2  Microstrip Antennas

2.1  Introduction

The rapidly growing communication industry demands for smaller and low profile antennas have put the microstrip antennas (MSA) to the forefront [12-16]. An MSA in its simplest form consists of a radiating patch on one side of a thin dielectric substrate backed by a ground plane. The radiating patch could be of any arbitrary shape but generally regular shapes are considered for ease of analysis and design. One of the simplest and widely used configurations is rectangular MSA as shown in figure 2.1. A rectangular patch is defined by its length $L$ and width $W$. For a simple microstrip line the width is much smaller than the wavelength. However, for the RMSA, the width is comparable to the wavelength to enhance radiation from edges. The thickness is much smaller than the wavelength.

Figure 2.1  Microstrip antenna
For the fundamental TM\textsubscript{10} mode, the length L should be slightly less than \(\lambda/2\), where \(\lambda\) is wavelength in dielectric medium. \(\lambda = \lambda_0/(\varepsilon_e)^{0.5}\), where \(\lambda_0\) is free space wavelength and \(\varepsilon_e\) is the effective dielectric constant. The value of \(\varepsilon_e\) is slightly less than \(\varepsilon_r\), because the fringing fields around the periphery of the patch are not confined in the dielectric substrate but are also spread in air as shown in figure 2.2.

The fundamental mode TM\textsubscript{10} implies that the field varies one \(\lambda/2\) cycle along the length, and there is no variation along the width of the patch. Along the width of the patch, the voltage is maximum and current is minimum due to the open end. It may be observed from figure 2.2 that the vertical components of the electric field at the two edges along the width are in opposite directions and hence cancel one another in the broadside direction, whereas the horizontal components are in same direction and hence combine in the broadside direction. Therefore edges along the width are termed as radiating edges. The fields due to the sinusoidal distribution along the length cancel in broadside direction, and hence edges along the length are known as non-radiating edges [17]. The fringing fields along the width can be modeled as radiating slots as shown in figure 2.3.
2.2 Advantages of Microstrip Antennas

The main advantages of microstrip antennas are:
a) The microstrip antennas are lightweight and have small volume and low profile planar configuration.
b) They can be made conformal to the host surface.
c) Their ease of mass production using printed circuit technology leads to a low fabrication cost.
d) They are easier to integrate with other MIC’s on the same substrate.
e) They allow to both linear and circular polarization.
f) They can be made compact for use in personal mobile communication.
g) They allow for dual and triple frequency operation.

2.3 Drawbacks of Microstrip Antennas

The main drawbacks of microstrip antennas are:
a) They are narrow band.
b) They are having low gain.
c) They are having low power handling capability.
d) They do not offer polarization purity.

In this thesis, efforts have been made to overcome above drawbacks using proper techniques. One of the serious limitations of the microstrip antennas is its narrow bandwidth. The impedance bandwidth of MSA is around 1% only [19] for thin substrates. The VSWR or impedance bandwidth of the MSA is defined as the frequency range over which it is matched with that of the feed line within specified limits. The bandwidth of the MSA is inversely proportional to its quality factor Q and is given by

\[ BW = \frac{(VSWR - 1)}{Q\sqrt{VSWR}} \]
In order to increase the bandwidth, Q should be decreased. The required value of Q can be obtained by varying dielectric constant $\varepsilon_r$ and thickness $h$. The Q increases with increase in dielectric constant $\varepsilon_r$. It decreases with increase in $h$. In order to increase the bandwidth, simply increasing the substrate height keeping dielectric constant low, will create following problems [18]:

1) Substrates thicker than $0.11\lambda_0$ for $\varepsilon_r = 2.2$ make the impedance locus of the probe fed patch antenna increasingly inductive in nature, resulting in impedance matching problem.
2) Thick substrate with microstrip edge feed will give rise to increased spurious radiation from the microstrip step in width and other discontinuities. Radiation from probe feed will also increase.
3) Higher order modes along the thickness may develop, giving rise to distortions in the radiation patterns and impedance characteristics. This is a limiting factor in achieving octave bandwidth.

