CHAPTER 3

INVESTIGATIONS ON THE DESIGN OF BROADBAND MONOPOLE MICROSTRIP ANTENNA
3.1 Introduction

The multi-resonant behaviour of a patch on a dielectric substrate yields a broad impedance-bandwidth. Wide impedance bandwidth (e.g., greater than or about 10% suitable for present-day cellular communication systems) for a compact microstrip antenna (MSA) is becoming an important topic in MSA design. The impedance bandwidth (-10 dB return loss) has been reported to be 10% or much greater for a compact design using a shorted patch with a thick air substrate [1,4]. This kind of broadband shorted patch antenna commonly referred to as a planar inverted-F antenna (PIFA) is conventionally fed by a probe [1]. Various feeding methods such as an aperture-coupled feed [2], a microstrip-line feed [3], a capacitively coupled [4] feed or an L-probe [5] feed for exciting shorted patch antennas for broadband operation have been demonstrated.

In general a combination of various approaches would lead to an optimum broadband configuration. The size of a half-wavelength (λ/2) rectangular microstrip antenna (RMSA) is too large in the ultra-high frequency (UHF) band. A shorted λ/4 RMSA has the same resonance frequency as that of a λ/2 RMSA, with half the area [6]. The resonance frequency reduces further as the width of the shorting plate decreases [6,7]. Similarly, compact MSA in circular and triangular configurations is realized by placing shorting posts at the zero potential lines [8,9]. A single shorting post yields a maximum reduction in the resonance frequency of the rectangular, circular, and triangular MSAs [10–12]. The compact antennas have also been realized by cutting slots in regularly shaped antennas. The requirements of these compact broadband MSAs will increase in the future due to the ever-growing miniaturization of communication systems. The bandwidth of the compact MSA has been increased in both planar as well as multilayer configurations [13–15], using thick air substrate [16–18].

Shorting pin loading for compact operation of MSAs is well known. By replacing the shorting pin with a chip resistor [19,20] of low resistance, the required antenna size can be significantly reduced for operating at a fixed frequency. In addition the antenna bandwidth can be greatly enhanced.
MSA in its regular shape cannot yield multi-octave bandwidth because of its resonant nature. Some modification of the MSA configuration is required to obtain an octave bandwidth. If a rectangular patch without the substrate and ground plane is fed at the edge by a coaxial feed with an orthogonal ground plane, then the patch will have an effective dielectric constant equal to 1 with large $h$. Both of these factors yield broad bandwidth. This modified configuration can be thought of as a planar rectangular monopole antenna [21]. Other configurations such as triangular, hexagonal, circular, and elliptical monopoles also yield broad bandwidth. An elliptical monopole with an ellipticity of 1.1 yields bandwidth of 1:11 for VSWR $\leq 2$ [22–24].

Section 3.2 describes fundamental of monopole microstrip antenna. Dual of monopole is described in section 3.3. In section 3.4 bandwidth improvement of monopole microstrip antenna is explained. Recent research on broadband monopole microstrip antennas is discussed in section 3.5. Proposed broadband monopole microstrip antenna with parametric study is presented in section 3.6 [31].

3.2 Monopole microstrip antenna

Planar monopole antennas are fabricated using the printed circuit technology. These antennas are suitable for use in the wireless communication devices.

To understand the characteristics of monopole antenna let us consider a dipole over a ground plane [25]. This resembles to a two-element array of the dipole and its image. The ground plane more than doubles the gain of the element by limiting the radiation in the back directions. We can expect a change in the input impedance as the dipole interacts with its image.
A vertical dipole excites currents in the ground plane, when transmitting, equivalent to its image. The image is vertical (Figure 3.1) and has the same phase as the dipole (even mode). The array radiates its maximum in the direction of the ground plane. The dipole also radiates its maximum pattern along the ground plane.

In general a monopole consists of a single conductor fed out of a ground plane from the centre conductor of a coax. When we include its image (Figure 3.1), the monopole equates to a dipole for analysis. The fields vanish below the ground plane and restricting the fields to the upper hemisphere doubles the gain over a dipole, since only half the input power of the dipole is needed to produce the same field strength.

