INVESTIGATION ON THE SIGNAL STRIP MODIFICATION OF A COPLANAR WAVEGUIDE (CPW) TRANSMISSION LINE

This chapter deals with the design and development of a Co-Planar Waveguide (CPW) fed planar antenna. The chapter begins with an elaborative study on open ended Coplanar Waveguide transmission line. A thorough parametric analysis of a CPW is presented to understand the radiation mechanism and radiation performance. From the analysis and detailed study it is found that, the open ended CPW fed transmission line can be transformed into a radiating structure by suitably modifying the signal strip. This modified CPW fed antenna operates in quad bands with omnidirectional radiation pattern with reasonable gain.
3.1 Introduction to Coplanar Waveguide (CPW) transmission line

A conventional Coplanar Waveguide (CPW) on a dielectric substrate consists of a center strip conductor with semi-infinite ground planes on either side separated by a small gap. CPW structures supporting quasi TEM mode have gained great attention in microwave and millimeter wave applications due to uniplanar structure. They are commonly used in Monolithic Microwave Integrated Circuits (MMIC). The CPW transmission lines have lower radiation loss and less dispersion than microstrip lines. Moreover, the characteristic impedance and phase velocity of CPW are less dependent on the substrate height and more dependent on the dimensions in the plane of the conducting surface [1]. Due to this exceptional behavior, CPW structures have been explored a lot for compatible modern wireless communication gadgets.

When a finite length open ended CPW is excited, a standing wave pattern is formed with the reflection coefficient nearly equal to one. The structure is not radiating electromagnetic energy. This is not the case for all frequencies. The device exhibit low values of reflection coefficient and matched at higher frequencies. Thus a particular Finite Ground Coplanar Waveguide (FGCPW) will not radiate electromagnetic energy at lower frequency band.

A conventional 50Ω coplanar waveguide transmission line with ground plane width \( W_g = 20\text{mm} \), ground plane length \( L_g = 20\text{mm} \), signal strip width \( W = 3\text{mm} \) and gap \( g=0.35\text{mm} \) is designed. The front view of such a FGCPW printed on a substrate of dielectric constant \( (\varepsilon_r) \) 4.4 and thickness \( (h) \) 1.6mm is shown in figure 3.1 (a). The side view and 3-dimensional view is also shown in figure 3.1 (b) and 3.1(c) respectively.
The measured reflection characteristic and impedance characteristics of this CPW structure is shown in figure 3.2 (a) and (b) respectively. From the figure it is found that the reflection coefficient is very high in the band and is behaving as an open ended transmission line with low radiation. The smith chart gives a clear picture on the impedance characteristics of this structure from 1GHz to 10GHz. From the figure it is evident that for all frequencies the locus of impedance curve is on the outer region of the smith chart. Thus this open ended structure is unmatched at these frequency regions with very low radiation. This is confirmed with the transmission characteristics shown in figure 3.2(c).
Figure 3.2  (a) Reflection characteristics of a FGCPW (b) Impedance Diagram (c) Transmission characteristics

(W_g=20mm, L_g=20mm, W=3mm, g=0.35mm, h=1.6mm and \( \epsilon_r=4.4 \))
It is reported that a transmission line can be converted to a radiating structure by creating discontinuity [2]. This possibility is explored in this thesis. An open ended CPW transmission line can be converted into an efficient radiator by improving the matching at the desired frequency. Matching and hence radiation can be achieved by introducing discontinuity on the structure. In a normal CPW, discontinuity can be created in three different ways.

a) Modifying the signal strip.

b) Modify the ground plane.

c) Modify the signal strip and ground plane simultaneously so that both will contribute to impedance match and effective radiation.

In this chapter a quad band antenna designed by modifying the signal strip is discussed elaborately. The other two modification techniques are discussed in later chapters.

### 3.2 Coplanar waveguide fed Monopole antenna

Design and Development of a coplanar waveguide fed monopole antenna with modified signal strip is the main theme of the present chapter. Initially the coplanar waveguide transmission line is modified to fundamental quarter wavelength monopole [2] by increasing the signal strip length $L_1$ as shown in figure.3.3.
Figure 3.3 Finite Ground Coplanar Waveguide monopole antenna  
(a) Front view (b) side view  
\((W_g=20\text{mm}, L_g=20\text{mm}, L_1=18\text{mm}, W=3\text{mm}, g=0.35\text{mm}, h=1.6\text{mm} \text{ and } \epsilon_r=4.4)\)

The antenna is printed on a substrate of dielectric constant \(\epsilon_r=4.4\) and height \(h=1.6\text{mm}\). The signal strip width (w) and signal to ground gap (g) are selected for 50\(\Omega\) impedance match. The effect of signal strip length \((L_1)\) on the resonant frequency of the monopole antenna is shown in figure 3.4. The signal strip length is arbitrarily varied from 14mm to 22mm in this study. It is evident from the reflection characteristics that as the length increases the frequency shifts to lower side. The variation in resonant frequency, bandwidth and quality factor of the antenna is shown in table 3.1. At \(L_1=14\text{mm}\) the antenna is operating at 3.8625GHz with a 10dB bandwidth of 730MHz. This means the Q of the system is 5.2838. Moreover, the quality factor of this resonating system remains almost constant with strip length.
Table 3.1 Variation in resonant frequency, band width and quality factor with monopole length L1

<table>
<thead>
<tr>
<th>Monopole Length (L₁)mm</th>
<th>Resonant Frequency (GHz)</th>
<th>10 dB bandwidth</th>
<th>Quality factor</th>
<th>Guided wavelength (λg)</th>
</tr>
</thead>
<tbody>
<tr>
<td>14</td>
<td>3.8625</td>
<td>0.731</td>
<td>5.2838</td>
<td>0.26</td>
</tr>
<tr>
<td>16</td>
<td>3.5625</td>
<td>0.6813</td>
<td>5.2289</td>
<td>0.27</td>
</tr>
<tr>
<td>18</td>
<td>3.2875</td>
<td>0.6062</td>
<td>5.425</td>
<td>0.28</td>
</tr>
<tr>
<td>20</td>
<td>3.05</td>
<td>0.5563</td>
<td>5.482</td>
<td>0.29</td>
</tr>
<tr>
<td>22</td>
<td>2.8687</td>
<td>0.525</td>
<td>5.464</td>
<td>0.30</td>
</tr>
</tbody>
</table>

This monopole antenna of signal strip length L₁=18mm is resonating at 3.2875GHz. From the experiment it is found that this occurs when L₁ is quarter wavelength. The transmission characteristic of the antenna is shown in figure 3.5. For all cases it is found that maximum transmission occurs at the resonance. This confirms that this is working as a λ/4 monopole antenna.