Therefore to avoid above problems thicker substrate is not used to obtain wide bandwidth. The common techniques to improve bandwidth are as follows:

a) Planar multi-resonator configurations: The planner multiresonator gives bandwidth around 20%. The drawback of this configuration is its large planner size[20].

b) Electromagnetically coupled MSA: With this configuration we can obtain 10% to 30% bandwidth. The increase in the bandwidth is obtained due to increase in the overall height of the antenna, a decrease in the effective dielectric constant $\varepsilon_e$ and the multi-resonator effect [17].

c) Aperture coupled MSAs : Aperture coupled elements have been demonstrated with bandwidths up to 10 - 15% with a single layer[22]-[24], and up to 30-50% with a stacked patch configuration [25]-[28].

d) Impedance matching networks for broad band MSAs: The complexity and losses of the matching network generally limit the achievable bandwidth of the antenna to about 10% to 30% [18, 29]

e) Log periodic MSA configurations.
2.4 Wave Polarization

Microstrip antennas are having many advantages as compared to the conventional antennas. They are having few drawbacks such as narrow bandwidth, high cross polarization levels or low polarization purity, low gain and low power handling capability. How to increase bandwidth of microstrip antenna is already discussed in the previous chapter. In this chapter we shall discuss about polarization, the types of polarization and the various methods to achieve these polarizations.

Antenna polarization is a very important consideration when choosing and installing an antenna. Most communications systems use either vertical, horizontal or circular polarization. Knowing the difference between polarizations and how to maximize their benefit is very important to the antenna user. A linear polarized antenna produces linearly polarized wave, a circular polarized antenna produces circularly polarized wave. Recently, the microstrip antenna with switchable polarizations [30-37] has received much attention because it is useful for some specific application in wireless communication, such as multi-system operation, frequency reuse and reducing multipath fading. The various types of wave polarizations are explained in detail in the following section.

An important property of an electromagnetic wave is its polarization, a quantity describing the orientation of the electric field $E$ [38]. Consider a plane wave traveling out of the page (positive $z$ direction), as in Figure 2.4, with the electric field at all times in the $y$ direction. This wave is said to be linearly polarized (in the $y$ direction). As a function of time and position the electric field of a linearly polarized wave, as in Figure 2.9(a), traveling in the positive $z$ direction (out of the page) is given by

$$E_y = E_2 \sin(\omega t - \beta z)$$

In general, the electric field of a wave traveling in the $z$ direction may have both a $y$ component and an $x$ component, as suggested in Figure 2.4 (b). In this more general situation the wave is said to be elliptically polarized. At a fixed value of $z$ the electric
vector $E$ rotates as a function of time, the tip of the vector describing an ellipse called the **polarization ellipse**. The ratio of the major to minor axes of the polarization ellipse is called the **axial ratio** (AR). Thus, for the wave in Figure 2.4 (b), $AR = E_2/E_1$. Two extreme cases of elliptical polarization correspond to **circular polarization**, as in Figure 2.4 (c), and **linear polarization**, as in Figure 2.4 (a). For circular polarization $E_x = E_2$ and $AR = 1$, while for linear polarization $E_x = 0$ and $AR = \infty$.

In the most general case of elliptical polarization the polarization ellipse may have any orientation, as suggested in Figure 2.5. This elliptically polarized wave may be expressed in terms of two linearly polarized components, one in the $x$ direction and one in the $y$ direction. Thus, if the wave is traveling in the positive $z$ direction (out of the page), the electric field components in the $x$ and $y$ directions are

$$E_x = E_1 \sin(\omega t - \beta z) \quad (2.1)$$

$$E_y = E_2 \sin(\omega t - \beta z + \delta) \quad (2.2)$$

where $E_1 = \text{amplitude of wave linearly polarized in } x \text{ direction}$

