Development and analysis of the Compact Dual-band Coplanar Antenna is presented in this chapter. The device is developed based on the conclusions arrived in the previous chapter. The coplanar geometry is excited with a microstrip line at the corner of its wide centre strip for dual band operation. The resulting compact antenna has dimension of $\lambda/5 \times \lambda/14$, where $\lambda$ is the free space wavelength at the fundamental resonance.
6.1 Introduction

The conclusions of the investigations on coplanar waveguide presented in the previous chapter obviously indicate that a resonant mode can be easily achieved on a coplanar waveguide by exciting it with an offset feed scheme. The device exhibit good radiation characteristics. Moreover the device dimensions are very small compared to the operating wavelength. The compactness together with the excellent radiation characteristics suggests its application as an ideal antenna for compact devices like mobile phone, wireless LAN cards in laptops etc. Antennas for mobile terminals should meet certain criteria such as small size compared to the wavelength of operation, nearly omni-direction radiation coverage, reasonable gain, good efficiency etc. This novel design exhibits all the above criteria.

Previous discussions conclude that there exists a lateral current variation on the outer conductor of the SMA connector when the coplanar transmission line is excited with an offset feed scheme. It should be noted that this current variation is very important in order to sustain the resonance. A current path (a conducting path) between the two ground points should exist to excite the so called resonance. The radiation pattern study and the radiation mechanism at the fundamental resonance of the device indicate that a wide centre strip width is essentially required to obtain a broadside radiation with omni directional coverage. In order to proceed, a study with wide centre strip width is to be carried out. More over, a wide centre strip may give rise to an additional resonance on the structure. This chapter discusses the design and analysis of a compact antenna system by taking all these points into consideration. Resulting radiating configuration is an ideal candidate for compact mobile terminal applications. The
theoretical investigations, parametrical analysis and finally simple design equations are discussed in this chapter.

6.2 Offset fed coplanar geometry with wide centre strip width

A Coplanar waveguide geometry with wide centre strip width is analyzed in this section. The parametric analysis presented in the previous chapter concluded that the resonant frequency is decreasing as the centre strip width is increasing. In this section a wide centre strip width ‘w’ is employed for the study. Note that one should expect the same resonance and radiation characteristics presented in the previous chapter for this case also. Otherwise, the working of the device may not be same as that of coplanar waveguide geometry with offset feed.

When wide centre strip is employed, the device no longer behaves as a transmission line. The offset fed coplanar structure along with the parameters are depicted in Fig.6.1. The structure consists of a wide centre strip of l=9 mm, w=12.5 mm, g=0.5mm and c=5mm printed on an FR4 substrate of \( \varepsilon_r = 4.7 \) and thickness \( h = 1.6 \) mm.
The device is excited at the corner of the centre strip (P) and the two lateral ground strips are grounded. The ‘port for edge group’ scheme in IE3D is employed to excite the structure. According to the equivalent circuit representation of the offset feed explained in section 5.8 of chapter 5, a resonance should exist in this structure also, due to the potential difference between the two ground points. This potential difference is due to the asymmetry in the feeding, as explained in section 5.8. Figure 6.1 clearly shows the offset feed point P and the ground points G.
Fig. 6.2 depicts the simulated return loss characteristics of the device. The device resonates at 3.655 GHz with 7.6% return loss bandwidth.

![Graph showing return loss characteristics](image)

Return loss value at the resonance is ~19dB. The -10dB return loss bandwidth for this case is 280MHz. The resonant mode thus obtained for this case strongly indicates that the offset fed coplanar waveguide structure can excite the so called resonance by widening the centre strip. Here, the attempt to widen the centre strip is done purposefully in order to modify the radiation pattern to cater the need of mobile wireless gadget applications. Equivalent circuit representation presented in the previous chapter holds good for this case also.
Fig. 6.3 depicts the current distribution obtained from IE3D for this case. It can be seen that the wide centre strip width excites the same ‘U’ shaped resonant path on the coplanar structure. The resonance frequency in this case is due to the ‘U’ shaped resonant path on the centre strip.

It is observed from the figure that there is only negligible current variation on the middle parts of the three strips. Current densities are more intense at the edges of the strips. The ‘U’ shaped current path is very clear from the structure. The ‘U’ shaped resonant path on the centre strip is approximately equal to the half wavelength in the substrate at the resonant frequency. The ‘U’ shaped current distribution induces an oppositely flowing currents at the edges of the ground strips due to the strong coupling with centre strip and the two ground strips. Thus a horizontally polarized electromagnetic radiation is expected for this case. Width of the ground strips and the gaps has no significant influence on the resonant frequency, where as length of the waveguide and substrate materials strongly affect the resonant frequency.
Fig. 6.4 shows the 3D radiation pattern of the antenna at the resonant frequency. The radiation pattern is to be viewed according to the coordinate system mentioned in Fig. 6.1.

![3D radiation pattern](image)

The 3D radiation pattern shows that the maxima of the radiation pattern is along broad side directions. When the width ‘w’ is increased, the two radiating slots becomes far apart. This should produce a broad side radiation according to the discussions in Chapter 5, section 5.8.
The estimated gain of the antenna at the resonant frequency is 2.1 dBi. In this case the length 'l' of the structure is small compared to the operating wavelength. This will reduce the phase variation in the slots, producing good directive gain at resonance. The estimated radiation efficiency is 79% at the resonance.

During the above studies the two grounds are excited near the two slots. That means the current excitation on the ground strips are negligible at resonance. In order to attain maximum compactness for the final radiating structure an attempt were done to excite a current distribution on the ground strips along with the so called resonance due to the 'U' shaped resonant path on the centre strip, resulting a coplanar configuration as shown in Fig. 6.5.

![Diagram](image)

Fig. 6.5 Offset fed coplanar geometry with wide centre strip and ground connections at the extreme corners of the ground strips.
(a) Top view
(b) Cross sectional view

The dimension of the coplanar geometry is same as that used in the previous study. It should be noted that both the ground points are shifted symmetrically to the extreme ends of the ground strips. In compact antennas the only way to excite a low
frequency with in the small volume is to increase the current path on the structure by the meandering techniques.

The idea behind this attempt is the same current meandering technique applied in compact antenna designs. But the fundamental mode of the structure should remain as it is, otherwise this technique may result in new radiating modes. Fig. 6.6 depicts the estimated return loss characteristics of the device using IE3D for this case.

![Graph](image)

**Fig. 6.6 Return loss characteristics of offset fed coplanar geometry with wide centre strip width and ground connection at the extreme corners of the ground strip**

The structure resonates at 2.46 GHz with 16% return loss bandwidth. The bandwidth obtained from -10dB points is 400MHz. It should be noted that the resonant frequency has shifted to the lower side for this case. The return loss value at resonance is \(-19\)dB. This structure is very compact in terms of the wavelength at resonance. A
compact dimension of \( \lambda/13 \times \lambda/5 \) is achieved. The bandwidth is also higher than that obtained in the previous sections.