The input impedance decreases to half that of the equivalent dipole. We can form the image of the voltage source feeding the monopole in the ground plane. The voltage across the input of the equivalent dipole is twice that of the monopole to produce the same current. Therefore, the impedance of the monopole is half the impedance of the dipole. The large value of edge diffraction greatly limits the F/B (front to back radiation) ratio of a monopole when it is placed on a finite ground plane. Figure 3.2 shows the pattern of a monopole [25] when placed on $1\lambda$, $2\lambda$, and $10\lambda$-diameter circular ground planes. The back radiation can be reduced by placing the monopole over a ground plane with circular corrugations that forms a soft surface at the edge when the corrugations are slightly deeper than $\lambda/4$ [26]. When the corrugations are less than $\lambda/4$, the ground plane can support surface waves.
3.3 Cavity-backed slots (dual of a monopole)

A slot that radiates only on one side of the ground plane is the dual of a monopole [25]. As in the case of the monopole, restricting the radiation to above the ground plane doubles the gain. The voltage across the slot determines the field strength. Since the radiated power is only half that of the slot radiating on both sides and having the same peak fields, the input impedance doubles. The already high slot impedance becomes even higher. The cavity must present an open circuit at the slot, or its susceptance must combine with the slot susceptance to resonate. Normally, it is a quarter-wavelength deep. Since many cavities form a box, the waveguide mode determines the propagation constant used to determine the depth.
3.4 Bandwidth improvement using Monopole antenna

It has been observed that the bandwidth of the MSA increases with an increase in the substrate thickness and a decrease in the dielectric constant of the substrate [1,27]. For a thick substrate with a low dielectric constant, a bandwidth of 5% to 10% is obtained. Further increase in the substrate thickness decreases the efficiency of the MSA and increases cross-polar levels.

![Figure 3.3: (a) Microstrip antenna suspended in air, (b) modified Microstrip antenna with side feed, and (c) planar monopole antenna](image)

In [21] it is explained that a long coaxial probe is required to feed the radiating metallic patch suspended in air at a large height $h$ as shown in Figure 3.3(a). The large $h$ increases the probe inductance, and the input impedance becomes too inductive to obtain impedance matching. This large inductive input impedance can be taken care of by feeding the patch with a shorter probe of length $p$ as shown in Figure 3.3(b). In this case, the patch is fed along the periphery and an additional perpendicular ground plane is required. If $h$ is very large, the bottom ground plane would have a negligible effect and hence can be removed. This configuration
becomes similar to that of a planar monopole antenna, as shown in Figure 3.3(c) [21–24]. The planar disc monopole antennas yield a very large-impedance bandwidth, which can be explained in the following two ways:

1. A monopole antenna generally consists of a thin vertical wire mounted over the ground plane, whose bandwidth increases with an increase in its diameter [29]. A planar monopole antenna can be equated to a cylindrical monopole antenna with a large effective diameter.

2. The planar monopole antenna can be viewed as a MSA on a very thick substrate with $\varepsilon_r = 1$, so a large bandwidth is expected. In the radiating metallic patch, various higher order modes will get excited. Since all the modes will have a larger bandwidth, these will undergo a smaller impedance variation. The shape and size of these planar antennas can be optimized to bring several modes within the VSWR = 2 circle in the Smith chart, leading to very large-impedance bandwidth.

3.5 Recent research on planar rectangular monopole antenna

In [21] it has been shown that the increase of $h$ in Fig. 3.4(a) will increase the percentage bandwidth. The side and the front views of a rectangular radiating patch with $L = W = 12$ cm made of a copper plate of thickness 0.1 cm with two orthogonal ground planes are shown in Figure 3.4(a,b). The patch is fed with a 50-$\Omega$ SMA connector of probe length $p$ through a fixed ground plane and the orthogonal ground plane is moveable. For the moveable ground plane spacing $h = 3$ cm from the radiating patch and the probe length $p = 0.4$ cm, the measured input impedance and VSWR plots are shown in Figure 3.4(c,d). Multiple loops occur due to the excitation of various higher order modes of RMSA (Rectangular microstrip antenna). The impedance plot shows less inductive shift due to the smaller value of the feed probe length $p$. Hence the value of $p$ is increased to 1 cm to shift the impedance plot in the clockwise direction.
Figure 3.4: (a) Side and (b) front views of modified RMSA with orthogonal ground planes. Measured (c) input impedance and (d) VSWR plots.

