of designed structure are shown in Fig. 4.14. The basic wideband structure displays consistent gain values in the complete wideband range along with 5.8 dB of highest gain value at 9.4 GHz and 1.8 dB of lowest gain value at 4.2 GHz, whereas in band notched antenna, the simulated gain has decreased in stop band frequency region (3.1 GHz-5.8 GHz) with minute alterations in passband region. It is seen from measured curve that band notching does not affect the passband gain of the antenna. The maximum level of decrement in gain value is seen in frequency region covering 3.7 to 4.9 GHz, where it has dropped below 0 dB and lowest gain value is found at 4.3 GHz where reduction of 6.1 dB is observed from 1.9 dB to -8 dB. Further the measurement and simulation gain plots show little discrepancies because of losses during testing and liberties while manufacturing.

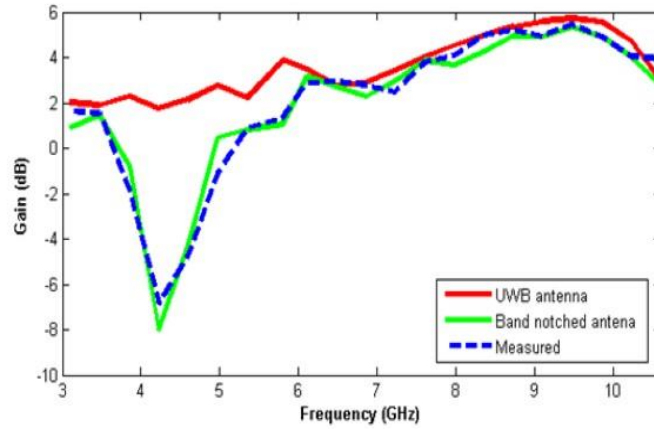


Fig. 4.14 Measured and simulated gain curves of designed structures

The measurement and simulated radiation efficiency curves of designed structure are shown in the Fig. 4.15. UWB antenna has highest efficiency value of 0.95 at 3.8 GHz along with lowest value of 0.68 at 10.5 GHz, whereas in case of band notched antenna the simulated radiation efficiency has reduced in frequency band covering 3.1 to 5.8 GHz with small modifications in passband region. It is found from measurement plot that band notching does not affect the passband radiation efficiency of the antenna. The maximum level of reduction is observed at 4.3 GHz where it has reduced by 0.82 from 0.87 to 0.05. The simulation and measurement efficiency plots are almost similar with slight variations.

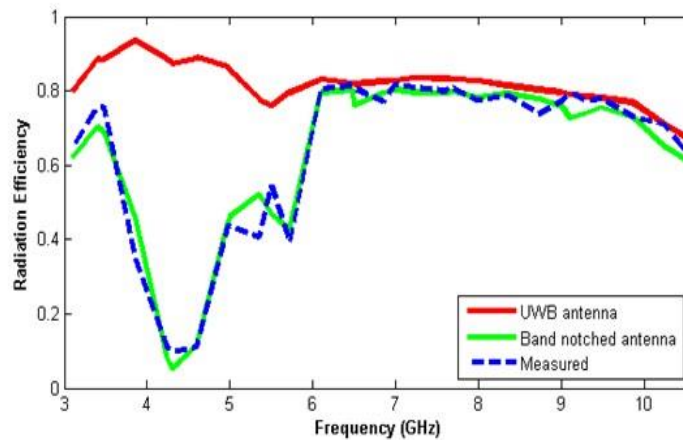


Fig. 4.15 Measurement and simulation Radiation Efficiency plots of designed structures

Table 4.2 Contrast between the designed band notched structure specifications with various latest structures

Ref.	Size (mm ²)	Notched bands or frequencies	Max. VSWR in rejected bands	Real and imaginary part of equivalent circuit impedance in rejected bands at resonance	Minimum (Gain, Radiation efficiency) in rejected bands
[122]	(48 × 36)	WiMAX, WLAN	(5.3,8.3)	Re [z] (137 Ω , 12 Ω) Im [z] (-87 Ω , -15 Ω)	N/A
[123]	(30 × 39.3)	WLAN	N/A	N/A	(-5.8 dB, 17%)
[124]	(33 × 32)	WiMAX, WLAN	N/A	N/A	(-2.3 dB, N/A) and (-3 dB, N/A)
[125]	(40×34)	WiMAX, WLAN, ITU	(6.5,4,2.5,4)	N/A	(0.1 dB, N/A), (-0.8 dB, N/A), (2 dB, N/A) and (-1 dB, N/A)
[126]	(48×32)	WiMAX, WLAN	(8,12)	N/A	(-2.1 dB, 30%) and (-0.8 dB, 57%)
[127]	(44×35)	WiFi, WiMAX, WLAN	(12,7.7,9,5.1)	N/A	(-6.1 dB, N/A), (-3.1 dB, N/A), (-4 dB, N/A) and (-3.6 dB, N/A)
[128]	(36×38)	3.75/5.43/7.87/8.6 2/9.87	N/A	Re[z] (150 Ω , 0 Ω , 10 Ω , 25 Ω , 125 Ω) Im[z] (-50 Ω , 0 Ω , 10 Ω , 0 Ω , -50 Ω)	(-4.5 dB, N/A), (0 dB, N/A), (-1 dB, N/A), (-2.5 dB, N/A) and (-0.8 dB, N/A)
[129]	(80×80)	2.3/3.2/5.2/5.5/7.5	(12,11,7,12,6)	N/A	N/A
[130]	(38×30)	C band, WLAN, DSRC	(13,7,4.1)	N/A	N/A
[131]	(40×30)	WiMAX, HiperLAN	(14,16)	N/A	(-10 dB, 20%) and (-8 dB, 30%)
Proposed Work	(30 × 30)	WiMAX, WLAN	18.7	Re [z] (585 Ω) Im [z] (-357 Ω)	(-7 dB, 5%)

Table 4.2 displays the differentiation of several specifications of proposed structure such as overall size , notched bands, largest VSWR, minimum efficiency and gain values in the rejected bands and the imaginary and real component values obtained from circuital model in rejected bands at resonance frequency with latest structures in writings. It is found that the designed structure shows maximum VSWR (18.7), minimum radiation efficiency (5%), maximum deviation of Re [Z] and Im [Z] from 50 Ω and zero (585 Ω , -357 Ω) , least size (30×30) and optimum value of gain (-7 dB) at rejected band, indicating good notch performance in comparison with other discussed

manuscripts in the Table 4.2. Hence it is viewed as a suitable contender for various wireless applications.

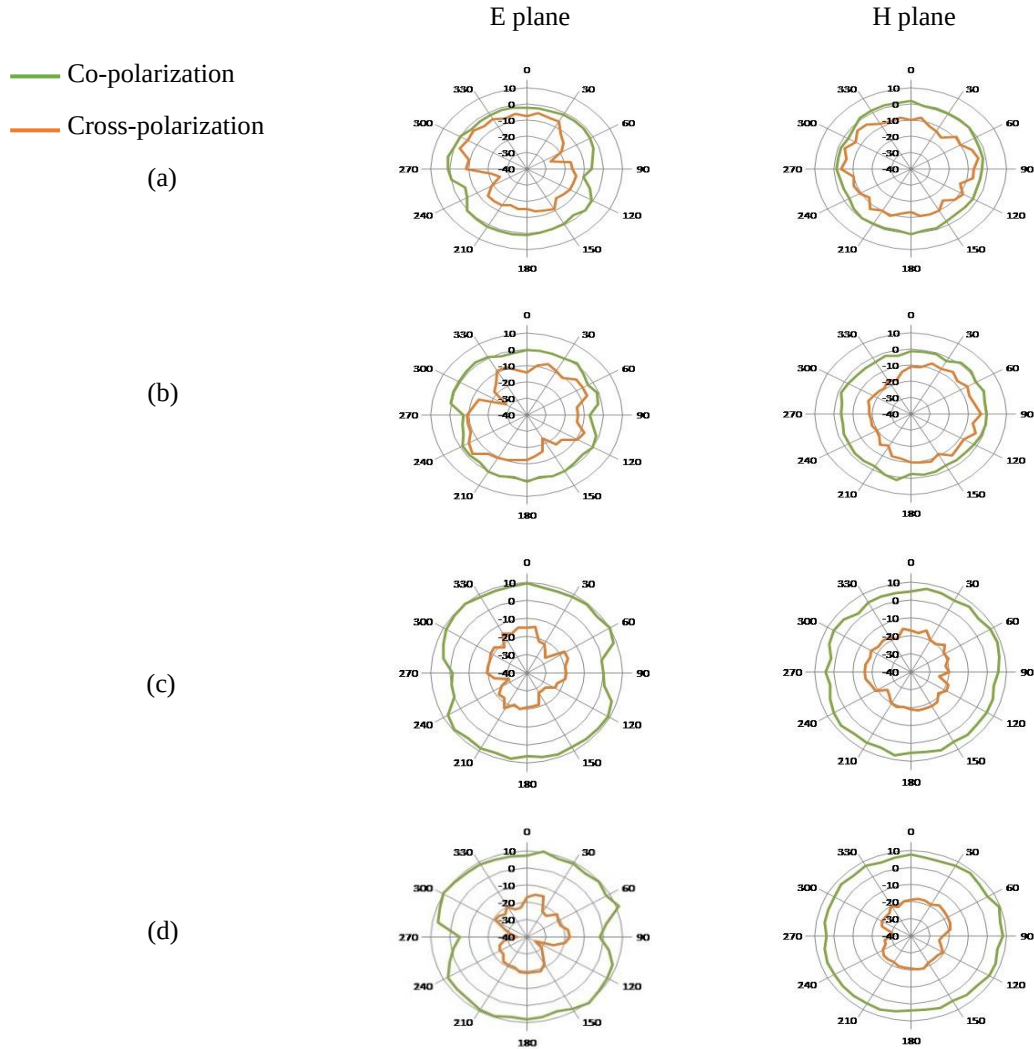


Fig. 4.16 Displays measured radiation characteristics of band notched structure at four resonating frequencies (a) 3.4 GHz (b) 4.3 GHz (c) 7 GHz (d) 9 GHz

Based on the placement of designed antenna in the coordinate system, in present work yz plane is the H plane ($\Phi=90^\circ$) while xz plane represents the E plane ($\Phi=0^\circ$). Figure 4.16 illustrates the standardized radiation features of designed band notched structure showing cross and co polarization amplitudes at two stop band frequencies (3.4 GHz, 4.3 GHz) and two pass band frequencies (7 GHz, 9 GHz). It is observed that the pass band frequencies are quite unaffected by the negative effects of notching as co polarization amplitudes larger than 8 dBi along with cross polarization amplitudes smaller than -17 dBi are observed in pass band frequencies, whereas in

stop band frequencies quite lower co polarization level and higher cross polarization level are observed as shown in the Table 4.3. Further at passband frequencies nearly omnidirectional H pattern and quasi omnidirectional E pattern is found. Moreover, the co and cross polarization components at E field pattern at various pass band frequencies are higher than the H field pattern components.

Table 4.3 Showing the cross and polarization amplitudes at various resonance frequencies

Frequency (GHz)	MLM of band notched UWB antenna			
	E plane		H plane	
	Co-polarization (dBi)	Cross-polarization (dBi)	Co-polarization (dBi)	Cross-polarization (dBi)
3.4	1.12	-3.01	2.09	-3.51
4.3	0.73	-2.81	1.88	-3.22
7	10.12	-21.92	8.19	-17.61
9	11.45	-23.24	9.51	-18.70

The correlation factor between the transmitted and received antenna can be calculated using (4.7) below.

$$F = \max_{\tau} \left| \frac{\int_{-\infty}^{+\infty} S(t) r(t-\tau) dt}{\int_{-\infty}^{+\infty} S(t)^2 dt \cdot \int_{-\infty}^{+\infty} r(t)^2 dt} \right| \quad (4.7)$$

Here $S(t)$ and $r(t)$ denotes the transmitted and received antenna signals and τ represents the delay factor. As compared to many telecommunication systems, in IR-UWB systems, it is required to achieve good correlation within the transmitting and receiving signals so to have maximum information transfer with minimum distortions due to radio channel. To calculate the fidelity parameter, firstly the input and received signals are studied for two antenna configurations i.e. side by side and face to face. In this study the transmitting and receiving antennas are placed 250 mm apart from each other and simulated results are obtained using CST microwave studio. Now for excitation purpose, the default Gaussian pulse in CST with spectrum ranging from 3.1 to 10.6 GHz is utilized as shown in the Fig. 4.17 (a). As observed below the received signals in Fig. 4.17 (b) and Fig. 4.17 (c) are broadened, still a good correlation is observed within the transmitted and received signals. By utilizing the above equation (4.7), the antenna fidelity factor for two orientations is calculated. The Fig. 4.17 shows the normalized wave signals used for calculating the antenna fidelity factor. Table 4.4 gives fidelity factor values for two antenna orientations.

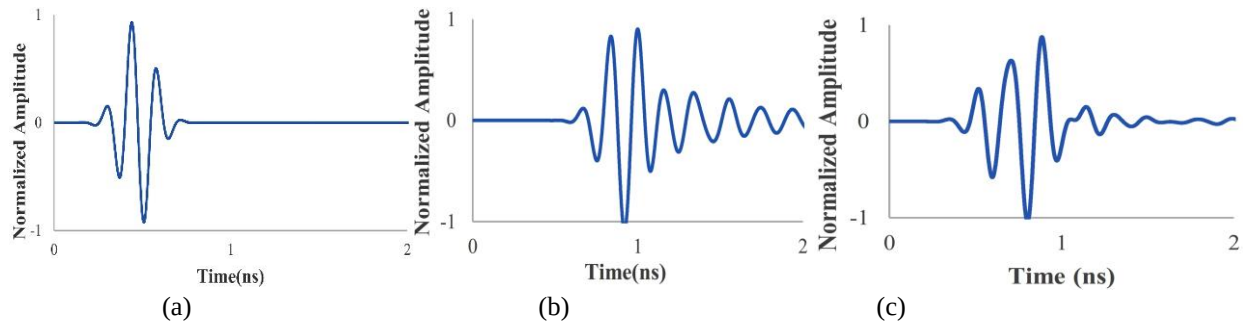


Fig. 4.17 (a) Input excitation signal (b) Received signal (side by side) (c) Received signal (face to face)

Table 4.4 Antenna fidelity factor for two antenna scenarios

Scenarios	Side by Side	Face to face
Fidelity factor	0.9236	0.8805

4.3 SUMMARY

The prominent characteristics of the current chapter are:

1. The designed structure shows overall dimension of $30 \times 30 \times 1.6 \text{ mm}^3$ using FR4 substrate.
2. A set of inverted L and parallel I shaped resonating stubs are utilized as notching structures.
3. The antenna shows broad band stop features in the frequency region covering 3.1 to 5.8 GHz and further demonstrates acceptable performance by generating highest gain, radiation efficiency, wave ratio and attenuation levels of -7 dB, 0.05, 18.7 and -1.07 dB in the rejected band.
4. The designed structure also displays satisfactory passband attributes by generating stable radiation characteristics, reflection coefficient under -10 dB along with highest radiation efficiency and gain values of 0.82 and 5.8 dB.
5. Further the rejection method utilized is quite easy and flexible as bandwidth of notched region can be tuned effortlessly by altering the lengths as well as coupling gaps of the proposed notching structures.

In the next chapter a dual sided broadband metamaterial unit cell is designed for various S, C, X and Ku microwave bands.

CHAPTER 5

METAMATERIAL CELL FOR VARIOUS MICROWAVE BANDS

The current chapter discusses the designing, fabrication and testing of a broadband, novel and conformal DNG metamaterial unit cell. Further, there are different approaches for designing a unit cell such as resonant approach, transmission line approach and hybrid approach. In this chapter the “resonant designing approach” is utilized where two independent resonating structures namely the top layer and the bottom layer are unified together to develop a combined dual sided metamaterial unit cell.

5.1 GEOMETRICAL PROGRESSION OF DESIGNED METAMATERIAL

The designed metamaterial comprises of a single negative (SNG) metamaterial as topmost surface along with a DNG metamaterial as the bottommost surface. The geometrical progression of proposed metamaterial is given in the Fig. 5.1. The impact on the designed metamaterial at various stages of geometrical progression on the scattering parameters (S_{11} , S_{21}), along with the various extracted specifications (ϵ , μ , n) is further analyzed. The ultimate impact of various phases of geometrical progression over scattering parameters along with extracted parameters is displayed in the Fig. 5.2, are also scrutinized to know the supporting science behind and the fundamental metamaterial’s functioning mechanism. Using the principles as well as concepts of these metamaterials, it is widely notable truth that ϵ and μ having negative values generates inductive and capacitive effects.

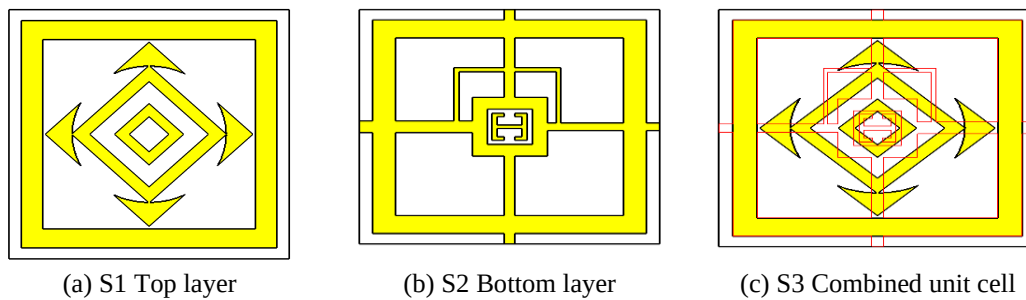


Fig. 5.1 Geometrical progression of designed metamaterial

In the beginning (S1) various resonance metal formations such as a square shaped ring in addition to a centering rhombus shaped design attached to four triangle shaped radiating designs is

created on a conformal FR4 substrate having a small dimension of 12 mm $\times$ 12 mm. The unique layer creates resonant characteristics at 4.6 GHz and 8.3 GHz of frequencies and the band stop features are observed in the frequency region covering 5.1 GHz -7.68 GHz as well as 9.21 GHz - 16 GHz as displayed in the Fig 5.2 (a) and Fig. 5.2 (b), in the S_{21} curve of unit cell. The accompanying ϵ area with negative values in 4.7 to 6.4 GHz and 8.5 to 10.3 GHz of frequency region (Fig. 5.2 (c)) as well as 12.9 GHz -16 GHz of zone with negative values of μ is further analyzed (Fig. 5.2 (d)), but the ultimate improved design does not display DNG features.

At the next phase (S2) various metal formations including a square shape ring combined to a plus shape metal strips in company with an altered H shape resonating structure in middle is produced on another conformal FR4 substrate. The individual layer creates resonant characteristics at 11.4 GHz only (Fig. 5.2 (a)) and shows stopband features in 2 to 10 GHz and 12.1 to 16 GHz of frequency regions (Fig. 5.2 (b)). Accompanying ϵ area with negative values in 2 GHz -16 GHz of frequency region (Fig. 5.2 (c)) as well as 7.9 GHz -11.5 GHz of zone with negative values of μ is found (Fig. 5.2 (d)), further 2.9 GHz -13.5 GHz of frequency region is observed to have negative n values (Fig. 5.2 (e)). The concluding upgraded design offered DNG features in frequency region covering 7.9 to 11.5 GHz.