Analysis of surface current distribution and radiation pattern has to be carried out to confirm that the radiation mechanism at resonance is same as that of the fundamental mode of the coplanar waveguide structure with offset feeding scheme. Fig. 6.7 depicts the surface current distribution on the coplanar patches at the resonant frequency.

![Surface current distribution](image)

Fig. 6.7 Surface current distribution obtained on the offset fed coplanar waveguide structure with wide centre strip width and ground points at the extreme corners of the ground strip

The current distribution clearly shows that the two additional current paths are excited on the two lateral ground strips. An 'I.' shaped current path and 'reflected L' shaped current paths are thus formed on the two ground strips in addition to the 'U' shaped current path on the centre strip ('reflected L' shaped current path does not refer to the current reflections from discontinuities). It can be seen that the 'U' shaped current path is approximately half wavelength in the substrate and 'I.' and 'reflected L' current paths are quarter wavelength each in the substrate. Thus the total current variation on the surface of the structure is approximately full cycle variation in the dielectric.
current distribution shown in figure indicates that a strong coupling still exists between the centre strip and the two lateral ground strips. The fringing fields at the two slots are the cause of radiation for this configuration. The effect of substrate and other structural parameters are also similar to the characteristics presented in chapter 5, section 5.4. Since the feed asymmetry exists for this case, the equivalent circuit representation discussed in section 5.8 of chapter 5 holds good for this case also. Thus it can be concluded that the feeding asymmetry in coplanar structures can be used effectively to design an antenna. The feed point asymmetry is the only crucial point to excite resonance on the coplanar waveguide structure. The other parameters only govern its resonance and radiation characteristics.
Development and Analysis of a Compact Microstrip-fed Dual-band Coplanar Antenna

Fig. 6.8 shows the 3D radiation pattern of the device at the resonant frequency. Maximum radiation is found to be at the broad side directions.

The maxima at the broadside direction strongly support the fact that the radiation mechanism of this configuration is same as that explained in chapter 5, section 5.8. The two slots are the source of radiation for this case also. The only difference is that unused metallic strip is used by the surface current for more compactness. The estimated gain and efficiency of the device are 1.8dBi and 76% respectively at resonance.
6.3 Dual-band coplanar antenna configuration

The above investigations on coplanar waveguide structure concludes that a coplanar waveguide structure can excite a resonant mode in the structure if an offset feed scheme is employed, resulting a compact radiating structure with good radiation characteristics. More compactness can be achieved by feeding the ground potential to the lateral ground strips at the extreme ends.

In chapter 5 an SMA connector was used to excite the structure. The centre strip width of the coplanar waveguide was small for that case. Moreover there was a lateral variation of surface current on the outer conductor of the SMA connector. One can say that this current variation may result in an imperfect transition from the mode excited in offset fed coplanar waveguide structure to the transverse electromagnetic mode (TEM) of SMA connector. A lateral variation of surface current on the outer conductor of the SMA connector is to be avoided. Moreover when the width of the centre strip is large, the SMA connector cannot be used to excite the system. Thus a new feeding system is required to excite the coplanar geometry. A microstrip line is employed as the feed for this configuration. The microstrip line feed, importance of the ground plane in the present design, conducting pins for providing grounds to the ground strips etc. are described in this section.
6.3.1 Microstrip line feed

The microstrip line consists of a small printed strip called signal strip on the top layer of a substrate with a ground plane on the bottom layer. Width of the signal strip, substrate dielectric constant and the thickness of the substrate govern its characteristic impedance. A microstrip line feed is selected purposefully to excite the coplanar antenna. Since the width of the centre strip is wider than conventional signal strips in coplanar wave guides, there is a possibility to obtain a new resonant mode from the same structure with same feeding arrangement. Thus the microstrip feed line is placed at the extreme corner on the centre strip of the coplanar geometry. Fig. 6.9 shows the top view and cross sectional view of the microstrip line employed in the design.

![Ground plane truncation](image)

![Microstrip line feed used to excite the coplanar antenna](image)

(a) Top view  
(b) Cross sectional view

The ground plane of the microstrip line has length ‘L’ and width ‘W’ printed on the bottom layer of the substrate. A signal strip of width ‘w1’ designed for 50 Ω characteristic impedance is printed on the top layer of the substrate. Length of the signal
strip is also 'I'. The ground plane truncation is clearly shown in the figure. Typical values of the microstrip line parameters are \( L = 40\text{mm} \), \( W = 53\text{mm} \), \( w_1 = 3\text{mm} \) on FR4 substrate of thickness \( h = 1.6\text{ mm} \) and dielectric constant \( \varepsilon_r = 4.7 \). The microstrip line is excited using a 50 \( \Omega \) SMA connector. The proximity of the ground truncation to the coplanar geometry is very critical to excite the two resonant bands of the antenna. More explanations on the effect of ground plane on the final antenna configuration will be discussed in the parametric analysis section later in this chapter.

6.3.2 **Ground plane and its importance in the present design**

The ground plane of the microstrip line has another important job other than working as the ground plane of microstrip line feed. In order to sustain the resonance on the coplanar geometry a common conducting path between both the ground points (G) are very essential. This is achieved through the ground plane at the truncation edge. Two conducting pins or vias extending from the corners of the lateral strips to the ground plane truncation set up a current path between the two ground points. The current path on the resulting structure is depicted in Fig. 6.10.
The coplanar geometry is at the top layer of the substrate. But for the simplicity of understanding the operation they are placed on the same plane (XY plane) with the ground plane. The current path on the coplanar antenna structure can be split into two categories. One is called excitation path and other one the antenna resonance path. The excitation path is highly required to establish the antenna resonance path. It should be noted that the current on the ground plane is flowing through the ground plane edge at the ground truncation. The virtual voltage sources V1 and V2 due to the path difference 'w' between the two ground points of the vias are slightly at different potentials. V is the excitation signal at the tip of the microstrip line truncation and the internal resistance Zo is equal to the characteristic impedance of the feed line. As explained in section 5.8 of chapter 5, an excitation path is established in the structure. The Ex component of the electric field launched in the two slots produce the antenna resonant path and finally the radiation from the antenna structure. It is clear from the figure that the current path on
the ground plane is extremely important to sustain this mode of operation. But the current variation on the ground plane is only confined at the ground truncation edge, thus it does not alter the fundamental job of ground plane as the ground of the microstrip line. Since the current variation on the width of the coplanar strips near the ground truncation and on ground planes edges are opposite in phase at the excitation path of the antenna, there will not be radiation due to it. The radiation, at the lower resonance is solely due to the fringing fields at the two gaps of the coplanar structure. This will enhance the polarization purity of the excited electromagnetic wave from the antenna.