The results are summarized in Table 3.1. The measured BW for VSWR \( \leq 2 \) is from 858 MHz to 988 MHz. For analysis purpose [21,28], this structure can be thought of as a MSA on an air dielectric with an additional perpendicular ground plane.
Table 3.1: Resonance Frequency and BW of RMSA with \( L=W=12 \) cm for different \( h \)

<table>
<thead>
<tr>
<th>Sl. No.</th>
<th>( h (\text{cm}) )</th>
<th>From (MHz)</th>
<th>To (MHz)</th>
<th>BW (%)</th>
<th>Theoretical resonance Frequency (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>3</td>
<td>858</td>
<td>988</td>
<td>14.1</td>
<td>937(^a)</td>
</tr>
<tr>
<td>2</td>
<td>6</td>
<td>752</td>
<td>934</td>
<td>17.7</td>
<td>789(^a)</td>
</tr>
<tr>
<td>3</td>
<td>18</td>
<td>515</td>
<td>1081</td>
<td>70.3</td>
<td>484(^a), 523(^b)</td>
</tr>
<tr>
<td>4</td>
<td>Infinity</td>
<td>501</td>
<td>1154</td>
<td>81.2</td>
<td>0(^a), 523(^b)</td>
</tr>
</tbody>
</table>

\(^a\) Calculated using (3.1)
\(^b\) Calculated using (3.8)

Since the dielectric medium for all the cases under consideration is air, the effective dielectric constant is equal to 1. The theoretical resonance frequency for the fundamental mode can be calculated using,

\[
f_0 = \frac{c}{2L_e} \quad (3.1)
\]

Where,

\( L_e = \) effective resonant length; \( c = \) velocity of light in free space.

The theoretical resonance frequency, calculated using (3.1) for \( p = 1.0 \) cm, is 937 MHz, which is close to the measured center frequency of 923 MHz. Next, the effect of increasing \( h \) on bandwidth of the antenna is considered. For different values of \( h \) with \( p = 1.0 \) cm, the measured lower and upper frequencies (\( f_L \) and \( f_H \)) corresponding to VSWR = 2 are given in Table 3.1. With an increase in \( h \) from 3 cm to infinity (\( \infty \)), the percentage bandwidth of the antenna increases from 14.1% to 81.2%. As \( h \) increases, the measured lower resonance frequency decreases because of the increase in \( \Delta L \) due to the large fringing fields. For smaller values of \( h \), there is a reasonable agreement between the theoretical frequency obtained from (3.1) and the measured center frequency. As \( h \) increases, the theoretical frequency is close to the measured lower frequency corresponding to VSWR = 2.

For two different values of \( h \) [i.e., 18 cm (large) and \( \infty \) (bottom ground plane removed)], the measured input impedance and VSWR plots are shown in Figure 3.5. As \( h \) increases from 18 cm to \( \infty \) the measured lower frequency decreases.
slightly from 515 MHz to 501 MHz. For these two values of \( h \), the input impedance and VSWR plots are nearly the same. Therefore, for a large \( h \) tending to infinity, the MSA configuration reduces to that of a planar monopole antenna.

In this case, the approximate lower frequency corresponding to VSWR = 2 can be determined by using the monopole antenna concept described in Section 3.4.2.

![Figure 3.5: Measured (a) input impedance and (b) VSWR plots of RMSA for two values of \( h \): (- - -) 18 cm and (——) \( \infty \).](image)

In wireless mobile communication dipole antennas and loop antennas [32,33] could radiate omni-directional patterns. However, the total volume size of the combined dipole and loop antennas is larger than the monopole antenna of a 1/4 wavelength.

In [34] monopole microstrip antenna for dual frequency operation was proposed where either sideband (1.8/3.3 GHz) can be tuned suitably using independent matching technique. The designed antenna was fabricated on an FR4 microwave substrate (\( \varepsilon_r = 4.4 \)) with an area of 25 x 20 mm\(^2\) and a thickness of 1.6 mm equally divided into a ground plane section and a radiator section. For low frequency (1825 MHz) tuning in the reference antenna structure a coupling strip (CD section) was extended from the right edge of the antenna feed line. This structure is called a Type 1 antenna, as is shown in Fig. 3.6(b). Fig. 3.6(b) shows an increase in the amount of coupling for the coupling strip, which allows the imaginary part of the original impedance distribution curve to reduce, and a resonance ring occurs. This coupling strip forced the impedance of the low
frequency to draw into the circle at VSWR = 2, achieving impedance matching (Figure 3.6).

