INVESTIGATIONS ON FREQUENCY AGILE BEHAVIOR OF
NARROWBAND AND WIDEBAND MICROSTRIP PATCH ANTENNAS
BY EMPLOYING VARIABLE HEIGHT GROUND PLANES

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Rahul Bakshi
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The Undersigned Faculty Committee Approves the

Thesis of Rahul Bakshi:

Investigations on Frequency Agile Behavior of Narrowband and Wideband
Microstrip Patch Antennas by Employing Variable Height Ground Planes.

Satish Kumar Sharma, Chair
Department of Electrical and Computer Engineering

Sunil Kumar
Department of Electrical and Computer Engineering

Samuel K. Kassegue
Department of Mechanical Engineering

May 10, 2010
Approval Date
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DEDICATION

My Family & Almighty
ABSTRACT OF THE THESIS

Investigations on Frequency Agile Behavior of Narrowband and Wideband Microstrip Patch Antennas by Employing Variable Height Ground Planes

by

Rahul Bakshi
Master of Science in Electrical Engineering
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Microstrip patch antennas have been studied and designed extensively over the years which find many applications in wireless communications, satellite communications and radar applications. These antennas have features such as simple, robust, low profile, cheaper and versatile for various modes of operations. However, the patch antennas are inherently narrow band, hence require wide-bandwidth and/or frequency agile property in order to meet the need of modern wireless communications applications.

This thesis presents investigation results on the effect of employing variable height ground plane and copper ribbon switches on the frequency agile behavior of wideband microstrip patch antenna. The effect of ground plane height variation not only alters operating frequency, but also, Gain and impedance bandwidth. First a narrow band patch antenna and its array configuration will be studied for its frequency agility and far-field performance by varying the ground plane, followed by the proposed wideband patch antenna, which is basically a U-slot loaded modified E-shape microstrip patch antenna (USLMES). The narrowband patch antenna achieves 33% frequency agility which increases to 43% for the array configuration. The proposed USLMES antenna is excited using the notch feed mechanism by a 50 Ω coaxial probe outside the patch surface so that coaxial probe does not contribute significantly to the peak-cross-polarization levels. The parametric study results are presented for the wideband patch antenna design and important parameters have been noted. The proposed wideband patch antenna offers impedance (S_{11} = −10 dB) and 3dB gain bandwidths of at least 35% (3.09GHz to 4.42GHz) and approaches 38% (3.09GHz to 4.59GHz) based on experimental measured data with stable radiation patterns and acceptable cross-polarization levels. Frequency agility is also achieved from 3.02GHz to 4.95 GHz by turning different combination of copper ribbon switches ON/OFF in the later part of the thesis. The prototype antennas were fabricated and experimentally verified for both frequency agility and wideband patch performance verifications. The simulated performance is in reasonable agreement with the measured results. The proposed wideband microstrip patch antenna can be used as the radiating elements in base station applications.
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CHAPTER 1

INTRODUCTION

Microstrip patch antennas have become significantly popular after getting attention in the 1970s. They are widely known for their simplicity, ease of fabrication, low profile, conformable to planar and nonplanar surfaces and mechanical robustness. Further, when a particular patch shape and radiating mode are selected, they are very versatile in terms of resonant frequency, polarization, radiation pattern and impedance [1, 2]. There advantages and disadvantages have been cited in [3]. Much attention was given to the design of reconfigurable microstrip patch antennas that could provide various reconfigurable frequency states. Antennas with capability to change its resonant frequency adaptively are known as frequency agile antennas which is a type of reconfigurable antennas.

In order to investigate the frequency agility and polarization diversity of a microstrip patch antenna, several researchers have studied and experimentally verified different mechanisms in the past. The author in [4] has investigated the effect of putting PIN diode on the radiating element and based on the ON/OFF states, the antenna is reconfigured to exhibit linear, left or right hand circular polarizations. The author in [5] shows frequency and polarization reconfiguration with the use of PIN diodes between the shorting post and radiating element. A design is presented in [6] where the author uses PIN diodes between the two orthogonal slots on the antenna geometry to produce LHCP and RHCP by turning the diodes ON/OFF.

MEMS switches have been incorporated to produce frequency agility by putting them on the feeding arm of a diagonally fed square patch exhibiting circular polarization using a stub [7]. The author in [8] incorporated RF-MEMS switches between the 3×3 microstrip patch array to exhibit multiband frequency operation. A rectangular spiral antenna is presented in [9] which show left hand circular polarization (LHCP) and right hand circular polarization (RHCP) with ON/OFF states of the RF-MEMS switches. A rectangular patch antenna is tuned at L-band using 8 varactors in [10].
The slot antenna having polarization diversity capability between RHCP and LHCP or between linear polarization (LP) and circular-polarization (CP) is presented in [11]. The PIN diodes are installed on the perturbations of the slot ring antenna and reconfiguration is achieved by turning them ON/OFF. A novel compact circular-polarization (CP) operation of the square microstrip antenna with four slits and a pair of truncated corners is proposed in [12] which show a reduction of 36% in the antenna size as compared to the conventional corner-truncated square microstrip antenna at a given operating frequency.

The author in [13] introduced a novel reconfigurable patch antenna with switchable slot (PASS) to realize dual frequency operation, dual band circularly polarization and polarization diversity with a single feed. A dual band circularly polarized antenna is realized with the PASS concept in [14] where the switch is fixed at the center of the slot that is installed in the patch. This antenna has been designed for future Mars rover mission and operates at two discrete frequencies with ON/OFF states of the switches. A novel reconfigurable microstrip patch antenna loaded with variable RF-MEMS capacitors on the co-planar waveguide (CPW) stub is presented in [15]. Frequency reconfigurability is achieved by varying the actuation voltage of the MEMS capacitors from 0-11.9V, which influences the resonant frequency of the circuit.

Ground plane reconfiguration has been studied in [16] where the author detunes the resonant frequency of 3 elements array of patch by varying the substrate height from 210μm to 600μm between the radiating element and the ground plane. In [17], the author presented a new type of RF MEMS device for microstrip line phase shifters and frequency agility where the patch antenna achieves a 6% controlled variation in resonant frequency shift by employing electrostatic potential on the ground plane by varying the height from 0 to 138μm.

1.1 PROPOSED RESEARCH GOALS

In this thesis, frequency agile behavior of a narrowband rectangular patch antenna and its array in 1×4 configuration are presented first. The frequency agility is achieved by varying the height of the ground plane below the patch. The narrowband rectangular patch antenna is fed by a transmission line and shows about 33% frequency agility can be achieved by ground plane height variation. The array is fed-in using a corporate-feed network with coaxial probe on its feeding arm. The array shows around 43% frequency agility. Next, a
U-slot loaded modified E-shape wideband patch antenna (USLMES) fed by a coaxial probe is proposed with finite size ground plane and substrate. Its frequency agility behavior is investigated with different air gaps between the FR-4 substrate and ground plane. Three USLMES patch antenna prototypes were fabricated and measured to validate the results of the simulation.

1.2 MICROSTRIP PATCH ANTENNA

The following section presents the characteristics of microstrip patch antennas (MPA’s), current work to implement frequency reconfiguration and scope of thesis in reference to other methods to achieve frequency agility.

1.2.1 Patch Antenna Geometry

Figure 1.1 shows the geometry of a microstrip patch antenna with radiating edges on one side and ground plane on the other. The patch is generally made of conducting material such as copper or gold and can take any possible shape. The dielectric has a dielectric constant which decides the performance of the microstrip patch antenna (MPA). If the dielectric material is rigid as in our case (FR-4), the radiating patch can be photo etched on its surface.

![Figure 1.1. Geometry of a microstrip patch antenna.](image)

The microstrip patch antenna consist of very thin \( t \ll \lambda_0 \) where \( \lambda_0 \) is the free space wavelength) metallic strip (patch) placed a small fraction of a wavelength \( h \ll \lambda_0 \), usually
0.003 \( \lambda_0 \leq h \leq 0.05 \lambda_0 \) above a ground plane. The patch primarily radiates because of the fringing fields present at the two slots between the patch and the dielectric substrate. The dielectric substrate’s dielectric constant is usually in the range of \( 2.2 \leq \varepsilon_r \leq 12 \). For good antenna performance, a thick dielectric substrate having a low dielectric constant is desirable since this provides better efficiency, larger bandwidth and better radiation characteristics [2].

### 1.2.2 Shapes of Microstrip Patch Antennas

Microstrip patch antennas, as stated above have a radiating element that is photoetched on the dielectric substrate. The radiating element can be of square, rectangular, thin strip (dipole), circular, elliptical, triangular or any other shape. Square, rectangular and circular are the most common shapes because of the ease of analysis and fabrication, and their attractive radiation characteristics, especially low cross-polarization level. The rectangular patch is unquestionably the most extensively used geometry because it offers largest impedance bandwidth. Circular and elliptical geometries offer relatively smaller gain and impedance bandwidth because they are smaller in size as compared to a rectangular patch. Triangular patch can be utilized to produce dual polarization but exhibits high cross-polarization due to its asymmetrical geometry. They will also show lower gain and impedance bandwidth as compared to rectangular and circular patches. Circular rings have the smallest size in the family and exhibits very low impedance bandwidth. Figure 1.2 illustrates the different shapes.

![Shapes of microstrip patch antennas](image)

**Figure 1.2.** Shapes of microstrip patch antennas.
These shapes represent the most basic structure of a microstrip patch antenna which exhibits narrowband performance. Therefore, further investigations are needed in order to make them wideband antennas. One approach is to change the structure of the patch itself by cutting slots through its geometry. This technique has been used exclusively in this thesis to achieve wideband performance with respect to conventional narrowband patch antenna.

1.2.3 Advantages and Disadvantages

Microstrip patch antennas offer several advantages as compared to other conventional antennas. Some of them are broadly classified below [3]:

- Low cost and ease of fabrication
- Light weight, low volume, low profile planar configurations, which can be made conformal to host surfaces.
- Can be made thin hence they do not perturb the aerodynamics of the aircraft.
- Can be mounted of missiles, rockets and aircraft easily
- The antennas have low scattering cross section
- Can be easily reconfigured for multiband operations.
- No cavity back required.
- Can be fed with planar feed lines during fabrication itself
- Mechanically robust and rigid

Microstrip patch antennas also have some drawbacks which are listed as under [3].

- Narrow bandwidth, which is of the order of few percent depending upon the substrate dielectric constant and thickness (around 2.5% for single layer).
- Most microstrip patch antenna radiates in half plane.
- Very high losses resulting from surface wave excitation and conductor, dielectric losses, hence lower gain and directivity.
- Practical limitation of maximum Gain (30 dBi)
- Lower power handling capability.
- Poor isolation between the feed and radiating element.

1.2.4 Transmission Line Model of Patch Antennas

The transmission line model (TLM) of a microstrip patch antenna is shown in Figure 1.3.
As explained before, a patch is a finite structure of length $L$ and width $W$. So when in operation, the electric field lines undergo fringing at the edges. It can be observed in Figure 1.3 that most of the field lines reside inside of the substrate and some are in air. Thus the transmission line supports quasi-transverse electromagnetic waves (Q-TEM) and not pure transverse electric-magnetic (TEM) mode. The fringing depends on the ratio of $L/h$ of the patch and the height $h$ of the substrate of dielectric constant $\varepsilon_r$.