Since the thickness of the substrate is very small compared to the operating wavelength, there will be a strong coupling between the centre strip and ground plane at some frequencies. At a higher frequency, when the width of the centre strip is approximately equal to half of the wavelength in the dielectric a new mode will be excited on the structure. The pictorial representation of this case is shown in Fig. 6.11.

![Fig. 6.11](image)

Fig. 6.11 Current path at the second resonance on the coplanar antenna geometry when excited by a microstrip line at the corner of the centre strip

The fringing fields from the edge of the centre strip to the ground plane is the cause of radiation at this frequency. This will produce a vertically polarized
electromagnetic radiation (along Y direction) from the structure. That means the radiation at the second resonance of the coplanar antenna can be approximated to the radiation from a slot excited with half wave sinusoidal distribution, lying on XY plane of the coordinate system mentioned in the figure. It can be concluded that the ground plane of the feed line has a significant role in exciting the two resonant frequencies in the coplanar antenna. But the current distribution on the ground plane for the two modes are confined mostly to the ground truncation edge only, and it will not affect the entire circuit performance when the antenna is integrated with RF circuits, where the same ground plane is shared by the circuit.

6.3.3 Conducting pins or vias

Two conducting pins made of copper are used to connect the two ground strips to the truncated edge of the ground plane. Diameters of the conducting pins are 1 mm throughout the study. The conducting pin has a height exactly equal to the height of the substrate and the ends of the pins are soldered to the ground strip and the ground plane respectively. Instead of conducting pins via holes can be used. Performance of the via holes are almost similar to the conducting pins. This has been confirmed through simulation studies.
6.3.4 Antenna configuration

The final microstrip line excited dual-band coplanar antenna configuration is depicted in Fig. 6.12.

![Diagram of the dual-band microstrip line fed coplanar antenna](image)

Fig. 6.12 Geometry of the dual-band microstrip line fed coplanar antenna printed on a dielectric substrate
(a) Cross section view
(b) Top view

It consists of the three coplanar strips and the corner fed microstrip line signal strip of 50 Ω characteristic impedance on the top layer, a truncated ground plane near the feed line end and two conducting pins for connecting ground strips to the ground. The device is printed on a standard substrate material. Substrate materials with low dielectric constants are preferred for the design in order to avoid surface wave excitations, which will reduce the antenna efficiency and distort the radiation patterns.
The centre strip of the coplanar geometry has length 'l' and width 'w' excited by the microstrip line of length 'L' and width 'wl'. The two lateral strips are separated from the centre strips by the gap 'g'. Width of the each lateral strip is 'c'. The cross-sectional view clearly shows the via connecting the top lateral strips to the ground through the substrate. The top view of the antenna is in the XY plane of the coordinate system. Position of the feed line is fixed at the extreme corner of the centre strip to excite the first two modes of the antenna. The antenna should exhibit good impedance matching for the two resonances, when the feed line is placed at this location. The ground truncation is represented by dotted lines in the figure. A 50 Ω SMA connector is used to excite the system. When the antenna is excited, two distinct resonant modes are excited, which generate two wide bands with orthogonal polarizations. The gap 'g' should be small compared to the wavelength corresponding to the lower resonance to obtain good electromagnetic coupling between the centre conductor and the lateral strips.

Fabrication of the antenna was done using standard photolithographic techniques. Care should be taken while doing photolithography for obtaining a uniform and exact ground truncation as shown in the figure, other wise the impedance bandwidth and matching at the two resonant frequencies may be degraded. Exact aligning of the top layer and bottom layer masks is the crucial part to be considered during fabrication for achieving good accuracy. The conducting pins should be perfectly soldered to the structure. A dry soldering or imperfect connection of conducting pins result spurious resonances in the structure, other than the two expected bands. The characteristics of the device will be explained in the forthcoming sections of the chapter.
6.4 FDTD analysis of coplanar antenna

The theoretical analysis of the antenna is carried out using FDTD method. MATLAB based in-house codes were developed for analyzing the antenna. Perfect Matched Layer ABC is employed in FDTD. Return loss characteristics of the antenna and the field distribution in the structure are computed using FDTD.

6.4.1 Description of the problem

The 2D view of the FDTD computation domain defined for the microstrip line fed dual-band coplanar antenna is depicted in Fig. 6.13.

![Fig. 6.13 2D view of FDTD computation domain of microstrip line fed coplanar antenna](image)

Consider a microstrip fed dual-band coplanar antenna with the structural dimensions as $l=9$ mm, $w=12.5$ mm, $c=5$ mm, $g=0.8$ mm, $L=40$ mm, $W=53$ mm when printed on FR4 substrate of $\varepsilon_r=4.7$, and thickness $h=1.6$ mm. Width of the signal strip is $3$ mm for $50$ $\Omega$ characteristic impedance. The FDTD analysis will compute the
frequency response of the antenna by exciting it with a Gaussian pulse. The return loss characteristic of the antenna was first calculated. The E-field distributions on the top, middle and bottom layers of the antenna structure are then computed using sinusoidal excitation at the centre frequencies of the two resonant bands. In the figure the antenna geometry is surrounded by few cells of air to simulate the real condition in which the antenna is placed in air. The antenna geometry and air cells are surrounded by the PML layer.

Luebber's feed model was employed to excite the microstrip line feed of the antenna. $D_z$ component value is assigned as excitation signal at the port. $E_z$ value is automatically computed from $D_z$ value in the FDTD loop. Using the $H_x$ and $H_y$ components in the Yee cells around the excitation field the source current is calculated. Return loss of the antenna is calculated at the excitation point. The feed model employed to simulate the system is shown in Fig. 6.14.

![Perfect Electric faces](image)

Fig.6.14 Luebbers feed model employed in the FDTD computation of coplanar antenna.

In the figure the darkened portions are PEC regions in order to obtain a gradual transition from the field excitation point to the signal strip. The PEC condition is
achieved by assigning $E_z$ and $E_x$ values of the corresponding Yee cell faces are as zeros. In this computation the substrate layer is discretized as 3 cells in the $Z$-direction.

The conducting pins used to realize the prototype of the antenna is of cylindrical in shape. Fig. 6.15 shows the modeling of conducting pins in FDTD.

![Conducting pin and Stacked Yee cells with PEC](image)

Fig.6.15 Modeling of conducting pin in FDTD computation

In order to make the computation efficient in terms of time the two conducting pins are modeled as stacked Yee cells in the $Z$ direction by assigning with PEC at all the faces ($E_x=E_y=E_z=0$). This will not affect the accuracy of the computation since the frequency of operation of the device is not very high.