In type 2 structure a tuning pad (EF section) is introduced in Figure 3.6 (c) between upper and lower spacing of coupling strip. This compensates the high inductive reactance required at 3.3 GHz as the same frequency falls into the centre of the smith chart shown in Figure 3.7 (a,c). The coupling strips have a comparatively strong electric current (Figs. 3.8 a,b) that ensures low frequency tuning (3.7 a,b). Hence the high frequency matching of the Type 1 antenna as shown in Fig. 3.7 (c) deteriorated faster than did the reference antenna. The tuning pad shows a comparatively strong electric current distribution only when operating at high frequency, as shown in Fig. 3.7 (c).

To improve the frequency tuning range the coupling strip and the tuning pad are placed on different sides in Type 3 antennas as shown in Fig. 3.6 (d). This type of structure increases the selection space of low frequency parameter w, and high frequency parameter d as shown in Fig. 3.9 (a).

After optimizing the low frequency peak of Type 3 antennas to 2.943 GHz, the low frequency tuning range became 1.825 GHz - 2.943 GHz, with a frequency ratio achieving 1.6. Hence the proposed control mechanism achieved a maximal tuning range with minimal area including the finite ground.
Figure 3.6: (a) Reference antenna; (b) Type 1; (c) Type 2; (d) Type 3; (e) Type 4; (f) proposed antenna structure for dual band operation.
(a) Return Loss (dB) vs Frequency (GHz)

- --- Reference Antenna
- --- Type 1
- --- Type 2
- --- Type 2 (mea.)

(b) Smith Chart

- 1.75 GHz ~ 2 GHz
- 1825 MHz

- Reference Antenna
- Type 1
Figure 3.7: (a) The simulation of the antenna development from reference antenna to Type 1 and Type 2 antennas; (b) Smith chart for the low frequency matching; (c) Smith chart for the high frequency matching.

Figure 3.8: The current distribution at (a) 1.825 GHz; (b) 3.3 GHz.
Figure 3.9: (a) The parameters of tuning the frequency bands and their corresponding controlling mechanism (b) The simulation result of the antenna frequency tuning
Table 3.2: The performance with different tuning parameters for the low bands.

<table>
<thead>
<tr>
<th>Antenna Type (simulation)</th>
<th>$L$ (mm)</th>
<th>$w_s$ (mm)</th>
<th>$d$ (mm)</th>
<th>The Central Lower Band ($f_L$) Frequency (MHz)</th>
<th>$S_{11}$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Type 2</td>
<td>25</td>
<td>0</td>
<td>7</td>
<td>1825</td>
<td>-22.8</td>
</tr>
<tr>
<td>Type 2–2</td>
<td>20</td>
<td>1</td>
<td>5.5</td>
<td>1921</td>
<td>-19.55</td>
</tr>
<tr>
<td>Type 2–3</td>
<td>17.5</td>
<td>2</td>
<td>5</td>
<td>2042</td>
<td>-20.6</td>
</tr>
<tr>
<td>Type 2–4</td>
<td>15</td>
<td>2.5</td>
<td>3</td>
<td>2194</td>
<td>-26.7</td>
</tr>
<tr>
<td>Type 2–5</td>
<td>10</td>
<td>5</td>
<td>4</td>
<td>2474</td>
<td>-19.3</td>
</tr>
<tr>
<td>Type 3</td>
<td>10</td>
<td>6</td>
<td>4</td>
<td>2943</td>
<td>-19.9</td>
</tr>
<tr>
<td>Type 3–1</td>
<td>8</td>
<td>7</td>
<td>3</td>
<td>3157</td>
<td>-19.8</td>
</tr>
</tbody>
</table>

The relevant antenna performance parameter data are summarized in Table 3.2. Instead of coupled wires or pads an alternative approach of tuning mechanism by using discrete components such as varactors is shown in Fig. 3.6 (e). This design (type 4 antennas) improves antenna impedance matching without inserting a tuning pad that may contribute to a certain amount of insertion loss.