Since we have a mixture of waves traveling in substrate and air, we have an effective dielectric constant $\varepsilon_{reff}$. This effective dielectric constant is a function of resonant frequency. The value of $\varepsilon_{reff}$ is less than $\varepsilon_r$ because the fringing fields around the periphery of the patch are not confined in the substrate but are spread in the air also as shown Figure 1.3. The expression for $\varepsilon_{reff}$ is given as [1]:

$$\varepsilon_{reff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ 1 + 12 \frac{h}{W} \right]^{\frac{1}{2}}$$

(1.1)

where

- $\varepsilon_{reff}$ = Effective dielectric constant
- $\varepsilon_r$ = Dielectric constant of substrate
- $h$ = Height of dielectric substrate
- $W$ = Width of the patch

Figure 1.4 shows the normal component of the electric field at the two edges along the width are in opposite direction and thus out of phase since the patch is $\lambda/2$ long and hence they cancel each other in broadside [2]. The tangential components which are in phase combine resulting in giving maximum radiated field normal to the surface of the structure.
The fringing fields at the width can be modeled as radiating slots thus behaves electrically large as compared to its original dimensions. The following equations have been taken from [1].

The length of the patch is given as:

\[
\Delta L = 0.412h \left( \frac{\varepsilon_{\text{reff}} + 0.3}{\varepsilon_{\text{reff}} - 0.258} \right) \left( \frac{W}{h} + 0.8 \right)
\]  

(1.2)

Therefore, the effective length of the patch \( L_{\text{eff}} \) can be written as:

\[
L_{\text{eff}} = L + 2\Delta L
\]  

(1.3)

For a given resonance frequency \( f_0 \), the effective length is given as:

\[
L_{\text{eff}} = \frac{c}{2f_0\sqrt{\varepsilon_{\text{reff}}}}
\]  

(1.4)

For a rectangular Microstrip patch antenna, the resonance frequency for any \( TM_{mn} \) mode is given as:

\[
f_0 = \frac{c}{2\sqrt{\varepsilon_{\text{reff}}}} \left[ \left( \frac{m}{L} \right)^2 + \left( \frac{n}{W} \right)^2 \right]^{\frac{1}{2}}
\]  

(1.5)

where, \( m \) and \( n \) are modes along \( L \) and \( W \), respectively.
For efficient radiation, the width $W$ is given as:

$$W = \frac{c}{2f_0\sqrt{\varepsilon_r + \frac{1}{2}}}$$

(1.6)

1.3 TYPES OF RECONFIGURABLE MICROSTRIP PATCH ANTENNAS

With the increasing growth of modern wireless communication systems, wise usage of resources and hardware implementation on circuit board has become imperative. This has put challenging demands for the antenna designs. In this context, much attention has been given to reconfigurable patch antennas because they offer versatility in terms of multiband and wideband operations, radiation pattern and polarization. Reconfigurable microstrip patch antennas are broadly classified in three categories.

- Frequency reconfigurable MPAs
- Radiation pattern reconfigurable MPAs
- Polarization reconfigurable MPAs

Antennas with capability to change its resonant frequency adaptively are known as frequency reconfigurable antennas. These antennas can serve more than one application through their reconfigurability mechanism. The aim of these antennas is solely to provide multi-antenna functionality at the desired communication location [18-21]. Moreover a frequency reconfigurable antenna can serve advantages of frequency reuse for enhancing the system performance or integrate the transmitter and receiver functions in one setup. They can effectively reduce the unwanted interference and jamming.

For the above antenna, the radiation pattern remains unchanged when the frequency is tuned back and forth in the frequency band. However we have other types of reconfigurable antennas where the resonant frequency remains unchanged while radiation pattern changes with respect to desired system requirements. These are known as radiation pattern reconfigurable antennas. These antennas are capable of steering their radiation pattern at different directions and are presented in [9, 22-24]. The third type of reconfigurable antenna is polarization reconfigurable antennas. These antennas exhibit switching between linear polarizations to left hand circular polarization (LHCP) or right hand circular polarization (RHCP) for digital multimedia broadcasting systems [4-7]. They are an important area of study in mobile communications as they provide improvement is signal reception where
multipath fading is involved. Reconfigurable antennas find application in satellite communication and collision avoidance radars.

1.4 Ground Plane Frequency Reconfiguration

Reconfigurable antennas presented above have been designed by incorporating switching components like PIN diodes switches [4-6, 11-14], RF-MEMS switches [7-9], varactor diodes [10] and ferroelectric varactors [20] on the antenna’s geometry. In most of the above, it is seen that the placement of number of switches on the antenna radiating edges deteriorates its performance. Moreover biasing of the loaded diodes between the patch and ground plane requires complex circuitry. Therefore, an alternative like ground plane reconfiguration is very desirable that offers simple frequency detuning technique as compared to the above. It avoids complex circuitry, biasing of external components and operates on the inherent characteristics of the patch itself.

This thesis focuses on the ground plane reconfiguration methodology to reconfigure microstrip patch antennas for its frequency agility. Since the microstrip patch antenna consists of a metallic conducting radiating strip on the top separated by a dielectric material with ground plane at the bottom, it ideally behaves as a capacitor. Therefore, by varying the distance between the metallic patch and ground plane, a change in overall capacitance of the system is achieved. This change in capacitance influences other parameters such as resonant frequency and impedance bandwidth. The analytical equation for the resonant frequency is expressed as:

$$f_r = \frac{C}{2 L_e \sqrt{\varepsilon_r}}$$  \hspace{1cm} (1.7)

where $L_e$ is the effective length and $\varepsilon_r$ is the effective dielectric constant of the patch antenna. It can be seen from the equation that the resonant frequency $f_r$ is inversely proportional to the $L_e$ and $\varepsilon_r$. The effective dielectric constant depends on the substrate material incorporated between the patch and the ground plane. Therefore as the height between the patch and ground plane is varied, the effective permeability of the substrate changes which influences the resonant frequency of the antenna. In addition, the total length of the antenna also increases by a factor of $\Delta L$ (Figure 1.4, p. 7), which due to fringing fields, affects the resonant frequency.
This concept of ground plane reconfiguration was first introduced around 1980’s when author J. S. Dahele and Kai Fong Lee presented frequency tunability phenomenon of dual circular stacked microstrip patch antenna discs in [25]. Figure 1.5 shows the geometry of the circular discs with the air-gap concept.

![Figure 1.5](image)


The stacked circular discs in Figure 1.5(a) is made frequency agile/tunable by putting air-gaps heights $\Delta_1$ and $\Delta_2$ between the upper and lower discs and/or between the lower disc and the ground plane in Figure 1.5(b). The width of upper air-gap controls the upper resonant frequency while the lower air-gap affects both the upper and lower resonant frequency. By varying the air-gaps thickness, the effective permeability of the system changes inducing change in resonant frequency.

The authors followed the air-gap variation principal in [26] and showed the effect of air-gap variations on circular-disc, annular ring and stacked dual-frequency microstrip patch antenna for their frequency agility by exciting different modes. A comparison of different air-gap height variation $\Delta$ for circular disc is shown in Table 1.1.

The variation of air-gaps between the circular patch and ground plane excites the three modes $TM_{11}$, $TM_{21}$ and $TM_{31}$. The resonant frequency changes with the increase in air-gap heights for all the three modes with impedance bandwidth increasing almost twice.

Further, a frequency tunable mechanically actuated microstrip patch antenna is presented in [27] where an electrostatic potential is employed to deflect the ground plane to
Table 1.1. Measurement of Resonant Frequency and Impedance Bandwidth for Circular Disc When Subject to Air-Gap Variation from $\Delta = 0$ to $\Delta = 1.0$ mm

<table>
<thead>
<tr>
<th>Air-Gap $\Delta$</th>
<th>$f_{nm}$ MHz</th>
<th>% BW</th>
<th>$f_{nm}$ MHz</th>
<th>% BW</th>
<th>$f_{nm}$ MHz</th>
<th>% BW</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\Delta = 0$</td>
<td>1128</td>
<td>0.89</td>
<td>1286</td>
<td>1.48</td>
<td>1350</td>
<td>2.07</td>
</tr>
<tr>
<td>$\Delta = 0.5$ mm</td>
<td>1879</td>
<td>0.85</td>
<td>2136</td>
<td>2.15</td>
<td>2256</td>
<td>2.61</td>
</tr>
<tr>
<td>$\Delta = 1.0$ mm</td>
<td>2596</td>
<td>0.77</td>
<td>2951</td>
<td>1.63</td>
<td>3106</td>
<td>2.02</td>
</tr>
</tbody>
</table>

create air-gaps ranging from 50-600μm. The resonance frequency shifts from 8.13GHz for a height of 150μm to 8.8GHz at 550μm.

An analytical formulation for calculating the shift in resonant frequency by plugging air-gaps between the circular microstrip patch antenna and ground plane is studied in [28]. By varying the air-gap height from $h_1 = 0.508$mm to $h_2 = 1.092$mm, a change in resonant frequency from 16GHz to 18.5GHz is seen.

Ground plane reconfiguration has been studied in [16] where the author detunes the resonant frequency of 3 elements array of patch by varying the substrate height from 210μm to 600μm between the radiating element and the ground plane. Figure 1.6 shows its frequency agility performance.

In [17], the author presented a new type of RF MEMS device for microstrip line phase shifters and frequency agility where the patch antenna achieves a 6% controlled variation in resonant frequency shift by employing electrostatic potential on the ground plane by varying the height from 0 to 138μm. The resonant frequency shifts from 16.9GHz with no air-gap to 17.9GHz with air-gap of 138μm. Figure 1.7 shows the frequency agility of the microstrip patch antenna. Thus it is seen that the ground plane reconfiguration offers promising results and does not alter the radiating element geometry. It in turn, confines its operation at the bottom part of the antenna geometry to provide maximum radiation performance from the patch surface. Moving the ground plane down also makes the antenna electrically large, which increases the aperture area to achieve more gain.

It has been seen that reconfigurable microstrip patch antennas studied so far, implemented either by employing shorting pins, RF MEMS switches, varactor diodes or

ground plane reconfiguration [19], show discrete frequency reconfigurable states. In practical application, a wideband antenna is very desirable. Therefore in the next section, investigations on the methods to increase the impedance bandwidth of wideband antennas and their need in the modern wireless communication systems are shown and discussed.

1.5 NEED FOR WIDEBAND MICROSTRIP PATCH ANTENNAS

Microstrip patch antennas are employed at various sites where size, weight, performance and cost are key requirements. These antennas however suffer on account of narrow bandwidth. Researchers have continuously put their efforts to increase the impedance bandwidth on numerous occasions. This is because with the increasing progress in modern wireless and communication system, an antenna as an end Tx/Rx component should serve for multi-operational applications. Investigations on wideband antennas are also important because they offer several advantages over conventional narrowband antenna.

The author in [28] presented a U-shape slot loading on the rectangular patch fed by a coaxial probe to achieve impedance bandwidth between 20-30%. Impedance matching networks have been investigated in [29] where impedance bandwidth has been increased by a factor of 3.9 using Fano-broadband matching techniques. E-shape patch antenna was introduced in [30] with impedance bandwidth of 30% fed by a coaxial probe. Another E-shape patch presented in [31] offers 19.5% impedance bandwidth with microstrip transmission line as feeding mechanism. Multi-Layer stack elements have been explored in [32] to increase the impedance bandwidth.

An impedance bandwidth up to 20.7% is achieved by employing shorting pins and shorting wall on the quarter wave patch in [33]. This technique helps in reducing the size of the patch significantly. The same concept of shorting wall is incorporated on triangular shape patch to achieve impedance bandwidth of 30% in [34].

However, here in this thesis, a U-slot loaded modified E-shape (USLMES) wideband patch antenna is presented that offers an impedance bandwidth of at least 35% with relatively smaller ground plane. The U-slot is seen imperative parameter to bring the wideband performance. In addition, the variation of air-gap height between the patch surface and ground plane is seen affecting the impedance bandwidth of the USLMES. Further, care has been taken to include the coaxial probe outside the patch area to avoid high cross-
polarization generation due to the coaxial probe. The patch is printed on low cost FR-4 substrate which is a real microwave substrate hence soldering of the coaxial probe is easily possible than when done on a foam substrate. The USLMES patch antenna can find application as base station antennas in addition to other such wireless communication applications.