$E_2 = \text{amplitude of wave linearly polarized in } y \text{ direction}$

$\delta = \text{time-phase angle by which } E_y \text{ leads } E_x$

Combining (2.1) and (2.2) gives the instantaneous total vector field $E$:

$$E = x E_1 \sin(\omega t - \beta z) + y E_2 \sin(\omega t - \beta z + \delta) \quad (2.3)$$

Figure 2.4 Linear, Elliptical and Circular polarization for wave propagation out of page
At \( z = 0 \), \( E_x = E_1 \sin(\omega t - \beta z) \) and \( E_y = E_2 \sin(\omega t - \beta z + \delta) \) \hspace{1cm} (2.4)

Expanding \( E_y \) yields

\[
E_y = E_2 (\sin \omega t \cos \delta + \cos \omega t \sin \delta)
\]

\hspace{1cm} (2.5)

From the relation for we have \( \sin \omega t = E_x / E_1 \) and \( \cos \omega t = \sqrt{1 - (E_x / E_1)^2} \)

Introducing these in (2.4) eliminates \( \text{rot} \), and on rearranging we obtain

\[
\frac{E_x^2}{E_1^2} - \frac{2E_xE_y \cos \delta}{E_1E_2} + \frac{E_y^2}{E_2^2} = \sin^2 \delta
\]

\hspace{1cm} (2.6)

Or

\[
aE_x^2 - bE_xE_y + cE_y^2 = 1
\]

\hspace{1cm} (2.7)

where

\[
a = \frac{1}{E_1^2 \sin^2 \delta} \hspace{1cm} b = \frac{2 \cos \delta}{E_1E_2 \sin^2 \delta}
\]

\[
c = a = \frac{1}{E_2^2 \sin^2 \delta}
\]

Equation (2.7) describes a (polarization) ellipse, as in Figure 2.4. The line segment \( OA \) is the semi major axis and the line segment \( OB \) is the semi minor axis. The tilt angle of the ellipse is \( \tau \). The axial ratio is

\[
AR = \frac{OA}{OB} \hspace{1cm} (1 \leq AR \leq \infty)
\]

(2.8)

For \( E_1 = 0 \), the wave is linearly polarized in the \( y \) direction. For \( E_2 = 0 \), the wave is linearly polarized in the \( x \) direction. If \( \delta = 0 \) and \( E_1 = E_2 \), the wave is also linearly polarized but in a plane at an angle of 45\(^\circ\) with respect to the \( x \) axis

\( (\tau = 45^\circ) \)

For \( E_1 = E_2 \) and \( \delta = \pm 90^\circ \) the wave is circularly polarized. When \( \delta = +90^\circ \) the wave is \textit{left-circularly polarized}, and when \( \delta = -90^\circ \) the wave is \textit{right-circularly polarized}. For the case \( \delta = +90^\circ \) and for \( z = 0 \) and \( t = 0 \) we have from (2.1) and (2.2) that
$E = \hat{y}E_2$, as in Figure 2.6 (a). One-quarter cycle later ($\omega t = 90^\circ$) $E = \hat{x}E_1$ as in Figure 2.6(b). Thus, at a fixed position ($z = 0$) the electric field vector rotates clockwise (viewing the wave approaching). According to the IEEE definition, this corresponds to left-circular polarization. The opposite rotation direction ($\delta = -90^\circ$) corresponds to right-circular polarization.

Polarization ellipses, as a function of the ratio $E_1/E_2$ and phase angle $\delta$ (wave approaching), are shown in Figure 2.6. In special cases, the ellipses become straight lines (linear polarization) or circles (circular polarization).