6.4.2 FDTD flow chart

The flow chart of PML based FDTD computation is shown in Fig. 6.16. The $\Delta x$, and $\Delta z$ in the computation domain are taken as 0.5 mm. $\Delta y$ is 1mm in the computation, this will bring down the total number of Yee cells to a great extent. These values are less than $\lambda/20$ at the maximum frequency of the computation, and gives good accuracy for the computed values. Five air cells are assigned at each side of the CPW structure to simulate the practical condition in which the antenna is in contact with

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surrounding air. A layer of cells just above the three coplanar strips and underneath the bottom layer are assigned with effective dielectric constant value as the average value of substrate dielectric constant and air to facilitate fringing field effect at the three radiating slots for lower and higher resonances in the computation. 10 cells are assigned for PML ABC at each side of the problem space. This is sufficient to effectively absorb the power at the boundary. According to the stability criteria, the calculated time step is $\Delta t = 0.83$ picoseconds.

There are 142 cells in the X direction, 87 cells in Y direction and 33 cells in Z direction of the 3D FDTD computational domain.

Fig. 6.16 FDTD Flow chart for the analysis of coplanar antenna
6.4.3 Input Gaussian pulse

The Gaussian pulse used to excite the computation domain is shown in Fig. 6.17. A narrow pulse of 10 picoseconds duration is used to excite the structure.

![Gaussian pulse used in the FDTD computation](image)

The Gaussian pulse value is employed as the voltage source for calculating the source 'dz', which is assigned as the excitation signal derived using Kirchhoff's voltage law.
6.4.4 Computed time domain characteristics at feed point

The computed time domain response of the Ez component at the feed point is depicted in Fig. 6.18.

![Time domain response of the Ez field at the port](image)

The electric field component has settled down at around 5000 time steps. The Luebbers feed model employed in the computation effectively brings down the computation time steps. Note that the Ez value at any instant of time in the port is the mixture of reflected as well as incident excitation field. The time domain response of the electric field Ez is proportional to the voltage at the port.
A prototype of the same device used in FDTD computation is fabricated using standard photolithography process. A standard 3.5 mm SMA is connected to the feed line to excite the device. The antenna was tested using HP8510C vector network analyzer. Photograph of the device is shown in Fig. 6.19. The two conducting pins connected to the extreme ends of the lateral ground strips are clear in the figure.

![Photograph of the device](image)

**Fig. 6.19.** Photograph of the compact microstrip-fed dual-band coplanar antenna for 2.4/5 GHz bands

The following section describes the results obtained using FDTD computation. Experimental results are also provided in order to substantiate the computed results.

### 6.4.5 Return loss characteristics

When the launched Gaussian pulse is complexly settled down in the computation domain the return loss value of the device is calculated. Incident voltage is the time domain Gaussian pulse, where as reflected voltage at the port is the difference of time domain values of $E_z$ field at the port and the Gaussian pulse. The time domain values are first converted to frequency domain by taking FFT of the values and then

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return loss is calculated. The computed, measured as well as HFSS simulation results are shown in Fig. 6.20.

![Fig. 6.20 Computed, measured and HFSS simulated return loss characteristics of the dual-band microstrip line fed coplanar antenna when \( l=9 \) mm, \( w=12.5 \) mm, \( c=5 \) mm, \( g=0.8 \) mm, \( L=40 \) mm, \( W=53 \) mm when printed on FR4 substrate of \( \varepsilon_r=4.7 \), and thickness \( h=1.6 \) mm.](image)

The computed return loss characteristics are almost in agreement with the measured values. The return loss characteristics computed using FDTD shows that the antenna resonates at 2.28 GHz and 5.55 GHz with -30.5 dB and -16 dB return loss for lower and higher resonances respectively. The antenna has 17% band width (-10dB bandwidth: 410 MHz) in the lower band and 20% band width (-10dB bandwidth: 1.1 GHz) in the higher band. The measured curve shows that the antenna resonates at 2.34 GHz and 5.26 GHz respectively. The measured bandwidth is 14% in the lower band (2.19-2.52 GHz) and 22% (4.84-6.07 GHz) in the higher band. The FDTD results differ from the measured values by 2.6% in the lower band and 4.3% in the higher band. More accurate results may be obtained by incorporating the effect of SMA connector.
Return loss characteristics of the antenna is simulated using Ansoft HFSS. The validity of the results obtained from FDTD computation and measurement are thus confirmed. It can be seen that HFSS results are almost same as measured values. This again confirms that the error in the FDTD and measured results are due to the contribution of SMA connector.

6.4.6 Computed electric field values at the top layer, middle layer and bottom layer of the substrate in the two bands

The return loss characteristics of the coplanar antenna discussed above concludes that the device exhibit two wide resonant bands at around 2.4 GHz and 5GHz. The wide band width from such a small radiating geometry demands its use in many commercial applications, where the size of the antenna is a major concern. The gain, radiation pattern, efficiency are to be studied in order to completely accept this as a new efficient radiating system. Following section gives more insight into the radiation mechanism of the device at the two resonant bands. The computed E-field distribution at the two resonant frequencies, at the three layers of the antenna are presented to confirm that the radiation from the device at the lower band is from the two gaps of the coplanar geometry, where as the radiation at the higher band is due to the fringing fields at the gap separating the width of the centre strip and the truncated ground plane.

The computed E-fields at top layer, middle layer and bottom layer of the substrate obtained by giving the sinusoidal excitations at the two resonant frequencies are presented to confirm it.
The $E_x$, $E_y$ and $E_z$ components at 2.28 GHz are shown in Fig. 6.21 (a)-(i). This is the first resonant frequency of the antenna obtained using FDTD computation.

A. Top layer

![Fig. 6.21 (a)](image)

![Fig. 6.21 (b)](image)
B. Middle layer
C. Bottom layer

Fig. 6.21 (g)

Fig. 6.21 (h)
Fig. 6.21 (a), (d) and (g) shows the computed Ex-fields at the top, middle and bottom layers of the antenna. The FDTD computed results shows that Ex values at the two slots are dominating at the first resonance. At the top layer a maximum of 0.5 V/meter is observed when the system is excited by 1V/meter (maximum amplitude of the gaussian pulse). Ex values are slightly less at the middle layer and bottom layer of the structure at locations corresponding to the slots. The Ex values at middle and bottom layers are ~0.1V/meter. It should be noted that the Ex fields in the slots are in phase components. This is same as the results observed in the case of offset fed coplanar waveguide discussed in chapter 5.
The computed $E_y$ components presented in Fig. 6.21 (b), (e) and (h) clearly shows that there is only negligible $E_y$ component in the structure at the first resonance. In the top layer of the antenna ~0.05V/meter is observed at the gap separating ground truncation and the right lateral ground strip. This is due to a weak coupling between the right lateral strip and ground truncation edge. This condition is also similar to the FDTD results observed in the case of offset fed coplanar waveguide presented in chapter 5.