1.6 Software Tools Used

Ansoft Corporation’s Designer [35] was initially used to design the U-slot loaded modified E-shape (USLMES) patch antenna. It performs the calculations by breaking up the geometry onto triangles and rectangles which is collectively known as Mesh on an infinite dielectric substrate. The ground plane can be made finite. Additionally, to investigate the performance of USLMES as a 3D finite structure with finite boundaries, Finite Element Method (FEM) based 3-D full wave analysis software Ansoft HFSS v.11 [36] was used. The HFSS is an acronym for High Frequency Structural Simulator and is a Maxwell equation solver. Ansoft HFSS has a frequency domain solver that computes numerical techniques to find approximate solutions by solving complex partial differentials equations and integral equations. HFSS can plot the S-parameters, current distribution, E and H field, radiation patterns and Gain for an antenna. The USLMES was further verified by simulating it using the Computer Simulation Technology’s Microwave Studio tool, referred to as CST Microwave Studio [37]. CST Microwave Studio is a time domain solver. It is based on the Maxwell’s equations’ integral form and Fourier transform.

1.7 Organization of Thesis

The thesis is divided into the following chapters. Chapter 2 presents the analysis of ground plane height control method to investigate the frequency agility of a narrowband rectangular patch antenna with its far-field performance. This chapter explains the very basic idea of ground plane reconfiguration technique used in this thesis to achieve frequency agile behavior. Chapter 3 investigates the frequency agile behavior and other properties of single element patch antenna configured in an array fashion of 1×4 patch array configuration. It is seen that the array continues to behave as an array antenna even towards the end of the air-gap height variation. Chapter 4 presents the geometry, design and parametric study of the proposed U-slot loaded modified microstrip patch antenna (USLMES). Chapter 5 presents
the analysis of ground plane control method to investigate the frequency agility of the proposed USLMES and its far-field phase shift performance. Chapter 6 presents the experimental verification and comparison of measured and simulated results for the proposed USLMES. The conclusion and future study has been presented in Chapter 7.
CHAPTER 2

FREQUENCY AGILE BEHAVIOR OF
MICROSTRIP PATCH ANTENNA

A microstrip patch antenna can be designed for any shape and size as per the requirement of the application. Among all the existing geometrical shapes of the patch antenna, rectangular shape is the most widely studied. This section will focus on the design of narrowband rectangular patch antenna following its behavioral response to ground plane height control reconfiguration method.

Before designing a microstrip patch antenna, certain fundamental requirements should be considered for the optimal performance. The antenna size plays an important role, so care must be taken to keep it small and return loss minimum as possible. The radiation pattern performance and gain requirements should be achieved in the desired area of frequency band. Sometimes it is seen that the gain becomes zero at some frequency. This could arise due to currents cancelation on the patch surface leading to less or no gain at that frequency. Care should be taken while exciting modes in the patch antenna. Higher order modes are very undesirable that can arise due to wrong position of the feed or other parameters. The position of feed is a very critical parameter while designing a patch antenna. A probe feed at undesired position deteriorates the radiation pattern by putting its own radiation. Hence in the following section, proper care has been taken to design a rectangular microstrip patch antenna.

2.1 GEOMETRY AND CONCEPT

The geometry of a rectangular patch antenna fed by a 50Ω transmission line through a quarter wave transformer is shown in Figure 2.1(a, b) where L, W, QTL and TL are patch’s length, width, quarter wave length and transmission line length, respectively. A quarter wave transformer is essentially a λ/4 transmission line incorporated between the transmission feed line and patch to match the feed line input impedance to the patch input impedance. As shown in Figure 2.1, the rectangular patch antenna is designed on a FR-4 microwave
Figure 2.1. Geometry of a microstrip transmission line fed rectangular patch antenna. (a) Top view, (b) Side view, and (c) Concept of ground plane variation from the FR-4 substrate. Patch design parameters are: \( L=24 \), \( W=32 \), \( Q_{TL}=16.75 \), \( T_L=4.25 \text{mm} \).

The ideal reference rectangular patch antenna shown in Figure 2.1 does not have any air-gap initially. The antenna parameters are optimized to bring resonance below \(-10 \text{dB}\) at 2.85GHz. The Reflection coefficient magnitude \( S_{11} \) (dB) for the rectangular patch antenna is shown in Figure 2.2. The magnitude of reflection coefficient at 2.85GHz is \(-41.02 \text{dB}\). This reference rectangular patch antenna will serve as the initial design to implement ground plane reconfiguration by varying the ground plane heights below the patch surface.

Figure 2.3 shows the radiation pattern plot for the reference rectangular patch antenna. The co-polarization and cross-polarization gain of the antenna is 2.03dBi and
Figure 2.2. Reflection coefficient magnitude $S_{11}$ (dB) for rectangular patch antenna.

Figure 2.3. Gain radiation pattern at 2.85GHz for the reference rectangular patch antenna.
-38.26dBi. The two co-polarization cut-planes (Etheta phi=0deg and Ephi phi=90deg) overlap each other significantly exhibiting stronger radiation performance at broadside θ=0°. It is seen that cross-polarization cut planes are sufficiently low and does not overlap with the main broadside co-pole beam.

This reference rectangular patch antenna is studied further to achieve frequency reconfiguration by varying the substrate height $h_a$ below the patch. As stated earlier the height $h_a$ represents air-gap variation. Figure 2.4 shows the discrete frequency reconfigurable states of the rectangular patch antenna when subjected to different air-gaps height ($h_a$) variation from the reference (0mm) (no air-gap) to 2mm. The resonant frequency, which is a function of effective dielectric constant, resonates initially at 2.85GHz followed by multiple resonances at 3.09GHz, 3.21GHz, and 3.375GHz and so on until 3.95GHz, all with narrow impedance matching bandwidth. Therefore, patch offers a frequency agility of 2.85GHz to 3.95GHz, which accounts to around 33% frequency agility. In all the cases, patch shows impedance matching better than $S_{11} = -10$ dB.

![Figure 2.4. Reflection coefficient magnitude S11 (dB) versus frequency (GHz) of the patch antenna for different air-gap height (ha) variations.](image)
Figure 2.5 shows the Gain (dBi) variation of the patch antenna with different air-gap heights $h_a$. The linear increase in Gain from 2dBi to almost 7dBi is attributed to the patch’s characteristic of behaving as an electrically large aperture antenna as the resonance frequency increases from 2.85GHz to 3.95 GHz because the physical dimensions of the patch remain the same.

Figure 2.5. Gain (dBi) versus frequency (GHz) of the patch for different air-gap height variations.

Figure 2.6 shows the center frequency shift for different air-gap height $h_a$ variation. As we can see from Figure 2.4 (p. 19), the resonant frequency is seen shifting to the higher end, and center frequency shows a linear increase with increase in air-gap height. At no-airgap (0 mm reference) the center frequency is at 2.85GHz. The Center Frequency shifts to 3.21GHz at 0.2mm air-gap height. The CF shifts to 3.955GHz at air-gap height of 2mm.

Figure 2.6. Center frequency shift (GHz) for different air-gap height variation.
Figure 2.7 shows the percentage bandwidth for different air-gap height variation. The reference (0mm) has percentage bandwidth of 3.15%. With the increase in air-gap thickness to 0.1mm the percentage bandwidth decreases to 2.58%. A further increase in air-gap height to 0.2mm decreases the percentage bandwidth to 2.49%. However with the increase in air-gap to 0.4mm, the percentage bandwidth increases to 3.73%. The percentage bandwidth continues to increase and achieves maximum of 4.92% at 1.2mm air-gap. A further increase to 2mm decreases the percentage bandwidth to 4.29%. This analysis exhibits the narrowband performance of the rectangular patch antenna.

![Figure 2.7. Percentage bandwidth (%) for different air-gap height variation.](image)

Antenna gain [38] is defined as the radiation intensity of an antenna in a given direction to the radiation intensity that would be produced by a hypothetical isotropic (ideal) antenna that radiates equally in all directions and has no losses. Mathematical interpretation is expressed as:

$$G = \frac{4\pi U(\theta, \phi)}{P_{in}}$$  \hspace{1cm} (2.1)

where U is the radiation intensity defined as “the power radiated from an antenna per unit solid angle and P_{in} is the power supplied to the antenna.

Total gain can be disintegrated into co-polarization gain and cross-polarization gain. The co-polarization gain is the component of electric field vector in the desired polarization...
and the cross-polarization gain is defined as the electric field vector orthogonal to the desired polarization.

Figure 2.8 shows the peak cross-polarization levels for the rectangular patch antenna when subjected to different air-gap height $h_a$ variations. The reference (0mm) antenna has co-polarization gain of 2.03dBi and cross-polarization gain of $-37.39$ dB at 2.85GHz which is 39.42dB down from the main beam. As already seen above, the co-polarization gain increases with increase in air-gap height. Thus at air-gap 0.1mm, the antenna resonates at 3.10GHz with the co-polarization gain of 3.29dBi and cross-polarization gain of $-36.54$ dB which is 39.83dB down. It is noticed that with the increase in air-gap height, the peak cross polarization level decreases which is attributed to increase in co-polarization gain since the antenna behaves as electrically large aperture with the increase in resonant frequency. At air-gap height of 2mm, the co-polarization gain is 7.17dBi and cross-polarization gain is $-25.15$ dB which is 32.32dB down from the main beam.

![Figure 2.8. Peak cross polarization level (dB) for different air-gap height variation.](image)

The far-field phase shift property of rectangular microstrip patch antenna is discussed next. It is a well known fact that the resonant nature of microstrip patch antenna can be utilized to change the phase of the radiated fields [16]. The change in resonant behavior can be due to dimensional change in the geometry or by a reactive loading of its cavity such as an aperture on its ground plane [16]. Here the phase properties have been investigated by
varying the air-gap height $h_a$, which changes the resonant behavior of the antenna. Figure 2.9 shows the far-field phase and achieved far-field phase shift for the radiated fields. For all the air-gap height $h_a$ variation, the antenna return loss is better than $S_{11}<-10$dB. It can be noted that about 32° phase shift is achieved, which is due to variation of the substrate’s effective permittivity attributed to the change in airgap height variation.

In this chapter, a narrowband rectangular patch antenna was presented that exhibited frequency agile behavior by variation of ground plane at different heights. The change in resonant frequency is seen sensitive to other antenna properties like gain, center frequency, percentage bandwidths, peak cross-polarization level and far-field phase properties. This antenna is suitable for applications that require discrete frequency operations. In the following section, a wideband microstrip patch antenna is proposed that offer wideband performance for frequency reconfigurable states.
CHAPTER 3

FREQUENCY AGILE BEHAVIOR OF
MICROSTRIP PATCH ANTENNA ARRAY

In the previous chapter we have studied the frequency agile behavior of the single rectangular patch antenna. However in this chapter, we will study the frequency agile behavior of the rectangular patch antenna in an array form. Usually the radiation pattern of a single radiating element is relatively wide, and each element provides low directivity (gain). However, in many applications, it is necessary to design antennas with high directive characteristics (high gains) to meet the demands of long distance communications. This can only be accomplished by increasing the electrical size of the antenna. Enlarging the dimensions of a single elements often lead to more directive characteristics. Another way to enlarge the dimensions of the antenna, without necessarily increasing the size of the individual elements, is to form an assembly of radiating elements in an electrical and geometrical configuration [1]. This new antenna, formed by multielements, is referred to as an array. In most cases, the elements of an array are identical. This is not necessary, but it is often convenient, simpler, and more practical. The individual elements of an array may be of any form (in our case rectangular). The total field of the array is determined by the vector addition of the fields radiated by the individual elements. This assumes that the current in each element is the same as that of the isolated elements. This is usually not the case and depends on the separation between those elements. To provide very directive patterns, it is necessary that the fields from the elements of the array interfere constructively (add) in the desired directions and interfere destructively (cancel each other) in the remaining space.