Two wide operating bands at about 900 and 1900 MHz covering GSM 850/900/1800/1900/UMTS penta-band WWAN operation in the mobile phone was reported in [35] using single-strip monopole antenna. The monopole antenna is divided into two sections (Figure 3.10) a front section and an end section due to the chip inductor embedded in between. The embedded chip inductor compensates the increased capacitance seen at the feeding point with the decreasing strip’s resonant length. On the back side of the circuit board, a system ground plane of 100x60 mm$^2$ is printed.
Figure 3.10: Geometry of the proposed single-strip monopole antenna embedded with a chip inductor for penta-band WWAN operation in the mobile phone.

Fig. 3.11 shows the measured and simulated return loss for the proposed antenna. Two wideband resonant modes at about 900 and 1900 MHz are obtained. The first and second modes provide wide bandwidths of 155 MHz (810–965 MHz) and 515 MHz (1675–2190 MHz) to cover the GSM850/900 and GSM1800/1900/UMTS bands, respectively, for WWAN operation
Fig. 3.12 shows the simulated return loss as a function of the inductance $L$ of the chip inductor. The increasing inductance causes the lowering of both the first two modes, and a frequency ratio of about 1 to 2 is maintained. It was noted that when a larger inductance is used, the obtained bandwidth of the lowest mode is decreased with the decreasing of its resonant frequency. Fig. 3.13 shows the simulated return loss for the proposed antenna, the case replacing the chip inductor using a connecting strip of 1.0X1.0 mm$^2$ (Ref 1), and the case with a chip inductor of 10 nH placed near the feeding point (Ref 2). It is seen that the lowest mode of Ref 1 occurs at about 1100 MHz, higher than those of the proposed antenna and Ref 2. In addition, the second mode of Ref 1 occurs at about 3300 MHz; that is, the first two modes have a frequency ratio of about 1 to 3.
Figure 3.12: Simulated (HFSS) return loss of the proposed antenna as a function of the inductance $L$ of the chip inductor.

Figure 3.13: Comparison of the simulated (HFSS) return loss for the proposed antenna, the case without the chip inductor (Ref 1), and the case with a chip inductor of 10 nH placed near the feeding point (Ref 2).
For Ref 2, with the chip inductor having a smaller inductance of $L = 10\,\text{nH}$ than that in the proposed antenna, the lowest mode can also occur at about 900 MHz, similar to the proposed antenna. It was mentioned that when the chip inductance has an inductance of 27 nH, the same as that in the proposed antenna, the lowest mode can be shifted to lower frequencies at about 700 MHz. Also, the second mode of Ref 2 occurs at about 2800 MHz, about three times that of the lowest mode. That is, both the first two modes of Ref 1 and Ref 2 have a frequency ratio of about 1 to 3, similar to the traditional monopole. Hence, an additional strip is required to generate a resonant mode at about 1900 MHz to cover the GSM1800/1900/UMTS bands. A longer strip is also required for Ref 1 to generate a resonant mode for operating in the 900 MHz band.

A monopole antenna capable of multiband operation, covering the 900 MHz band global system for mobile communication (GSM), 1800-MHz-band digital communication system (DCS), 1900-MHz-band personal communication system (PCS), and 2050-MHz-band universal mobile telecommunication system (UMTS) reported in [36]. The proposed antenna shown in Figure 3.14 has a planar rectangular radiating patch in which a folded slit is inserted at the patch’s bottom edge. The folded slit separates the rectangular patch into two sub patches, one smaller inner sub patch encircled by the larger outer one. The proposed antenna is then operated with the inner sub patch resonating as a quarter-wavelength structure and the outer one resonating as both a quarter-wavelength and a half-wavelength structure. In this novel planar monopole antenna design antenna height is less than $0.04\lambda_0$ (the total antenna height is only 12 mm for operating at the 900-MHz band). The radiating rectangular patch has dimensions of $10 \times 30 \,\text{mm}^2$ and is placed on top of the ground plane with a distance of 2 mm.

The open end of the folded slit at the patch’s bottom edge is placed close to the feed point, and the other end inside the patch is also designed to be close to the feed point. In this case, the smaller inner sub patch is encircled by the outer one, which leads to two possible excited surface current paths inside the rectangular patch. The longer path starts from the feed point and follows the folded slit to the open end of the slit at the patch’s bottom edge, while the shorter one is from the feed point to the end of the inner sub patch encircled by the folded slit.
Figure 3.14: Geometry and dimensions of the proposed low-profile planar monopole antenna for GSM/DCS/PCS/UMTS operation.