$E_z$ components presented in Fig. 6.21 (c), (f) and (i) indicate that the $E_z$ value is 180 degree out of phase in the two slots in top, bottom and middle layers of the structure at the first resonance. 0.1V/meter of electric field is observed on top and middle layers. But the $E_z$ field is very feable in the bottom layer of the substrate. The 180 degree out of phase $E_z$ values will be cancelled at the far field and does not contribute to the radiated field from the structure.

The interpretation of field values presented above concludes with two important points. Radiation at the lower resonance is mainly due to the contribution of in phase $E_x$ components at the two slots. This inference is exactly similar to the results obtained for the electric field distribution in the case of a coplanar waveguide excited with corner feed. Owing to the similarity of radiation mechanism of coplanar antenna at the fundamental resonance to the radiation mechanism of an offset fed coplanar waveguide, the present design is not a wide monopole element with two shorted parasitic strips. In
the case of monopoles the radiation should be due to the current distribution on the top patch and its image on the ground plane (ground plane of the microstrip line in this case).

**Li-field components at the second resonance**

According to the discussions in section 6.3.2, a half wave length slot with sinusoidal field variation lying in the XY plane of the antenna structure, approximating the fringing field from the edge of centre strip width to the ground truncation is the cause of radiation at the second resonance. This indicates that Ex field values should be absent in the radiating slot at this frequency. Where as Ey and Ez field values can exist. The Ex, Ey and Ez components at 5.55 GHz are shown in Fig. 6.22 (a)-(i). This is the second resonant frequency of the antenna obtained using FDTD computation. The field components at the three layers of the substrate are presented. The field components are computed by exciting the antenna using sinusoidal excitation function with f=5.5 GHz.

**A. Top layer**

![Fig. 6.22 (a)](image-url)
Fig. 6.22 (b)

Fig. 6.22 (c)

Development and Analysis of a Compact Dual-band Coplanar Antenna
Middle layer

Figure 6.22 (d)

Figure 6.22 (c)
Chapter 6

C. Bottom layer

Fig. 6.22 (f)

Fig. 6.22 (g)

Development and Analysis of a Compact Dual-band Coplanar Antenna
Fig. 6.22 (a)
Fig. 6.22 (a), (d) and (g) shows the Ex component of electric fields present in the top, middle and bottom layers of the antenna structure at the second resonance. It is clear from the figure that Ex component has ~0.2 V/meter amplitude in the two slots of the coplanar geometry at the top layer. But they are out of phase components and cancels at the far field, resulting no radiation. In the middle and bottom layer Ex component has only a feasible value at the location underneath the two slots and are out of phase each other. It is worth noting that during all these cases the Ex value at the location of the slot approximating the fringing field from centre strip edge to the microstrip line ground truncatin is zero. This indicates that Ex component is not contributing radiation from the coplanar antenna in the second resonant band.

The computed Ey components on the top, middle and bottom layer of the antenna structure at the second resonance is presented in Fig. 6.22 (b), (e) and (h). A half wave variation of Ey field component is observed at the second resonance. A peak amplitude of 0.15 V/meter is observed at the top layer. Ey component on the radiating structure is found to be present only at the slot. This support the concept present in section 6.3.2 that the radiation at the second resonance is solely due to the radiation from the fringing field along the width of the centre strip to the ground plane edge.

Ez components in the antenna structure at the second resonance is depicted in Fig. 6.22 (c), (f) and (i). Ez component has got only very small value on top, middle and bottom layers. It is worth noting that Ey component is the only strong field component present in the structure on the three layers at the second resonance. Moreover, the electric
field components present in feed lines are mutually opposite during all the cases and no radiation is expected from the feed structure. That is the antenna at the second resonance will produce a vertically polarized electromagnetic radiation due to the Ey component of the fringing field along the width of the centre strip edge to the ground plane edge. Since the field variation in the radiating slot is approximately $\lambda d/2$, the radiation behavior will not be same as that obtained for the first resonance.
6.4.7 Conclusions

This part of the chapter describes the important points brought out from the theoretical investigation of the dual-band coplanar antenna using FDTD method.

- The return loss characteristics in the two bands indicate that the antenna excites two resonant bands with wide impedance band width.
- The proposed dual-band antenna supports the concept of resonance due to the offset feeding in coplanar waveguides.
- Microstrip line fed coplanar antenna configuration looks similar to a wide monopole antenna with shorted parasitic patches. But according to the FDTD results this apprehension is no longer valid. The computed E-field values at the fundamental resonance are not same as that usually present in the monopoles.
- The computed E-field at the second resonance on the antenna shows that Ey component is very strong at the slot approximating fringing field from width of the centre strip to the bottom ground plane truncation.
- A half wave length variation of E-field is observed at the radiating slot in the second resonance.
- The electromagnetic field is horizontally polarized (X direction) in the lower band of the antenna.
- The electromagnetic field is vertically polarized (Y direction) in the second band.
Since top layer and bottom layer of the antenna structure are supporting the same E-field distribution for the radiating field components in both the bands, a bidirectional radiation pattern is expected from the structure in the two operating bands.

PML based FDTD approach have given a clear insight into the problem. The return loss characteristics, radiation mechanism and the polarization of the EM energy radiated from the structure at the two resonances are clearly understood from the FDTD analysis.

6.5 Parametric analysis

In this section the parametric analysis of the device is discussed in detail. The structural parameters of the antenna are varied in order to study its effect on the two resonant frequencies and thus to confirm the two modes of the proposed antenna. The resonant frequency and impedance band width are the main parameters studied. Parametric analysis also leads to the formulation of simple equations to design the proposed antenna for any two bands of operation. In order to perform the parametric analysis the coplanar antenna with structural parameters as \( l=9 \) mm, \( w=12.5 \) mm, \( c=5\) mm, \( g=0.8\) mm, \( L=40\) mm, \( W=53\) mm when printed on FR4 substrate of \( \varepsilon_r=4.7\), loss tangent 0.02 and thickness \( h=1.6\) mm is used. Width of the signal strip is 3 mm for 50 \( \Omega \) characteristic impedance.
6.5.1 Effect of length \( l' \) on resonant frequencies and bandwidth

Fig. 6.23 shows the effect of length \( l' \) of the coplanar antenna on the return loss characteristics.

The length \( l' \) is varied from 9 mm to 15 mm. In this case, the resonant frequency in the lower band changes from 2.33 GHz to 1.77 GHz, while the higher resonance varies from 5.2 GHz to 5.18 GHz only. Here, all other parameters of the antenna are kept constant. It can be concluded that the lower resonance decreases as \( l' \) increases and the upper resonance remains unchanged. Because as \( l' \) increases symmetrically for the three strips, the resonant length for the first band increases four times and thus the resonant frequency in the first band decreases more rapidly. Since width of the centre strip is kept
constant during this study the second resonance remains almost unchanged. Second resonance is solely due to the width ‘w’ of the centre strip.