Ideally this can be done, but practically it is only approached. In an array of identical elements, there are five controls that can be used to shape the overall pattern of the antenna [1]. These are:

- The geometrical configurations of the overall array (linear, circular, rectangular, spherical, etc.)
- The relative displacement between the elements
• The excitation amplitude of the individual elements.
• The excitation phase of the individual elements
• The relative patterns of the individual elements

**3.1 FEED NETWORKS**

The array as discussed above is very versatile and used to synthesize a required pattern that cannot be achieved with a single antenna element. But this can be only achieved if they are properly fed and the fields form individual elements do not cancel. Therefore the elements can be fed by a single line as shown in Figure 3.1(a) or by a multiple lines in feed network arrangements, as shown in Figure 3.1(b). The first is referred to as a series feed network as a single line feeds the entire set of elements, while the second is known as a corporate-feed network [1]. In this thesis, array is designed by using corporate feed network. This is because with corporate feed -network, additional control can be achieved on the radiating elements (amplitude and phase). In addition, both planar and linear array architecture can be implemented using a corporate feed network.

![Figure 3.1. Feed arrangements for microstrip patch array antenna (a) Series feed and (b) Corporate feed. Source: C. A. Balanis, Antenna Theory Analysis and Design, 3rd ed. New York: John Wiley & Sons, 2005.](image)

**3.2 MODIFIED RECTANGULAR PATCH GEOMETRY**

In the last chapter we studied the effect of ground plane height movement on the frequency agile behavior of the rectangular patch antenna. We however noticed that the co-polarization gain of the rectangular patch was around 2.85dBi. In this section, an effort has been made to increase the co-polarization gain of the rectangular patch by further optimizing
its parameters such as length and width and cutting inset slot on the feeding arm. The inset slot is an important parameter that is used to match the input impedance of patch to the input impedance of the feeding transmission line. The rectangular patch is fed through a 50Ω coaxial probe. Figure 3.2(a) shows the geometry of the modified single element rectangular patch which will be utilized to make an array in a linear 1×4 patch configuration.

![Figure 3.2. Modified rectangular patch (a) Top view and (b) Side view.](image)

Figure 3.2(b) shows the side view of patch antenna. We will see in section 3.4 that the co-polarization gain of modified rectangular patch antenna is significantly higher and serves as one of the building block for the linear array geometry presented in the next section.

### 3.3 Array Geometry

Figure 3.3(a) shows the geometry of a 1×4 rectangular patch linear array with a corporate-feed network excited by a coaxial probe. The advantage of this feeding network is that we have to feed the antenna only through one port, and thus minimizing the array complexity in terms of feeding at four points for practical implementation. In this study, the patch dimensions are those of modified rectangular patch that is 40 mm to 23 mm (L×W) designed on FR-4 substrate ($\varepsilon_r = 4.4$, $\tan \delta =0.02$) with thickness $h_s = 60$ mil backed by a ground plane. The inset slot is cut in each of the rectangular patch element to match the impedance of the transmission line to the patch input impedance. The width of the quarter wave transformer has been optimized to match the 50Ω input impedance of feeding line to the input impedance of the individual radiating element. The total dimension of the entire array design is 105×220 mm².
Figure 3.3. (a) Top view of the 1×4 element linear array, (b) Side view, (c) Concept of ground plane reconfiguration.

Figure 3.3(b) shows the side view of the array antenna with the array elements on the top followed by FR-4 substrate of thickness $h_s = 60$ mil and backed by a ground plane. Figure 3.3(c) shows the concept drawing of the ground plane reconfiguration. This is explained in section 2.2 by changing the effective permittivity of the dielectric substrate which induces change in overall capacitance of the system. The resonant frequency being a function of dielectric constant changes each time the air-gap height $h_a$ changes.

### 3.4 RESULTS AND DISCUSSIONS

The modified rectangular patch antenna shown in Figure 3.2 (p. 26) is designed to resonate at 3.00GHz. Figure 3.4 shows the plot for reflection coefficient magnitude $S_{11}$ (dB) with impedance matching of better than $<-10$dB at $-26.50$ dB for the modified rectangular patch antenna.

Further, Figure 3.5 shows the gain radiation pattern for modified rectangular patch antenna at 3.00 GHz. The co-polarization and cross-polarization gain of the antenna is
Figure 3.4. Reflection coefficient magnitude $S_{11}$ (dB) for modified rectangular patch antenna.

Figure 3.5. Gain radiation pattern at 3.00 GHz for the modified rectangular patch antenna.
5.46dBi and −18.58dBi. The two co-polarization cut-planes (Gaintheta-phi = 0 deg and Gainphi- phi = 90 deg) overlap each other significantly exhibiting stronger radiation performance at broadside θ=0°. It is seen that cross-polarization cut planes are sufficiently low and does not overlap with the main broadside co-pole beam. Thus this 5.46dBi co-polarization gain of single modified rectangular patch antenna increases our confidence to expect higher gain value for its array configuration.

Having studied the performance of modified rectangular patch antenna for its reflection coefficient magnitude S11 (dB) and gain pattern, we are now interested in studying the array behavior of this patch antenna in a linear 1×4 array configuration for its frequency agile behavior.

Referring to Figure 3.3 (p. 27), initially the array is designed to resonate at 3.00GHz. The reflection coefficient magnitude S11 (dB) for the reference array design with no air-gap height variation is shown in Figure 3.6. The array antenna is matched better than −10dB with the best reflection coefficient magnitude of −17.2dB. Figure 3.7 shows the gain radiation pattern for the reference array antenna. The co-polarization gain in Gaintheta-phi = 0 deg cut plane is 11.295dBi which is higher than the single element modified rectangular patch antenna co-polarization gain of 5.46dBi. Figure 3.7 also reveals that the configuration is behaving as an array configuration because the pattern is highly directive. This can also be confirmed by seeing the Gainphi-phi = 90 deg cut plane which has a very narrow, highly directive beam. The cross-polarization gain is −16.549dBi and the co-polarization and cross-polarization are far apart from each other.

Further, Figure 3.8 shows the effect of varying the air-gap height h_a from h_a= 0 mm to h_a= 2.0 mm on the frequency agile behavior of the array antenna. As expected, we see a series of multiple resonances from 3.0GHz to 4.3GHz. This happens, because resonant frequency being function of effective dielectric constant of the substrate material, changes each time the antenna is loaded with different airgap heights. Figure 3.8 reveals the frequency agile behavior of microstrip patch antenna array accounting around 43% of frequency agility.

For an array of n elements, the far-zone field consists of the product of the field of a single element, at a selected reference point and the array factor of that array [1]. This array factor depends on the number of elements in an array and configures the E-field parameters
Figure 3.6. Reflection coefficient magnitude $S_{11}$ (dB) for reference ($h_a=0\text{mm}$) array antenna.

Figure 3.7. Gain radiation pattern of reference ($h_a = 0\text{mm}$) array antenna.
Figure 3.8. Reflection coefficient magnitude $S_{11}$ (dB) of the microstrip patch antenna array at different airgap height $h_a$ variations.

Accordingly, Figure 3.9 shows the co-polarization cut planes Gaintheta-phi = 0 deg for the different air-gap height $h_a$ variations. It should be noted that as the height of the substrate increases from $h_a = 0$ mm to $h_a = 2.0$ mm, the co-polarization cut plane Gaintheta-phi = 0 deg shifts from the broadside ($\theta = 0^\circ$) towards the left side. This is accounted to the fact that as the height $h_a$ increases, the array behaves electrically large for the same physical dimensions.

Figure 3.10 shows Gainphi-phi = 90 deg cut-plane gain patterns for the different airgap height $h_a$ variations. It is seen that Gainphi-phi = 90 deg cut plane shows highly directive pattern throughout the airgap thickness variation. Table 3.1 presents the co-polarization gain values for Gaintheta-phi = 0 deg and Gainphi-phi = 90 deg at broadside $\theta = 0^\circ$ and peak co-polarization gain for Gaintheta-phi = 0 deg at shifted scanned angles with reference to $\theta = 0^\circ$.

It can be noticed from Table 3.1 that both the co-polarization cut plane Gaintheta-phi = 0 deg and Gainphi-phi = 90 deg show the same co-polarization gain values at broadside $\theta = 0^\circ$. However, the cut plane Gaintheta-phi = 0 deg scans through $0^\circ$-$15^\circ$ with respect to $\theta = 0^\circ$ and exhibits higher peak co-polarization gain as the height increases to...
Figure 3.9. Co-polarization cut plane Gaintheta-phi = 0 deg for the different air-gap height $h_a$ variations.

Figure 3.10. Co-polarization cut-plane Gainphi-phi = 90 deg for the different air-gap height $h_a$ variations.
Table 3.1. Gaintheta-Phi = 0deg and Gainphi-Phi = 90deg Co-Polarization Gains for Different Variations of Heights at Broadside (θ=0 deg)

<table>
<thead>
<tr>
<th>Air-gap Height $h_a$ (mm)</th>
<th>Gaintheta-phi = 0 deg (dBi) at θ=0 deg</th>
<th>Gainphi-phi = 90 deg (dBi) at θ=0 deg</th>
<th>Peak co-polarization gain of Gaintheta-phi = 0 deg (dBi) at shifted angles (Degrees)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>11.295</td>
<td>11.295</td>
<td>11.295/0°</td>
</tr>
<tr>
<td>0.2</td>
<td>12.165</td>
<td>12.165</td>
<td>12.46/10°</td>
</tr>
<tr>
<td>1.0</td>
<td>11.224</td>
<td>11.224</td>
<td>13.65/15°</td>
</tr>
<tr>
<td>1.4</td>
<td>10.819</td>
<td>10.819</td>
<td>13.84/15°</td>
</tr>
<tr>
<td>1.6</td>
<td>11.530</td>
<td>11.530</td>
<td>14.36/15°</td>
</tr>
<tr>
<td>2.0</td>
<td>9.080</td>
<td>9.080</td>
<td>14.46/15°</td>
</tr>
</tbody>
</table>

$h_a = 2.0\text{mm}$. But importantly the array antenna continues to behave as an array antenna at different airgap heights along with being frequency agile.

Figure 3.11 shows the center frequency shifts for the single rectangular patch and the array at different airgap height $h_a$ variations. It is seen that the center frequency shifts towards higher frequency values for the same airgap distance but remains in accordance with the single element center frequency behavior shifts. This can be explained with reference to Figure 3.7 (p. 30) where the array shows 43% frequency agility and single element shows around 33% frequency agility for the same air-gap height variations. Finally, Figure 3.12 shows the far-field phase and achieved far-field phase shift for the radiated fields. It can be seen that the achieved phase shift is a delay as it goes in the negative values.

Figure 3.11. Center frequency shift for a single element and the array for different airgap height $h_a$ variations.
Hence in this chapter, we were able to see the frequency agile behavior of an array antenna when subjected to different air-gap height variations. The co-plane cut plane Gainphi-phi = 90 deg of the radiated fields of the array antenna preserved the behavior of the array antenna by showing highly directive beam during all the airgap height variations. This study also focused on frequency agile behavior for discrete reconfigurable states and thus in the following chapters, we will lay emphasis on wideband antenna for its wideband reconfigurable states. A novel microstrip patch antenna is presented next that shows wideband reconfigurable states which will be shown in Chapter 5.
CHAPTER 4

NOVEL U-SLOT LOADED MODIFIED E-SHAPE (USLMES) PATCH ANTENNA

In the previous sections, we investigated the performance of a narrow band rectangular patch antenna and array for their frequency agility and other properties under the influence of ground plane height variation. The resonant frequency showed considerable shifts through the frequency band but offered discrete reconfigurable states. In order to fulfill the ever increasing requirement to incorporate one antenna for varied applications, a wideband antenna with larger impedance bandwidth is a must. In modern day wireless and communication systems, wideband antennas are very desirable because they offer various advantages over narrowband antennas. They address wide range of applications namely radars, base stations, multi-band satellite communications, electronic warfare and multifunctional systems [32]. In addition, wideband antennas possesses approximate or exactly the same operating characteristics over a very wide passband [39].