At last phase (S3) as given in the Fig. 5.1 (c), the separately improved designs including S2 and S1 are finally merged with each other to acquire broader NRI as well as DNG features developing a dual side metamaterial on similar substrate. The final design showed broad pass band features from 10.4 GHz -12.7 GHz and band stop characteristics in the frequency range covering 2 to 4.1 GHz, 4.4 to 7.4 GHz, 8.5 to 9.9 GHz and 14.2 to 16 GHz. It also gave negative ϵ values from 2 GHz -16 GHz along with negative values of μ from 3.8 to 4.3 GHz and 6.8 to 14.5 GHz. The concluding integrated unit cell design showed 4.6 GHz enhancement in the complete bandwidth of DNG region (as seen in the S2 phase) and further produced 13 GHz of broad NRI region.

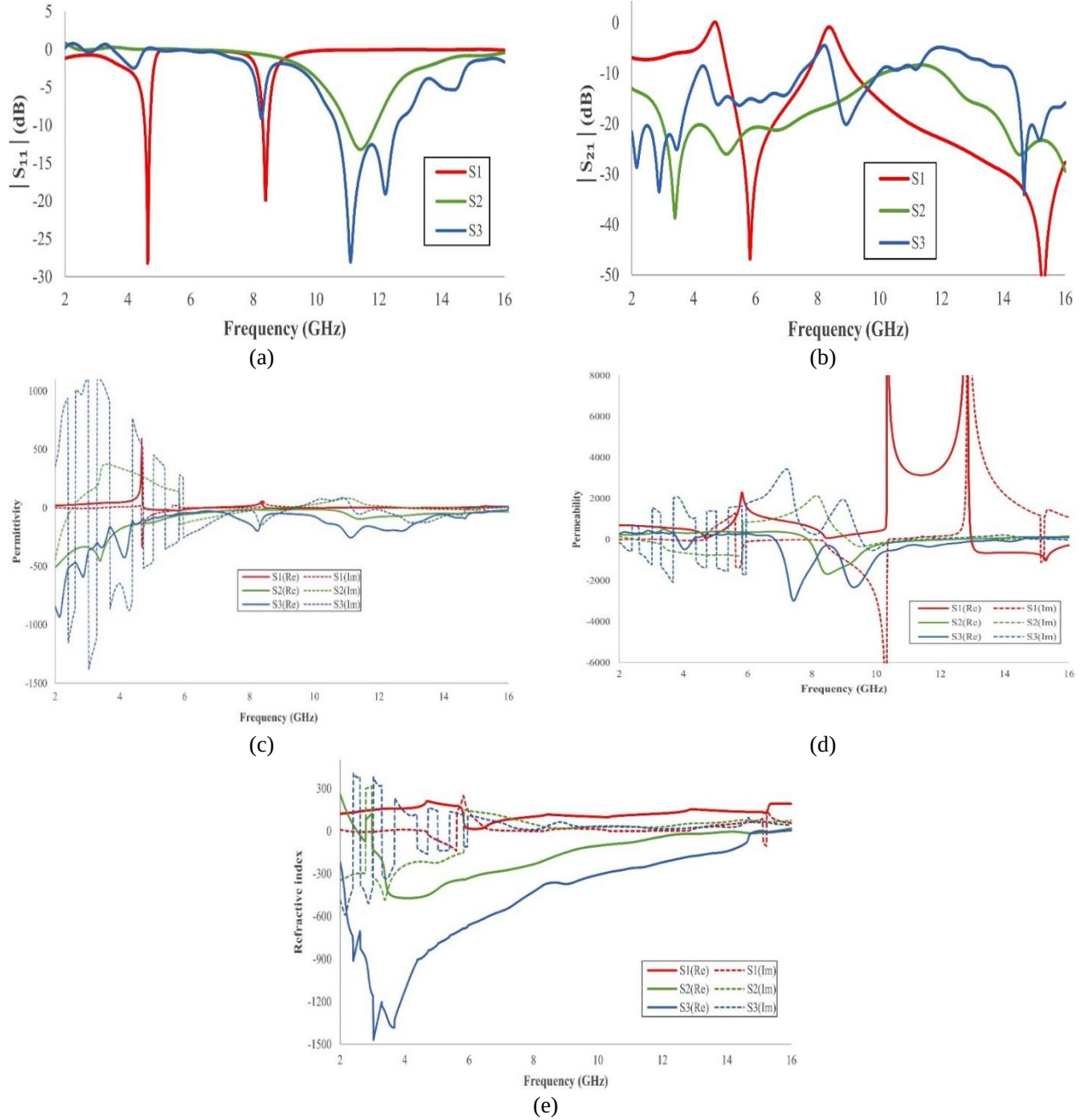


Fig. 5.2 Impact of geometrical progressions on different metamaterial specifications (S_{11} , S_{21} , ϵ , μ and n)

Additionally, to know the broadband NRI features accurately, the complex ϵ and μ of united unit cell is given in the Fig. 5.2 (c) and Fig. 5.2 (d). By carefully observing real portion of the two plots, it is found that from 3.8 GHz -4.3 GHz and 6.8 GHz -14.5 GHz of frequency region both the ϵ and μ parameters are observed to have negative values concurrently, hence resulting in a dual n negative area as observed in the Fig. 5.2 (e). Further, it is seen that NRI area is adequately broader

than total μ negative region, in which dual negative features are found. In the frequency regions covering 2 to 3.7 GHz, 4.4 to 6.7 GHz and 14.6 to 16 GHz, the permeability shows positive values however ε values are observed to be negative in addition negative n values are still observed, simply a single n negative region. This shows a very peculiar characteristics of merged unit cells where a n negative region can be accomplished, in the absence of both ε and μ negative regions, just by analyzing the imaginary and real sections of both extracted parameters including μ and ε [132]. It's found that single n negative area fulfills criteria of $(\mu\varepsilon' < -\mu'\varepsilon)$ hence n negative region is achieved regardless of having negative values of both ε and μ (here ε and μ specifies real components and ε' , μ' shows imaginary components). Besides, as observed in the single negative zone although n negative features can be found, yet large ε' and μ' values result in significant losses, hence not considered for experimental objectives.

The final alterations in various specifications including S_{11} , S_{21} , ε , μ and n along with structural progression at each and every phase are compiled in the Table 5.1. From Table 5.1, it is found that at all steps the inductive and capacitive regions are noticed; yet the NRI and DNG features are produced at S2 phase only. Further the dielectric gaps within resonating structures results in capacitive regions and circulating loops on unit cell surface generates the inductive regions.

Table 5.1 Effective parameters (S_{11} , S_{21} , ε , μ and n) of metamaterial at different phases

Stages	Resonance frequencies in S_{11} plot (GHz)	Stop band regions in S_{21} plot (GHz)	Frequency region with negative ε (GHz) (Capacitive region)	Frequency region with negative μ (GHz) (Inductive region)	Frequency region with negative n (GHz) (NRI region)	Frequency region with both ε and μ negative (GHz) (DNG region)	Bands with DNG characteristics
S1	4.6, 8.3	5.1-7.68, 9.21-16	4.7-6.4, 8.5-10.3	12.9-16	-	-	-
S2	11.4	2-10, 12.1-16	2-16	7.9-11.5	2.9-13.5 (10.6 GHz)	7.9-11.5 (3.6 GHz)	C and X
S3	10.4-12.7 (11.55 as center frequency)	2-4.1, 4.4-7.4, 8.5-9.9, 14.2-16	2-16	3.8-4.3, 6.8-14.5	2-15 (13 GHz)	3.8-4.3, 6.8-14.5 (8.2 GHz)	S, C, X and Ku

5.1.1 Optimized structure of proposed unit cell

The concluding DNG metamaterial comprising both the upmost and bottommost layers are given in the Fig. 5.3 (a) and Fig. 5.3 (b). The corresponding improved structural specifications of proposed cell are given in the Table 5.2. For simulating a metamaterial cell, perfect magnetic as well as electric boundaries are utilized in both the z and x axis along with open boundaries in the y axis. The metamaterial is placed in the middle of waveguide ports along y-axis along with perfect boundary conditions at remaining axis suggesting the propagation of electromagnetic waves through y-axis. The simulating design prototype along with perfect boundaries is displayed in the Fig. 5.3 (c). The FR4 substrate with 4.4 of relative permittivity and 0.25 mm of height is utilized for manufacturing and designing of proposed metamaterial since it is low cost, conformal and well accessible in the marketplace. All the copper components of designed unit cell have thickness of 0.035 mm.

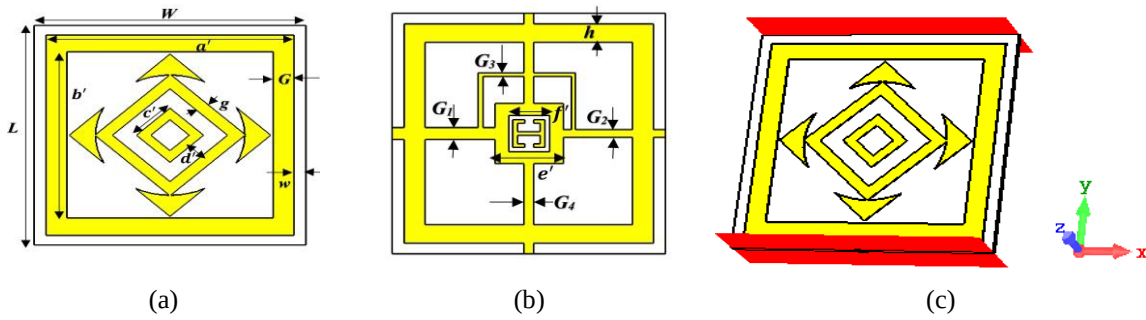


Fig. 5.3 Structure of improved metamaterial (a) Top view (b) Bottom view (c) Simulation prototype along with perfect boundaries

Table 5.2 Enhanced measurements of designed metamaterial

Parameters	Dimensions (mm)	Parameters	Dimensions (mm)
L	12	G_1	0.6
W	12	G_2	0.4
g	0.57	G_3	0.25
w	0.53	G_4	0.44
G	0.92	h	0.9
a'	10.94	b'	9.1
c'	2.4	d'	0.75
e'	3	f'	1.8

It is analyzed from S_{11} plot of designed cell (Fig. 5.4 (a)) that it displays a broad pass band region covering 10.4 GHz to 12.7 GHz (2.3 GHz) of frequency band along with 11.55 GHz of middle

resonant frequency. From analyzing S_{21} curve, it's found that designed metamaterial behaves as broadband band gap structure having 8.3 GHz of a huge stop band region in the RF bands of S, C, X and Ku. The upgraded metamaterial displays stopband characteristics in 2 to 4.1 GHz (2.1 GHz), 4.4 to 7.4 GHz (3 GHz), 8.5 to 9.9 GHz (1.4 GHz) and 14.2 to 16 GHz (1.8 GHz) of frequency regions. Additionally, in the correspondent phase plots of S parameters as described in the Fig. 5.4 (b), at various resonant frequencies the 0° reflection in phase plots can be seen. Now to demonstrate the resonance nature and to recognize the physical characteristics of metamaterial, the electric as well as magnetic field concentration are studied at various resonance frequencies of 2.5 GHz, 7.4 GHz and 15.5 GHz, which are selected from band stop areas of S_{21} plot.

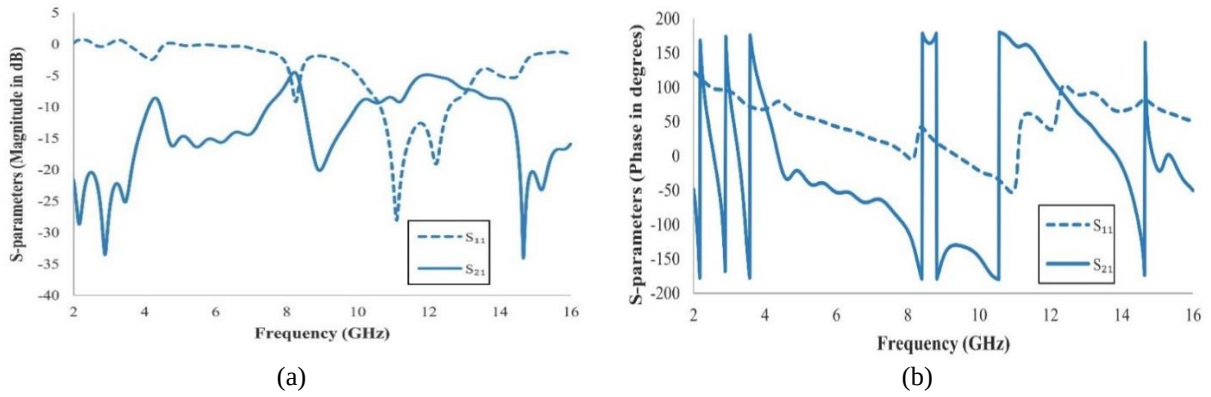


Fig. 5.4 Scattering parameters of proposed unit cell (a) Magnitude plots (b) Phase plots

5.2 INVESTIGATION OF PROPOSED METAMATERIAL

5.2.1 Electric and magnetic field concentration at various resonance frequencies

It can be noticed from Fig. 5.5 that both the upmost and lowest layers of designed unit cell are equivalently responsible for showing the magnetic resonant feature at various resonating frequencies. At resonance frequency of 2.5 GHz the magnetic field distribution at bottom area of metamaterial is larger when contrasted with the higher regions. The density of magnetic field is found to be largest and consistently distributed at 7.4 GHz of resonance frequency and is mostly produced by altered rhombus shape resonance structure at the topmost layer of metamaterial. Further at 15.5 GHz frequency, magnetic field density is localized to few specific regions of the metamaterial, although the distribution of field is larger when contrasted with the field distribution at 2.5 GHz of frequency. The field concentration is found to be largest at 7.4 GHz and smallest at 2.5 GHz in both the layers of unit cell. Hence, it can be concluded that total magnetic field

concentration is achieved mostly because of the self and mutual inductance of various resonating structures existing in both the layers of unit cell.

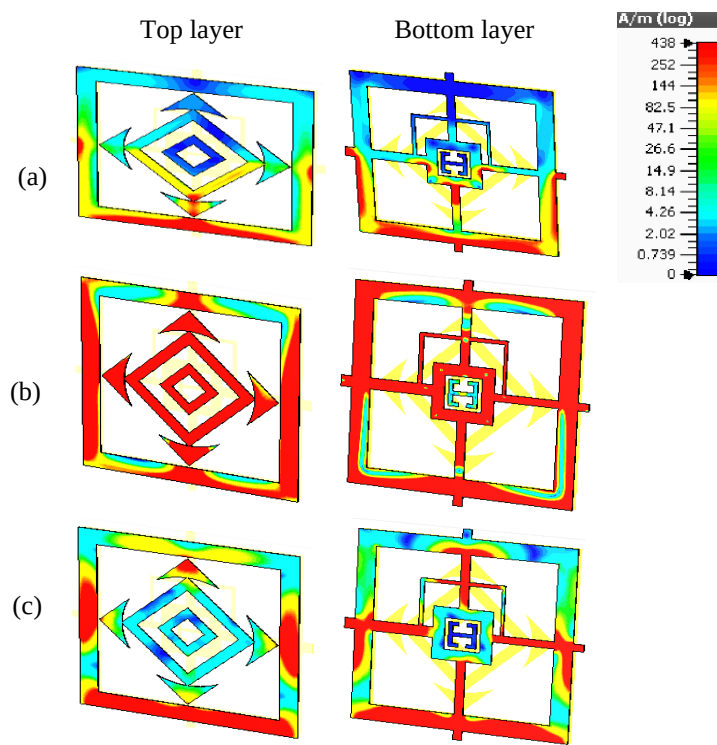
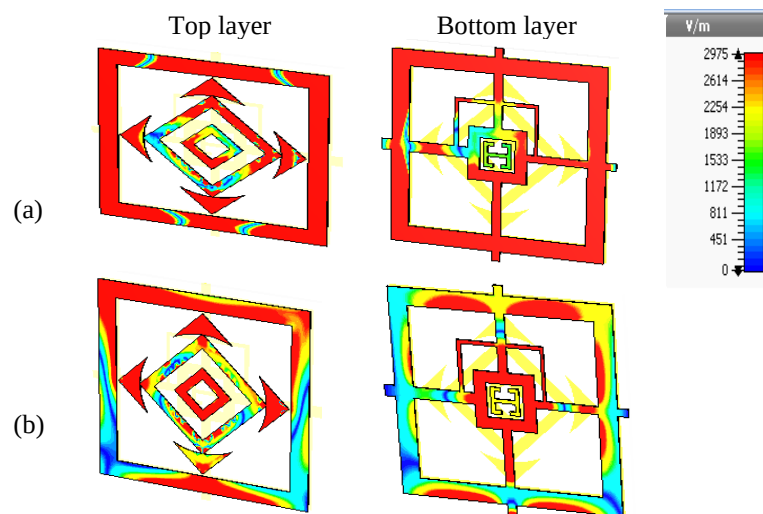


Fig. 5.5 Magnetic field distribution at various resonance frequencies of (a) 2.5 GHz (b) 7.4 GHz
(c) 15.5 GHz



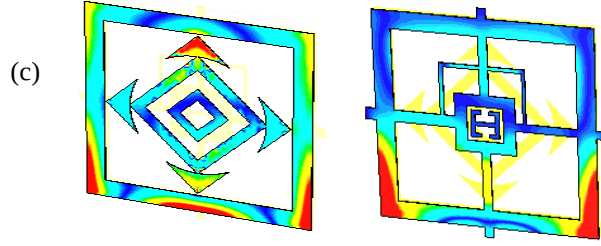


Fig. 5.6 Electric field distribution at various resonance frequencies of (a) 2.5 GHz (b) 7.4 GHz
(c) 15.5 GHz

Electric field concentration at different resonating frequencies including 2.5, 7.4 and 15.5 GHz is displayed in the Fig. 5.6. Similarly, here also both topmost as well as bottommost layers of unit cell are equivalently responsible for electric resonant character of metamaterial, although the distribution of electric field is quite larger in comparison to magnetic field in both layers at various resonance frequencies. The distribution of electric field is found to be largest at 2.5 GHz and is highly intensive in the outermost borderlines of unit cell along with the square shape ring structure. However, the higher field distribution at 7.4 GHz can be seen in few specific regions, that mostly includes the top portion of the metamaterial. The smallest field distribution is observed at 15.5 GHz. Here largest field distribution is seen at 2.5 GHz while the smallest density is seen at 15.5 GHz as seen in the Fig. 5.6 (b) and Fig. 5.6 (c).

By scrutinizing the analyses done above, it can be concluded that the stop bands generated at 2.5 GHz and 15.5 GHz of resonance frequencies is mainly because of electric resonant nature and the stop band at 7.4 GHz is produced because of electric as well as magnetic resonant nature of the metamaterial. The resonance nature found at various frequencies is again validated in section 5.2.3 afterwards, in which negative ϵ having -481.8 and -28.7 values is seen at 2.5 GHz and 15.5 GHz frequency along with simultaneous negative ϵ and μ of -65.7 and -2896.5 values is found at 7.4 GHz of resonance frequency.