Figure 3.4 Variation of Reflection(S11) Characteristics of FGCPW monopole antenna with monopole length L₁
(Wg=20mm, Lg=20mm, W=3mm, g=0.35mm, h=1.6mm and ϵr=4.4)
The variation in resonant frequency with dielectric constant and height of the substrate is shown in Table 3.2. While varying the dielectric constant the height of the substrate is taken as 1.6mm and while varying the height of the substrate dielectric constant is taken as 4.4. As the dielectric constant and height increases the resonant frequency shift to lower region and vice versa.

<table>
<thead>
<tr>
<th>Dielectric Constant</th>
<th>Obtained Resonant frequency(GHz)</th>
<th>Height of substrate (mm)</th>
<th>Obtained Resonant frequency(GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>3.8625</td>
<td>1.2</td>
<td>3.3688</td>
</tr>
<tr>
<td>3</td>
<td>3.5812</td>
<td>1.4</td>
<td>3.3188</td>
</tr>
<tr>
<td>4</td>
<td>3.3625</td>
<td>1.6</td>
<td>3.2563</td>
</tr>
<tr>
<td>5</td>
<td>3.1563</td>
<td>1.8</td>
<td>3.225</td>
</tr>
<tr>
<td>6</td>
<td>3.0063</td>
<td>2</td>
<td>3.1688</td>
</tr>
</tbody>
</table>
The impedance characteristic of the coplanar waveguide fed monopole antenna is shown in figure 3.6. From the Smith chart it is evident that the real part of impedance is nearly equal to 60Ω at the resonant frequency. Moreover, the reactance is found to be nearly equal to zero and becomes inductive or capacitive as we move away from resonance. From the experimental and simulation studies it is confirmed that the resonance occurs when the monopole length is almost equal to $\lambda_g/4$, where $\lambda_g = \frac{\lambda}{\varepsilon_{re}}$ is the wavelength in the substrate with $\varepsilon_{re}$ as the effective dielectric constant in the substrate $\varepsilon_{re} = \frac{(\varepsilon_r + 1)}{2}$.

The simulated surface current distribution of the antenna at resonance is shown in the figure 3.7. It is obviously noticeable that there is a quarter wavelength variation of field along the signal strip length $L_1$. The current is maximum near the ground plane and minimum at the tip of the monopole. It is also noted that there is
no significant current variation on the finite ground plane at the resonant frequency. Nevertheless there is a very small current variation at the edges along the width of the ground plane which varies with the ground plane dimension.

![Picture](image)

**Figure 3.7** Current distribution of FGCPW monopole antenna 
\((W_g=20\,mm, \, L_g=20\,mm, \, L_1=18\,mm, \, W=3\,mm, \, g=0.35\,mm, \, h=1.6\,mm \text{ and } \varepsilon_r=4.4)\)

### 3.3 Top loaded monopole antenna

Increasing the ground plane dimension and signal strip dimension will affect the overall compactness of the antenna. As we already discussed that by increasing the monopole length will decrease the resonant frequency but the overall size increases. In order to bring the resonance to lower frequency without altering the size, the signal strip is meandered. A strip of length \(L_2\) is added to the monopole antenna resulting a structure as shown in figure 3.8 (a). The bending will provide additional resonating length in the given area. With bending the unutilized area in the substrate is made available for the radiation phenomenon. Conventional monopole antenna is resonating at 3.26GHz. Bending the monopole to one side will shift the resonant frequency from 3.26GHz to 2.14GHz as in figure 3.9 (a).
Moreover there is a possibility for resonance at higher frequency. The impedance characteristic of the antenna is shown in figure 3.9 (b). The impedance is found to be matched at lower frequency. But at higher frequency the imaginary part is found to be inductive and hence it is not matched at that frequency as in the figure. Here the overall dimension of the antenna remains unchanged.

Again another strip of same length $L_2$ is added to the other side to make it a symmetrically meandered (T-shaped) antenna as shown in figure 3.8 (b). The unused space in the substrate is effectively made use in this design. The reflection characteristic of the antenna with symmetric bend is shown in figure 3.9. By symmetrically extending the strip $L_2$ as in the figure will not affect earlier resonance much. There is no considerable improvement in the reflection characteristics for the symmetrically meandered structure rather than a small frequency shift to lower value which provide a further compactness. But while looking to the impedance characteristics it is found that at higher frequency the imaginary part is nearly zero hence, the higher resonance is made visible and is found to be matched in this case.

![Figure 3.8](image)

**Figure 3.8** Finite Ground plane monopole antenna with (a) Meandered to one side (b) Symmetrical meandered

(W$_g$=20mm, $L_x$=20mm, $L_1$=18mm, $L_2$=10mm, W=3mm, $g$=0.35mm, $h$=1.6mm and $\varepsilon_r$=4.4)
Figure 3.9  (a) Reflection Characteristic (b) Input Impedance of Finite Ground plane monopole antenna with Meandered to one side and Symmetrical meandered

\( W_g=20\, \text{mm}, \, L_g=20\, \text{mm}, \, L_1=18\, \text{mm}, \, L_2=10\, \text{mm}, \, W=3\, \text{mm}, \, g=0.35\, \text{mm}, \)  
\( h=1.6\, \text{mm} \) and \( \epsilon_r=4.4 \)
The current distribution of the antenna which is meandered to one side at 2.14GHz is shown in figure.3.10 (a). Like a monopole antenna discussed before there is a quarter wavelength variations in current along the length. Moreover, the intensity of current is found to be higher on the lower edge of the bend structure compared to the upper edge. The current distribution of the symmetrically bend structure is shown in figure.3.10 (b). Here the directions of current in the two horizontal arms are opposite. But it is found that the current distribution is much more uniform on both the signal strip and ground plane for a symmetrically bended antenna. In both cases (Meandered to one side and symmetrically bend) there is a quarter wavelength variation of current along the monopole. The total area of the antenna remains intact in both the cases.