Influence of ‘l’ on the bandwidth performance is also very important. Return loss characteristics of the fabricated antenna shows that the two bands are wide. The influence of length ‘l’ on the -10dB return loss bandwidth at the two resonant bands are shown in Fig. 6.24.

![Graph showing bandwidth performance](image)

**Fig. 6.24** Effect of dimension ‘l’ on the bandwidth at the two resonant bands of the microstrip line fed dual-band coplanar antenna

It is noted that bandwidth of the antenna’s second resonant band increases slightly as l increases. When ‘l’ changes from 9 mm to 15 mm the band width in the lower band remains unchanged, but in the second band it varies from 19% to 23%. This is due to the increase in surface area of the centre strip as l increases. It is a well proved fact that thick dipole elements yield more band width. The principle holds good for this case also.
6.5.2 Effect of centre strip width ‘w’ on resonant frequencies and band width

The centre strip width ‘w’ is a common dimension for the two resonances. The ‘U’ shaped current path in the lower band uses ‘w’ to sustain the resonance. In the higher band width ‘w’ is the sole cause of resonance in the structure. Influence of dimension ‘w’ on the two resonant frequencies are depicted in Fig. 6.25.

![Figure 6.25](image)

Fig. 6.25 Effect of dimension ‘w’ on the return loss characteristics of the microstrip line fed dual-band coplanar antenna

Figure shows that when ‘w’ increases from 10 mm to 13 mm the lower resonance changes from 2.41 GHz to 2.32 GHz and the upper resonance changes from 5.5 GHz to 5.32 GHz. That is the higher resonance is more rapidly varying than the lower resonance due to the change in ‘w’. Because as ‘w’ increases, as explained, the resonant path for lower resonance increases and thus first resonance decreases. The ‘w’ contributes a small part in the total resonant length in the lower resonance, where as ‘w’
is the sole part of resonance in the higher band and thus influences more at the higher band.

The variation of bandwidth in the two bands with the change in ‘w’ is depicted in Fig. 6.26.

Fig. 6.26 Effect of dimension ‘w’ on the bandwidth at the two resonant bands of the microstrip line fed dual-band coplanar antenna

When ‘w’ changes from 10 mm to 13 mm return loss bandwidth in the first band increases as shown in the figure, where as the bandwidth in the second band remain unchanged. Because the increase in ‘w’ increases the surface area for the surface current on the centre strip at the lower resonant frequency, where as it doesn’t increase the surface area for the current on the strip corresponding to upper resonance, and thus the bandwidth remains unchanged in the second band.
6.5.3 Effect of lateral strip width ‘c’ on resonant frequencies and band width

The effect of the lateral strip width ‘c’ on the two resonant frequencies are depicted in Fig. 6.27. In order to facilitate the imbalance in the feed the two lateral ground strips are varied equally.

![Graph showing the effect of antenna length 'c' on return loss characteristics of the microstrip line fed dual-band coplanar antenna.](image)

**Fig. 6.27** Effect of antenna length ‘c’ on the return loss characteristics of the microstrip line fed dual-band coplanar antenna

The lateral strip width affects the first resonance very much, where as the second resonant frequency remains almost same. When ‘c’ varies from 5 mm to 7 mm the resonant frequency in the first band has changed from 2.39 GHz to 2.1 GHz. When ‘c’ increases symmetrically for the two strips, the resonant length for the first band increases two times, and thus the resonant frequency for the first band decreases. Since resonant length ‘w’ is not changed, the resonant frequency in the second band remains almost unchanged.
constant. The resonant frequency has only very small change when ‘c’ changes from 5 mm to 7 mm.

Influence of the lateral strip width on the band width performance of the antenna in the two bands are depicted in Fig. 6.28.

The impedance bandwidth in the two resonant bands are slightly affected when width of the two lateral strips are changed. When the width changes from 5 mm to 7 mm the bandwidth in the lower resonance changes from 9% to 12 %, where as the bandwidth in the higher resonance degrades slightly. When the lateral strip width is increased the surface current along the length of the lateral strips has got more surface area, which in turn will increase the band width in the lower band slightly. Since the surface current at the lateral strips have only negligible value at the higher resonance the effect is negligible for this case.
6.5.4 Effect of gap ‘g’ on resonant frequencies and band width

Gaps ‘g’ in the structure is the discontinuity forming radiation at the lower resonance of the structure. The ‘g’ determines the coupling between the centre strip and the two lateral strips. In order to achieve strong coupling the gap ‘g’ should be very small compared to the operating wavelength at the lower resonance. In the same time close proximity of ‘g’ may affect the resonant property of the higher band. Moreover the change in ‘g’ strongly affects the return loss. After exhaustive experimental and simulation studies ‘g’ is selected as, 

\[
g = 0.014\lambda_d
\]

where \(\lambda_d\) is the wave length in the dielectric corresponding to the lower resonance. This value gives good return loss response in the two resonant bands.

6.5.5 Effect of conducting pin dimension on resonant frequencies

The conducting pins or vias are head ache in almost all microwave circuits. Its length and diameter affects the performance of the circuit. The effect of conducting pin diameter on the two resonant frequencies is depicted in Fig. 6.29.
Conducting pins has inherently an inductance associated with it. The inductance increases as its length increase. But in this context the length of the conducting pin is equal to the substrate height, and is very small compared to the operating wave length at the lower resonant frequency. Length of the conducting pins is to be considered while calculating the lower resonance of the antenna. Variation study is done using HFSS simulation package. When the diameter of the conducting pin is increased from .5 mm to 2 mm, the resonant frequency is increased by only 20 MHz, at the lower band. Where as, at the higher band the resonant frequency is almost remains constant. The increased pin diameter reduces the resonant length, which in turn increases the resonant frequency in the lower resonance. Since the higher resonance is due to the width of the centre strip the conducting pin dimension has not affected the resonance.
Effect of ground plane parameters on the antenna performance is an important study as far as a complete communication system is concerned. In practice, the dimension of the ground plane is tailored according to the stringent specifications of communication systems. Following sections describe the effect of ground plane dimensions on the antenna characteristics.

6.5.6 Effect of ground plane length ‘L’ on the resonant frequencies and bandwidth

The effect of ground plane length ‘L’ on the resonant frequencies of the antenna are depicted in Fig. 6.30. It is observed that the ground plane dimension is not affecting the two resonant frequencies significantly. The lower resonance has increased by 50 MHz when the ground plane length is reduced from 40 mm to 10 mm.

![Fig. 6.30 Influence of ground plane length ‘L’ on the dual band coplanar antenna resonant frequencies](image-url)
The return loss value is found to be very poor at the higher resonance when length of the ground plane is 10 mm. But the resonant frequency is remains same. The influence of 'L' on the band width performance of the antenna in the two bands are depicted in Fig. 6.31.