5.2.2 Approximated circuital investigation

It is quite evident in the study carried out in the earlier sections that both the topmost and the bottommost layers of the proposed unit cell independently act as the LC resonance circuit, hence the final design of united metamaterial is represented as easy parallel LC resonance circuit having resonant frequency shown underneath in (5.1).

$$f_r = \frac{1}{2\pi\sqrt{LC}} \quad (5.1)$$

Where L and C denotes the equivalent inductance along with capacitance values in the approximated circuit of designed unit cell. As the entire dual sided metamaterial acts as a parallel LC resonance circuit, as a result we are largely concerned in finding the inductance as well as capacitance values in the estimated comparable circuit model and obtaining the best achievable agreement. On the bases of the information discussed above, an estimated circuit model of designed merged metamaterial is displayed in the Fig. 5.7.

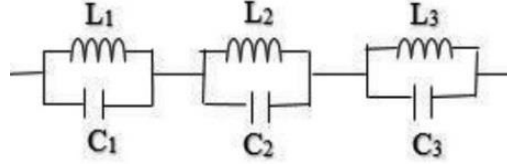


Fig. 5.7 Estimated circuit model of designed metamaterial ($L_1=0.001$ nH, $C_1=14.69$ pF, $L_2=0.393$ nH, $C_2=0.483$ pF, $L_3=0.168$ nH, and $C_3=1.135$ pF)

To find out the LC values in the estimated circuit of designed metamaterial, initially we determined the inductance and capacitance values at these resonant frequencies ($f_1=11.1$ GHz, $f_2=11.5$ GHz and $f_3=12.2$ GHz) by utilizing the input impedance as well as the normalized smith chart curves, else it is also observed that the computed values of LC elements at above frequencies is in accordance with the values obtained using (5.1). The computed LC values are ($L_1=0.014$ nH, $L_2=0.300$ nH, $L_3=0.150$ nH, $C_1=14.69$ pF, $C_2=0.639$ pF, and $C_3=1.135$ pF). Further the estimated circuit model of designed metamaterial is simulated by using a circuit simulator where the computed LC values are upgraded or improved to be in tune with the middle resonance frequency (11.55 GHz) of electromagnetic simulator. The curve obtained through circuit simulation is contrasted with the curve acquired by electromagnetic simulation and a satisfactory agreement within the working bandwidth as well as the middle resonant frequency is found as observed in the Fig. 5.8.

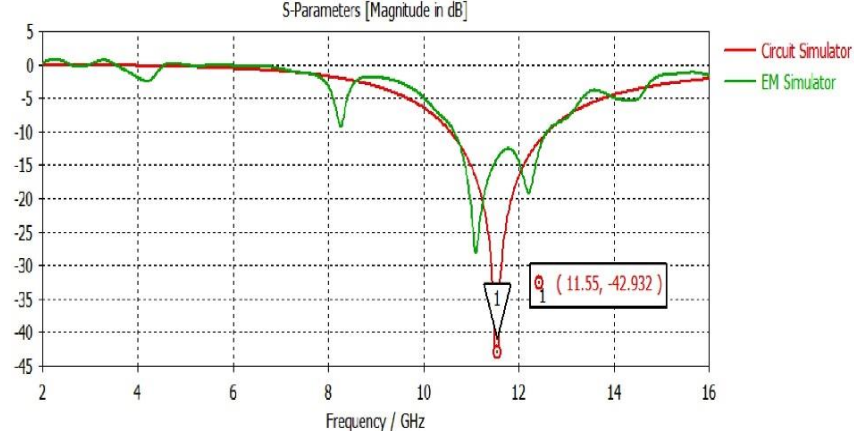


Fig. 5.8 Reflection coefficient plots acquired through both the circuital and electromagnetic simulation

Minute differences are found in both the curves in Fig. 5.8 mainly because of the resonance effects generated due to higher orders which is quite difficult to handle especially for an easy circuit like this. As we mostly focus on resonant frequencies for designing purpose, it is quite irrelevant to build an estimated circuitry with higher order [133].

5.2.3 Characterization of Unit cell

In this section various material characteristics of proposed unit cell including effective permittivity (ϵ), effective permeability (μ), wave impedance (z_m) and index of refraction (n) are derived from the acquired scattering parameters utilizing (5.2), (5.3), (5.4) and (5.5), which are based upon the Scattering parameter retrieval technique [134,135].

$$z_m = \sqrt{\frac{(1+S_{11})^2 - S_{21}^2}{(1-S_{11})^2 - S_{21}^2}} \quad (5.2)$$

$$n = \frac{1}{kq} \cos^{-1} \left[\frac{1}{2S_{21}} (1 - S_{11}^2 + S_{21}^2) \right] \quad (5.3)$$

$$\epsilon = \frac{n}{z_m} \quad (5.4)$$

$$\mu = nz_m \quad (5.5)$$

Here ϵ , n , μ , k , q and z_m denotes the permittivity, index of refraction, permeability, wave vector, substrate height and wave impedance of the metamaterial. Figure 5.9 displays effective parameters (permeability, permittivity along with index of refraction) against frequency curves of the designed metamaterial.

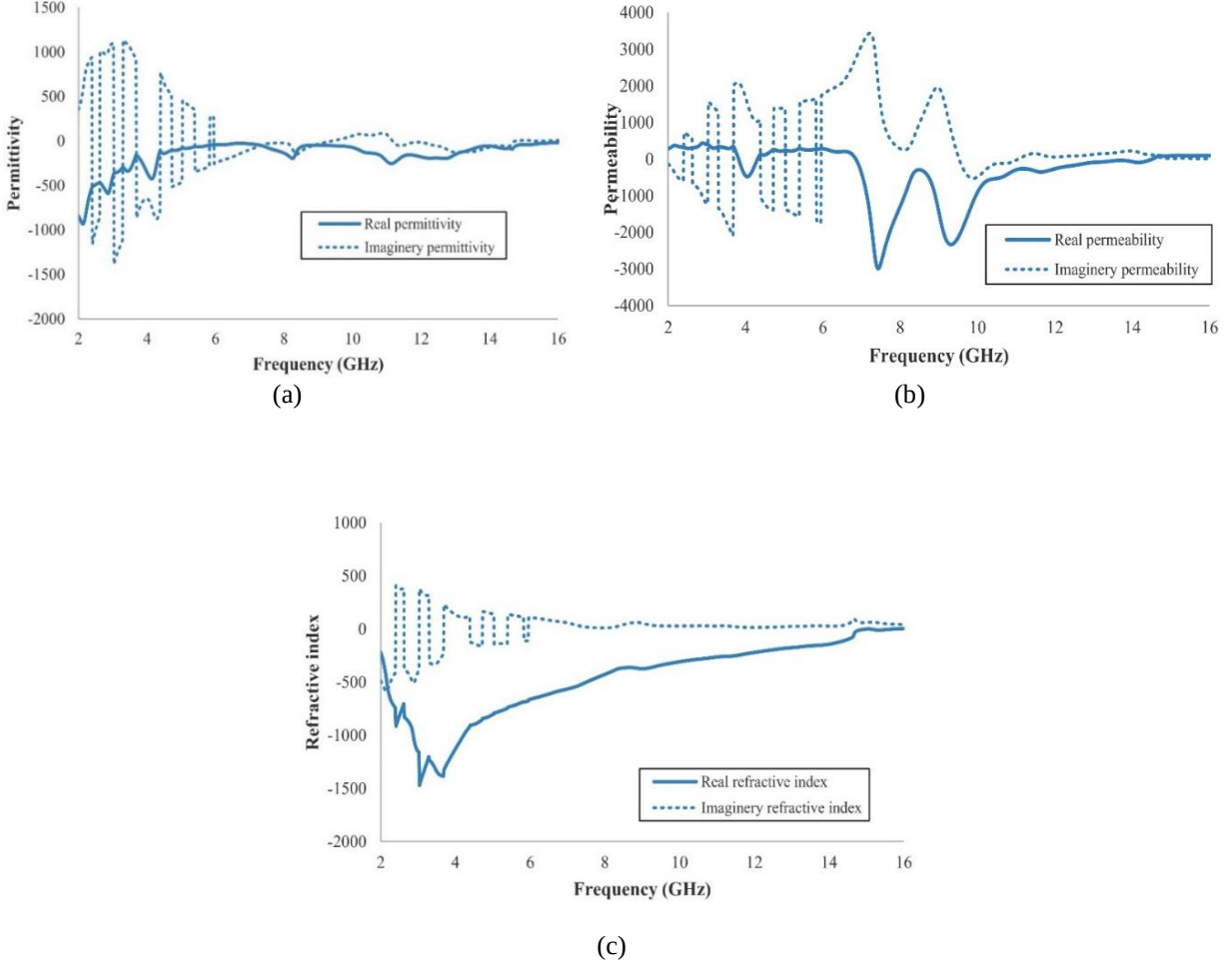


Fig. 5.9 Retrieved characteristics of designed metamaterial showing (a) electric permittivity (b) magnetic permeability (c) index of refraction

The real components of permittivity curve as seen in the Fig. 5.10 (a) display negative values in frequency region from 2 GHz to 16 GHz, hence enveloping the RF bands like S, C, X and Ku. The negative permeability values are observed in the frequency regions covering 3.8 to 4.3 GHz as well as from 6.8 to 14.5 GHz enveloping S, C, X and Ku microwave bands as seen in the Fig. 5.10 (b) also. From Fig. 5.12 (c) the negative n values can be seen in the frequency regions covering 2 GHz to 15 GHz along with S, C, X and Ku microwave bands. Table 5.3 displays effective variables of designed metamaterial together with values of S_{21} at the resonance frequencies of 3.9 and 9 GHz. The specific frequency areas showing negative μ and ϵ values at the same time called DNG regions are found in the Table 5.3. The DNG zone is found in frequency regions from 3.8 GHz to 4.3 GHz and 6.8 GHz to 14.5 GHz, as a result the designed metamaterial generates 8.2 GHz of huge DNG bandwidth covering S, C, X and Ku microwave bands. Further the proposed

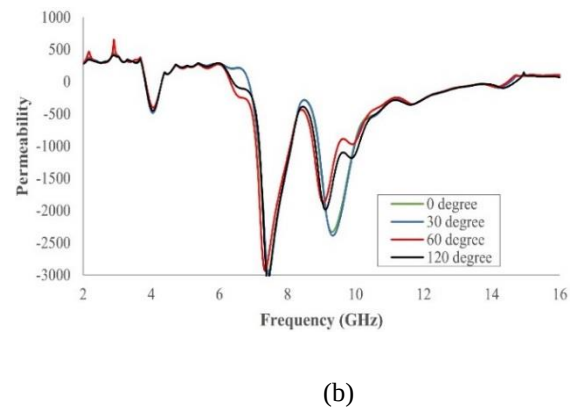
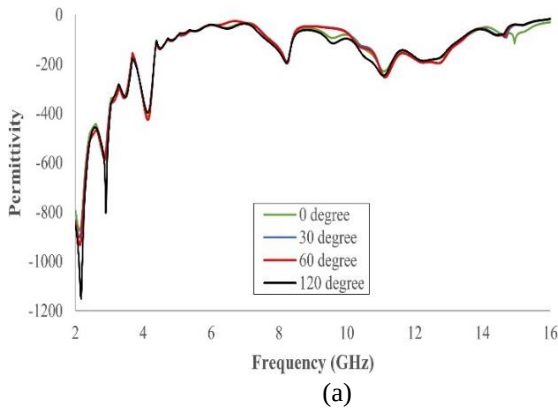
metamaterials display negative values of ϵ , μ and n in distinct frequency bands, hence it's quite noticeable that designed DNG metamaterial shows negative values at 3.9 GHz and 9 GHz for all the effective variables that are displayed in the Table 5.3.

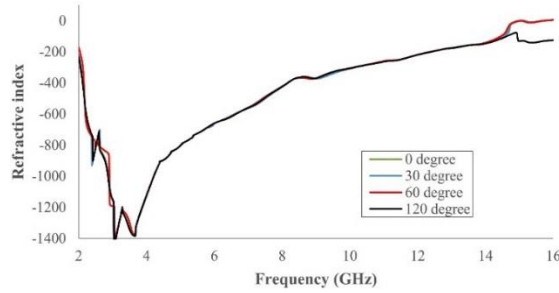
Table 5.3 Real components of three effective variables (ϵ , μ and n) together with S_{21} values

Effective parameters	Frequency range (GHz)	Covered bands	Values at 3.9 and 9 GHz
Transmission coefficient (S_{21})	2-4.1, 4.4-7.4, 8.5-9.9, 14.2-16 (8.3 GHz)	S, C, X, Ku	-13 dB, -20 dB
Effective permittivity (ϵ)	2-16 (14 GHz)	S, C, X, Ku	-346, -50
Effective permeability (μ)	3.8-4.3, 6.8-14.5 (8.2 GHz)	S, C, X, Ku	-356, -1484
Refractive index (n)	2-15 (13 GHz)	S, C, X, Ku	-1176, -373

5.2.4 Bending characteristics of proposed metamaterial

The impact of bending on the metamaterials various effective variables including ϵ , μ , and n , is studied by bending the designed metamaterial on top of a mathematically defined cylinder having 10 mm of fixed radius along with variable bending angle. The bending analysis is performed at three extents of bending namely 0° (no bending), 30° (small bending), 60° (average bending) and 120° (severe bending). It is found in the Fig. 5.10, that as we enhance the extent of bending minute alterations are observed in the permittivity values (still kept in negative region). Moreover, with increasing bending angle the permeability values are found to become more negative. The alterations seen in both permittivity and permeability values with changing bending angle causes variations in the refractive index values as well. It is found that aside from minute alterations in effective variables, the metamaterial is found to preserve the DNG features at various bending scenarios also.



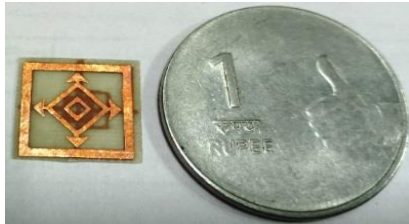


(c)

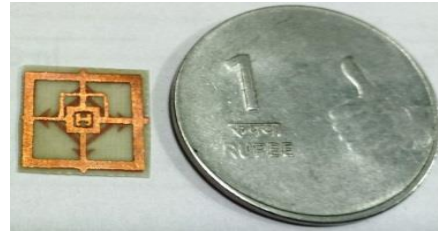
Fig. 5.10 Extracted parameters of designed metamaterial at various bending circumstances (a) effective permittivity (b) effective permeability (c) index of refraction

5.2.5 Fabrication and Measurement

A single metamaterial is manufactured for experimental reasons as described in the Fig. 5.11 (a) and Fig. 5.11 (b). S_{21} curve of designed metamaterial is experimentally measured covering the frequency region covering 2 to 16 GHz by utilizing E5063A VNA. The measurement assembly established for computing the S_{21} values of designed metamaterial are shown in the Fig. 5.12 (a). Further, the various radiating elements such as the patch, $50\ \Omega$ microstrip feedline behaving as feeding transmission lines are inserted on the similar substrate, hence the experimental testing can be performed by utilizing usual research laboratory equipment's in place of waveguides [136].



(a)



(b)

Fig. 5.11 Pictures of manufactured metamaterial (a) Upper layer (b) Lower layer

Measured and the simulated transmission coefficient curves of designed metamaterial display almost same band stop regions as shown in the Fig. 5.12 (b) enveloping the S, C, X and Ku bands. The little shift in the resonating frequencies along with some variations are found in the measured S_{21} plot is mainly because of errors occurred during the manufacturing and measuring processes.

Further, the proposed metamaterial produced acceptable measured outcomes that also confirms the experimental reliability of designed metamaterial.

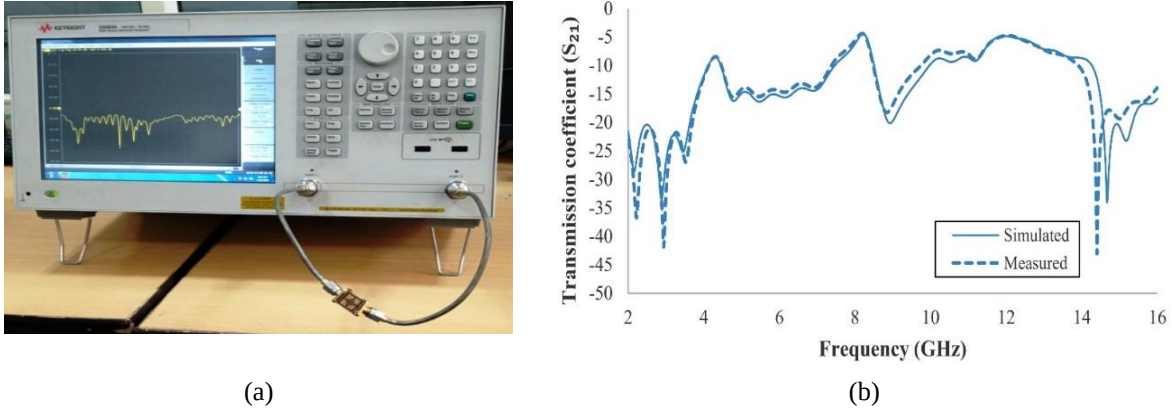


Fig. 5.12 (a) Measurement setup utilized for measuring purpose (b) Measured and simulated transmission coefficients of designed metamaterial

5.2.6 Differentiation with few latest metamaterials

To show the uniqueness of designed metamaterial it's different characteristics like complete volume, conformal nature, microwave bands enveloped, operating frequency range, overall shape together with the NRI and DNG bandwidths are contrasted with some of the latest metamaterial designs. It is found that maximum of the articles discussed have non-conformal nature, huge volume, lesser DNG bandwidths, further the DNG features are found at any two microwave frequency bands only. The proposed unique shape conformal unit cell shows smallest overall volume of 36 mm^3 , maximum DNG and NRI bandwidths of 8.2 GHz and 13 GHz, huge working bandwidth of 14 GHz, unique design and further displays DNG features at S, C, X and Ku microwave bands. It's quite clear from the Table 5.4 that designed metamaterial produced satisfactory outcomes and distinctly outperforms when compared to the latest metamaterial structures, all these confirms the given metamaterial as a strong contender for different S, C, X and Ku microwave region utilizations.