Researchers in the past have put several efforts with their work to come up with different geometries to increase the impedance bandwidth of the microstrip patch antenna. The U-slot loaded rectangular patch antenna is presented in [28] with 30% bandwidth. Author in [30] presented an E-shape patch with 30.3% impedance bandwidth with ground plane size of 140×120 mm^2 ground plane. The E-shape patch is further explored in [31] that achieves 19.5% impedance bandwidth with transmission line feed.

In this chapter, a wideband microstrip patch antenna is presented that utilizes a U-slot loading on the E-shape geometry studied from the literature. Both the U-slot loading and E-shape patch antenna studied in the past shows around 30% impedance bandwidth. But, here in this thesis, a U-slot loaded modified E-shape wideband patch antenna (USLMES) is presented that offers an impedance bandwidth of at least 35% with relatively smaller ground plane. Further, care has been taken to include the coaxial probe outside the patch area to avoid high cross-polarization generation due to the coaxial probe. The patch is printed on low cost FR-4 substrate which is a real microwave substrate.
4.1 ANTENNA GEOMETRY

The geometry of the proposed U-slot loaded modified E-shape (USLMES) notch fed patch on a microwave substrate FR-4 ($\varepsilon_r = 4.4$, $\tan \delta = 0.02$) of thickness $h_s = 30\text{mil}$ placed on a foam substrate ($\varepsilon_r = 1.06$) of thickness $h_a = 4.8\text{mm}$ excited by a notch fed through a $50\Omega$ coaxial probe is shown in Figure 4.1. The microwave substrate FR-4 is employed for ease of fabrication and SMA soldering to the patch surface. The patch parameters are: $L = 40\text{mm}$, $W = 60\text{mm}$, $E_W = 16.6\text{mm}$, $E_L = 17.2$, $E_{SW} = 7.2\text{mm}$, $N_W = 2.9\text{mm}$, $I_S = 0.2\text{mm}$, $N_{SL} = 6.5\text{mm}$, $F_W = 6.2\text{mm}$, $F_P = 6\text{mm}$, $U_W = 58\text{mm}$, $U_L = 8.6\text{mm}$, $U_{SW} = 5\text{mm}$, $h_a = 4.8\text{mm}$, and $h_S = 30\text{mil}$.

![Figure 4.1. Geometry of a U-slot loaded modified E-shape (USLMES) wideband patch antenna: (a) Top view, and (b) Side view.](image)

4.2 RESULTS AND DISCUSSIONS

The imperative parameters which were optimized extensively to bring matching level with respect to (w.r.t.) $S_{11} < -10\text{dB}$ are E-shape patch wings width ($E_W$), length of the E-wing ($E_L$), E-slot width ($E_{SW}$), Notch width ($N_W$), inset slot width ($I_S$), feed width ($F_W$), U-slot width ($U_W$), U-slot wing width ($U_{SW}$) and U slot length ($U_{SL}$). Thorough parametric study was
conducted to attain optimized values for the above design parameters. To understand the effect of USLMES patch antenna parameters on its impedance bandwidth, individual design parameter are varied, one at a time, or a set of parameters, while keeping all other parameters invariant. The simulation were generated using the Ansoft Designer and are shown in the figures. Ansoft Designer models infinite substrate while ground plane can be of finite size. For all the simulation, impedance bandwidth is defined with reference to $S_{11} = -10\text{dB}$.

Figure 4.2 shows the effect of varying the U-Slot edge distance (Z) to the USLMES edge. The distance Z is varied from 11.4mm to 16.4mm. At Z = 11.4mm the USLMES shows dual band response with first band operating in 3.05GHz to 3.2GHz (4.80 % bandwidth) and second in 3.90GHz to 4.35GHZ (10.90 % bandwidth). A further increase of distance to Z = 12.4mm shows dual band response from 3.05GHz to 3.25GHz (6.34% bandwidth) and 3.80GHz to 4.37GHz (13.95% bandwidth). At distance Z = 16.4mm, the USLMES shows a wideband response from 3.09GHz to 4.42GHz exhibiting 35.41% impedance bandwidth.

![Figure 4.2. Effect of U-slot distance to USLMES patch edge (Z) on the reflection coefficient magnitude $S_{11}$ (dB).](image-url)
Figure 4.3 shows the effect of varying U-slot width $U_W$ from 46-58mm. At width $U_W = 46$mm, the USLMES shows dual band response with first band operating in 2.90GHz to 3.18GHz (9.21% bandwidth) and second in 3.75GHz to 4.52GHz (18.62% bandwidth). Increasing U-slot width to $U_W = 50$mm again shows dual band response with decrease in first band bandwidth (3.0GHz to 3.25GHz, 8% bandwidth) and increase in second band bandwidth (3.65GHz to 4.48GHz, 20.41% bandwidth). A further increase in slot width to $U_W = 56$mm shows a wideband performance (3.08GHz to 4.43GHz, 35.91% bandwidth) but touches the $-10$dB line near 3.5GHz. Therefore on further investigation, U-slot width of $U_W = 58$mm is selected showing wideband response (3.09GHz to 4.42GHz, 35.41% bandwidth).

![Figure 4.3. Effect of varying U-slot width $U_W$ on the reflection coefficient magnitude $S_{11}$ (dB).](image)

The effect of varying the U-slot wing width is shown in Figure 4.4. The U-slot wing width $U_{SW}$ is varied from 3 to 8mm. It is seen to exhibit dual band response at $U_{SW} = 8$mm starting 3.02GHz to 3.3GHz (8.86% bandwidth) and 3.6GHz to 4.45 GHz (21.11% bandwidth). With further decreasing U-slot wing width to $U_{SW} = 7$ mm, USLMES shows a wideband response from 3.05GHz to 4.43GHz (36% bandwidth) but is seen touching the
Figure 4.4. Effect of varying the U-slot wing width $U_{SW}$ on the reflection coefficient magnitude $S_{11}$ (dB).

$-10\text{dB}$ line at 3.5GHz. The above problem is rectified at U-slot wing width $U_{SW} = 5\text{mm}$ with band operating in 3.09 GHz to 4.42GHz (35.41% bandwidth) with better bandwidth performance at $U_{SW} = 3\text{mm}$ (34.89%) also.

Figure 4.5 shows the effect of varying the U-slot length $U_{SL}$ from 2.0 to 1.6mm. By varying the U-slot length $U_{SL}$ from 2.0 to 1.6mm, the bandwidth increases from 33.65 % (3.2 to 4.50GHz) to 35.41 % (3.09 to 4.50GHz).

Figure 4.6 shows the effect of varying the USLMES patch wings width $E_W$ and notch width $N_W$. While investigating the above two parameters, the $E_{SW}$ is kept constant at 7.2mm. Therefore $E_W$ is varied from $E_W = 14$ to 16.6mm and $N_W$ from 5.5 to 2.9mm. At $E_W = 14\text{mm}$, $N_W = 5.5\text{mm}$, USLMES excites higher order modes at higher frequency, therefore on further increasing wing width $E_W = 16.6\text{mm}$ and reducing $N_W$ to 5.5mm, a wideband response is seen operating in 3.09GHz to 4.42GHz (35.41% bandwidth).

The effect of varying the USLMES patch length $E_L$ from 16 to 18 mm is shown in Figure 4.7. The impedance bandwidth increases from 31.96% (3.18GHz to 4.39GHz) at $E_L = 16\text{mm}$ to 33.82% (3.12GHz to 4.39GHz) at $E_L = 16.4\text{mm}$. A further increase of USLMES wing length to $E_L = 17.2\text{ mm}$ increases the impedance bandwidth to 35.41% (3.09GHz to 4.42GHz) which is better than impedance bandwidth of 34.41% (3.08GHz to 4.36GHz) at
Figure 4.5. Effect of varying the U-slot length $U_{SL}$ on the reflection coefficient magnitude $S_{11}$ (dB).

Figure 4.6. Effect of varying USLMES patch wings width $E_W$ and notch width $N_W$ on the reflection coefficient magnitude $S_{11}$ (dB).
Figure 4.7. Effect of varying the USLMES patch length $E_L$ on the reflection coefficient magnitude $S_{11}$ (dB).

$E_L = 18$mm. Therefore $E_L = 17.2$ mm is the best value for the bandwidth of the proposed USLMES.

Figure 4.8 shows the effect of varying the feed width $F_W$ and inset slot width $I_S$ from 5.2 to 6.2mm and 0.70 to 0.20mm, respectively. The USLMES shows dual band response at $F_W = 5.2$mm, $I_S = 0.70$mm operating in 3.08GHz to 3.18GHz (3.50% bandwidth) and 3.6GHz to 4.55GHz (23.31% bandwidth). The USLMES continues to show dual band response at $F_W = 5.4$mm, $I_S = 0.6$mm operating in 3.08GHz to 3.19GHz (3.50% bandwidth) and 3.58GHz to 4.53GHz (23.42% bandwidth). Further increasing feed width $F_W$ to 6.2mm and reducing inset slot width $I_S$ to 0.2mm, wideband response is achieved operating in 3.09GHz to 4.42GHz (35.41% bandwidth).

Figure 4.9 shows the effect of varying the finite ground plane size for $S_{11} = -10$dB. The dual band response at $G = 65 \times 65$mm$^2$ operating in 3.02GHz to 3.30GHz (8.86% bandwidth) and 3.6GHz to 4.43GHz (20.67% bandwidth) is optimized to wideband response at $G = 100 \times 120$ mm$^2$ operating in 3.09GHz to 4.42GHz (35.41% bandwidth).

The notch plays an important role in the feeding mechanism. It is interesting to see the effect of the notch on the impedance bandwidth of the USLMES. Figure 4.10 shows the band response with first band operating in 3.15GHz to 3.80GHz (18.70% bandwidth)
Figure 4.8. Effect of varying the USLMES feed width $F_W$ and inset slot width $I_S$ on the reflection coefficient magnitude $S_{11}$ (dB).

Figure 4.9. Effect of varying the ground plane size of the USLMES on the reflection coefficient magnitude $S_{11}$ (dB).
Figure 4.10. Effect of feeding the USLMES patch with a notch on the reflection coefficient magnitude $S_{11}$ (dB).

and second band operating in 3.95GHz to 4.20GHz (6.13% bandwidth). With the incorporation of the notch, a wideband response is achieved operating in 3.09GHz to 4.42GHz (35.41% bandwidth).

The effect of loading the USLMES patch with a U-slot is shown in Figure 4.11. It is seen that without the U-slot, USLMES shows no matching at all for the desired frequency range. This explains the role of U-slot on the proposed USLMES.

Figure 4.11. Effect of U-slot loading on the reflection coefficient magnitude $S_{11}$ (dB).