### 3.4 Dual band meandered monopole antenna

Without altering the overall compactness, the frequency can be further decreased by folding the signal strip as shown in figure.3.11. The antenna consists of a T-shaped monopole with symmetrical vertical strips ($L_3$) on both sides. These
two symmetrical vertical strips each of length $L_3$ is used to increase the current path without affecting the overall compactness and other desirable radiation characteristics. The bending should be done in such a way that the effective reactive coupling should not affect the overall impedance match. This bending in a given area will increase the resonating length (since the physical length increases) and shift the frequency downwards, and hence provide compactness.

Figure 3.11 Geometry of the symmetrical meandered dual band antenna
(a) Front View (b) Side view
($L_1=18\text{mm}, L_2=11.5\text{mm}, L_3=7\text{mm}, W=3\text{mm}, W_g=14\text{mm}, L_g=10\text{mm}, h=1.6\text{mm}, \varepsilon_r=4.4$ and $g=0.35\text{mm}$)

Thus top loading a monopole antenna make the higher resonance visible in the frequency region. Thus we can say that top loading can excite additional resonance. The addition of stubs $L_2$ and $L_3$ will give desired response for the antenna at the resonant frequency. The increased length $L_3$ causes a corresponding shift in the resonance towards the lower frequency. Simulated and experimental return loss of the antenna is shown in figure 3.12.
The lower band centered at 1.77GHz has a wide bandwidth from 1.47GHz-1.97GHz with a percentage bandwidth of about 34% covering DCS 1800 and DCS 1900/PCS. The measured impedance bandwidth of the upper band centered at 5.54GHz determined by 10-dB return loss, is good enough to cover the HIPERLAN (5150-5350MHz) band and ISM WLAN 5.8(5.725-5.875GHz) band, with wide bandwidth from 5.13GHz-6.48GHz.

The current distributions at the two resonant frequencies of the designed antenna are shown in figure 3.13. The current distributions on the horizontal arms (which provide the required resonating physical length) are equal and opposite and hence the radiated field due to this cancels at the far field. So the radiation is primarily due to the y-component and hence it is polarized along y-direction in the two bands. Moreover, the current distribution on the lateral ground planes is found to be symmetrical and as shown in the figure. It is also noted that the fundamental mode at 1.77GHz is due to the lengths $L_1+L_2+L_3$ which is nearly equal to $\lambda g/4$. 

![Figure 3.12 Measured and Simulated reflection characteristics of the dual band antenna](image-url)
This aspect is reconfirmed by conducting experiments with different parametric lengths. The analyses of dual band antenna with different antenna parameters are discussed in the next sections.

### 3.4.1 Effect of varying the strip length $L_1$ ($L_1+L_2+L_3$ is constant)

The distance between the top load and the ground plane is crucial in providing the impedance match. Since the length $L_1$ determines the distance between the top load and ground plane, it is important and should be optimized for providing better impedance matching at the resonant frequencies. The variation of the reflection characteristics of the antenna with the strip length $L_1$ by keeping the total length ($L_1+L_2+L_3$) constant is investigated initially. The reflection characteristic of the antenna is shown in figure.3.14. The first resonance frequency is not affected by the length $L_1$ since the total resonant length ($L_1+L_2+L_3$) remains constant, but the impedance matching gets affected. It is also noted that the higher resonance is very much affected by the strip length. As the length increases the higher frequency decreases and vice versa. From the impedance characteristics
shown in figure it is found that for lower frequency the impedance become more capacitive and the impedance loci is found to shrink as $L_1$ increased. For higher frequency the impedance loci is found to expand as $L_1$ is increased.

Figure 3.14 Variation of a) reflection characteristics b) impedance characteristics with strip length $L_1$ and by keeping the total length $L_1+L_2+L_3$ constant ($L_1=18\text{mm}, L_2=11.5\text{mm}, L_3=7\text{mm}, W=3\text{mm}, W_\pi=14\text{mm}, L_\pi=10\text{mm}, h=1.6\text{mm}, \varepsilon_r=4.4$)
3.4.2 Effect of varying the strip length $L_1$ ($L_1+L_2+L_3$ is not a constant)

![Variation of reflection characteristics with strip length $L_1$ and total length $L_1+L_2+L_3$ is not constant](image)

The variation of reflection coefficient with strip length $L_1$ without keeping the total length constant ($L_1+L_2+L_3$ =not constant) is also studied. Here the length $L_2$ and $L_3$ are constant throughout the variation. The corresponding reflection characteristic is shown in figure.3.15. Both the resonances are affected in this case. From the reflection characteristics it is found that as the length increases both the resonances decreases and vice versa. But the percentage shift in the higher frequency is larger as compared to the lower frequency.

The top loaded strips $L_2$ and $L_3$ is a part of radiating structure for lower frequency, but for higher frequencies these act as an open ended stubs whose length can affect the resonant frequency. Thus by varying the open ended stubs the higher frequency can be tuned.
From the above studies it is clear that the lower resonance is fully depending on the total length of the strip $L_1+L_2+L_3$ because it remains constant when total length is constant. But a thorough analysis should be performed to analyze the reason for each resonance.

3.4.3 Effect of varying the strip length $L_2$ ($L_1+L_2+L_3$ is a constant)

The variation in the reflection characteristics of the antenna with the strip length $L_2$ by keeping the total length constant is studied and its response is shown in the figure. As the length $L_2$ increases keeping the total length constant, both the resonant frequencies will almost tend to remain in the same position as their resonating length constant. But the separating distance between $L_1$ and $L_3$ will vary as $L_2$ changes. This change in stub length $L_2$ causes a corresponding change in reactance of the antenna and hence the resonant frequency changes as in the figure.

![Figure 3.16 Variation of reflection characteristics with strip length $L_2$ and by keeping total length $L_1+L_2+L_3$ is constant (L_1=18mm, L_3=7mm, W=3mm, W_g=14mm, L_g=10mm, h=1.6mm, $\varepsilon_r=4.4$)](image-url)
3.4.4 Effect of varying the strip length $L_2$ ($L_1+L_2+L_3$ is not a constant)

The variation in the reflection characteristics of the antenna with the strip $L_2$ without keeping the total length constant is studied and its response is shown in the figure.3.17. As the stub length $L_2$ varies the distance between the vertical strips ($L_1$ and $L_3$) changes. As mentioned in the previous section the length of this open ended stub can be used to tune the resonant frequency. As $L_2$ increases both resonating bands shifts to lower frequency value and vice versa. A length of $L_2=11.5$mm is chosen as optimum by considering the impedance match.