Table 5.4 Differentiation of designed metamaterial with different latest metamaterial designs

Ref.	Volum e size of unit cell (mm ³)	Flexible/ Non- flexible	Frequency regions with NRI characteristics (GHz)	Frequency regions with DNG characteristics (GHz)	Frequency bands with DNG characteristi cs	Working frequenc y range (GHz)	Unit cell shape
[137]	160	Non- flexible	8.3-11.37, 11.8- 15.55, 15.82-18 (8.2 GHz)	8.3-9.5, 11.8-12, 12.15-13.46, 15.53- 17.81(4.99 GHz)	X and Ku	1-18	Unified split P shaped
[138]	230.4	Non- flexible	8.31-15.43, 17.43- 18 (7.7 GHz)	8.8-11.4, 13.59- 13.67, 14.12- 16.86 (5.42 GHz)	X and Ku	1-18	Modified H shaped
[139]	129.6	Non- flexible	8.13-12.14, 13.01- 15.22, 16.73- 16.95 (6.3 GHz)	8.14-9.31, 13.80-15.22 (2.59 GHz)	X and Ku	1-18	Modified theta shaped
[140]	160	Non- flexible	3.48-7.09, 7.87- 10.04, 11.59-14 (8.3 GHz)	8.13-10.37, 13.38-14.0 (2.86 GHz)	C and X	2-14	Improved square Z shaped
[141]	160	Non- flexible	3.4-7.096, 7.8- 10.047, 11.5-14 (8.4 GHz)	8-10, 13.38-14 (2.62 GHz)	X and Ku	2-14	Modified Z shaped
[142]	230.4	Non- flexible	4.4-10.8, 11.3- 13.7 (8.8 GHz)	7.3-8.224, 9.554-10.632, 12.046-12.732 (2.688 GHz)	C and X	1-15	Two C shaped split circular resonators
[143]	115.2	Non- flexible	3.5-6.5 (3 GHz)	2.95-4, 4-4.9, 5- 5.57 (2.52 GHz)	S and C	0.5 -7	Two G shaped split square resonators
[144]	153.66 4	Non- flexible	1.91-7.33, 12.56- 14.30 (7.16 GHz)	6.21-6.39, 7.26- 7.92, 8.01-8.69, 12.17-12.56 (1.91 GHz)	C and X	0-15	Modified hexagonal structure
[145]	54	Flexible	8.29-11 (2.71 GHz)	8.29-10.11, 10.38-10.78 (1.42 GHz)	X	8-11	Novel structure
Propos ed design	36	Flexible	2-15 (13 GHz)	3.8-4.3, 6.8-14.5 (8.2 GHz)	S, C, X and Ku	2-16	Novel structure

5.3 SUMMARY

The prominent characteristics of the current chapter are:

1. A 12×12×0.25 mm³ novel dual sided unit cell is developed using conformal flame retardant 4 (FR-4) substrate.

2. It gave widest DNG and NRI bandwidths along with maximum number of microwave bands covered when contrasted with latest designs.
3. The complete analysis of the designed cell such as electric and magnetic field distributions along with bending and equivalent circuit analysis are done in this chapter.

In the next chapter two flexible broadband metamaterial antennas are designed utilizing an array of metamaterial cells as a reflector layer for reducing the antenna back radiations and gain enhancement.

CHAPTER 6

FLEXIBLE BROADBAND ANTENNAS USING METASURFACES

This chapter discusses the designing, fabrication and testing of the two conformal metamaterial structures i.e. the broadband antennas utilizing an array of metamaterial unit cells as a reflector layer for enhancing the several antenna parameters. There are mainly two methods for using the metamaterials for antenna parameter enhancement. First is to load the unit cells on the antenna design and second is to use the unit cells as a layer at certain distance with the antenna. The second method is utilized in this chapter as it gave higher gain and forward radiation values because of the constructive interference (reflected waves in phase) phenomenon.

6.1 CONFORMAL ANTENNA USING MU-NEGATIVE METAMATERIAL

6.1.1 Wideband antenna structure

The ultimate structure of designed wideband antenna that is experimentally tested and manufactured using FR4 substrate having 4.4 of relative permittivity and 0.15 mm of thickness is displayed in the Fig. 6.1. The improved measurements of proposed structure are also displayed in the Table 6.1. The primary design of proposed structure includes an ellipse shaped patch along with 50 Ω CPW feeding technique. The main motive for using this specific type of design is that elliptical patch antennas along with co planar wave guide feed provide higher operating bandwidth [146, 147] as well as good unification with different circuital modules [148, 149]. Moreover, to acquire wideband characteristics in proposed antenna, various alterations including a slot insertion in the ellipse shaped radiating patch and slanting the uppermost corner edges of designed CPW ground layer are also performed. The step by step structural progression of designed wideband structure is shown in the Fig. 6.2, however the corresponding reflection coefficient plot of above design is displayed in the Fig. 6.3. At initial phase, an ellipse shaped radiating patch having CPW feed (Fig. 6.2 (a)) generates three distinct working frequency regions, first region covering 3 to 4.5 GHz is created because of resonance characteristics at 3.5 GHz, next region covering 5.3 to 11.9 GHz is produced as a result of amalgamation of resonance frequencies at 6.5 and 11.2 GHz, however last region covering 13.5 to 16 GHz is mainly because of resonant features at 15.4 GHz as seen in the Fig. 6.3 also. At next phase (Fig. 6.2 (b)) alterations in radiating patch, leads to the integration of two bands at lower frequency regions and increment in operating bandwidth at upper

frequency regions as observed in the Fig. 6.3. At last and final phase, the slanting of uppermost corner edges of designed CPW ground layer (Fig. 6.2 (c)) causes increment in antenna bandwidth and a large working bandwidth covering 2 GHz to 16 GHz along with good matching characteristics at 3.5, 5.8, 10.2, 14.6 GHz of distinct resonant frequencies is displayed in the Fig. 6.3. Designed wideband structure is manufactured and corresponding reflection coefficient curve is experimentally measured using the KEYSIGHT E5063A (100 KHz -18 GHz) vector network analyzer along with electromagnetic chamber. The photograph of manufactured structure is given in the Fig 6.4 (a), further the measured and simulated reflection coefficient curves of the structure is displayed in the Fig. 6.4 (b).

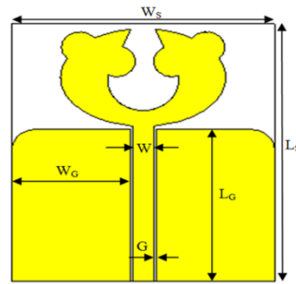


Fig. 6.1 Ultimate structure of designed wideband antenna

Table 6.1 Improved measurements of designed wideband antenna (mm)

Parameters	FR4
L_S	50
W_S	43
L_G	29.55
W_G	19
W	3.54
G	1

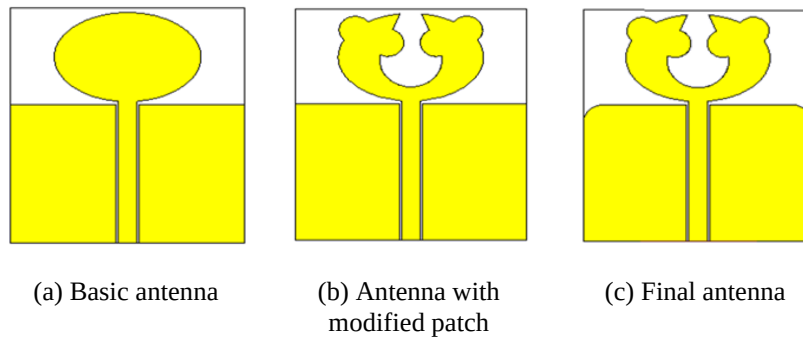


Fig. 6.2 Structural progression of designed wideband antenna

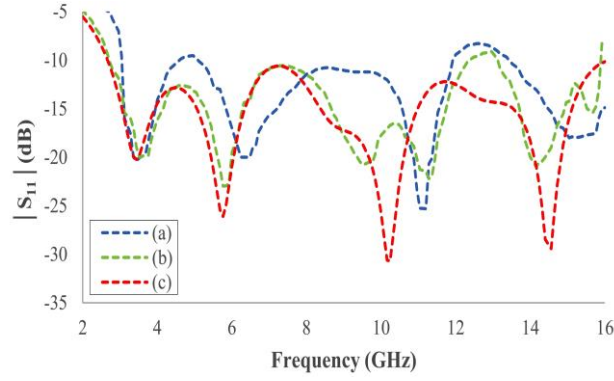


Fig. 6.3 Difference between the reflection coefficient curves of proposed structure in correlation with the Fig. 6.2

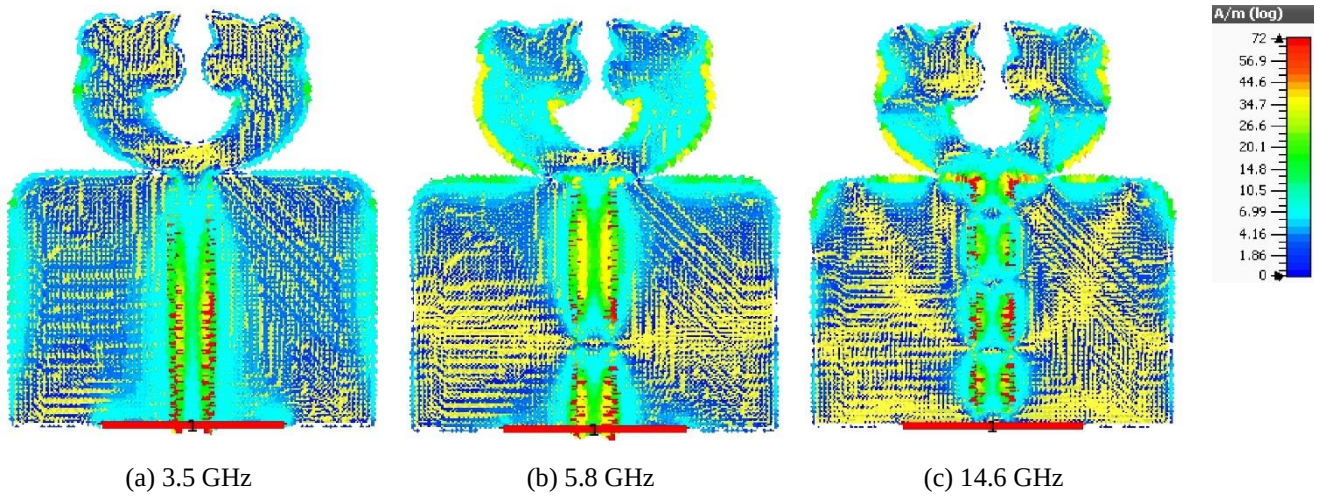
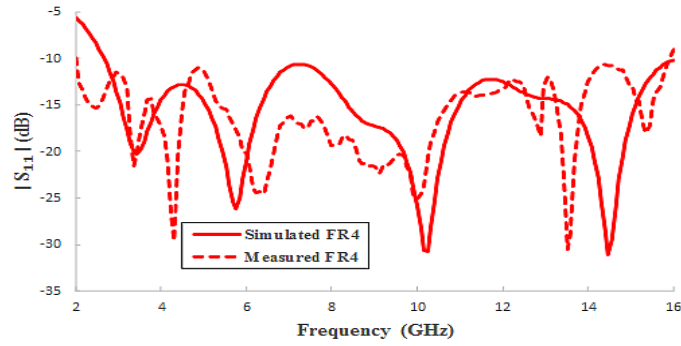


Fig. 6.4 Surface current distribution at three resonance frequencies

Figure 6.4 shows the simulated surface current distribution of proposed antenna at three of the resonance frequencies (3.5, 5.8 and 14.6 GHz). At the lowest frequency of 3.5 GHz, it is observed that the current distribution is stable and uniform across the entire antenna as compared to higher frequencies. However consistent higher concentration of current at the feedline can be seen at all the three frequencies resulting in higher electric fields. Now at the middle frequency the concentration of current has increased in the outer periphery of antenna with reduced concentration in the central area as compared to 3.5 GHz. At highest frequency of 14.6 GHz, it is clearly visible that it has maximum current concentration at the outer boundaries of antenna structure. Hence it can be concluded that at lower frequencies the current is uniformly distributed however as the frequency increases the current gets confined in the outer boundaries of the designed structure.



(a)



(b)

Fig. 6.5 (a) Picture of manufactured wideband antenna (b) Measured and the simulated reflection coefficient curve of manufactured antenna

As observed from the Fig. 6.5 (b), the wideband antenna using FR4 substrate shows wide simulated operating bandwidth covering 2.6 to 16 GHz of frequency range, along with measured operating bandwidth covering 2 to 15.9 GHz of frequency band. It's evident from curves discussed earlier that both simulated as well as measured outcomes are quite same along with minute shifting in lower limiting frequency because of manufacturing inaccuracies as well as computational intolerances.

6.1.2 Structural evolution of unit cell

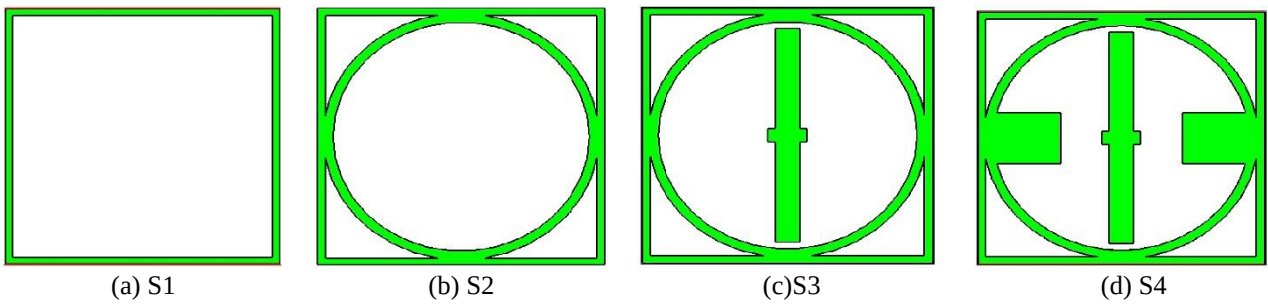


Fig. 6.6 Structural progression of designed unit cell

Initially (S1), a $14 \times 14 \text{ mm}^2$ sized square shaped copper patch with thickness 0.035 mm is proposed on FR-4 substrate as shown in Fig. 6.6 (a). It gave a single passband frequency at 5.9 GHz (Fig. 6.7 (a)). Further no mu-negative region was found at this stage of designing (Fig. 6.7 (b)).

In next stage (S2), a concentric ring with 7.1 mm and 6.6 mm of outer and inner radius is etched out from the square patch as observed in Fig. 6.6 (b). This results in two passband frequencies at 2.9 GHz and 10.3 GHz (Fig. 6.7 (a)). Moreover, the ring structure also generated a mu negative region from 9.5 GHz to 11 GHz (1.5 GHz) as observed in Fig. 6.7 (b).

In S3 stage, a plus shaped slot is inserted in the middle of the square patch which results in the shifting of lower resonance frequency obtained at S2 stage shown in Fig. 6.6 (c). The two passband frequencies are now observed at 5.5 GHz and 10.3 GHz. Further it also shifted the mu negative region to lower UWB range i.e. from 8.5 GHz to 10.2 GHz (1.6 GHz).

In the last stage (S4) as shown in Fig. 6.6 (d), a pair of U-shaped rectangular slots are inserted which gave two passband frequencies at 4.8 GHz and 9.6 GHz along with a mu negative region from 7.2 GHz to 9.2 GHz which is comparatively larger than the MNG regions obtained at S2 and S3 stages. Figure 6.7 shows the effect of various structural variations of unit cell on its reflection coefficient and permeability curves.

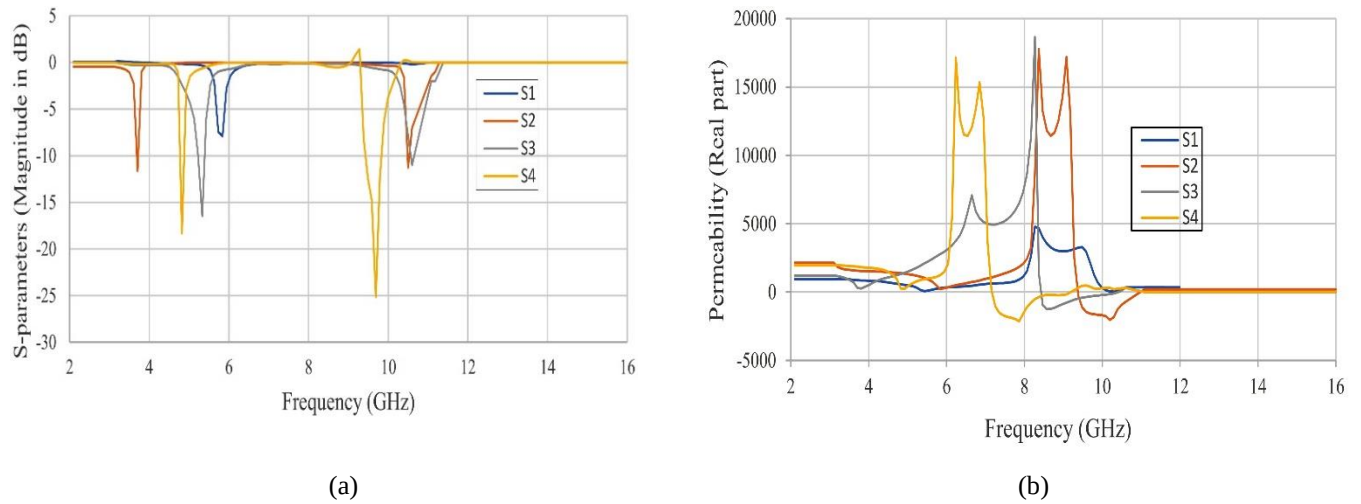


Fig. 6.7 Effect of unit cell variations on (a) S11 (b) Permeability

6.1.3 Design of mu-negative unit cell

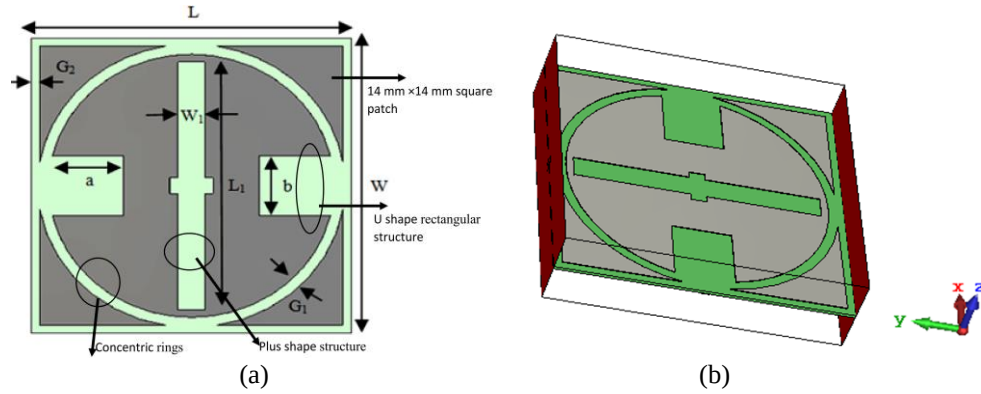


Fig. 6.8 (a) Upper layer of designed unit cell (b) The simulated model using PM and PE boundaries

The wideband structure discussed in the above section is designed without any ground layer at the backside as a result the physical body remains unprotected by the dangerous radiations emitted. Here we analyzed as well as manufactured an easy and novel structured conformal unit cell that not only increases the proposed antenna's forward radiation features and gain properties but also results in reduced backside radiations from designed wideband antenna. The designed unit cell design utilizes similar conformal FR4 substrate, that we utilized earlier to design the basic wideband antenna. The designed metamaterial structure contains a $14 \times 14 \text{ mm}^2$ sized square shape radiating patch in which various structures including a concentric ring with 7.1 mm and 6.6 mm of outer and inner radius, a plus shape structure in the middle along with a pair of U shape rectangular structures that are inserted close to the center plus shape structure. The upper layer of designed unit cell together with simulating model using perfect magnetic and electric boundaries is displayed in the Fig. 6.8 (a) and Fig. 6.8 (b). Further various improved measurements of proposed antenna in millimeter are: $L = 14.8$, $W = 14.8$, $G_1 = 0.5$, $G_2 = 0.42$, $L_1 = 12.45$, $W_1 = 1.24$, $a = 3.4$, $b = 3$. It is analyzed that the initial resonance frequency at 4.8 GHz is created mainly because of square shape radiating patch and the other resonance frequency at 9.6 GHz is produced due to various slots. The complete characterization of designed unit cell is partitioned in two distinct phases. At starting phase, the designed metamaterial is simulated to obtain scattering parameters (S_{11} and S_{21}). In the next phase the various effective material characteristics such as permeability (μ), permittivity (ϵ), wave impedance (z_m) and refractive index (n) are extracted employing the above-mentioned scattering parameters with the help of (6.1) as well as (6.2). Each and every structure in the metamaterial has a very crucial part in developing various distinct

features in the designed unit cell. The magnitude (dB) as well as phase (degree) plots from obtained scattering parameters of unit cell are displayed in the Fig. 6. 9 (a) and Fig. 6.9 (b). It's clear from S_{11} plot, that designed metamaterial demonstrates resonance characteristics at 4.8 GHz and 9.6 GHz of two distinct pass band frequencies. Moreover, from S_{21} plot, its analyzed that proposed metamaterial behaves as a double broadband bandgap material by displaying two different broadband stop regions covering 5 to 9.2 GHz and 10.3 to 16 GHz of frequency regions along with -49 and -42 dB of smallest S_{21} values obtained at 7.87 and 15 GHz of resonance frequencies. The corresponding S_{11} and S_{21} phase plots of designed unit cell are displayed in the Fig. 6.9 (b) in which 0° reflection in phase is found at 7.87 GHz.