The other design parameters did not show any significant effect on the impedance bandwidth and are kept unchanged throughout the parametric study. Table 4.1 shows the design parameters of the final proposed USLMES patch antenna.
Table 4.1. Final USLMES Patch Antenna Design Parameters

<table>
<thead>
<tr>
<th>PARAMETERS</th>
<th>VALUES</th>
</tr>
</thead>
<tbody>
<tr>
<td>USLMES wings width , $E_W$</td>
<td>16.6 mm</td>
</tr>
<tr>
<td>USLMES wings length , $E_L$</td>
<td>17.2 mm</td>
</tr>
<tr>
<td>Notch width , $N_W$</td>
<td>2.9 mm</td>
</tr>
<tr>
<td>Inset slot width , $I_S$</td>
<td>0.2 mm</td>
</tr>
<tr>
<td>Inset slot length , $N_{SL}$</td>
<td>6.5 mm</td>
</tr>
<tr>
<td>Feed width , $F_W$</td>
<td>6.2 mm</td>
</tr>
<tr>
<td>U Slot length , $U_{SL}$</td>
<td>1.6 mm</td>
</tr>
<tr>
<td>U Slot distance to USLMES patch edge (Z)</td>
<td>16.4 mm</td>
</tr>
<tr>
<td>U Slot Width , $U_W$</td>
<td>58 mm</td>
</tr>
<tr>
<td>U Slot wing width , $U_{SW}$</td>
<td>5.0 mm</td>
</tr>
<tr>
<td>Ground plane (X×Y mm$^2$)</td>
<td>100×120 mm$^2$</td>
</tr>
</tbody>
</table>

Since Ansoft Corporation’s Designer models electromagnetic structures with infinite substrate, the USLMES was further optimized as a 3D finite structure including finite ground plane and finite substrate using Finite Element Method (FEM) based 3-D full wave analysis software Ansoft HFSS [34] and CST’s Microwave Studio ver. 2009 v.11 [35]. Figure 4.12 shows the top view and 3D view of the final optimized USLMES patch in HFSS software.

Figure 4.12. (a) Top view of USLMES patch, and (b) 3D side view.
Figure 4.13 shows the simulated reflection coefficient magnitude $S_{11}$ (dB) for the proposed USLMES patch generated through the HFSS and CST. It can be noted that the return loss results using the HFSS and CST tools agree affirmatively showing USLMES operating from 3.09 GHz to 4.4GHz (35% bandwidth).

The USLMES is further investigated for its radiation performance by simulating its radiation patterns at three frequencies. The three frequencies are selected at the start (3.14GHz), middle (3.50GHz) and at the end (4.40GHz) of the impedance bandwidth. Figure 4.14 shows the radiation patterns for the USLMES at 3.14GHz. The co-polarization gain is 9.50dBi and cross-polarization gain is $-18.14$dB. Thus the peak cross-polarization level is 27.64dB down from the main beam. The 3dB beamwidth calculated is 63°. Further investigating the radiation pattern at the center of the band at 3.50GHz in Figure 4.15, we notice co-polarization gain of 10.64dBi and cross-polarization gain of $-19.218$dBi. It is seen that the peak cross-polarization level is 29.858dB down from the main co-polarization cut planes (“Phi” = $0^\circ$ and $90^\circ$). The 3dB beamwidth calculated for this case is 54°. Figure 4.16 shows the radiation pattern at 4.40GHz. The co-polarization gain is 9.31dBi and cross-polarization gain is 2.23dBi with peak cross-polarization level 7.08dB down from the two main co-polarization plane (Phi = $0^\circ$ and $90^\circ$). However, in all the above cases, cross-polarization levels are very low at broadside angle.
Figure 4.14. Simulated radiation pattern at 3.14GHz.

Figure 4.15. Simulated radiation pattern at 3.50GHz.
Figure 4.16. Simulated radiation pattern at 4.40GHz.

Figure 4.17 shows the broadside realized gain (dBi) of the proposed USLMES at different substrate heights $h_a$. It is seen in Figure 4.17 that the co-polarization gain decreases with increase in frequency of operation. The cross-polarization cut planes are relatively higher at 4.40GHz. The gain drop at 4.4GHz can be attributed to comparatively high cross-polarization than lower frequencies which arises due to large electrical length of the antenna towards end of the frequency. With the increase in electrical size, unwanted higher modes can generate causing cross-polarization to go up; however the cross-polarization level is still very good for most of the wireless communications. Further antenna shows 3dB gain bandwidth similar to the impedance matching bandwidth.

The radiation characteristics of USLMES are further verified by using CST’s Microwave Studio. A 3-D gain radiation pattern of the USLMES simulated at the three frequencies (3.14GHz, 3.50GHz and 4.40GHz) using CST Microwave Studio is shown in Figure 4.18, Figure 4.19, and Figure 4.20, respectively. Figure 4.18 shows the gain radiation patterns at 3.14GHz. The co-polarization gain is 9.65dBi which is in close agreement with HFSS generated co-polarization gain of 9.50dBi. Figure 4.19 shows the gain radiation pattern for 3.50GHz, with co-polarization gain of 10.70dBi as compared to co-polarization gain of
Figure 4.17. Simulated broadside realized gain (dBi) for the proposed USLMES at different airgap height $h_a$ variations.

Figure 4.18. 3D gain radiation pattern of USLMES patch generated by CST Microwave Studio at 3.14GHz.
Figure 4.19. 3D gain radiation pattern of USLMES patch generated by CST Microwave Studio at 3.50GHz.

Figure 4.20. 3D gain radiation pattern of USLMES patch generated by CST Microwave Studio at 4.4GHz.
10.64dBi generated by HFSS. Figure 4.20 shows the radiation pattern at 4.4GHz with co-polarization of 9.70dBi as compared to co-polarization gain of 9.31dBi generated through HFSS. In all the three, it is seen that the pattern shows highly directive pattern with maximum gain at the top with red color showing largest intensity. Hence it can be see that the results generated by the software tools (HFSS and CST) comply each other affirmatively. Further, the USLMES shows directional 3D pattern throughout the frequency band.

The E-field distribution on the surface of the USLMES at three frequencies are shown in Figure 4.21, Figure 4.22 and Figure 4.23. Figure 4.21 shows the E-field distribution at 3.14GHz. It can be seen that the entire surface of E-patch shows current distribution with maximum concentration around the U-slot edge facing the feed and edges of the patch. This surface of current distribution is shown in orange to red color. The green color signifies an area of relatively low current at the back of the U-slot. Further, Figure 4.22 shows the E-field distribution at 3.50GHz. It can be seen that the field intensity decreases at the back of the U-slot further and shifts to the front side and near the feed. The maximum field is seen around the inner edge surface of U-slot and notch in red color.

Figure 4.21. E-field distribution of USLMES patch at 3.14GHz.
Figure 4.22. E-field distribution of USLMES patch at 3.50GHz.

Figure 4.23. E-field distribution of USLMES patch at 4.40GHz.
Further investigation of E-fields at 4.40 GHz in Figure 4.23 reveals less concentration of E-fields at back of the U-slot and more on the front side of the U-slot. The maximum concentration is seen around the feed area and inset slot edges. This explains why there is a relatively higher level of cross-polarization at 4.40GHz, as the radiation is influenced by the feed area and transmission line arm.

The next chapter presents a comprehensive study on the effect of ground plane reconfiguration method on the frequency agile behavior of the proposed wideband U-slot loaded modified E-shape (USLMES) microstrip patch antenna by varying the thickness of substrate heights with respect to the patch surface. The air-gap height variations are achieved by employing different thicknesses of foam substrate. It will be seen that the USLMES patch shows wideband frequency agility states that can serve for variety of wireless communication applications such as the base stations.
CHAPTER 5

FREQUENCY AGILE BEHAVIOR OF U-SLOT LOADED MODIFIED E-SHAPE (USLMES) PATCH ANTENNA

In the following sections, USLMES will be studied for its frequency agile behavior by varying the height of ground plane. This variation affects other characteristics of the USLMES such as gain, impedance bandwidth and far-field phase properties of the radiated fields.

5.1 CONCEPT OF FREQUENCY RECONFIGURATION

Figure 5.1 explains the concept of frequency reconfiguration for the proposed wideband USLMES patch antenna. In this figure, we can see the side view of the USLMES patch which consists of the patch layer on the top followed by FR-4 substrate layer backed by foam substrate whose dielectric permeability is equivalent to air (\(\varepsilon_r = 1.06\)) and ground plane at the bottom. The theory of ground plane reconfiguration, as stated earlier for the narrowband rectangular patch antenna in Chapter 2, works with the movement of ground plane with respect to the patch surface. The resonant frequency is a function of effective dielectric constant of the substrate. With the change of substrate height, the effective dielectric permittivity changes which induces change in the resonant frequency. Different thicknesses of foam provide different air-gap height variation \(h_a\) to study the effect on the frequency agility of USLMES patch.

![Figure 5.1. Conceptual drawing of ground plane reconfiguration for the USLMES patch.](image)
5.2 RESULTS AND DISCUSSIONS

The simulation results for this part of the study were generated using the Ansoft Corporation’s finite element method (FEM) based high frequency structure simulator (HFSS) which models all finite dimensions including the substrate and ground plane. The simulation considers interaction of SMA placement close to the antenna, which in this case is outside of the patch. Figure 5.2 shows the simulation results for the USLMES when investigated for different air-gap heights between the FR-4 substrate and the ground plane. The structure initially operates (w.r.t. $S_{11} = -10\text{dB}$) at 4.2GHz at $h_a = 2.8\text{mm}$ height. With the increase in air-gap height ($h_a$) to 3.2mm, the frequency shifts to the lower end and operates at 3.4 GHz showing dual band performance with first band operating in 3.26GHz to 3.6GHz (9.91% bandwidth (BW)) and second operating in 3.91GHz to 4.34GHz (10.42% BW). As the height $h_a$ is increased further to 4.8mm, the frequency shifts more towards the lower end and operates as wideband antenna achieving 35% bandwidth (3.09GHz to 4.40GHz). By further increase in height to 6.4mm, the lower end frequency shifts to 3.0GHz with operational bandwidth of 34.71% (3.0GHz to 4.26GHz). By further increasing the air-gap ($h_a$) to 7.8mm, the USLMES patch resonates between 3.26GHz to 3.75GHz offering 13.98% impedance bandwidth. Thus, the effect of employing different height ground plane reinstates reconfigurable or frequency agile operational bands with single, dual and wideband responses.

Figure 5.2. Reflection coefficient magnitude $S_{11}$ (dB) for the USLMES for different air-gap height variations.
The USLMES patch is investigated for its broadside gain performance at different air-gap height $h_a$ variations. Figure 5.3 shows the broadside Gain at $(\theta=0^\circ)$ for the USLMES patch for different air-gap height variations. The gain remains above 5dBi throughout the operational band for all the cases. The best case from the air gap height study is the reference case with $h_a = 4.8$mm where the realized gain stays above 7dBi throughout the frequency bandwidth from 3.09GHz to 4.42GHz. A slight drop in gain at 4.4GHz is attributed to increase in the cross-polarization, which can be still acceptable for most of the wireless communication applications such as the base station antennas.

![Figure 5.3. Realized broadside gain (dBi) for the USLMES patch for different airgap height variations.](image)

Further, the USLMES is investigated for its center frequency shift under the influence of airgap height variation. From Figure 5.2 (p. 54), it can be seen that with the increase in airgap height, the resonant frequency shifts to the lower end. Figure 5.4 shows the center frequency shift v/s airgap height variation. It can be noted that the same effect is seen with center frequency shifting towards the lower end of the frequency band.

Figure 5.5 shows the effect of varying the airgap height $h_a$ on the USLMES impedance bandwidth. The percentage bandwidth is 5.65% at $h_a = 2.8$mm air-gap height. Further increasing the air-gap height to $h_a = 3.2$mm increase the impedance bandwidth to 7.86%. It is at $h_a = 4.8$mm (proposed USLMES) that the impedance bandwidth increases to
Figure 5.4. Center frequency shift v/s airgap height $h_a$ variation.