Figure 3.15: Measured and simulated return loss for the proposed antenna.
Fig. 3.15 shows the measured return loss of the proposed antenna. It is observed from the figure that two wide operating bandwidths are obtained. The lower bandwidth obtained 142 MHz and covers the GSM band (890–960 MHz).

On the other hand, the upper band has a bandwidth as large as 565 MHz and covers the DCS (1710–1880 MHz), PCS (1850–1990 MHz), and UMTS (1920–2170 MHz) bands. The measured data in general agree with the simulated results.

3.5.1 Lower Frequency calculations of the Planar Monopole Antennas

For a planar monopole antenna, the lower frequency corresponding to VSWR = 2 can be approximately calculated by equating its area (in this case, a rectangular disc monopole) to that of an equivalent cylindrical monopole antenna of same height \( L \) and equivalent radius \( r \), as described below [21, 24]:

\[
2\pi rL = WL \quad (3.2)
\]

\[
\Rightarrow r = \frac{W}{2\pi} \quad (3.3)
\]

The length of a monopole for real input impedance is given by [29]

\[
L = 0.24\lambda F \quad (3.4)
\]

Where,

\[
F = \left( \frac{L}{r} \right) \left( 1 + \frac{L}{r} \right) = \frac{L}{L + r} \quad (3.5)
\]

Using (3.4) and (3.5) we find,

\[
\lambda = \frac{L + r}{0.24} \quad (3.6)
\]

Therefore, the lower frequency \( f_L \) is given by:

\[
f_L = \frac{c}{\lambda} = \frac{30 \times 0.24}{(L + r)} = \frac{7.2}{(L + r)} \text{GHz} \quad (3.7)
\]

Equation (3.7) does not account for the effect of the probe length \( p \), which increases the total length of the antenna thereby reducing the frequency. Accordingly, this equation is modified to
\[
    f_L = \frac{7.2}{L + r + p} \text{GHz} \quad (3.8)
\]

where \( L, r, \) and \( p \) are in centimeters. The theoretical frequency of 483 MHz for \( h = \infty \) (monopole antenna) obtained using (3.8) is close to the measured \( f_L \) of 501 MHz. For \( h =18 \text{ cm} \), the theoretical frequencies obtained using the MSA and monopole concepts are very close to each other. Thus, an interesting transition in antenna characteristics (with respect to resonant frequency) was observed, as the ground plane spacing \( h \) is increased. For a smaller \( h \), the measured resonance frequency is close to the theoretical frequency determined by the expressions applicable to the MSA and for a larger \( h \), it was close to the frequency obtained using expressions for a monopole antenna.

### 3.6 Proposed Rectangular Monopole microstrip antenna for broadband operation

In section 3.5 it has been discussed in detail that the impedance-bandwidth of a monopole antenna can be enhanced by removing the movable ground plane. In this section an investigation was carried out on a proposed monopole RMSA [31].

![Figure 3.16: configuration of Rectangular Monopole microstrip antenna](image-url)
The radiating patch of monopole RMSA was fed with 50-Ω transmission line via an impedance transformer. The antenna is designed on a substrate of thickness \( h = 1.6 \) mm and dielectric constant \( \varepsilon_r = 4.4 \). The configuration of the antenna investigated is shown in Figure 3.16 and given as follows (all the dimensions are in mm),

\[
W_p = 16, \quad L_p = 14, \quad L_1 = 7, \quad M_1 = 3, \quad M_2 = 1, \quad M_3 = 3, \quad M_4 = 12, \quad M_5 = 1, \quad L_G = 9, \quad W_G = 38.
\]

3.6.1 Results obtained by different parametric investigations

The proposed antenna is simulated using Method-of-Moment-based IE3D software [30]. Return loss of the same antenna has been measured using Agilent vector network analyser N5230A. Simulated result shows that there exists dual band of wide operating impedance-bandwidth.