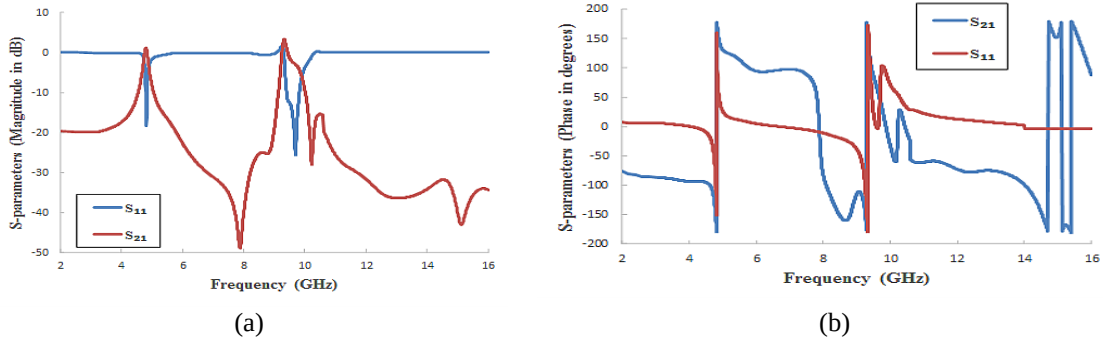


Fig. 6.9 Scattering characteristics of proposed metamaterial showing (a) Magnitude in dB (b) Phase in degrees

The complete material specifications (ϵ, μ, n, z_m) of designed unit cell, are obtained from scattering parameters by utilizing the (6.1) and (6.2) given below [150, 151].

$$z_m = \sqrt{\frac{(1+S_{11})^2 - S_{21}^2}{(1-S_{11})^2 - S_{21}^2}} \quad (6.1)$$

$$n = \frac{1}{kq} \cos^{-1} \left[\frac{1}{2S_{21}} (1 - S_{11}^2 + S_{21}^2) \right] \quad (6.2)$$

Here, q and k denote the substrate height along with wave vector. The effective permeability and permittivity are calculated utilizing these two formulas i.e. $\mu = nz_m$ and $\epsilon = \frac{n}{z_m}$.

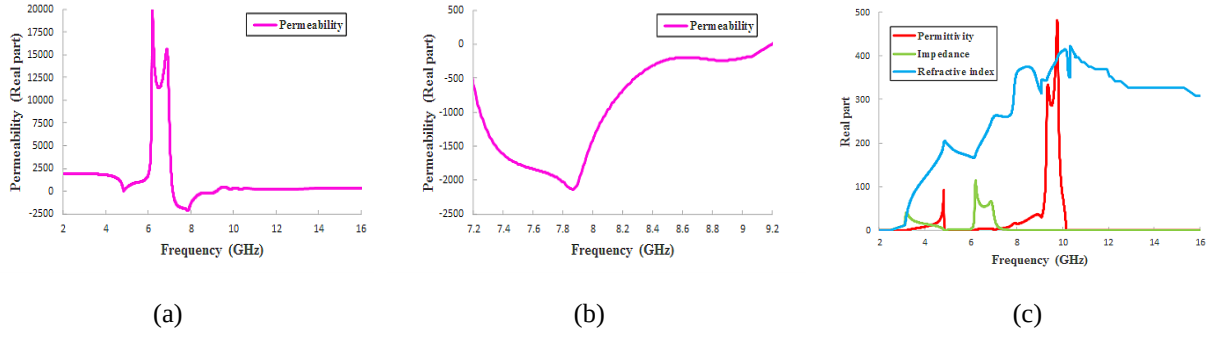


Fig. 6.10 (a) The effective permeability of designed metamaterial (b) Real component covering 7.2 to 9.2 GHz of frequency region (c) Material specification of unit cell

Now it is analyzed from the Fig. 6.10 (a) that real component of effective permeability (μ) displays magnetic lorentz characteristics and shows negative components from frequency region covering 7.2 to 9.2 GHz of frequency. The plot obtained in the Fig. 6.10 (a) is focused in the range from 7.2 to 9.2 GHz as displayed in the Fig. 6.10 (b), it's found that the negative as well as highest effective permeability value of -2121.6 is seen at resonance frequency of 7.87 GHz, that is precisely the similar frequency where smallest S_{21} value along with accurately 0° reflection in phase plot is observed in the Fig. 6.9 (a) and Fig. 6.9 (b). Moreover, real components of all effective parameters like refractive index (n), permittivity (ϵ), and wave impedance (z_m) are observed to be positive across complete working frequency region as seen in the Fig. 6.10 (c). Hence, it is analyzed from observed outcomes that in various band stop regions of S_{21} plot, the designed unit cell shows perfect reflection to the electromagnetic waves.

6.1.4 Unified antenna structure

This section discusses, the ultimate unified antenna proposed by combining the wideband antenna developed in the section 6.1.1 together with an array of mu negative metamaterial as examined earlier in section 6.1.2. The 3×3 array of mu negative metamaterial is chosen as the optimum array size to achieve maximum outputs as shown in the Fig. 6.8 (a), as it decently shields the entire area of designed wideband antenna and further the complete antenna volume requires to be compact in size so that it can be suitable for wearable utilizations. To achieve best performance from the designed unified antenna, the mu negative metamaterial layer cannot be positioned directly over the antenna structure without any intermediate space. Therefore, a fixed air space “ d ” or gap needs to be established in between the metamaterial layer and antenna structure and in

practical scenarios the foam substrate is utilized to obtain the desired air separation. The impact of alterations in the value of air space “ d ” ranging from 2 mm to 20 mm on the unified antenna’s (antenna along with metamaterial layer) S_{11} values is displayed in the Fig. 6.11 (b). It’s found that at 2 mm of air space value, the unified metamaterial antenna is not able to cover the complete ultrawideband frequency range. With increasing “ d ” value, the overall antenna bandwidth is also increased and further at air space value of 14 mm, the unified antenna covers the complete ultrawideband region covering 2.3 to 16 GHz. Further as we increased the air space above 14 mm, the acquired operating bandwidth remains unaffected. Hence it is observed that by increasing the air space value above 14 mm simply results in the increment in overall dimensions and hugeness of the unified antenna. Further it is found that keeping the metamaterial layer in close proximity to wideband antenna develops impedance mismatch conditions and however as we place metamaterial layer a greater distance from the unified antenna, the better matching conditions are achieved. At last, it is concluded that the best possible outcomes are achieved by using a 3×3 element metamaterial layer along with air separation value of 14 mm. The final manufactured unified antenna using expanded polystyrene (EPS) foam for providing air separation value of 14 mm is displayed in the Fig. 6.12 (a).

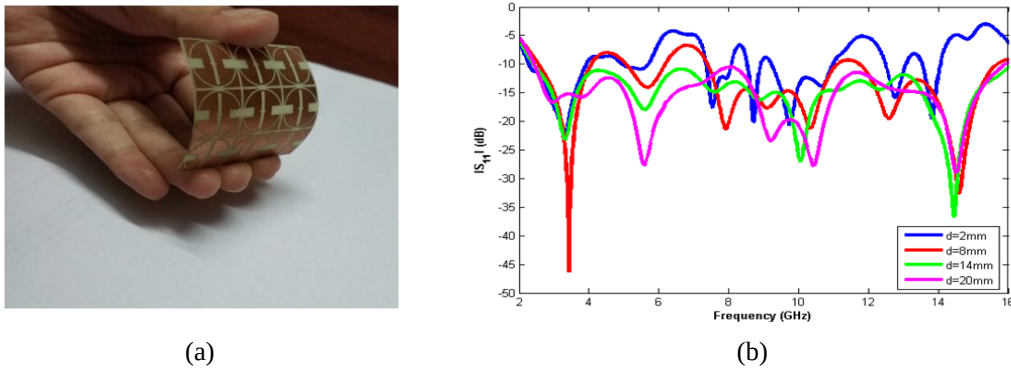


Fig. 6.11 (a) A 3×3 metamaterial layer (b) Impact of modifying air space “ d ” value on reflection coefficient curve of unified structure

Simulated reflection coefficient plots of both the wide band antenna along with the unified antenna and the measured reflection coefficient plot of unified antenna are displayed in the Fig. 6.12 (b). The unified antenna is observed to envelope the wide frequency ranges from 2.3 to 16 GHz along with satisfactory matching conditions (under -15 dB) amongst its various resonance frequencies of 3.4, 5.7, 10.2 and 14.5 GHz. Moreover, measured bandwidth plot is observed to enhance in the lower frequency region together with slight shift in the reflection coefficient values in the complete

working frequency range because of manufacturing, positioning and experimental inaccuracies. The simulated gain plots for both the wideband as well as unified structure together with measured gain plot of unified structure are shown in the Fig. 6.13. Gain values acquired at various resonance frequencies for the wideband antenna as well as the unified antenna are displayed in the Table 6.2. It is found that the metamaterial layer enhanced the gain values obtained at various resonance frequencies. At 3.4, 10.2 and 14.5 GHz of frequencies, above 2 dB gain enhancement is found, however at 5.7 GHz of resonance frequency, 1.8 dB gain enhancement is observed. Moreover, both the measured as well as simulated gain curves of proposed unified antenna are almost same with minute alterations.

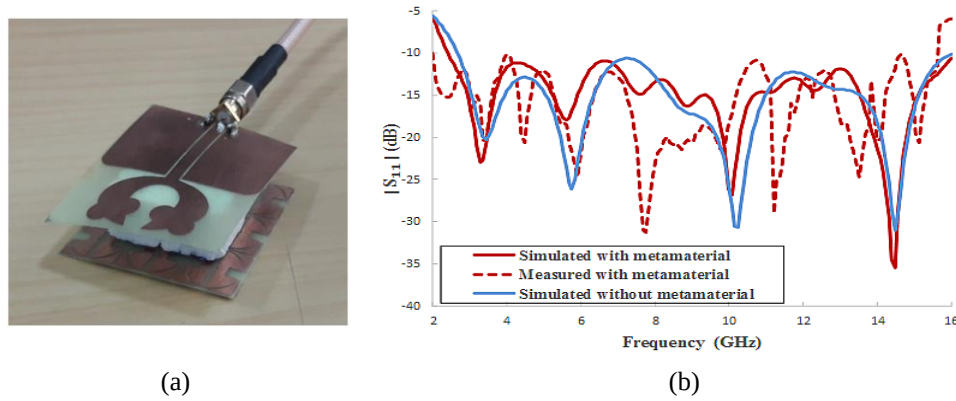


Fig. 6.12 (a) Manufactured unified antenna along with air gap of 14 mm (b) Reflection coefficient curve of designed wideband antenna without/with metamaterial layer

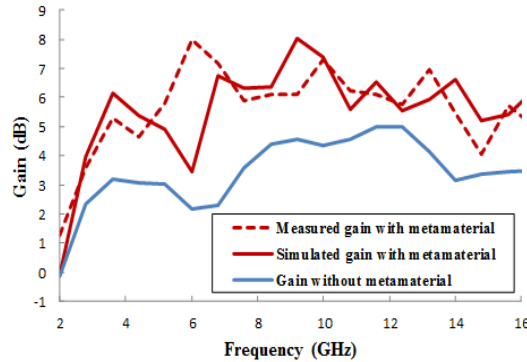


Fig. 6.13 Gain curve of designed wideband antenna without/with metamaterial layer

Table 6.2 Obtained values of gain for both the wideband and unified antenna

Frequencies (GHz)	Gain values for UWB antenna (dB)	Gain values for Integrated antenna (dB)
3.4	2.7	5.1
5.7	3.1	4.9
10.2	4.3	7.2
14.5	3	6.5

6.1.5 Analysis of wideband and unified wearable structures

6.1.5.1 Impact of bending analysis

In this section the wideband antenna along with the unified antenna in simulation model are placed on the cylindrical shape structure for analyzing its bending characteristics, however in practical scenarios the bending features are studied for the unified antenna structure only. The proposed structures are kept on a cylindrical surface with 10 mm of fixed radius and further the angle of bending is altered in the range of 30° to 180° to examine its impact on the S_{11} plots of proposed antennas. The bending characteristics of proposed antennas at 60° of distinct bending angle is displayed in the Fig. 6.14 (a) and Fig. 6.14 (b). Further impact of bending on S_{11} curves of proposed structures is shown in the Fig. 6.15 (a) and Fig. 6.15 (b). A uniform shifting in various resonant frequencies in the direction of lower frequency regions at different angles of bending is found as seen in Fig. 6.15 (a), however shifting in resonant frequencies is not observed in the Fig. 6.15 (b). Also, minute alterations in S_{11} values for both curves is found across the complete ultrawideband region.

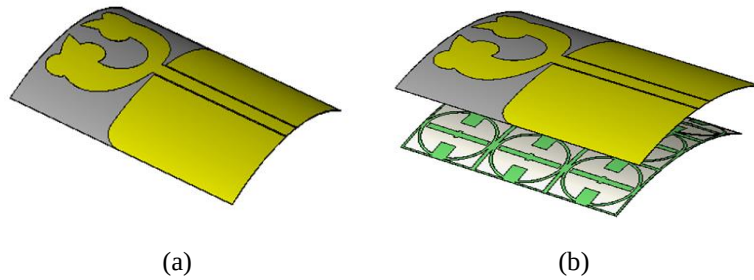


Fig. 6.14 Bending characteristics at bending angle of 60° (a) wideband structure (b) Unified structure having 14 mm of air space

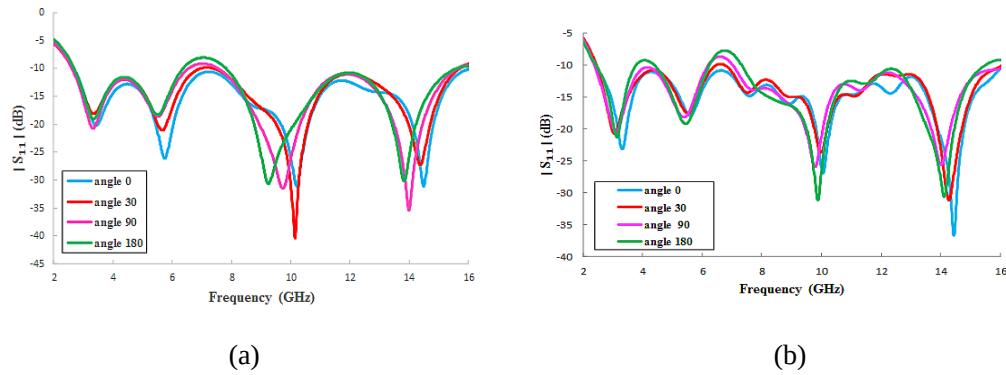


Fig. 6.15 Impact of bending on reflection coefficient curves of the (a) wideband structure (b) Unified structure

The measurement setup utilized for measuring the bending characteristics of proposed antenna is displayed in the Fig. 6.16 (a), here the unified antenna is bent on the cylindrical shaped foam structure. Further the measured S_{11} plots of the unified antenna without/with bending are displayed in the Fig. 6.16 (b). It is found that the measured reflection coefficient plot of proposed structure is almost same with the simulated plot, however minute shift in larger resonant frequency values is seen because of manufacturing inaccuracies.

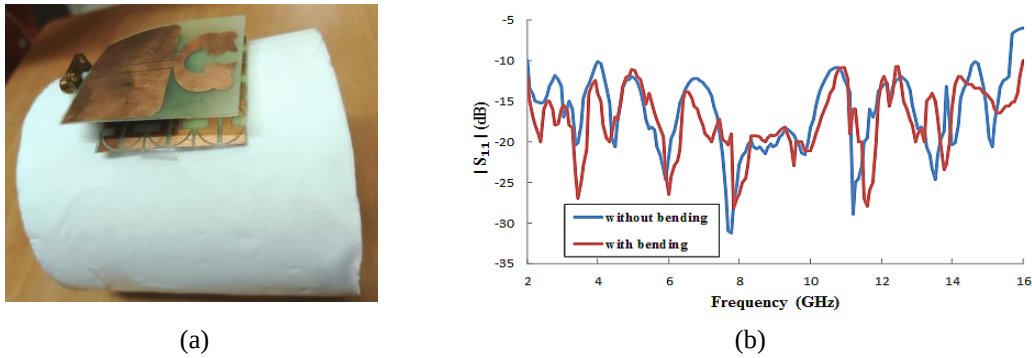


Fig. 6.16 (a) Unified antenna bent on the cylinder-shaped foam structure (b) Measured S_{11} curve of unified structure without/with bending

6.1.5.2 Human body impact

The physical body comprises of multiple layers that are developed due to various human organs like bones, stomach, skin, muscles and so on. Every distinct layer possesses its peculiar frequency

variable dielectric features as well as thickness. On the basis of the analysis presented by the Italian National Research Council, various electric specifications of human tissues at various resonant frequencies of 3, 5 and 10 GHz are displayed in the Table 6.3, that are effortlessly accessible in the internet [152]. To demonstrate the human body impact on various parameters of designed antenna, a multi layered human model comprising skin, fat and muscle tissue is designed as well as analyzed in the CST microwave studio. The proposed antenna features are analyzed in the existence of both the flat and cylindrical human model using skin, fat and muscle tissues with average width of 2, 5 and 10 mm. The wideband antenna using both cylindrical and flat phantoms at separation of 2 mm is displayed in the Fig. 6.17 (a) and Fig. 6.17 (b).