Figure 5.5. Effect of airgap height variation on percentage bandwidth.
Further increasing the height $h_a$ to $h_a = 6.8\text{mm}$, impedance bandwidth decreases to 24.81%. The impedance bandwidth further decrease to 15.8% at $h_a = 7.8\text{mm}$.

The USLMES patch is further studied for its far-field phase properties with respect to the different air-gap height $h_a$ variations. The phase of co-polarization cut planes (“$\Phi_i$” = $0^\circ$ and $90^\circ$) is dependent on the resonant nature of the microstrip patch antenna. With the variation of air-gap height $h_a$, the resonant frequency changes inducing change in the phase properties of the radiated fields. Thus it is interesting to note the effect of this dimensional change on the phase of the far-field radiation. Figure 5.6, Figure 5.7, and Figure 5.8 shows the effect of far-field phase variation of USLMES for different airgap height $h_a$ variation at the three resonant frequencies. In all of these figures, the $S_{11}$ was maintained at better than $-10\text{dB}$. Figure 5.6 shows the achieved far-field phase shift when the airgap height is varied from $h_a = 2.8\text{mm}$ to $h_a = 7.8\text{mm}$ at 3.14GHz. It can be noted that about $63.5^\circ$ phase shift is achieved.

Far-field phase shift of $53^\circ$ is achieved by airgap height variation at 3.50GHz. A further investigation finds about $46.37^\circ$ phase shift at 4.40GHz. It is seen that the achieved far-field phase (degrees) drops as the frequency of operation increases.

![Figure 5.6](image_url)  
*Figure 5.6. Far-field phase variation of USLMES patch for different air-gap height $h_a$ variation at 3.14GHz.*
Figure 5.7. Far-field phase variations of USLMES patch for different airgap height $h_a$ variation at 3.50GHz.

Figure 5.8. Far-field phase variations of USLMES patch for different air-gap height $h_a$ variation at 4.40GHz.
Thus the antenna can be reconfigured for a specific frequency range between the frequency range from 3.09GHz to 4.42GHz by employing a variable height ground plane (Figure 5.2, p. 54). The mechanism to implement a variable height ground plane with the desired airgap variation from 2.8mm to 7.8mm is a challenging task. This can be realized using the methods explored in [40-41] but comes at the price of complex fabrication and little airgap variation which will not be sufficient. Therefore, at this point, electronic method for varying the ground plane has not been completed and will be the subject for future studies. However, to prove the finding, three different antennas with three different air-gaps were fabricated and experimentally tested. These are discussed in the next chapter.
CHAPTER 6
PROTOTYPE ANTENNAS AND EXPERIMENTAL VERIFICATIONS

The U-Slot loaded modified E-shape (USLMES) microstrip patch antennas simulated in last chapters are fabricated and experimentally tested. The three USLMES were built on FR-4 substrate of thickness 30mil ($\varepsilon_r = 4.4$, tan $\delta = 0.002$) on foam substrate of thickness 3.2mm, 4.8mm and 6.4mm. The U-slot loaded modified E-shape (USLMES) microstrip patch antenna was fabricated and measured in the Antenna and Microwave Lab at San Diego State University (SDSU) which houses an Anritsu’s vector network analyzer (37269D), LPKF milling machine and anechoic chamber with the capability to measure both far-field and spherical-near field based radiation patterns (Orbit/FR). The photograph of LPKF Protomat S42 milling machine and Anritsu vector analyzer (37269D) is shown in Figure 6.1.

![LPKF Protomat S42](a) ![Anritsu vector network analyzer (37269D)](b)

**Figure 6.1.** (a) LPKF Protomat S42, and (b) Anritsu vector network analyzer (37269D).

Protomat S42 is a compact high speed printed circuit board plotter used to print single sided and double sided circuit layouts that are high in quality and chemical free. It plots the RF designs which are highly precise with unsurpassed accuracy by means of drilling and milling, without any industrial setup. Protomat S42 comes with Circuit Cam software which is used to import the layout designs from any commercial simulators software (in our case...
Ansoft Designer) and Board Master Software to control and monitor the movement of plotter [42].

Anritsu vector network analyzer (37269D) is used to measure the reflection coefficient magnitude $S_{11}$ (dB) of the fabricated USLMES patch. Radiation pattern measurements are performed in the anechoic chamber (Orbit/FR system). An anechoic chamber is a shielded room designed to attenuate electromagnetic energy. They provide shielding environment for radio frequency (RF) and microwaves by suppressing echo and isolating noise from the external environment [43].

6.1 FABRICATION PROCEDURE

In order to fabricate the USLMES patch as a real 3D finite structure, we have to follow series of steps, starting from raw simulated layout design model to a final prototype structure [44]. The procedure is described as follow:

- Bring up the simulated model design by opening it in the CAD software (in our case Ansoft Designer).

- Press layout icon and then export file and select file type GERBER (*.ger).

- In the GERBER export dialogue box under “Layer”, select the “Trace” layer (default), and deselect any others. This selects the USLMES top layer for exporting.

- Since the USLMES has a coaxial probe feeding at its notch, a via hole has to be drilled. For this select the “Layout” icon again in the Ansoft Designer and press export file. Select file type NC Drill file (*.ncd).

- Open these exported files in the Circuit Cam software installed on the computer with the milling machine by pressing “Select File”. Import and then selecting the gerber file. In our case, select the drill Gerber file also.

- Select “BorderOutline” from the menu to create a rectangle contour along the USLMES boundary where the milling machine will cut from the board.

- Press the “Rubout” tab to erase the unwanted area on the design surface (in our case USLMES patch) where we do not want conducting copper.

- Press “Insulate tab” in the Edit menu to specify the design format with sense of direction for the cutting tool to remove the insulating material.

- This file will be saved in *.cam file and *.lmd file format which will serve as a raw file for the circuit board plotter Board Master.

- Open Board Master Software and import the previously saved lmd file by clicking “Import”. 
• Switch on the Protomat S42 and place the safety board first on the working platform. Then place the desired dielectric board (FR-4 in our case) on the safety board. This helps in saving the tip of the milling bit if it accidentally hits the metal base of the machine.

• Locate the working area on the dielectric board where we want our design to be milled.

• Press milling top/bottom tab from the bar on the left corner of the Board Master and hit ALL+ icon followed by Start icon to execute fabrication.

• Cut out the fabricated design (in our case USLMES patch) from the board and feed it with coaxial probe.

Figure 6.2 shows the photograph of the one of the fabricated patch antennas on FR-4 substrate of $h_s = 30$ mil.

![Image of fabricated patch antenna](image)

**Figure 6.2. Photograph of the top view of the notch fed U-slot loaded modified E-shape (USLMES) patch antenna fabricated on FR-4 substrate of 30 mil thickness.**

### 6.2 Experimental Results

The U-slot loaded microstrip patch antenna with $h_a = 4.8$ mm (proposed USLMES) was measured for its reflection coefficient magnitude $S_{11}$ (dB) through Anritsu vector network analyzer (37269D) and is shown in Figure 6.3. The reflection coefficient magnitude $S_{11}$ (dB) graph is shown in Figure 6.4. The reflection coefficient results using the HFSS and CST tools agree well (3.09 GHz to 4.4GHz, 35%), whereas the measured data shows slightly wider matching bandwidth of 38% (3.09 GHz to 4.60 GHz). This disagreement at the end of the band is attributed to fabrication error due to uneven foam surfaces after the patch on FR-4 was placed on top of the ground plane separated by foam. In addition, the USLMES was further investigated for its frequency agile behavior at different air-gap height $h_a$ variations.
Figure 6.3. The photograph of the USLMES patch antenna connected to the Anritsu vector network analyzer for the reflection coefficient magnitude $S_{11}$ (dB) measurement.

Figure 6.4. Simulated and measured reflection coefficient magnitude $S_{11}$ (dB) for the proposed wideband USLMES patch antenna with $h_a = 4.8$mm.
For the experimental verification purposes, three different U-slot loaded modified E-shape (USLMES) patch antennas prototypes were fabricated with airgaps $h_a = 3.2\text{mm}$, $4.8\text{mm}$ and $6.4\text{mm}$ and tested using the Anritsu’s vector network analyzer after proper calibrations. The measured reflection coefficient shows agreement with the simulated reflection coefficient in Figure 6.5.

![Figure 6.5. Measured and simulated reflection coefficient magnitude $S_{11}$ (dB) for three fabricated USLMES versus frequency (GHz).](image)

In Figure 6.5 the simulated $S_{11}$ for the USLMES patch with $3.2\text{mm}$ airgap height shows dual band performance with first band operating in $3.26\text{GHz}$ to $3.6\text{GHz}$ (9.91% bandwidth) and second band operating in $3.91\text{GHz}$ to $4.34\text{GHz}$ (10.42% bandwidth). The measured $S_{11}$ results show reasonable agreement with first band operating in $3.26\text{GHz}$ to $3.71\text{GHz}$ (12.91% bandwidth) and second band $3.91\text{GHz}$ to $4.2\text{GHz}$ (7.15% bandwidth). The USLMES with $4.8\text{mm}$ (proposed, Figure 4.1, p. 36) air-gap height has simulated frequency band from $3.09\text{GHz}$ to $4.42\text{GHz}$ offering 35% impedance bandwidth. In comparison to this, the measured $S_{11}$ covers the above band and extends to $4.60\text{GHz}$ (38% bandwidth). The simulated $S_{11}$ for the $6.4\text{mm}$ airgap patch starts from $3.09\text{GHz}$ and ends at $4.26\text{GHz}$, whereas the measured $S_{11}$ starts at $3.09\text{GHz}$ and ends at $4.26\text{GHz}$ offering 35% impedance bandwidth. Therefore, the simulated and measured $S_{11}$ agree reasonably well over
the operational frequency bands. Slight disagreement in the $S_{11}$ for the three cases is attributed to fabrication errors because of uneven foam surfaces between the FR-4 and ground plane. Finally, it can be seen that the ground plane height variation offers frequency agility over a wide bandwidth. Basically, an antenna with dual band with reasonable bandwidths and/or with single wideband can be realized based on the communication need by ground plane height variation.

Having verified the frequency agile behavior of USLMES patch at three different airgap height $h_a$. The reference USLMES ($h_a = 4.8$ mm) patch is further measured for its radiation pattern performance in the anechoic chamber. Figure 6.6 shows the photograph of the USLMES patch antenna mounted on the receiver side with the Satimo Gain horn at the transmitter end.