![Figure 3.17](image)

**Figure 3.17** Variation of reflection characteristics with strip length $L_2$ and total length $L_1+L_2+L_3$ is not constant

(L$_1$=18mm, L$_3$=7mm,W=3mm,W$_w$=14mm,L$_g$=10mm,h=1.6mm,$\varepsilon_r$=4.4)

3.4.5 Effect of varying the strip length $L_3$ ($L_1+L_2+L_3$ is a constant)

The variation of reflection characteristics with strip length $L_3$ by keeping $L_1+L_2+L_3$ as constant is analyzed and shown in figure.3.18. Thus the lower
resonant frequency is not affected by its variation but the impedance matching is deeply affected. Thus it should be taken into consideration while optimizing the final design; that the total length is the main factor in determining the lower resonance. The higher resonance is fundamentally contributed by the total length but the interaction between strips $L_1$ and $L_3$ will also contributed towards this resonant frequency.

Figure 3.18 Variation of reflection characteristics with strip length $L_3$ by keeping total length $L_1+L_2+L_3$ as constant (W=3mm, Wg=14mm, $L_g=10$mm, h=1.6mm, $\varepsilon_r=4.4$)

3.4.6 Effect of varying the strip length $L_3$ ($L_1+L_2+L_3$ is not a constant)

The effect of the vertical strip $L_3$ on the reflection characteristics is studied and is shown in figure.3.19. As $L_3$ increases the resonant length corresponding to the lower frequency increases and the frequency shift towards the lower region and vice versa. There is a small effect on the resonant
frequency of the upper band and the impedance matching deteriorates as seen in the figure. As \( L_3 \) increases its coupling with strip \( L_1 \) and ground plane varies which affects the impedance characteristics. Thus the first resonance is mainly due to the total length \( L_1+L_2+L_3 \) and the second resonance is primarily due to the monopole length \( L_1 \).

![Figure 3.19](image)

**Figure 3.19** Variation of reflection characteristics with strip length \( L_3 \) and total length \( L_1+L_2+L_3 \) is not constant

\( (L_1=18\text{mm, } L_2=11.5\text{mm, } W=3\text{mm, } W_g=14\text{mm, } L_g=10\text{mm, } h=1.6\text{mm, } \varepsilon_r=4.4) \)

### 3.4.7 Variation on reflection characteristics with Ground plane length \( L_g \)

The ground plane parameters like length \( L_g \) and width \( W_g \) should be optimized properly. The effect of ground plane length on the radiation and impedance characteristics of the antenna is studied and its variation is shown in figure 3.20. It is found that there is not much variation in the resonant frequency with the ground plane length \( L_g \). However, the impedance match and hence the
antenna performance is affected slightly as in the figure. Thus the length of ground plane is chosen as \( L_g = 0.1\lambda_g \) by considering the compactness and impedance match at the resonant frequency.

![Figure 3.20 Variation of reflection characteristics with ground plane length \( L_g \) (\( L_1 = 18\text{mm}, L_2 = 11.5\text{mm}, L_3 = 7\text{mm}, W = 3\text{mm}, W_g = 14\text{mm}, h = 1.6\text{mm}, \varepsilon_r = 4.4 \))](image)

**3.4.8 Variation on reflection characteristics with Ground plane width \( W_g \)**

There should be a compromise between compactness and performance of any antenna. The width \( W_g \) of the ground plane is varied and its effects on reflection characteristics are also studied. Here, as the width increases the resonance shift to lower frequency and vice versa. But it should be noted that when \( W_g \) is decreased beyond \( 0.1\lambda_g \) the resonance frequency is severely affected as shown in figure.3.21. Moreover, it is meaningless to put a very large ground plane for a compact wireless device, if small ground plane itself is providing better performance. The radiation characteristics of a finite ground CPW should be analyzed to confirm the antenna performance.
3.4.9 Radiation pattern

The polarization of the antenna is along the Y-direction (Vertical) for both the resonant frequencies. From the current distribution shown in figure 3.13 it is clear that the field component along the strip $L_1$ is $y$-directed while that of $L_2$ is along $+x$ and $-x$ direction. Thus the $x$-directed component gets cancelled at the far field resulting the polarization along the $y$-direction for both the resonant frequencies.

The measured co-polar and cross-polar radiation pattern of the antenna at resonance for both E and H planes are shown in figure 3.22. The pattern is found to be omnidirectional. A constant gain pattern is obtained along the H-plane and along E-plane a pattern with two nulls at $90^0$ and $270^0$ degrees is obtained. The nulls are found to be above and below the lateral ground planes ie above the horizontal arm $L_2$ and on the connector side. The antenna shows good cross polarization level of nearly 15dB along the boresight direction for both the frequencies along the principal planes.
Investigation on the signal strip modification of a coplanar waveguide (CPW) transmission line

Figure 3.22 Measured Radiation pattern of the dual band antenna at (a) H-plane 1.77GHz (b) E-plane 1.77GHz (c) H-plane 5.5GHz (d) E-plane 5.5GHz

\( L_1 = 18\text{mm}, L_2 = 11.5\text{mm}, L_3 = 7\text{mm}, W = 3\text{mm}, W_g = 14\text{mm}, L_g = 10\text{mm}, h = 1.6\text{mm}, \varepsilon_r = 4.4\)
3.4.10 Gain and Efficiency

The measured gains of the antenna at two bands are shown in figure 3.23. The antenna shows an average gain of 3dBi in the lower band and 3.5dBi in the upper band. The measured efficiency of the antenna using Wheeler cap method is also shown in figure 3.24. The antenna is showing an average efficiency greater than 80% in both the bands.

![Graph](image1)

(a) 1.77GHz band

![Graph](image2)

(b) 5.5GHz band

Figure 3.23 Measured gain of the dual band antenna at (a) 1.77GHz (b) 5.55GHz
(L₁=18mm, L₂=11.5mm, L₃=7mm, W=3mm, W₉=14mm, L₉=10mm, h=1.6mm, εᵣ=4.4)
Investigation on the signal strip modification of a coplanar waveguide (CPW) transmission line

3.5 Design and Development of Quad band antenna

The signal strip of the Dual band antenna presented in the above section can be modified to resonate in additional application bands. Antenna resonating in four different frequency bands are discussed in this.