Table 6.3 Electrical features of body tissue at various resonant frequencies

Tissue name	Frequency (GHz)	Conductivity (S/m)	Effective Permittivity (ϵ)	Loss tangent (δ)
Skin (Dry)	3	1.7406	37.45	0.27848
	5	3.0608	35.774	0.3076
	10	8.0138	31.29	0.46038
Fat	3	0.13004	5.2239	0.14916
	5	0.24222	5.0291	0.17315
	10	0.58521	4.6023	0.22857
Muscle	3	2.1421	52.058	0.24655
	5	4.0448	49.54	0.29353
	10	10.626	42.764	0.44666

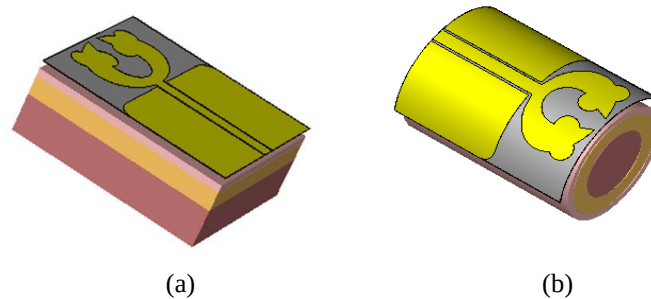


Fig. 6.17 Designed wideband antenna placed on multilayered human phantom (a) Flat model (b) Cylindrical model

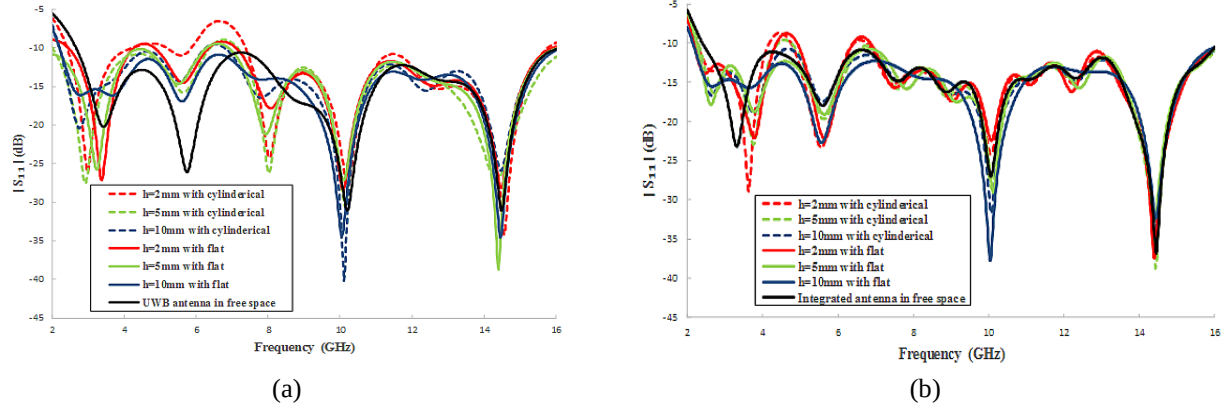


Fig. 6.18 Impact on reflection coefficient curves of both antennas when using cylindrical and flat human models (a) wideband antenna (b) unified antenna

Here the separation “ h ” within the proposed antenna and multilayered human models is altered over the cylindrical as well as flat phantoms to examine their counterpart impact on the S_{11} values of the designed antenna. For assessment motive the following three distances are chosen such as 2, 5 and 10 mm. It is observed that once the wideband antenna is kept on the human models as observed in the Fig. 6.18 (a) the enhancement in reflection coefficient values is seen in the lower frequency regions under 8 GHz, however over 8 GHz only minute alterations in reflection coefficient values is found. Further same trend in reflection coefficient plot is observed in case of unified structure also as displayed in the Fig. 6.18 (b), although extent of enhancement in S_{11} values is comparably little. Moreover, it is found that both antenna structures using flat human models generates smaller S_{11} values along with satisfactory matching features across the entire ultrawideband region when contrasted with cylindrical human models. As a result, the multilayered flat models are preferred more for carrying out future analysis. The measurement set up utilized for measuring the reflection coefficient values of unified antenna along with physical body is displayed in the Fig. 6.19 (a) and Fig. 6.19 (b). The measured S_{11} plot of unified structure without/with physical body is displayed in the Fig. 6.20. A good match between the simulated and measured S_{11} plots is found till 6 GHz, but over 6 GHz, a uniform shift in resonant frequencies in the direction of upper frequency region is seen. Impact on different antenna properties such as radiation efficiency and gain are also studied in the Table 6.4 in two scenarios, when both antennas are kept on flat human model and in the free space at three distinct distances.

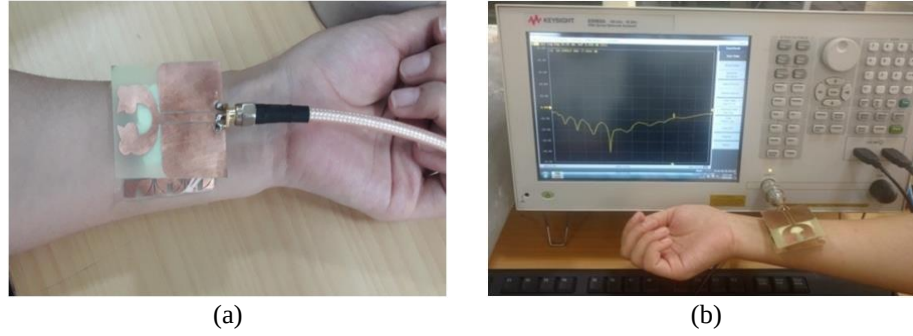


Fig. 6.19 Measurement setup utilized for measuring S_{11} values over physical body

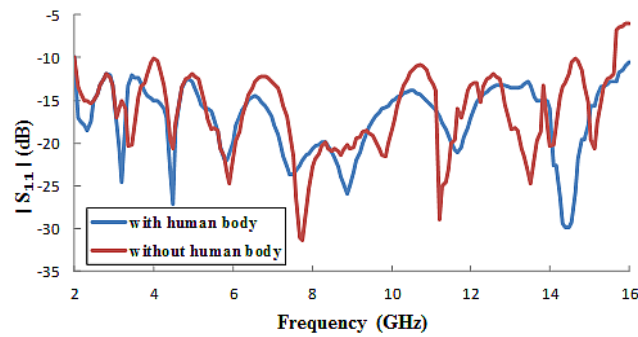


Fig. 6.20 Measured reflection coefficient plot of unified antenna without/with physical body

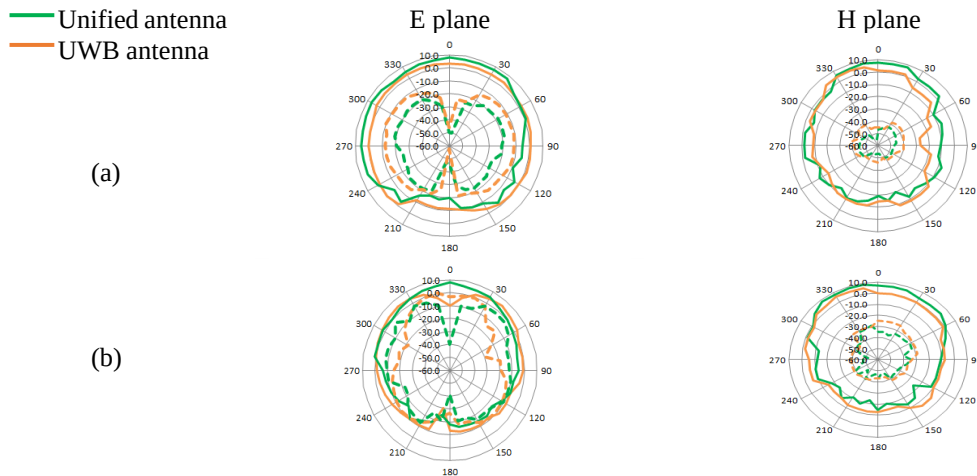
Table 6.4 Highest Radiation Efficiency and Gain values obtained by the wideband and unified antenna when placed on flat human model and in the free space at three distinct distances

	Distance from phantom (mm)	UWB antenna		Integrated antenna	
		Maximum Gain (dB)	Maximum η (%)	Maximum Gain (dB)	Maximum η (%)
Without Human phantom	-	5.1	89	8	90
With Human phantom	At 2	2.78	21	8	77
	At 5	6.5	60	8	78
	At 10	7.3	80	8	85

It is observed from Table 6.4 that the highest radiation efficiency as well as gain values of designed wideband antenna is considerably influenced when it is placed nearby to the physical body. The decrement in radiation efficiency and gain values by 68% and 2.32 dB is seen as we placed the wideband antenna at 2 mm of separation from human phantom. Moreover, it is found that wideband antenna at 10 mm of distance from physical body gave best possible radiation efficiency and gain values. However, in unified antenna case, the physical body when placed even at 2 mm

of close proximity shows very minute impact on antenna characteristics. However, in case of unified antenna the values of radiation efficiency and gain acquired with human model minutely varies in the complete range covering 2.3 GHz to 16 GHz of frequency region yet comparatively larger value of highest radiation efficiency and gain is achieved even in close proximity of physical body in comparison with free space.

The measured radiation features (cross and co polarization components) of proposed wideband and unified antenna showing H and E plane patterns when both of these are placed at separation of 2 mm from flat human model at various resonance frequencies is displayed in the Fig. 6.21. Based on the antenna orientation with respect to the axes, here yz plane represents H plane ($\Phi=90^\circ$) and xz plane represents the E plane ($\Phi=0^\circ$). It is observed that the co polarization level rises and cross polarization level reduces along with similar maximum radiation direction in both H and E plane, when the proposed wideband antenna is kept on the metamaterial layer. Moreover, the extent of increase or decrease in cross and co polarization magnitudes is observed to be larger in the E plane patterns as contrasted to the H plane pattern. Largest co polarization values of 9.8 dBi is observed at 10.2 GHz of frequency in unified antenna's H plane pattern along with smallest cross polarization value of -42.8 dBi is found at 3.4 GHz of frequency in unified antenna's H plane pattern. Moreover, in unified structure case, enhanced radiations are found in the direction of maximum forward radiation, hence it can be concluded that unified antenna aided in decreasing the magnitude of side and back lobes in basic wideband antenna's radiation characteristics.



(c)



Fig. 6.21 Measured radiation characteristics of wideband and unified antennas displaying cross and co polarization components in solid and dotted lines (a) 3.4 GHz (b) 5.7 GHz (c) 10.2 GHz. The different parameters of designed unified antenna are contrasted with few existing articles reported for wearable utilizations as observed in the Table 6.5. It is found that the majority of articles demonstrate confined working bandwidth region excluding [149]. The designed unified antenna shows $50 \times 44.4 \text{ mm}^2$ of optimal profile, 13.7 GHz of huge working bandwidth along with 8 dB of gain value. Moreover, in majority of articles excluding [148, 154, 156], the values of radiation efficiency are not even discussed, however the designed antenna showed 85% of higher radiation efficiency values as compared to others. Hence, the proposed antenna produced satisfactory and optimal outcomes and it can be reviewed as a suitable contender for wearable utilizations.

Table 6.5 Differentiation of designed unified antenna characteristics with various latest structures

Ref.	Size (mm^2)	Bandwidth achieved or resonating frequencies (GHz)	Maximum gain on phantom (dB)	Maximum radiation efficiency on phantom (%)	Minimum value of SAR achieved with phantom (W/kg)
[153]	(54×24)	1.97	5.021	NA	0.525
[154]	(68×38)	2.45	6.88	76	0.244
[155]	(80×61)	(2-12)	7.2	NA	0.52
[156]	NA	(1.8-4.5)	NA	NA	NA
[157]	(96×101)	2.4	4	NA	NA
[158]	(90×90)	2.45 GHz, 5.8	7.04	NA	0.054
[159]	(50×50)	2.4	NA	NA	0.0721
[160]	NA	2.43	NA	NA	0.010
[161]	(50×40)	2.45, 2.6, 5.5	1.5	25	0.015
[162]	(58×40)	2.44	-0.63	19.37	0.20
Proposed Work	(50×44.4)	(2.3-16)	8	85	0.10

denoted by solid lines. From reflection coefficient curve, it is found that designed metamaterial displays pass band resonance features at 4.7, 7.2, 9.9 and 12.2 GHz and from transmission coefficient curve the proposed metamaterial is observed to behave as broad bandgap structure showing band stop characteristics from 2 to 4.5 GHz, 5 to 6.4 GHz, 7.5 to 8.6 GHz, 10.8 to 11.5 GHz and 13.1 to 16 GHz of frequency region, hence providing 8.6 GHz of large stopband features wrapping the S, C, X and Ku bands. Moreover, in accompanying phase plots of metamaterial the phase reversal is found at various resonant frequencies.

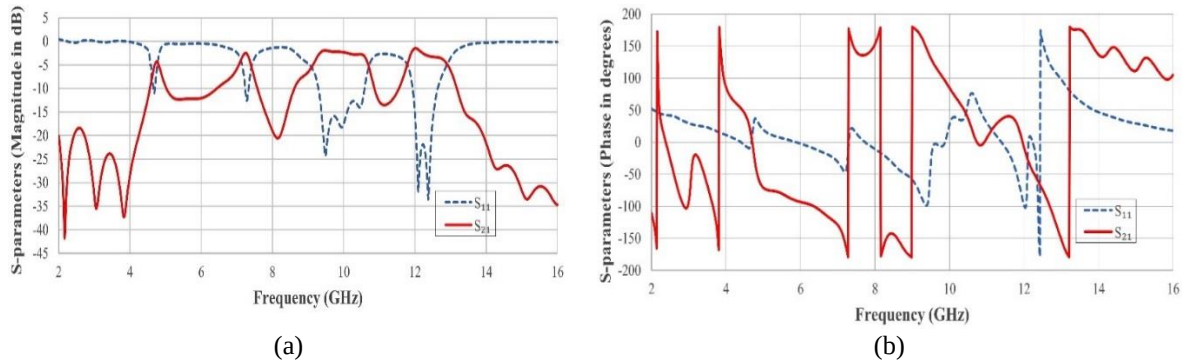


Fig. 6.23 Scattering parameters of designed unit cell

6.2.1.2 Circuital analysis of unit cell

The designed metamaterial is represented as an easy LC resonance circuit with resonant frequency calculated by using (6.4) given below.

$$f_r = \frac{1}{2\pi\sqrt{LC}} \quad (6.4)$$

here C and L indicates the equivalent capacitance and inductance elements in the proposed metamaterial's approximated circuit. On the basis of above examination, an equivalent circuitual model of proposed metamaterial is displayed in the Fig. 6.24.

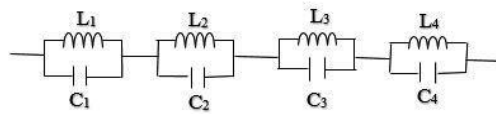


Fig. 6.24 Circuitual model of designed unit cell ($L_1=0.028$ nH, $C_1=40$ pF, $L_2=0.047$ nH, $C_2=10.2$ pF, $L_3=0.10$ nH, $C_3=2.58$ pF, $L_4=0.0062$ nH, and $C_4=27.40$ pF)

To find out the LC elements value in the unit cell's equivalent circuitual model, initially the elements value is determined at various passband resonance frequencies of 4.7, 7.2, 9.9 and 12.2

GHz by utilizing the impedance curves or smith charts, moreover the LC element values determined at above resonant frequencies are in accordance with (6.4) also. First the circuital model of designed metamaterial along with calculated LC values is simulated utilizing a circuit simulator, further the values are improved and calibrated to match up the working bandwidth and middle resonant frequencies of electromagnetic simulator. The simulation outcomes for both EM and circuital simulator is compared and satisfactory match within working bandwidth along with various passband frequencies of 4.7, 7.2, 9.9, 12.2 GHz is obtained as observed in the Fig. 6.25.

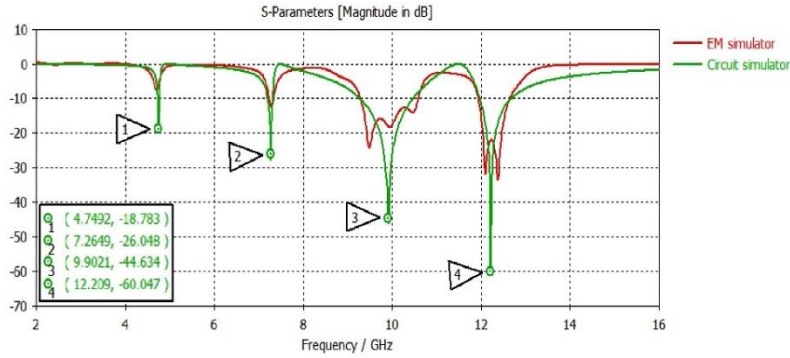


Fig. 6.25 Reflection coefficient plots acquired using both circuit and EM simulator

The outcomes acquired using the EM and circuital simulator demonstrates proper uniformity, although minute variations seen are due to higher level resonant effects.

6.2.1.3 Testing and manufacturing

A single metamaterial cell along with its 4×4 array structure is manufactured for the testing purpose as displayed in the Fig. 6.26. The S_{11} and S_{21} plots of designed metamaterial are determined utilizing the E5063A VNA covering 2 GHz to 16 GHz of frequency region. The measurement arrangement utilized to compute the S_{21} values of metamaterial are shown in the Fig. 6.27 (a). Moreover, radiating patch, various radiator structures along with 50Ω feedline used for excitation purpose are placed on similar substrate, hence the overall testing can be performed by familiar lab equipment's in place of utilizing waveguide structures.

Figure 6.27 (b) demonstrates simulated and measured transmission coefficient plots of designed metamaterial. Both these plots demonstrate nearly identical band stop regions in S, C, X and Ku frequency regions. Further minute variations along with shift in resonant frequencies is found in the measured plot because of manufacturing and testing inaccuracies. The proposed metamaterial

generates acceptable measured outcomes which verifies the experimental viability of designed unit cell.