![Figure 6.6. Photograph of USLMES patch measurement set-up in the anechoic chamber.](image)

The Gain radiation pattern measurement at 3.14GHz, 3.50GHz and 4.40GHz are shown in Figure 6.7(a-f). The measured co-polarization gain at 3.14GHz is 9.55dBi which is in good agreement with the simulated co-polarization gain of 9.50dBi as shown in Figure 6.7(a, b). Further at 3.50GHz, the measured co-polarization gain is 10.07dBi which is again in good agreement with the simulated co-polarization gain of 10.64dBi in Figure 6.7(c, d). At 4.40GHz, the measure co-polarization gain is 7.65dBi which is in close agreement with the simulated co-polarization gain of 9.31dBi. The measured 3dB beamwidths at 3.14GHz, 3.5GHz, and 4.40GHz are 58.54°, 58.12°, and 69.80°, respectively. Table 6.1 summarizes the simulated and measured results.
Figure 6.7. Simulated and measured gain radiation pattern, (a) Simulated gain pattern at 3.14GHz, (b) Measured gain pattern at 3.14GHz, (c) Simulated gain pattern at 3.50GHz, (d) Measured gain pattern at 3.50GHz, (e) Simulated gain pattern at 4.40GHz, (f) Measured gain pattern at 4.40GHz.
(c) [Graph with legend: Curve Info]

(d) [Graph with legend: Curve Info]
Table 6.1. Comparison of Simulated and Measured Co-Polarization Gain (dBi) and Peak Cross-Polarization Level (dB) for the Proposed USLMES Patch Antenna

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Simulated Co-Polarization Gain (dBi)</th>
<th>Simulated Peak Cross-Polarization Level (dB)</th>
<th>Measured Co-Polarization Gain (dBi)</th>
<th>Measured Peak Cross-Polarization Level (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.14</td>
<td>9.50</td>
<td>27.64</td>
<td>9.55</td>
<td>20.40</td>
</tr>
<tr>
<td>3.50</td>
<td>10.64</td>
<td>29.858</td>
<td>10.07</td>
<td>21.70</td>
</tr>
<tr>
<td>4.40</td>
<td>9.31</td>
<td>7.08</td>
<td>7.65</td>
<td>11.59</td>
</tr>
</tbody>
</table>

Further, Figure 6.8 shows the comparison of the simulated and measured broadside realized gain for the proposed antenna. The measured results comply with the simulated results affirmatively. The gain drop at 4.40GHz can be attributed to comparatively high cross-polarization than lower frequencies which arises due to large electrical length of the antenna towards end of the frequency. With the increase in electrical size, unwanted higher mode can generate causing cross-polarization to go up; however the cross-polarization level is still very good for most of the wireless communications. Further the antenna shows 3dB gain bandwidth similar to the impedance matching bandwidth.

![Figure 6.8. Comparison of the simulated and measured broadside realized gain (dBi) vs. frequency (GHz) for the proposed \(h_a = 4.8\text{mm}\) USLMES patch antenna.](image-url)
CHAPTER 7

FREQUENCY AGILE BEHAVIOR OF USLMES BY INCORPORATING COPPER RIBBON SWITCHES

One approach to achieve frequency reconfiguration is by putting switches, diodes on the geometry of the radiating element. This approach has been reported in literature and has been verified by various authors [4-6, 11-14]. This chapter presents the effect of employing copper ribbon that behaves as ON switches when connected and OFF when disconnected. The presented reference USLMES is modified to incorporate 1.6 mm × 1.6 mm copper switches. These are realized by pasting copper strips at various desired locations indicating ON configuration and removed in OFF configuration.

7.1 MODIFIED USLMES ANTENNA GEOMETRY

Figure 7.1 shows geometry of the modified USLMES with copper ribbon type switches at desired locations. Two conductive arms are incorporated inside the E-patch slot width area to compensate switches of 1.6 mm × 1.6 mm dimensions. These are numbered as switch 4, switch 5, switch 6 and switch 7. All the USLMES parameters remain the same as mentioned in the previous section. Three switches are also incorporated inside the U-slot numbered switch 1, switch 2 and switch 3. They are seen affecting the frequency reconfigurable characteristics of the USLMES under turning ON/OFF condition. The other parameters of USLMES are same as described in the previous chapters. The reference USLMES has $h_a = 4.8$mm.

7.2 SIMULATION RESULTS

Figure 7.2 shows the reflection coefficient magnitude of the reference USLMES with different combinations of switches in operation as shown in Table 7.1. Initially all the switches 1, 2, 3, 4, 5, 6 and 7 are OFF (Case 1) and the USLMES shows impedance bandwidth of 8.86% with band operating in 3.02GHz - 3.3GHz. With turning ON of the switches 1,3,4,5,6 and 7 (Case 2), the modified USLMES shows dual band performance with
Figure 7.1. Geometry of modified USLMES with different copper ribbon switches configurations (a) Top and, (b) Side view.

Figure 7.2. Simulated reflection coefficient magnitude $S_{11}$ (dB) for reference USLMES patch for different combinations of copper ribbon switches turned ON/OFF.
Table 7.1. Copper Ribbon Switches Configurations for Modified USLMES Patch Antenna

<table>
<thead>
<tr>
<th>S. No.</th>
<th>Switch Configuration</th>
</tr>
</thead>
<tbody>
<tr>
<td>Case1</td>
<td>All Switches OFF</td>
</tr>
<tr>
<td>Case2</td>
<td>#1,#3,#4,#5,#6 and #7 ON</td>
</tr>
<tr>
<td>Case3</td>
<td>#2,#4,#5,#6 and #7 ON</td>
</tr>
</tbody>
</table>

first band operating in 3.3GHz-3.55GHz (8.82% impedance bandwidth) and second band operating in 3.85GHz - 4.65GHz (18.82 % impedance bandwidth). At this point the switches # 1 and # 3 are turned OFF and switch # 2 is turned ON with switches 4, 5, 6 and 7 (Case 3). The USLMES shows resonance below −10dB from 3.60GHz - 4.95GHz (31.57 % impedance bandwidth). Thus the combination of 7 switches incorporated on the geometry of modified USLMES contribute to operation in different frequency bands and can be reconfigured for various applications between 3.02GHz-4.95GHz.

7.3 EXPERIMENTAL VERIFICATION

The modified USLMES patch was further fabricated to record the measured data. Figure 7.3 shows the photograph of fabricated modified USLMES for the three cases discussed previously.

![Figure 7.3. Photograph of the modified USLMES patches with 3 different copper ribbon switches configurations.](image)

The fabricated prototypes of modified USLMES were measured for their reflection coefficient magnitude $S_{11}$ (dB) in the Antenna and Microwave Laboratory (AML) at SDSU. Figure 7.4 shows the measured $S_{11}$ for the USLMES patch with the above 3 cases.
Figure 7.4. Measured reflection coefficient magnitude $S_{11}$ (dB) for the reference USLMES with different combinations of copper ribbon switches turned ON/OFF.

The measured $S_{11}$ (dB) for Case 1 shows band of operation from 3.02GHz-3.35GHz (10.36 % impedance bandwidth). With switches 1, 3, 4, 5, 6 and 7 turned ON (Case 2), modified USLMES patch shows dual band operation with first band operating in 3.25GHz-3.6GHz (10.21% impedance bandwidth) and second band operating in 4.00GHz-4.85GHz (19.20% impedance bandwidth). With turning OFF of switches #1 and #3 and turning ON of switches #2, 4, 5, 6 and 7 (Case3), the modified USLMES shows wideband response with band operating in 3.80GHz-4.95GHz (26.28 % impedance bandwidth). Therefore it is seen that the measured results are comparable to simulated results except for the band between 3.60GHz-3.80GHz. This is attributed to fabrication errors because of adhesives and wear tear of copper robbons. The modified USLMES patch was further measured for its radiation pattern in the anechoic chamber. The radiation pattern for the three cases was measured and showed common resemblance and plots. Hence here, only radiation pattern of Case 2 is presented. Figures 7.5, 7.6 and 7.7 show the measured radiation patterns at 3.30 GHz, 3.50 GHz and 4.30GHz. At 3.30 GHz, the co-polarization gain is 6.50dBi and cross-polarization gain is −1.79dBi with peak cross-polarization level of 8.29dBi. The co-polarization gain at
Figure 7.5. Measured radiation pattern of modified USLMES for case 3 copper ribbon switch configuration at 3.30GHz.

Figure 7.6. Measured radiation pattern of modified USLMES for case 3 copper ribbon switch configuration at 3.50GHz.
Figure 7.7. Measured radiation pattern of modified USLMES for case 3 copper ribbon switch configuration at 3.50GHz.

3.50GHz is 7.83dBi and cross-polarization gain is −1.64dBi. At 4.3GHz, the co-polarization gain is 5.23dBi and cross-polarization gain is 0.677dBi. The radiation patterns at all three frequencies show acceptable performance.
CHAPTER 8

CONCLUSION AND FUTURE STUDY

The purpose of this thesis was to design and present microstrip patch antennas that show frequency agile behavior for discrete as well as wideband frequency reconfigurable states by employing variable ground plane heights and ribbon type switches. Chapter 1 talked about the theory of microstrip patch antenna, advantages and disadvantages, various shapes and literature of how researchers have been able to achieve frequency reconfiguration so far. The concept of ground plane reconfiguration is elaborated to explain the basis of frequency reconfiguration technique used in this thesis.

The design of a narrowband rectangular patch antenna is presented in Chapter 2 which exhibits frequency reconfiguration by the shift of ground plane height below the patch surface. It is seen that the patch antenna was able to achieve 33% frequency agility with shift in resonance frequency from 2.85GHz to 3.95GHz. The far-field phase shift of the co-polarization cut planes is discussed that show 32° shift in the radiated field.

Chapter 3 presents investigations on microstrip patch antenna linear array in 1×4 elements configuration for its frequency agile behavior by varying the airgap heights. It is seen that the array antenna achieved frequency agility of about 43% with resonances varying from 3GHz initially to 4.3GHz. It is noted that the array beam (cut plane Ephi-phi=90deg) continued to show highly directive behavior pointing at θ=0° at all times. The Far-field phase properties are discussed.

The proposed wideband U-slot loaded modified E-shape microstrip patch antenna with $h_u=4.8$mm (reference throughout the thesis) is presented in chapter 4. This chapter explains the thorough and exhaustive parametric study done to obtain the final desired values for the proposed U-slot loaded modified E-shape (USLMES) patch antenna. Further, it is shown why it was important to feed the USLMES through notch and how the U-slot played an important role to achieve wideband performance. The reference USLMES patch achieves simulated reflection coefficient magnitude $S_{11}$ (dB) of at least 35%. The simulated gain radiation patterns for reference USLMES at 3.14GHz, 3.50GHz and 4.40GHz show co-
polarization gain of 9.50dBi, 10.64dBi and 9.31dBi, respectively. The E-fields showed directional properties throughout the frequency band.

Chapter 5 presented the frequency agile behavior of the proposed U-slot loaded modified E-shape patch antenna. By plugging different thickness of air-gap height (foam) from $h_a = 2.8\text{mm}$ to $h_a = 7.8\text{mm}$, the resonant frequency shifts is noted showing dual, multiband and wideband performances. The broadside realized gain is plotted for the USLMES patch at all of the above air-gap heights. The reference $h_a = 4.8\text{mm}$ shows gain above 7dBi throughout the band. Frequency agile study is followed by study on center frequency shift, percentage bandwidth and far-field phase shift for the USLMES at different airgap heights. The shift of co-polarization cut plane phase at three resonant frequencies for all the airgap height variation was shown which showed that, about 81.71°, 73.33° and 55.88° phase shifts at 3.14GHz, 3.50GHz and 4.40GHz, respectively was achieved.

In order to verify the frequency agile behavior of the proposed USLMES patch discussed in Chapter 5, Chapter 6 presented the fabrication and experimental verification results. Three different USLMES prototypes were fabricated in the Antenna and Microwave Lab that houses an Anritsu’s vector analyzer (37269D), LPKF Protomat S42 milling machine and an anechoic chamber by Orbit/FR system. It was seen that experimentally measured values agrees with the simulated values affirmatively. In addition the measured reflection coefficient magnitude for the ref USLMES extended to 38% (which was 35% in simulation otherwise). The measured broadside realized gain also showed close agreement with the simulated broadside realized gain.

Chapter 7 discusses frequency reconfiguration achieved through employing copper ribbon type switches on the radiating element. The frequency was seen shifting from 3.02GHz to 4.95GHz when different combination of switches was turned ON/OFF. The prototype antenna was fabricated with different combination of switches. The measured results are seen in consensus with the simulated results other than slight disagreement due to fabrication errors.

For future study, it is desired to investigate a simpler mechanism for the ground plane height actuation by applying electrostatic voltage or other alternative methods such that desired levels of air-gap variations could be obtained. This will offer the ground plane movement in real time. However this is a very challenging task as movement of ground plane
may require high actuation voltage followed by issues like repeatability, very limited ground plane height variation and complex fabrication of the reconfigurable ground plane structures.
REFERENCES