From the studies explained in the section above section, it is clear that in order to excite an additional resonance it is necessary to create an additional resonating path or to excite the higher harmonics. It is better to create additional resonating path for the new resonance because it is difficult to independently control the higher harmonic radiation with the fundamental one. Thus the top loaded signal strip (L₃) is modified by increasing the length (L₃+L₄) on one side of the top load and thereby creating asymmetry in the structures as shown in figure.3.25. This asymmetry can generate a resonance in addition to the existing two resonances. Now the resonances are centered at 1.61GHz, 2.4GHz and
5.8GHz respectively. The comparison of resonant frequencies of the monopole antenna, symmetrically meandered antenna and asymmetrically loaded antenna are shown in figure 3.26.

Thus the simple monopole antenna is resonating at 3.286GHz. There is only one resonance in the entire band. This curve is marked as ‘a’. When the antenna is symmetrically top loaded, it resonates at two frequencies (b). It resonates at 1.77GHz and 5.54GHz respectively. However when it is asymmetrically loaded it resonates a three frequencies (c). The three modes are at 1.61GHz, 2.4GHz and 5.8GHz respectively.

Figure 3.25  Geometry of the unsymmetrical triple band antenna (a) Front view (b) side view
(L1=18mm, L2R=L2L=11.5mm, L3=4mm, L4=8mm, W=3mm, Lg=10mm, WgR= WgL=14mm, g=0.35mm, h=1.6mm, εr=4.4)
Investigation on the signal strip modification of a coplanar waveguide (CPW) transmission line

Figure 3.26 Reflection characteristics comparison of single band, dual band and triple antenna

(L1=18mm, L2R=11.5mm, L3=4mm, L4=8mm, W=3mm, Lc=10mm, WgR=WgL=14mm, g=0.35mm, h=1.6mm, εr=4.4)

By the addition of the strip L4 the resonance at 1.77GHz of the dual band antenna explained in the earlier section due to the length (L1+L2L+L3) is now shifted down to 1.61GHz, corresponding to a resonant length (L1+L2L+L3+L4). It is noted that this total length is nearly equal to the quarter wavelength corresponding to 1.61GHz.

By the addition of this strip another resonance is excited at 2.4GHz. The frequency corresponding to the resonance is in the application region(ISM 2.4) but is poorly matched. This resonance is nearly equal to half wavelength corresponding to the resonant length (L1+L2R+L3).

The higher resonance at 5.54GHz is also shifted to 5.8GHz as shown in the figure which is corresponding to a length (L1+L2R+L3) of 0.9λg. Thus further design
Chapter -3

should carry out by taking at most care to keep the existing matched resonances as such and to provide impedance matching for the resonance at 2.4GHz.

In order to provide better insight to the antenna performance the current distribution on the antenna should be analyzed. The current distribution of the triple band antenna is shown in figure.3.27.

Figure.3.27 Current distribution of the Dual band antenna (a) 1.61GHz (b) 2.4GHz (c) 5.8GHz

(L₁=18mm, L₂R=L₂L=11.5mm, L₃=4mm, L₄=8mm, W=3mm, L₉=10mm, W₉R=W₉L=14mm, g=0.35mm, h=1.6mm, εᵣ=4.4)
A quarter wavelength variations are found along the resonating length for the first two resonances as in the figure. The first resonant path is due to the longer asymmetrical path and the second is due to smaller path. A three quarter variation of current is observed along the shorter asymmetrical path and larger asymmetrical path. Since these two resonances are close to each other they merge together to give a wide bandwidth corresponding to the third resonance as in figure.3.27 (C).

In addition to the three resonances of the antenna the next aim is to generate an additional resonance at 900MHz together with better match for the present bands; preserving the compactness. It is really challenging to excite a lower frequency without trading the compactness of the antenna.

Keeping the overall area intact a slit ‘ABCDEFGH’ is introduced in the antenna as shown in figure.3.28. It should be noted that the present modification is made without altering the already existing three resonant paths. That is the existing resonant paths are there and a new discontinuity is created. This forces the current to flow through a longer path around the slit (Through ABCDEFGH) and produce an additional lower resonance at 900MHz as shown in figure.3.29. The subsequent resonances are at 1.74GHz, 2.44GHz and 5.5GHz. It is interesting to note that in addition to the excitation at 900MHz the structure provides a well matched resonances at the existing resonant frequencies centered at 1.74GHz, 2.44GHz and 5.5GHz.
The GSM900 band (850MHz-960MHz) exhibits 2:1 VSWR impedance bandwidth and is wide enough to cover the GSM application band. The resonance at 1.61GHz of the triple band antenna is now shifted to 1.74GHz with a 2:1 VSWR band width from 1.56GHz-1.92GHz covering the DCS-1800 application band. The resonance at 2.4GHz of the triple band antenna is matched by adding the slit. The antenna offers a 2:1 VSWR band width from 2.39GHz to 2.49GHz which is wide enough to cover the ISM-2.4 band. Moreover, this antenna covers the 5.2/5.8GHz WLAN band with 2:1 VSWR bandwidth from 5.07GHz-6.23GHz. The measured and simulated reflection characteristics and of the antenna are shown in figure.3.29.
Figure 3.29  Experimental and Simulation results of the quad band antenna (a) Reflection characteristics and (b) VSWR (L_1=18mm, L_{2R}=L_{2L}=11mm, L_3=4mm, L_4=8mm, L_5=31mm, L_6=10mm, W_{gR}=W_{gL}=14mm, W_s=3mm, CD=10mm, BC=29mm, DE=ha=1mm, h=1.6mm, \epsilon_r=4.4 and g=0.35mm)

3.5.1 Measured Variation on reflection characteristics with strip length (L_3+L_4)

An exhaustive parametric study has been performed to investigate the effect of various parameters on antenna performance. The variation in reflection characteristics with strip length L_3+L_4 (on the Left arm – with L_3 as constant) is shown in figure 3.30. It is found that by increasing the strip length the second
resonance frequency is mainly affected and shifts towards the lower frequency region and vice versa. Similarly as the length increases the inductive reactance corresponding to the third resonance increases, affecting the overall impedance match. By considering the radiation and impedance characteristics corresponding to the DCS-1800 application band the strip length is optimized at 12mm. From the studies it is confirmed that the second resonance frequency can be easily tuned by adjusting $L_3+L_4$.