If the wave is viewed receding (from negative $z$ axis in Figure 2.6), the electric vector appears to rotate in the opposite direction. Hence, clockwise rotation of $E$ with the wave approaching is the same as counterclockwise rotation with the wave receding. Thus, unless the wave direction is specified, there is a possibility of ambiguity as to whether the wave is left- or right-handed. This can be avoided by defining the polarization with the
aid of helical antennas. Thus, a right-handed axial-mode helical antenna radiates (or receives) right circular (IEEE) polarization. A right-handed helix, like a right-handed screw, is right-handed regardless of the position from which it is viewed. There is no possibility here of ambiguity.

\[ \omega t = 0 \]

\[ \omega t = 90^0 \]

Figure 2.6 Instantaneous orientation of electric field vector \( E \) at two instants of time for a left-circularly polarized wave which is approaching (out of page).

Following section describes applications of microstrip antennas and how to produce various types of polarizations using the microstrip antennas.

2.5 Loaded Microstrip Antenna

Modern communication systems, such as those for satellite links (GPS, vehicular-Jar, etc.), for mobile communication, and for emerging applications, such as Wireless local-area networks (WLANs), often require compact antennas at low cost [12-16]. Further, due to their lightness, microstrip antennas are well suited for airborne applications, such as synthetic aperture radar (SAR) systems and scatterometers. In addition to compactness, the antenna may be required to provide circular polarization as in satellite links. In some applications, operation at two or more discrete bands and an
It may not be possible to achieve these objectives from the basic microstrip antennas having regular shapes that have been discussed in earlier chapters. The range of applications of microstrip antennas and their performance can be improved considerably by suitably loading them. Some examples of loaded microstrip antennas are given in next chapters, where loadings of the basic shape was used to obtain circular polarization, frequency tuning, broadbanding, impedance matching, higher gain, and so on. It is the goal of this chapter to discuss loading in a general way to obtain characteristics such as size reduction, dual-frequency operation, polarization control, radiation pattern control, and frequency agility [18].

A survey of various papers published on the applications of microstrip antennas shows that shorts, stubs, and slot loadings have been used for the most part. Therefore, we consider only the effect of these loadings although loading can take various forms such as stub loading, slot or notch loading, short circuits or vias, parasitic coupling, substrate loading, superstrate cover, resistors, capacitors, and diodes.

A microstrip antenna can easily be made to resonate at many frequencies associated with various modes. The characteristic of each mode is different and is determined by the resonant frequency, radiation pattern, polarization, bandwidth and so on. For a given feed location, if the patch is now loaded with a short, the field and current distributions for various modes will be disturbed and therefore their characteristics will change. This change will depend on the amount of load and the mode under consideration. For example, a short placed at the nodal line of a mode will hardly affect its characteristics, whereas another mode with electric field maximum at the short will experience the maximum change. If used properly, the loading effect can be used to obtain a desirable change in the antenna characteristics.
First, some applications of loaded microstrip antennas to realize polarization diversity, frequency agility, and radiation pattern control are discussed. Next, loading will be used to reduce the size of circularly polarized antennas, and dual-frequency operation antennas. Various types of compact antennas are listed. The planar inverted-F antenna, a compact antenna suggested, is described in detail.

2.6 Polarization Diversity Using Microstrip Antennas

Polarization diversity [25] of reception is important to counter the effect of fading in communications, especially in mobile communications. The polarization of a microstrip antenna can be selected by making a proper choice for the feed location. For example, the square microstrip antenna shown in Figure 2.7 [39] can be used to transmit or receive vertical polarization or horizontal polarization depending on which feed is used and which shorting posts are not used. The two polarizations are associated with the modes (1, 0) and (0, 1) excited by the respective feeds. Simultaneous operation in two polarizations at the same frequency is possible if both feeds are used. If the square patch is replaced by a rectangular patch antenna, then it is possible to have dual-frequency and dual-polarization operation. The square patch antenna of Figure 2.7 when loaded with posts as indicated can be used for generating desired circular polarization by a proper selection of feed. The posts located along the diagonal help in the excitation of both the modes with a single feed. Proper location of the posts ensures equal amplitude and a phase difference of 90° for the two modes.