It clear from the figure that the ground plane length 'L' significantly affects the bandwidth performance of the antenna in the two bands. When 'L' varies from 10 mm to 40 mm, the band width of the antenna in the lower band changes from 10 % to 14 %, where as in the higher band it changes from 16 % to 22 %. It is also noted that the bandwidth in the lower band remains constant above L=30 mm. It can be concluded that the ground plane is not affecting the resonant frequencies of the antenna significantly, but the impedance bandwidth characteristics in the two bands are affected. The worst case bandwidth (when L=10 mm) is 10% in the lower band and 16 % in the higher band.

![Diagram](image)

Fig. 6.31 Influence of ground plane length 'L' on the band width characteristics of the dual band antenna
6.5.7 Effect of ground plane length 'W' on the two resonant frequencies and band width

The influence of ground plane width 'W' on the two resonant frequencies of the antenna is depicted in Fig. 6.32. Width of the ground plane cannot be reduced below the total width of the coplanar radiating structure.

Fig. 6.32 Influence of ground plane width 'W' on the dual band antenna resonant frequencies

Since the lateral ground strips are connected to the ground truncation at the extreme corners, the left and right portions of the ground plane width extending from the conducting pins are symmetrically reduced to study the effect of ground plane width 'W' on the resonant frequencies. The dimension 'W' is not affecting the two resonant frequencies.
The bandwidth performance of the antenna in the two resonant bands with the various ground plane width is depicted in Fig. 6.33.

![Graph](image-url)

**Fig. 6.33 Influence of ground plane width 'W' on the bandwidth performance of the dual band antenna**

When W varies from 37 mm to 53 mm the bandwidth in the lower resonance varies from 11 to 14 %, whereas in the higher resonance the bandwidth has a small shift from 20 % to 22 %. Since the resonant frequencies of the antenna is not affected by the ground plane dimensions the device is suitable for integration in compact wireless modules where the size of the ground plane will be very small. It should be noted that the antenna with small ground plane dimensions reduces the bandwidth performance of the antenna.
6.5.8 Effect of dielectric constant $\varepsilon_r$ on resonant frequencies

The dielectric constant of the substrate material has influence on the resonant frequencies. When the antenna including the feed structure is fabricated on a dielectric substrate, the change in dielectric constant will affect the characteristic impedance of the feed line, which in turn will reduce the return loss. Due to the scarcity of different laminates, simulation results are presented to describe the effect of substrate dielectric constant on the two resonances. Fig. 6.34 depicts the effect of dielectric constant on the resonant frequencies of the antenna.

When $\varepsilon_r$ varies from 4.7 to 6 the lower resonant frequency reduces by 90 MHz, whereas the higher resonance reduces by 200 MHz. In the case of lower resonance the electric field components are not confined within the substrate. The average dielectric
constant of air and substrate material has to be considered while calculating the resonant frequency in the lower band. Since the gap ‘g’ is very small, the small variation in the substrate dielectric constant will not affect the lower resonance very much. At the second resonance the fringing field from width of the centre strip to the ground truncation has more influence (because of large gap) with the substrate material. This in turn will change the second resonance.

The parametric analysis presented above has brought out certain important points. The important points arrived from the parametric analysis are given below.

- The lower resonant frequency is influenced by length and width of the centre strip and the two lateral ground strips.
- The higher resonance is solely due to the width of the centre strip
- The conducting pin length should be accounted while calculating the resonant frequency in the lower band
- The rectangular geometry effectively increases the surface area for the meandered surface current on the there strips, and there by exhibiting wide band width in the lower band
- Length of the centre strip increases the band width in the higher band
- The ground plane dimensions of the antenna has no significant effect on the antenna resonant frequency. But the impedance band width in the two bands are greatly affected by the length and width of the ground plane.
Effective dielectric constant value is to be used for calculating the lower and higher resonant frequencies of the antenna.

6.6 Far field radiation and polarization

The far field radiation measured using a standard horn antenna is depicted in Fig. 6.35.

![Graph showing far field radiation of dual-band coplanar antenna received by a standard horn antenna.](image)

**Fig. 6.35 Far field radiation of dual-band coplanar antenna received by a standard horn antenna.**

When the two slots of the coplanar antenna are perpendicular to the E-plane (Horizontal polarization) of the horn antenna, the horn antenna receives maximum power. Thus horizontally polarized EM energy is received. ~18 dB polarization discrimination is achieved at the lower resonant frequency. In the higher band, the antenna produces a vertically polarized EM energy. In this case the slot separating the
centre strip to the ground truncation is parallel to the H-plane of the horn antenna. Polarization discrimination for this case is only -10dB.

### 6.7 Principal plane radiation patterns

The principal plane radiation pattern of the typical antenna at 2.19, 2.34 and 2.52 GHz are shown in Fig. 6.36.

![Fig. 6.36 (a) E and H plane radiation pattern of the dual-band coplanar antenna at 2.19 GHz](image)

![Fig. 6.36 (b) E and H plane radiation pattern of the dual-band coplanar antenna at 2.34 GHz](image)
As expected the radiation pattern in the E plane has figure of eight shape, whereas the H plane radiation patterns are nearly non-directive. The antenna has almost identical radiation pattern throughout the 2:1 VSWR band width in the first band. Large cross-polarization levels are observed in the radiation patterns. The weak coupling of the width of centre strip and the two lateral strip sides to the ground plane increases the cross-polarization level in the lower band. The non-directive behavior of radiation pattern suggests its use in mobile wireless gadgets.
Fig. 6.37 (a), (b) and (c) shows the E and H plane radiation patterns of the antenna at 4.84, 5.26 and 6.07 GHz respectively.
The radiation patterns in the higher band are found to have good radiation coverage in both the principal planes. The radiation patterns are nearly identical throughout the 2:1 VSWR bandwidth. Measured radiation patterns show large cross polarization values in the higher band. This may be due to the fringing field components present in the two slots separating the centre strip from the two lateral strips at the higher resonant frequency.
6.8 Gain and radiation efficiency in the two bands

Gain of the coplanar antenna in the lower resonant band is shown in Fig. 6.38.

The antenna exhibit almost constant gain through out the band. Maximum observed gain is 3.7 dBi at 2.5 GHz. Gain of the antenna on different ground planes are also measured. It is found that there is no significant degradation in gain even when the ground plane length is reduced to 20 mm. The measured gain of the antenna in the second band is shown in Fig. 6.39. Gain of the antenna is not constant though out the impedance band. It can be seen that the antenna gain is increasing at the higher end of the impedance band. The gain in the higher band varies from 3 dBi to 4 dBi. From the FDTD simulation it is found that the amplitude of fringing field is almost constant at higher band. So the gain increases with frequency.
The estimated radiation efficiencies using IE3D simulation package is 85% in the lower band and 73% in the higher band when the antenna is fabricated on FR4 substrate. Note that the loss tangent of the FR4 substrate is 0.02, which reduces the radiation efficiency. The estimated efficiency is found to be increasing when the antenna is designed on RT duroid substrate. In this case the efficiency is 89% and 79 % respectively in the lower and higher bands of the antenna.