![Figure 3.30 Variation in reflection characteristics of the antenna with variation in $L_3+L_4$](image)

$L_1=18$mm, $L_{2R}=L_{2L}=11$mm, $L_4=31$mm, $L_g=10$mm, $W_{gR}=W_{gL}=14$mm, $W_s=3$mm, $CD=10$mm, $BC=29$mm, $DE=HA=1$mm, $h=1.6$mm, $\epsilon_r=4.4$ and $g=0.35$mm)
3.5.2 Measured Variation on reflection characteristics with slot length CD

![Graph showing variation in reflection characteristics](image)

The measured variation in the return loss characteristics of the antenna with respect to the slit perimeter ‘ABCDEFGH’ (by varying length CD) is shown in figure 3.31. In this the length of slot perimeter varies and all other parameters remain unaffected. As slit perimeter increases, the lower resonance at 900MHz shifts down, worsening the impedance match. All other resonances remain unaltered as their resonating lengths are unaffected. Thus by adjusting the slit length we can tune the lower resonant mode. The width of the slot is selected as 1mm throughout the study. If the slit is opened or closed at both ends the resonating length will get halved and there will not be any resonance at lower frequency. So it is very important to keep the ends of the slits complimentary to each other, that is while one end is opened the other end should be closed (so that the total perimeter of the slot is constant).
3.5.3 Measured Variation on reflection characteristics with right arm strip $L_3$

Variation on reflection characteristics with strip length $L_3$ (on the right arm) is studied and is shown in figure 3.32. From the figure it is clear that the strip length affects the third and fourth resonances significantly. As the length increases the third and fourth resonance centered at 2.44GHz and 5.5GHz are shifted to lower resonant frequencies and vice versa. Since variation in $L_3$ results in variation of the effective slit length; the lower resonance at 900MHz is also affected slightly. The second resonance is not affected since the total length $L_1+L_2+L_3+L_4$ remains constant.

![Figure 3.32 Variation in reflection characteristics of the antenna with strip length $L_3$](image)

Figure 3.32 Variation in reflection characteristics of the antenna with strip length $L_3$

$L_1=18\text{mm}, L_{2R}= L_{2L}= 11\text{mm}, L_4=8\text{mm}, L_5=31\text{mm},$

$L_g=10\text{mm}, W_{gR}= W_{gL}=14\text{mm}, W_s=3\text{mm}, CD=10\text{mm}, BC=29\text{mm},$

$DE=HA=1\text{mm}, h=1.6\text{mm}, \epsilon_r=4.4$ and $g=0.35\text{mm}$)
3.5.4 Measured Variation on reflection characteristics with strip length $L_1$

The variation in the reflection characteristics of the antenna with respect to the strip length $L_1$ is shown in figure 3.33. It is found that the strip length affects the fourth resonance significantly. For a small change of strip length $L_1$ can contribute to a larger shift in resonant frequency for the fourth resonance. The impedance matching of the third resonance is also affected by the length. The first two resonant frequencies are found to remain almost constant.

![Figure 3.33 Variation in reflection characteristics of the antenna with strip length $L_1$](image-url)

$S_11$(dB) versus Frequency(GHz) for different values of $L_1$:
- $L_1=17$
- $L_1=18$
- $L_1=19$

Parameters:
- $L_2=R=11mm$
- $L_3=4mm$
- $L_4=8mm$
- $L_5=31mm$
- $L_g=10mm$
- $W_{g1}=W_{g2}=14mm$
- $W_s=3mm$
- $CD=10mm$
- $BC=29mm$
- $DE=HA=1mm$
- $h=1.6mm$
- $\epsilon_r=4.4$
- $g=0.35mm$
3.5.5 Measured Variation on reflection characteristics with Ground Length $L_g$

In the performance of compact antennas the effect of ground plane is also very crucial. The variation on reflection characteristics with ground length is shown in figure 3.34. There is a considerable change in the return loss characteristics of the antenna with ground length. As $L_g$ changes the coupling between signal strip and ground plane varies which results a corresponding change in the resonant frequency. Considering the application band and compactness of the antenna, the length of the ground plane is optimized as $0.27\lambda_g$. Where $\lambda_g$ is the dielectric wavelength corresponding to lowest frequency.

![Figure 3.34 Variation in reflection characteristics of the antenna with ground length $L_g$](image)

Figure 3.34 Variation in reflection characteristics of the antenna with ground length $L_g$

$L_1=18\text{mm}, L_{2R}=L_{2L}=11\text{mm}, L_3=4\text{mm}, L_4=8\text{mm}, L_5=31\text{mm}, W_{gR}=W_{gL}=14\text{mm}, W_s=3\text{mm}, \text{CD}=10\text{mm}, \text{BC}=29\text{mm}, \text{DE}=\text{HA}=1\text{mm}, h=1.6\text{mm}, \epsilon_r=4.4 \text{ and } g=0.35\text{mm}$
3.5.6 Measured Variation on reflection characteristics with Ground Width $W_{gL}$ and $W_{gR}$

The variation in reflection characteristics with Ground width $W_{gL}$ and $W_{gR}$ are studied and is shown in figure 3.35. The lower resonance at 900MHz is not at all affected by ground widths. The impedance matching at third resonance is severely affected but there is not much variation in the second resonant frequency. The fourth resonant frequency shows drastic variation with the ground width. As the ground width increases the frequency shift towards lower band with a corresponding deterioration in the impedance matching. Considering all the aforementioned observation the ground plane width is optimized at $0.38\lambda_g=14\text{mm}$.