![Simulated and measured return loss of the proposed antenna](image)

Figure 3.17: Simulated and measured return loss of the proposed antenna

Figure 3.17 shows the simulated and measured return loss versus frequency of the proposed monopole antenna. A -10dB bandwidth of 2.76 GHz is obtained from 4.66 to 7.42 GHz. Minimum return loss obtained in the same frequency range is about -38.35 dB at 6.78 GHz and about 45.7% bandwidth around the center frequency of 6.04 GHz. Another wide operating bandwidth of 1.18 GHz is
obtained from 9.76 to 10.94 GHz, which is 11.4% bandwidth around the center frequency of 10.35 GHz. In the same frequency range, minimum return loss of -17.16 dB is obtained at 10.18 GHz. The measured result shows similar but wider bandwidth as compared with the simulated result. A bandwidth of about 3.17 GHz (4.19–7.36 GHz) is obtained which is 54.90% around the center frequency of 5.775 GHz. Return loss of about -28 dB is obtained at 6 GHz.

Again a wide bandwidth of about 2.43 GHz (8.695–11.125 GHz) is obtained, which is 24.52% around the center frequency of 9.91 GHz. Return loss of about -18.8 dB is obtained at 10.6 GHz.

Figure 3.18: (a) Simulated and measured radiation patterns at 6.01 GHz and (b) simulated and measured radiation patterns at 10.6 GHz
Table 3.3: Observation with $M_5$ variations

<table>
<thead>
<tr>
<th>Sl. No.</th>
<th>$M_5$ (mm)</th>
<th>Lowest -10dB frequency (GHz)</th>
<th>% of Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>1</td>
<td>5.04</td>
<td>39.14</td>
</tr>
<tr>
<td>2.</td>
<td>3</td>
<td>6.82</td>
<td>25.66</td>
</tr>
</tbody>
</table>

Figure 3.19: Simulated Return Loss for different $M_5$

The simulated and measured E-plane radiation patterns at 6.01 GHz and at 10.6 GHz are shown in Figures 3.18(a,b), respectively. The radiation pattern shows normalized gain in dBi with respect to elevation angle measured in degree. The measured 3 dB Beam width is about 70° obtained at 6.01 GHz and 45° obtained at 10.6 GHz. Table 3.3 shows that due to the smaller value of $M_5$ keeping other dimensions unchanged the lower -10 dB frequency reduces in addition to the increased % of bandwidth. Thus smaller value of $M_5$ (Figure 3.19) results in compact, broadband monopole microstrip antenna. Now a comparative study of the same antenna was carried out due to change in $M_2$ keeping $M_5$ unaltered. Table 3.4 shows that maximum impedance bandwidth that is around 40% can be achieved when $M_2 = 2$mm. But for two different values of $M_5$ if $M_2$ reduces keeping all other dimensions unchanged the lower -10dB frequency also decreases (Figure 3.20-3.21).
Simulated absolute gain vs. Frequency plot of the proposed antenna is shown in Fig. 3.22. Peak gain of about 1.12 dBi was obtained at 10 GHz.
Table 3.4: Observation with $M_3$ variations keeping $M_5$ constant

<table>
<thead>
<tr>
<th>Sl. No.</th>
<th>$M_5$ (mm)</th>
<th>$M_2$ (mm)</th>
<th>Lowest -10dB frequency (GHz)</th>
<th>% of Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>1</td>
<td>1</td>
<td>5.04</td>
<td>39.14</td>
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<td>2.</td>
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<td>2</td>
<td>5.23</td>
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<td>3.</td>
<td>1</td>
<td>3</td>
<td>8.94</td>
<td>23.18</td>
</tr>
<tr>
<td>4.</td>
<td>3</td>
<td>1</td>
<td>6.82</td>
<td>25.66</td>
</tr>
<tr>
<td>5.</td>
<td>3</td>
<td>2</td>
<td>7.81</td>
<td>42.4</td>
</tr>
<tr>
<td>6.</td>
<td>3</td>
<td>3</td>
<td>8.68</td>
<td>32.1</td>
</tr>
</tbody>
</table>

Figure 3.22: Simulated Gain vs Frequency plot of the proposed antenna

3.7 Conclusion
A single layer single feed with reduced ground dimension printed antenna that exhibits wide operating impedance-bandwidth around two center frequencies is simulated using IE3D software, realized and radiation pattern was measured for validation. Radiation patterns show acceptable amount of -3dB beam width.
3.8 REFERENCES