6.9 Design procedure

The experimental as well as theoretical investigations gave insight to radiation mechanism and effect of various antenna parameters on the radiation characteristics. Inferences of these investigations leads to the formation of design equations for the
Development and Analysis of a Compact Microstrip-fed Dual-band Coplanar Antenna

microstrip fed dual-band coplanar antenna. The following equations can be used to design the antenna with good radiation characteristics.

1. Select any substrate with relative dielectric constant \( \varepsilon_r \) and thickness \( h \), and calculate the width \( (w_1) \) of the microstrip transmission line for 50 \( \Omega \) characteristic impedance.

2. Calculate width of the centre strip \( (w) \) using the following equation.

\[
w = \frac{c}{2f_2 \sqrt{\varepsilon_{re}}} \tag{1}
\]

where \( c \) is the velocity of light and \( f_2 \) is the second resonant frequency.

Since the field components are not confined to the substrate alone the effective dielectric constant \( (\varepsilon_{re}) \) has to be used in calculations instead of relative permittivity of the substrate.

\[
\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} \tag{2}
\]

3. The length \( l \) of the three rectangular strips is then calculated as

\[
l = \frac{0.15c}{f_1 \sqrt{\varepsilon_{re}}} \tag{3}
\]

Where \( f_1 \) is the first resonant frequency

4. Width of the lateral conductors \( (C) \) is calculated using the equation given below.
\[ C = \frac{c}{2f_1\sqrt{\varepsilon_{re}}} - \left(\frac{4l + 2h + w}{2}\right) \]  

(4)

Where \( h \) is the thickness of the dielectric substrate.

5. Gap separating centre strip from the lateral strips is then calculated

\[ g = \frac{0.014c}{f_1\sqrt{\varepsilon_{re}}} \]  

(5)

Where \( c \) is the velocity of electromagnetic signal in free space.

The constants 0.15 and 0.014 in equations (3) and (5) respectively are obtained after exhaustive experimental and simulation studies.

6. Ground plane dimensions are calculated using the following equations.

\[ L = \frac{0.12c}{f_1\sqrt{\varepsilon_{re}}} \]  

(6)

\[ W = \frac{0.98c}{f_1\sqrt{\varepsilon_{re}}} \]  

(7)

The constants 0.12 and 0.98 are derived empirically after studying the effect of ground plane on the two resonant frequencies.

7. The two extreme corners of the lateral conductors are connected to ground plane of the microstrip line using vias or conducting pins.
In order to confirm the validity of above equations various antennas were designed using the above equations. The return loss characteristics of an antenna designed on RT duroid ($\varepsilon_r=2.2$ and $h=1.5$ mm) substrate for 2.4/5GHz WLAN operation using the above equations are shown in Fig. 6.40. A photograph of the fabricated antenna is shown in Fig. 6.41. The antenna is resonating at 2.54 GHz and 5.59 GHz with 13% and 19% return loss bandwidth in lower and higher bands respectively.

Fig. 6.40 Return loss characteristics of the antenna printed on RT duroid substrate for 2.4/5GHz WLAN applications
The antenna offers two wide resonant bands around 2.54 GHz and 5.59 GHz. Another antenna on FR4 substrate with \( l = 23 \) mm, \( w = 40 \) mm, \( c = 11 \) mm, \( g = 1 \) mm \( w1 = 3 \) mm on a ground plane of \( L = 30 \) mm, \( W = 70 \) mm gives two resonances at 0.89 GHz and 1.91 GHz with 10\% band width in the lower band and 14\% band width in the higher band. Accuracy of the above design equations are within 3\%.

The measured and FDTD results of the antennas designed using the above design equations on different substrates are summarized below in Table 6.1.
The measured gain of the antenna is slightly increased when the antenna is fabricated on RT Duroid substrate. The low dielectric loss property of RT droid substrate increases the radiation efficiency slightly, which in turn increases the gain of the antenna.

6.10 Comparison with rectangular microstrip antenna (RMSA)

The printed antenna technology has gained popularity after the introduction of microstrip antennas. Owing to its conformal nature design and analysis of microstrip antennas have been addressed in literature very much. In this section the coplanar antenna is compared with a rectangular microstrip antenna in order to highlight its feasibility as efficient microwave antenna.
<table>
<thead>
<tr>
<th>STANDARD RECTANGULAR MICROSTRIP ANTENNA (RMSA)</th>
<th>Coplanar Antenna</th>
</tr>
</thead>
<tbody>
<tr>
<td>Half wave length resonant structures. Large size, nearly λd/2 size long and width larger than λd/2 to obtain maximum efficiency</td>
<td>Extremely compact. Nearly λ/5 x λ/14 size.</td>
</tr>
<tr>
<td>Simple and low fabrication cost</td>
<td>Easy to fabricate and low fabrication cost</td>
</tr>
<tr>
<td>Easy integration with microwave circuits</td>
<td>Coplanar antennas can be easily integrated with microwave circuits.</td>
</tr>
<tr>
<td>Narrow bandwidth</td>
<td>Broadband</td>
</tr>
<tr>
<td>7.2 dBi gain</td>
<td>~3.7 dBi gain</td>
</tr>
<tr>
<td>Substrate losses are high</td>
<td>Substrate losses are minimum due to the weak interaction of field components with substrate material</td>
</tr>
<tr>
<td>Uni-polar radiation pattern</td>
<td>Bi-polar radiation pattern</td>
</tr>
<tr>
<td>Vias are essential for active antenna applications</td>
<td>The DC ground potential is available on the top layer on the lateral ground strips, avoids the need of vias in many active antenna applications</td>
</tr>
</tbody>
</table>
6.11 Conclusions

A novel planar antenna is designed using the concept of resonance and radiation in offset fed coplanar waveguides. The FDTD analysis has made an insight into the radiation phenomena of the coplanar antenna. Coplanar antenna element exhibit similar properties observed in the case of coplanar waveguides with offset feed discussed in Chapter 5. The radiating system is extremely compact. The radiation mechanism and resonance of the coplanar waveguide is exhaustively explained in the thesis using the FDTD and experimental observations. Parametric analysis is also done to optimize the antenna. Measured radiation patterns of the antenna are broad and demand its application in compact wireless modules. Ground plane dimensions of the antenna are not significantly affecting the resonant frequency in the two bands. But bandwidth in the two bands is strongly affected by the ground plane dimensions. Measured peak gain of the coplanar antenna is $\sim$4 dBi in the two bands. The design relations are also developed to synthesize the coplanar antenna dimensions for any two bands.