![Figure 3.35: Variation in reflection characteristics of the antenna with ground width $W_{gL}$ and $W_{gR}$](image)

$L_1=18\text{mm}, L_{2L}=L_{2R}=11\text{mm}, L_3=4\text{mm}, L_4=8\text{mm}, L_5=31\text{mm}, L_g=10\text{mm}, W_s=3\text{mm}, CD=10\text{mm}, BC=29\text{mm}, DE=HA=1\text{mm}, h=1.6\text{mm}, \epsilon_r=4.4$ and $g=0.35\text{mm}$)
From the exhaustive parametric analysis performed above the working principle of the antenna at four frequency bands are analyzed. The design criterion for the quad band antenna based on the above observations is explained in this section. The first resonance centered at 900MHz is due to the total perimeter of slit. Slit perimeter can be calculated as,

\[
\text{Slit perimeter} = \frac{(0.4*\lambda_1)}{\sqrt{\varepsilon_{\text{reff}}}}
\]

Where, \( \varepsilon_{\text{reff}} = (1+\varepsilon_r +1)/3 \) is the effective dielectric constant of substrate, and \( \lambda_1 \) is the wavelength corresponding to the first resonant frequency. It is presumed that the capacitive coupling between the strips (ABCD & EFGH) across the slit effectively increases the overall length of the antenna, thereby reducing the physical length to 0.4 times the guided wavelength.

The second resonance centered at 1.74GHz occurs due to the combined effect of the length \( L_1+L_{2L}+L_3+L_4 \) and strip length ABCD (Fig.3.28). The strip length ABCD will adjust in accordance with \( L_1+L_{2L}+L_3+L_4 \) because \( L_{2L} \) is equal to \( L_{2R} \) and width \( W_s \) of the signal strip is constant. and therefore it can be calculated as below.

\[
L_1+L_{2L}+L_3+L_4 = \frac{(0.35*\lambda_2)}{\sqrt{\varepsilon_{\text{reff}}}}
\]

Where \( L_{2L} = 2.75 \) \( L_3 \) is chosen for minimum capacitive coupling between \( L_1 \) and \( L_3 \). The additional capacitive coupling contributed by top loading results in increasing the physical length, hence assuring compactness.

The third resonance is due to \( L_1+L_{2R}+L_3 \) (Fig.3.32). It is interesting to note that the matching is severely affected by \( L_1 \) (Fig.3.33) and \( L_3+L_4 \) (Fig.3.30) without affecting the resonant frequency. On increasing \( L_1 \) the real part of impedance is decreased causing a corresponding deterioration in
impedance match. As $L_3 + L_4$ increases the inductive reactance increases with corresponding increase in the imaginary part of impedance contributing to better impedance matching. The design criterion for the third resonance is,

$$L_1 + L_2 R + L_3 = \left(\frac{0.4*\lambda_3}{\epsilon}\right) \sqrt{\frac{1}{R_{\text{eff}}}}$$  \hspace{1cm} (3)

The fourth resonance centered at 5.5GHz is affected by the variation in $L_1$ (Fig.7), $L_3$ (Fig.6) and $L_3 + L_4$ (Fig.5). A significant change in resonance with ground dimensions is also observed (Fig.8-9). The shift in resonance is due to the change in reactance, resulting from a change in the coupling between the ground and the signal strip. Hence the fourth resonance is due to two parallel symmetrical paths $L_1 + L_2 R + L_3$ and $L_1 + L_2 L_3 + L_4$. These two resonances merge to give a broad band at 5.5GHz. The path lengths are found to be equal to three quarter wavelengths at the corresponding resonant frequency. Since a significant variation is only due to $L_1$, the length should be selected in such a way so as to cover the application band as given below.

$$L_1 = \left(\frac{0.49*\lambda_4}{\epsilon}\right) \sqrt{\frac{1}{R_{\text{eff}}}}$$  \hspace{1cm} (4)

The decrease in the physical lengths of the antenna at the different resonant frequencies is due to the top loading effect, by virtue of which the antenna appears electrically larger [3].

From the parametric studies, it is inferred that the ground plane affects the impedance matching of the resonant modes. By considering the compactness and antenna performance the ground plane length and width are optimized as follows.

$$L_g = \left(\frac{0.27*\lambda_4}{\epsilon}\right) \sqrt{\frac{1}{R_{\text{eff}}}}$$  \hspace{1cm} (5)
\[ W_{gl} = W_{gr} = \left( 0.38 \times \frac{\lambda^4}{\sqrt{\varepsilon_{\text{reff}}}} \right) \]  

(6)

To validate the design equations for various quad band antennas are designed for different dielectric substrates. The above equations are validated by characterizing antenna characteristics. The predicted antenna resonances are reasonably good agreement with the obtained results as shown in Table 3.3.

<table>
<thead>
<tr>
<th>Substrate (Dielectric constant)</th>
<th>Obtained Frequency(GHz)</th>
<th>Simulated Frequency(GHz)</th>
<th>% Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rogers RT/duroid 5880 (2.2)</td>
<td>0.95 1.70 2.30 5.42</td>
<td>5.55 5.55 4.34 4.23</td>
<td></td>
</tr>
<tr>
<td>Taconic RF-35 (3.5)</td>
<td>0.92 1.76 2.42 5.12</td>
<td>2.22 2.22 0.83 1.53</td>
<td></td>
</tr>
<tr>
<td>Alumina (9.4)</td>
<td>0.98 1.94 2.36 5.78</td>
<td>8.88 7.77 1.66 11.15</td>
<td></td>
</tr>
<tr>
<td>Rogers RO6010 (10.2)</td>
<td>0.98 1.93 2.39 5.77</td>
<td>8.88 7.2 0.41 10.96</td>
<td></td>
</tr>
</tbody>
</table>

The measured co and cross polarized received power of the antenna along with return loss is shown in Figure 3.36. Since the Y-component dominates over the X-component for all the four bands, the antenna is polarized along the Y direction. The antenna shows good polarization purity in the first and fourth band. The polarization is moderate on the second and third band. The first resonance at 900 MHz is due to the perimeter of the slit \( ABCDEFGH \). Since the direction of current is opposite in its parallel arms it is seen that the X-component gets cancelled in the far field resulting the polarization along Y direction. In the second band centered at 1.74GHz, the Y-component dominates the X-component resulting in polarization along Y-direction.

But for the third band centered at 2.44GHz, the polarization purity is poor, because here the X (due to L_{2R}) and Y (due to L_1+L_3) components have
approximately equal contribution. At 5.54GHz the symmetrical contribution from both horizontal strips makes the X-component cancel in the far field resulting in polarization along the Y-direction.