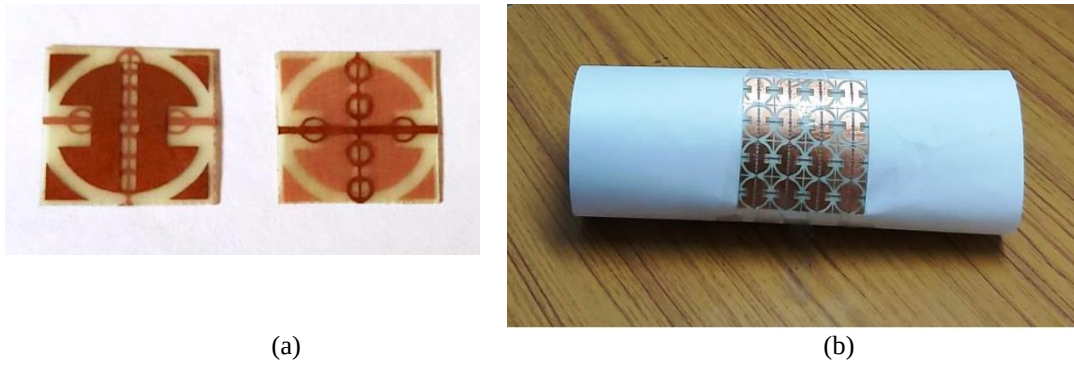


Fig. 6.26 Pictures of manufactured (a) Upper and bottom layer of metamaterial cell (b) 4×4 array arrangement

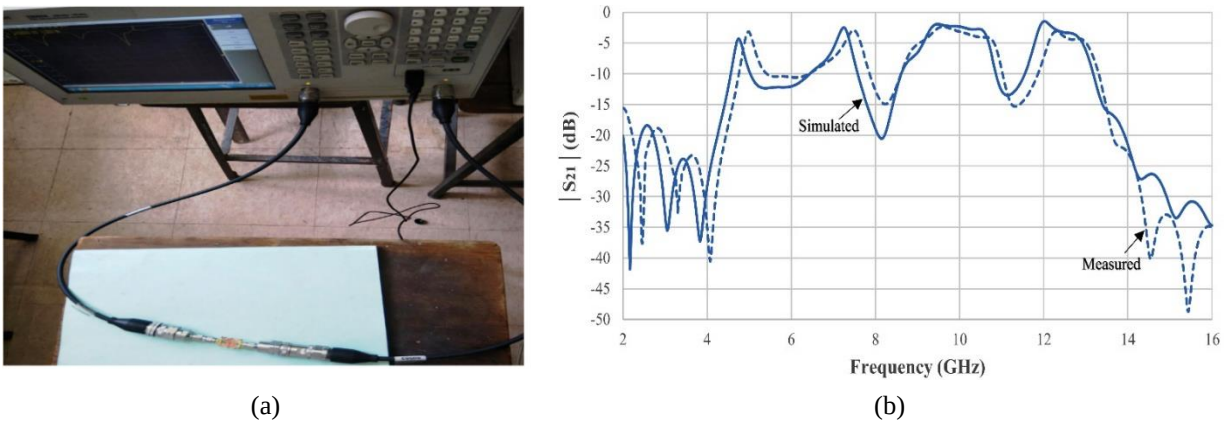


Fig. 6.27 (a) Measurement setup utilized for measuring various metamaterial properties (b) Measured and simulated transmission coefficient plots of designed metamaterial

6.2.2 Unification of designed metamaterial layer with wideband antenna

The present work largely pays attention towards the development and investigation of proposed metamaterial structure, hence to show its applicability for enhancing different antenna parameters like impedance bandwidth, gain values as well as radiation properties, the designed metamaterial layer is combined to the wideband structure discussed in the section 6.1.1 earlier.

The array configuration is fixed to 4×4 size, as it properly envelopes the wideband antenna. Further 12 mm of optimal air gap is kept in between the wideband structure and reflecting layer to acquire higher antenna performance. Fig. 6.28 (a) and Fig. 6.28 (b) studies impact on antenna's gain and

reflection coefficient plots due to varying air gap “h” values i.e. 2,6,10,12,16 and 20 mm. The band stop features of proposed metamaterial layer especially when kept nearer to wideband antenna is found in Fig. 6.28 (a). Operating in an antenna’s reactive near field zone shows conflicting consequences on antenna functioning. It is clearly seen in the reflection coefficient plot below, that impedance mismatch conditions are found at air gap value of 2 and 6 mm which results in smaller reflection coefficient, gain and impedance bandwidth. Further under air gap value of 12 mm, the reflection coefficient plots are observed to be over -10 dB at few frequency regions with smaller operating bandwidth. 12 mm of air gap value is observed to achieve consistent below -10 dB bandwidth covering 2 GHz to 16 GHz of frequency range and over 12 mm no considerable enhancement in bandwidth range is found. Likewise, from the gain curves in the Fig. 6.28 (b), the maximum satisfactory enhancement in gain values is found at 12 mm of spacing. Hence, based on this parametrical study the air gap value of 12 mm is selected as the most appropriate value.

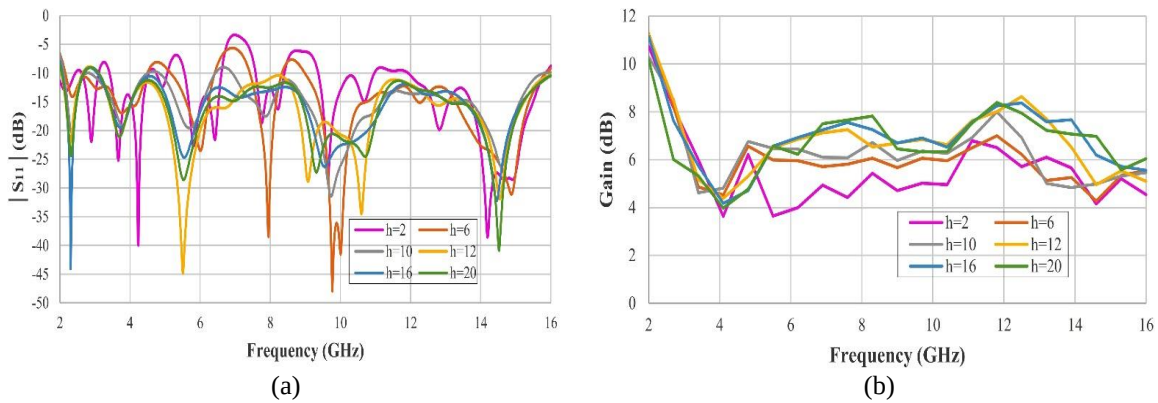


Fig. 6.28 Impact on unified structure performance (a) Reflection coefficient (b) Gain plot obtained for various air gap values

To demonstrate the impact of the number of unit cells on the antenna parameters (gain and bandwidth), the metamaterial antenna with fixed air separation as well as variable array configurations (1*2, 2*2, 3*3, 4*4 and 5*5) is studied. It is found that as we increase the number of unit cells the antenna gain also increases. As gain parameter mainly relies on the area of the antenna hence, with increment in the cell number, the size and area of antenna also increases resulting in higher gain values which further enhances the overall forward radiation characteristics.

Further minute alterations in the S_{11} curve is observed with increasing number of unit cells, a very minute enhancement in bandwidth range is seen at lower resonance frequencies, moreover the 4*4 and 5*5 array configurations gave exactly similar impedance bandwidths covering 2 GHz to 16

GHz as both of these configurations completely covers the wideband antenna. Figure 6.29 displays the impact on antennas gain and reflection coefficient curves because of different array configurations.

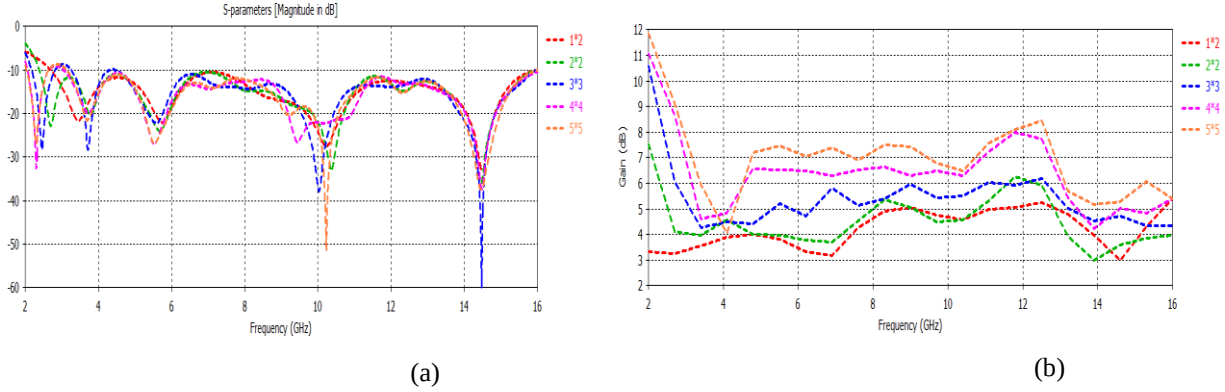
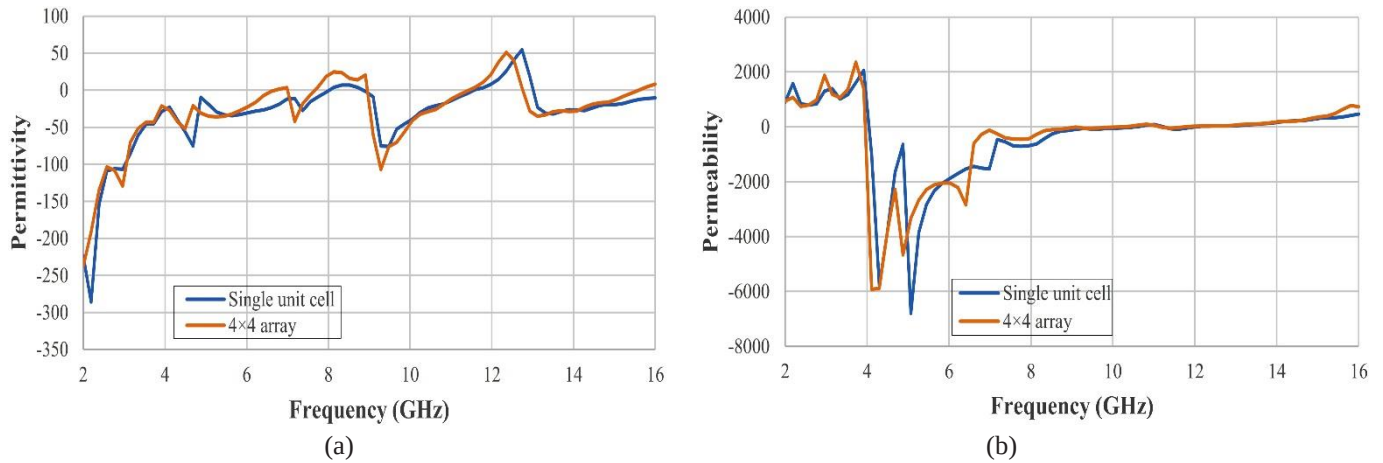
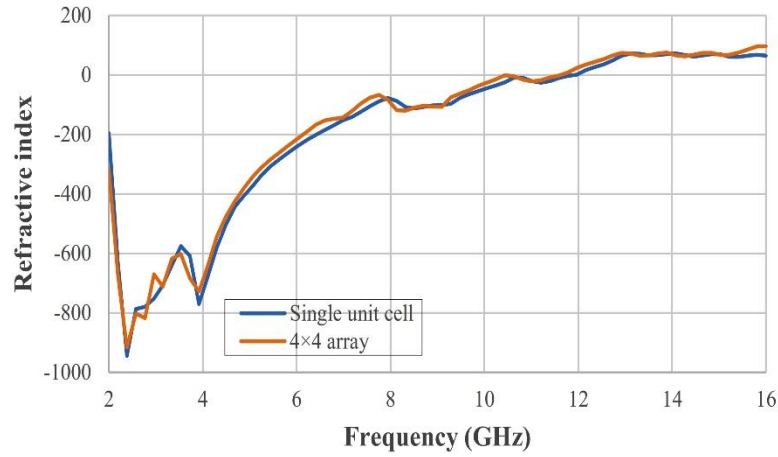


Fig. 6.29 (a) Antenna gain curves for various array configurations (b) S_{11} curves of proposed antenna for different array formations

Figure 6.30 displays simulated effective parameters (ϵ , μ and n) of 4×4 array formation as well as the basic metamaterial cell. The interchangeable coupling and interactions within various radiating patch components inside an array formation cause alteration in the array performance when compared to the basic metamaterial cell. The various effective properties along with DNG features of both solo metamaterial cell and array configuration are differentiated in the Table 6.7.





(c)

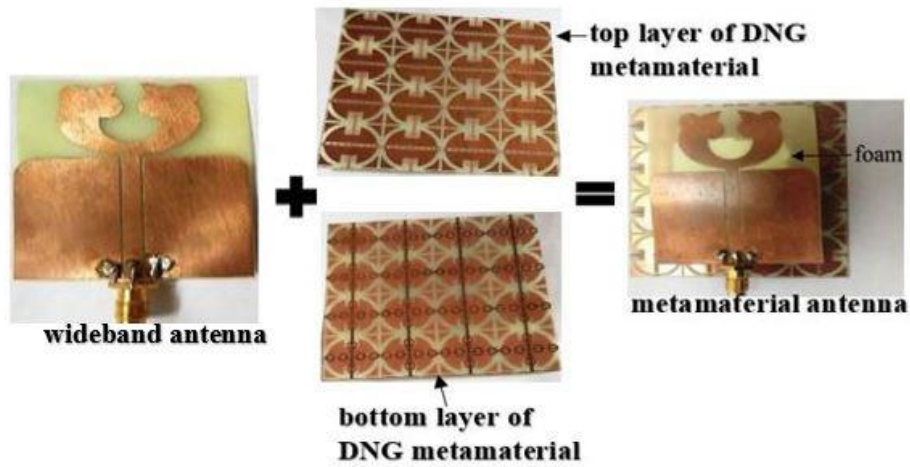
Fig. 6.30 Effective properties of both single metamaterials along with 4×4 array formation (a) permittivity (b) permeability (c) index of refractive

Table 6.7 Comparison between basic unit cell and 4×4 array

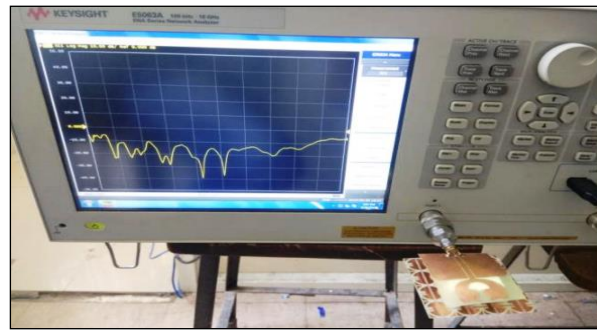
Structure	Effective parameters	Frequency range (GHz)	Covered Bands	Bands with DNG characteristics	Real values at 4.3 and 11.3 GHz
Unit cell	ϵ	2-8, 8.8-11.6, 13-16 (11.8 GHz)	S, C, X, Ku	C and X	-49.649, -25.6
	μ	4-10.5, 11.2-12 (7.3 GHz)	C, X		-5804.5, -80.812
	n	2-12 (10 GHz)	S, C, X		-586.65, -25.633
4×4 array	ϵ	2-6.5, 7.1-7.6, 8.9-11.31, 12.7-15.5 (10.21 GHz)	S, C, X, Ku	S, C and X	-45.72, -3.83
	μ	3.9-10.2, 11.1-11.7 (6.9 GHz)	S, C, X		-5889.6, -39.57
	n	2-11.5 (9.5 GHz)	S, C, X		-537.89, -12.79

6.2.2.1 Measurement and manufacturing of unified structure

The photographs of manufactured antennas with both conformal wideband structure as well as unified antenna together with their experimental arrangement is displayed in the Fig. 6.31 (a) and Fig. 6.31 (b).



(a)



(b)

Fig. 6.31 (a) Picture of manufactured structures (b) Measurement arrangement utilized for testing purpose

The measured as well as simulated reflection coefficient curves of proposed wideband structure along with unified structure are displayed in the Fig. 6.32 (a). The wideband structure is found to generated 13.4 GHz of consistent simulated bandwidth covering 2.6 GHz to 16 GHz of frequency range along with under -20 dB reflection coefficient values at 3.4, 5.7, 10.2 and 14.5 GHz of resonant frequencies, although in measured curve 0.2 GHz of reduction in operating bandwidth range is found. Moreover, almost similar measured and simulated bandwidth of 14 GHz covering 2 GHz to 16 GHz of frequency range is observed in the unified antenna case, minute shift in resonant frequencies in measured curve is seen because of manufacturing and calibration inaccuracies. While differentiating measured outcomes of both structures, a 0.8 GHz (2.8-2 GHz) of increase in operating bandwidth range is observed as we introduced the metamaterial layer. The measured as well as simulated gain curves of both unified and wideband structure are displayed in

the Fig. 6.32 (b). Above 2 dB enhancement in gain values at various resonant frequencies after introducing metamaterial layer is observed. Further, both measured and simulated gain plots of two structures are quite similar with minute fluctuations. The measured gain values for both structures acquired at various resonant frequencies is given in the Table 6.8.

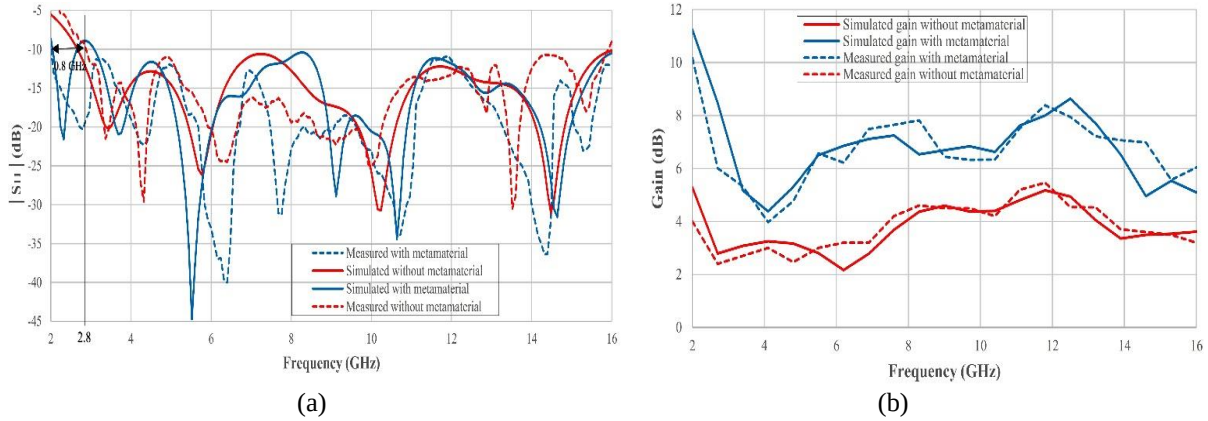


Fig. 6.32 (a) Reflection coefficient plots for both the structures (b) Gain curves for both structures

Table 6.8 The measured values of gain for both structures at various resonant frequencies

Frequencies (GHz)	Gain values for wideband antenna (dB)	Gain values for metamaterial antenna (dB)	Gain enhancement (dB)
3.4	2.7	5.42	2.72
5.7	3	6.60	3.6
10.2	4.27	6.38	2.11
14.5	3.5	6.98	3.48

E and H plane with respect to antenna geometry mainly depends on the placement of the antenna in the coordinate system or the antenna orientation with respect to the axes. In the present work yz plane is the H plane ($\Phi=90^\circ$) while xz plane represents the E plane ($\Phi=0^\circ$). Figure 6.33 displays the radiation characteristics of H and E principle planes for both the unified and wideband structure at various resonant frequencies of 3.4, 5.7 and 10.2 GHz. The refinement in the radiation features towards forward direction of unified structure are observed along with lesser side and back lobes. 4.8 dBi of enhancement in the magnitude of main radiating lobe is found as we introduce the metamaterial layer. Moreover, the enhancement in magnitude of main lobes is observed to be larger in H plane with respect to E plane.