![Graph showing S_21 and S_11 vs Frequency](image)

**Figure 3.36** Comparison of co polar and cross polar level of quad band antenna with reflection characteristics

(L_1=18mm, L_2R=L_2L=11mm, L_3=4mm, L_4=8mm, L_5=31mm, L_6=10mm, W_{GR}=W_{GL}=14mm, W_c=3mm, CD=10mm, BC=29mm, DE=HA=1mm, h=1.6mm, \epsilon_r=4.4 and g=0.35mm)

The measured 2D radiation patterns of the antenna in the four resonant bands are shown in Figure 3.37. The antenna shows good radiation characteristics in the entire band of operation. The antenna gives good polarization purity in all the operating bands. The pattern is non-directional in H-plane and directional in E-plane.
(a) E-plane 900MHz                                (b) H-plane 900MHz

(c) E-plane 1.77GHz                               (d) H-plane 1.77GHz

(e) E-plane 2.4GHz                                   (f) H-plane 2.4GHz
The simulated 3D radiation pattern of the antenna is shown in figure 3.38. A good omnidirectional radiation pattern is obtained at 900MHz and 1.77GHz band. From the pattern it is clear that the antenna has a constant gain along the XZ plane and null along the positive and negative Y-direction. The pattern is somewhat disturbed in the higher frequencies at 2.4GHz and 5.45GHz.
Figure 3.38 Simulated 3D radiation pattern of the quad band antenna (a) 900MHz (b) 1.77GHz (c) 2.4GHz (d) 5.45GHz

$L_1=18\,\text{mm}, L_{2R}=L_{2L}=11\,\text{mm}, L_3=4\,\text{mm}, L_4=8\,\text{mm}, L_5=31\,\text{mm},
L_g=10\,\text{mm}, W_{gR}=W_{gL}=14\,\text{mm}, W_s=3\,\text{mm}, CD=10\,\text{mm}, BC=29\,\text{mm},
DE=HA=1\,\text{mm}, h=1.6\,\text{mm}, \varepsilon_r=4.4\,\text{and}\, g=0.35\,\text{mm})

The current distributions on the quad-band antenna at different bands are shown in fig.3.39. These current distributions again confirm our earlier arguments about the different resonances. The current distribution in lower resonance (900MHz) is shown in fig.3.39 (a). It is obvious that the intensity of current is found to be maximum around the slit perimeter. The direction of
current is found to be opposite in both the parallel arms and it re-confirms the resonating current path described in earlier section. From fig.3.39 (b) it is found that the current distribution is maximum in the main arm (L_1) together with the lateral strip (L_{2L}+L_3+L_4). Similarly the intensity of current distribution for third frequency is found to be in the other arm (L_1+L_{2R}+L_3) as shown in fig.3.39(c). The current distribution for the fourth resonance shown in fig.3.39 (d) is found to be a variation through both the parallel opposite arms (L_1+L_{2R}+L_3 and L_1+L_{2L}+L_3+L_4). Since the arms are nearly equal in length the two resonances will come close and merge together to give a wideband response.

Figure 3.39 Simulated current distribution of the quad band antenna (a) 900MHz (b) 1.77GHz (c) 2.4GHz (d) 5.45GHz

(L_1=18mm, L_{2R}=L_{2L}=11mm, L_3=4mm, L_4=8mm, L_5=31mm, L_{6}=10mm, W_{GR}=W_{GL}=14mm, W_s=3mm, cd=10mm, be=29mm, de=ha=1mm, H=1.6mm, \epsilon_r=4.4 and g=0.35mm)
Figure 3.40 Measured Gain of the quad band antenna in four bands
$L_1=18\text{mm}, L_2=11\text{mm}, L_3=4\text{mm}, L_4=8\text{mm}, L_5=31\text{mm},$
$L_6=10\text{mm}, W_{GR}= W_{GL}=14\text{mm}, W_s=3\text{mm}, cd=10\text{mm}, bc=29\text{mm},$
$de=ha=1\text{mm}, H=1.6\text{mm}, \epsilon_r=4.4 \text{ and } g=0.35\text{mm}$

Gain of the antenna is measured using gain transfer method and is shown in figure 3.40. The peak gains of the antenna in the four bands are 1.25dBi, 1.94dBi, 1.11dBi and 3.71dBi respectively. The gain is considerably constant in the H-plane and hence is useful for omni-directional communication.

The comparison on the radiation characteristics of the dual band monopole antenna and Quad band monopole antenna is shown in Table 3.4.
Table 3.4. Comparison of Dual and Quad band monopole

<table>
<thead>
<tr>
<th>Sl. No</th>
<th>Antenna</th>
<th>Resonant Frequency</th>
<th>Band Width</th>
<th>Application Band</th>
<th>Area</th>
<th>Polarization</th>
<th>Gain</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>Dual band monopole Antenna</td>
<td>1.77GHz 5.54GHz</td>
<td>1.47-1.97 5.13-6.48</td>
<td>DCS 1800/1900 HIPERLAN ISM WLAN 5.8</td>
<td>32x31x1.6 cubic cm</td>
<td>Linear</td>
<td>3.5dBi</td>
</tr>
<tr>
<td>2</td>
<td>Quad band monopole Antenna</td>
<td>900MHz 1.74GHz 2.44GHz 5.5GHz</td>
<td>840-970MHz 1.56-1.92GHz 2.39-2.49GHz 5.07-6.23GHz</td>
<td>GSM 900 DCS-1800 ISM 2.4 5.2/5.8GHz WLAN</td>
<td>32x31x1.6 cubic cm</td>
<td>Linear Linear Linear Linear</td>
<td>1.25dBi 1.94dBi 1.11dBi 3.71dBi</td>
</tr>
</tbody>
</table>

3.6 Important conclusions from this chapter

- The signal strip of a coplanar waveguide transmission line can be effectively modified to act as a radiator.
- An additional resonance in a CPW fed monopole antenna can be generated by suitably top loading it.
- Folding technique can be effectively used to make the antenna more compact.
- Asymmetric loaded CPW antenna can be used for triple band operation.
- The introduction of slit will generate an additional lower resonance without increasing the overall compactness of the antenna.
- The presented quad band antenna is providing good impedance matching with independent control on resonances.
- Design equations are developed and are validated on different dielectric substrates.
Chapter - 3

References


....50GR.....