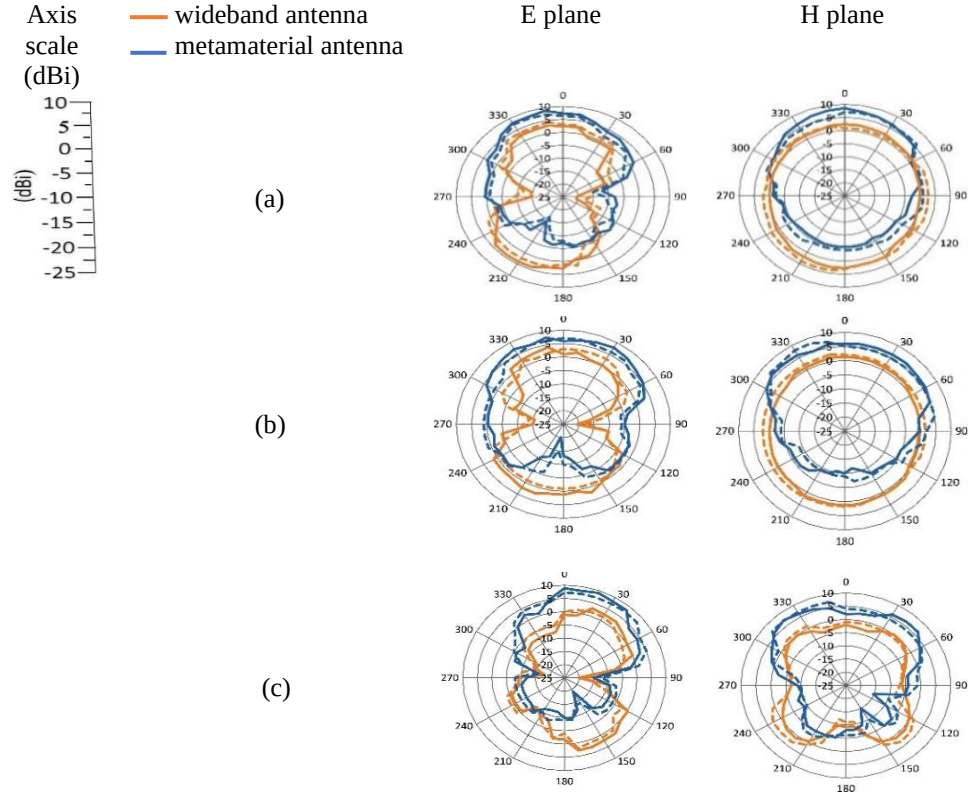


Fig. 6.33 Simulated (dotted lines) and measured (solid lines) patterns (H and E) for both the structures at various resonant frequencies (a) 3.4 GHz (b) 5.7 GHz (c) 10.2 GHz

To demonstrate the bending impact on the proposed unified antenna's performance, 4×4 metamaterial array combined with wideband design called unified antenna is placed on a cylindrical structure with fixed radius of 10 mm and angle of bending is altered from 30° to 180° to show its impact on reflection coefficient curve of unified antenna. The methodical H plane bending (bending across antenna width) of designed antenna is shown in the Fig. 6.34. Figure 6.34 (a) displays 0° bending conditions, where antenna is simply kept on an analytically defined virtual cylindrical structure. While showing bending effects with cylindrical structure, three essential parameters need to be considered i.e. S denoting length of arc, r indicates cylinder radius and θ denotes the angle of bending. Here the parameters S and r are fixed and only θ is altered. Figure 6.34 diagrams demonstrates the designed structure showing 30° , 60° and 120° of bending. The measured reflection coefficient curves of unified structure at 30° and 0° bending angles is displayed in Fig. 6.35 (a). Figure 6.35 (b) shows simulated bending curves of unified structure at various bending scenarios. As the proposed antenna being conformal shows very minute changes in its complete operating bandwidth due to bending, although at greater bending extents little shift in larger resonant frequencies is found.

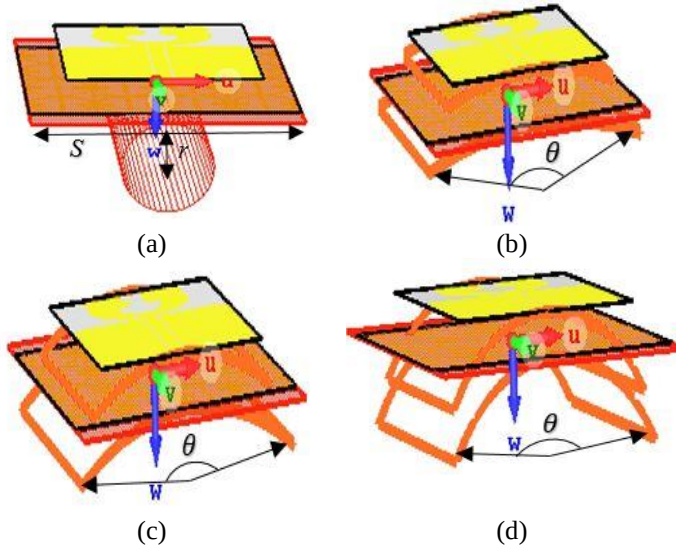


Fig. 6.34 The bending of unified structure across width dimension at various angles of bending
(a) 0 degree (b) 30 degree (c) 60 degree (d) 120 degree

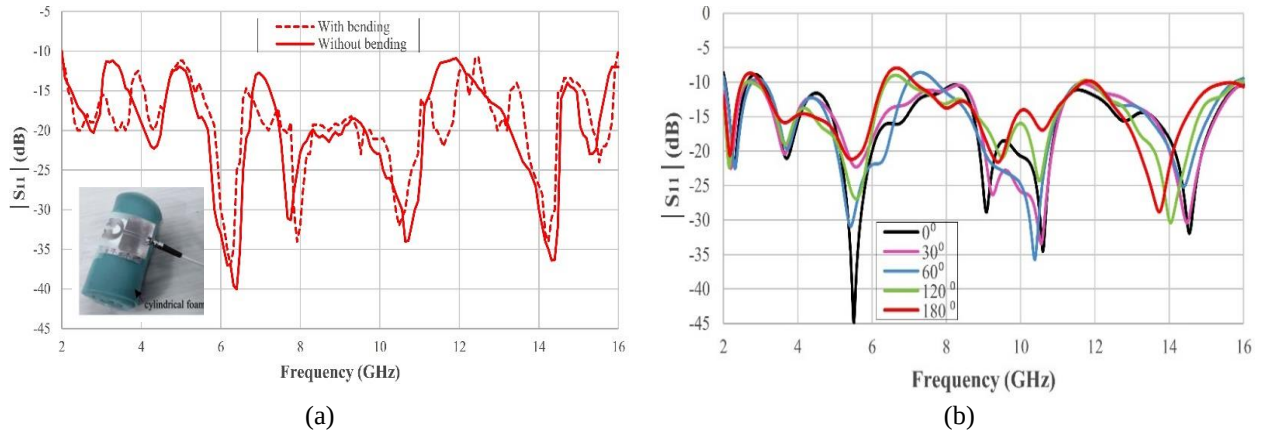


Fig. 6.35 (a) Measured reflection coefficient plots of unified structure without/with bending (b)
Impact of bending on reflection coefficient curve of unified structure

6.2.2.2 Differentiation with few current structures

Here the designed unified structure is contrasted with various formerly introduced designs taken from literary texts as displayed in the Table 6.9. In comparison to other manuscripts, proposed work has metamaterial cell with smallest volume, acquires 14 GHz of largest working bandwidth along with conformal nature. It also shows largest gain value with exception of [165], that shows 1.7 dB larger gain along with lesser profile, but [165] has 0.7 GHz of small operating bandwidth

that is 13.3 GHz smaller than the antenna presented in this work. The compromise within bandwidth and gain values is clearly visible in all the articles examined in the Table 6.9, however the proposed structure obtains optimum bandwidth and gain values.

Table 6.9 Differentiation of designed structure with earlier reported designs

Ref.	Unit cell size/ volume (mm ³)	-10 dB bandwidth (GHz)	No. of cells used	Substrate material	Total size/ volume (mm ³)	Max gain (dB)	Flexible /Non-Flexible (F/NF)	Unit cell structure
[163]	NA	3.1-13.5 (10.4)	3	FR4	25×22×1.6 (880)	4	NF	Split ring resonators (SRRs)
[164]	NA	3.9-13.2 (9.3)	2	FR4	25×15×1.6 (600)	4.8	NF	Concentric structure
[165]	7×7×1.6/78.4	3.1-10.71 (7.61)	(2×4) array	FR4	26×12×1.6 (499.2)	3.9	NF	SRRs
[166]	NA	4-12.5 (8.5)	4	FR4	25×15×1.6 (600)	4.5	NF	Spiral structure
[167]	NA	24-30 (6)	1	FR4	5×6×0.8 (24)	3	NF	Complimentary SRRs
[168]	NA	4.45-5.71 (1.26)	(4×4) array	FR4	44 × 35 × 4 (6160)	6.6	NF	Spiral structure
[169]	5.3×5.3×0.5/14.045	9.65-10.1 (0.45)	(3×3) array	Rogers RT/dur oid 5880	89×50×4.075 (18133.7)	8.24	NF	Novel structure
[170]	18×20×3.048/1097.28	5.47-7.25 (1.78)	(2×2) on top and (2×5) on the bottom	Rogers 350B	137 × 77 × 3.048 (32153.3)	9.2	NF	Novel structure
[171]	NA	5.4-6.1 (0.7)	2	FR4	52.5×67.1 ×10.96 (38609.34)	11.9	NF	SRRs
[172]	6×6×1.58/56.88	7.5-10 (2.5)	(9×13) array	FR4	NA	NA	NF	Modified alphabetical structure
Proposed work	14.8×14.8×0.25/54.76	2-16 (14)	(4×4) array	FR4	59.2×59.2 ×12 (42055.6)	10.2	F	Novel structure

6.3 ANTENNA CHARACTERISTICS IN TIME DOMAIN

The correlation factor between the transmitted and received antenna can be calculated using (6.1) below.

$$F = \max_{\tau} \left| \frac{\int_{-\infty}^{+\infty} S(t) r(t-\tau) dt}{\int_{-\infty}^{+\infty} S(t) dt \cdot \int_{-\infty}^{+\infty} r(t) dt} \right| \quad (6.1)$$

Here $S(t)$ and $r(t)$ denotes the transmitted and received antenna signals and τ denote the delay factor. As compared to many telecommunication systems, in IR-UWB systems, it is required to achieve good correlation within the transmitting and receiving signals so to have maximum information transfer with minimum distortions due to radio channel. To calculate the fidelity parameter, firstly the input and received signals are studied for two antenna configurations i.e. side by side and face to face. In this study the transmitting and receiving antennas are placed 250 mm apart from each other and simulated results are obtained using CST microwave studio. Now for excitation purpose, the default Gaussian pulse in CST with spectrum ranging from 3.1 to 10.6 GHz is utilized as shown in the Fig. 6.36 (a). As observed below the received signals in Fig. 6.36 (b) and Fig. 6.36 (c) are broadened, still a good correlation is observed within the transmitted and received signals. By utilizing the above equation (6.1), the antenna fidelity factor for two orientations is calculated. Figure 6.36 shows the normalized wave signals used for calculating the antenna fidelity factor. Table 6.10 gives fidelity factor values for two antenna orientations.

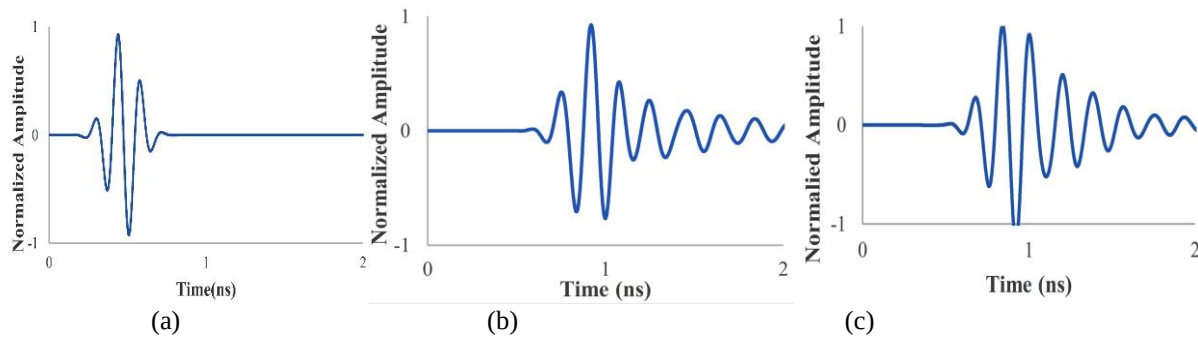


Fig. 6.36 (a) Input excitation signal (b) Received signal (side by side) (c) Received signal (face to face)

Table 6.10 Antenna fidelity factor without human phantom

Scenarios	Side by Side	Face to face
Fidelity factor	0.9516	0.9438

Now the analysis is also done with flat human phantom to see its effect on antenna fidelity factor. Here the flat human phantom is placed at a distance of 2 mm from the antenna. The Fig. 6.37 shows the normalized wave signals used for calculating the antenna fidelity factor with flat human phantom at a distance of 2 mm. Further Table 6.11 shows the antenna fidelity factor values with flat human phantom at 2 mm of distance.

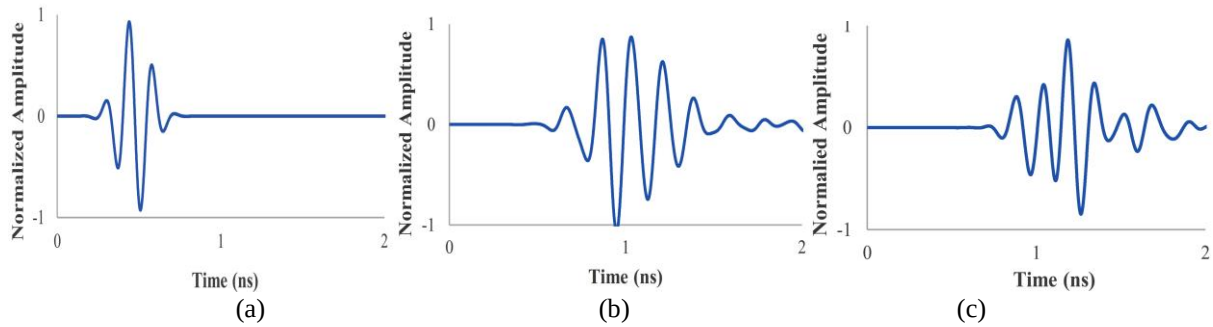


Fig. 6.37 (a) Input excitation signal (b) Received signal (side by side) (c) Received signal (face to face)

Table 6.11 Fidelity factor with human phantom at 2 mm distance

Scenarios	Side by Side	Face to face
Fidelity factor	0.8901	0.8578

6.4 SUMMARY

The extensive study of two flexible structures including the bending, physical body and varying air gap effects, the enhancement in gain, bandwidth and radiation pattern values along with specific absorption rate analysis is performed in this chapter. These structures utilize similar broadband conformal FR4 substrate with overall size of $50 \times 43 \times 0.15 \text{ mm}^3$ together with co planar wave guide feed for both the single sided MNG and dual sided NRI metamaterial layer. Table 6.12 below displays the differentiation between the two proposed metamaterial structures.

Table 6.12 Comparison between two proposed metamaterial structures

Wideband metamaterial antennas	Bandwidth obtained (GHz)	Maximum gain (dB)	Co-polarization level at all resonating frequencies (dB)	Cross polarization level at all resonating frequencies (dB)	Metamaterial array size	Volume/Material of single unit cell (mm^3)
MNG metamaterial based	2.3-16 (13.7)	8	Above 6.2	Below -7	3×3	$14.8 \times 14.8 \times 0.15$ (32.856)/FR4
NRI metamaterial based	2-16 (14)	10.2	Above 7.3	Below -19.6	4×4	$14.8 \times 14.8 \times 0.25$ (54.76)/FR4

Next in the last chapter of the thesis that includes the ultimate conclusion drawn and the future scope of the work presented.

CHAPTER 7

CONCLUSIONS AND FUTURE WORK

7.1 CONCLUSIONS

From the various designed structures and their analysis conclusions are drawn in this section. First of all, in the present work all the antennas designed are novel with unique structures. Complete design analysis is done for better understanding the physics behind the design. All the antennas are compared with the latest designs and are found better. For designing the UWB antennas the method of “combining resonant structures” along with CPW feeding is preferred more. Moreover, all the antenna gave optimum performance in terms of both bandwidth and gain, which is not taken care by most of the researchers earlier. The three wideband antennas designed covered 3.1 to 10.6 GHz, 2.3 to 16 GHz and 2 to 16 GHz of frequency range and also gave maximum gain values of 6.1 dB, 8 dB and 10.2 dB. The fabricated structures also produced acceptable measured outcomes validating the experimental usefulness of designed antennas. For all the flexible structures designed in this work FR-4, being easily available in the market, is used and preferred as compared to the other substrates. Further the fabrication process used is also quite simple and cheap which includes screen printing followed by the photolithography. As metamaterials are extensively explored in the past literature, still the unit cells designed in this work are completely new and their complete design analysis including electric, magnetic and circuital analysis is done. As NRI metamaterials are found to be potential contender in the various applications, therefore more focus is given to achieve the DNG and NRI features and “combined metamaterial” unit cells are designed. The two flexible unit cells gave wide DNG regions of 6.1 GHz and 8.2 GHz along with NRI regions of 10 GHz and 13 GHz. The final performance of the unit cells is compared with recent literature and were found better. Further all the designed structures are compared with recent work done by the researchers and present work is found better.

In third chapter, a compact UWB antenna with good matching characteristics at its five resonance frequencies is designed which gave optimum gain and radiation efficiency values of 6.1 dB and 94 %. In fourth chapter a band reject antenna using stubs as notching elements in designed for WLAN and WiMAX rejection and it presented good notching features with -7 dB gain and 18.7 VSWR values in stop band region. In fifth chapter a flexible resonant unit cell is designed which gave very large DNG and NRI bandwidths of 8.2 GHz and 13 GHz. In final chapter two flexible

metamaterial antennas operating in the frequency range from 2 to 16 GHz are designed which achieved high gain values of 8 dB and 10.2 dB along with satisfactory radiation characteristics.

7.2 FUTURE SCOPE OF WORK

The metamaterials designed in this work are only used for the performance enhancement of the UWB antennas. However, metamaterials are found to have great application scope such as specific absorption rate minimization, designing of antennas, band gap structures, photovoltaic cells, absorbers [173], high resolution lenses, filter designs and in cloaking devices. Therefore, the unit cells can be used for other applications as well in the future. Further other flexible and low-cost substrates such as paper, being easily available and environment friendly, can be explored in the future work with different fabrication techniques. Further as we have designed novel NRI metamaterials in the present work, in future we can go one step further by designing the “chiral metamaterial structures”. Chiral metamaterials have recently attracted considerable attention due to their simpler route to negative refraction in which NRI features are obtained without having both permittivity and permeability negative in the same frequency band with the help of “chirality” feature in them.

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LIST OF SCI PUBLICATIONS

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