![Figure 2.7 dual-feed antenna for horizontal and vertical][39]
It is also possible to obtain four different polarizations-horizontal, vertical, right-hand circular, and left-hand circular-with a single feed. This is shown in Figure 2.8 for a square patch antenna. The feed is located along the diagonal and the shorting pins are located along the center lines. The polarization of the signal is decided by the selection of the shorting pins, and can be explained on the basis of change in resonant frequency produced by the loading effect of the short.

![Diagram of a single-feed square patch antenna with four pairs of posts for obtaining four different polarizations.](image)

**Figure 2.8** Single-feed square patch antenna with four pairs of posts for obtaining four different polarizations.

The square patch, without shorting pins, supports both the (1, 0) and (0, 1) modes with x-oriented and y-oriented polarizations. The resonant frequency for both the modes is the same. Because the feed probe is located along the diagonal, both the modes are excited with equal amplitude and phase. By adding shorting pins along the center line \( x = a/2 \) (Figure 2.8), the resonant frequency of the (0, 1) or y-oriented mode can be raised without affecting the other mode. Therefore, the mode with desired polarization (x or y) can be selected by shifting the resonant frequency of the undesired mode far above that of
the desired mode. The required large frequency shift is obtained by placing the shorting pins at or near the edges of the patch as shown in the figure.

Circular polarization is obtained by exciting both the x- and y-polarized modes with equal amplitude and 90° phase difference. This can be accomplished by raising the resonant frequency of one mode slightly above the other and operating at a frequency midway between the two frequencies. Then the input impedance of one mode is inductive and the other mode is capacitive. By adjusting the difference between the resonant frequencies, both the modes can be excited with equal amplitude and 90° phase difference.

The pair of posts inside the patch raises the resonant frequency of one of the modes only. Figure 2.9 [39] shows the measured axial ratio of a typical patch antenna as the separation between a pair of shorting posts is varied.

![Figure 2.9](image-url) Measured axial ratio as a function of post spacing [39] s/a for a square patch antenna on a 1.6-mm Teflon fiberglass substrate.
When \( s/a = 0 \), the posts are at the center and they don’t affect either mode. Because both the modes are excited with equal amplitude by the diagonal feed, the polarization is linear oriented along the feed diagonal. When \( s/a = 0.09 \), the resonant frequencies of the two modes are offset enough to obtain a phase difference of about 90° and the antenna is circularly polarized. As the posts are moved further apart, that is, as they approach the respective edge, the resonant frequency of the vertically polarized mode is further increased and the polarization of the antenna becomes horizontal linear with large axial ratio as shown in the figure. The input impedance of the antenna changes with the movement of the posts, but the VSWR remains good for all senses of polarization. Although one post is sufficient to raise the resonant frequency of any mode, two symmetrically located posts are used so that the antenna structure is symmetric with respect to loading, and the cross-polarized component in the radiation pattern due to loading is minimized. The shorting posts can be realized by microwave switching diodes for precise control of the resonant frequency and polarization switching applications.

2.7 Single-Feed Circularly Polarized Microstrip Antennas

For a circularly polarized radiation, a patch must support the orthogonal fields of the equal amplitude but in phase quadrature. This requirement is accomplished by slightly perturbing a patch at appropriate locations with respect to the feed. However, these perturbation configurations have very narrow axial ratio (AR) and voltage standing wave ratio (VSWR) bandwidths. However, a successful attempt has been made in the present work to improve the axial ratio within the impedance bandwidth for VSWR \( \leq 2 \). In this chapter, we discuss the analysis and optimized designs of single-feed modified square microstrip antennas for enhancement of axial ratio and VSWR bandwidths. The effect on the input impedance and axial ratio for different values of width by keeping length constant of nearly square microstrip antenna is described. Improvement in the axial ratio and VSWR bandwidths on thick dielectric substrate with different values of probe diameter is investigated in the present work. Our analysis is along the lines of the discussions in [40] on the singly-fed microstrip antennas for circularly polarized radiation.
2.7.1 Diagonally-Fed Nearly Square Microstrip Antenna

In this section, a method for designing a circularly polarized microstrip antenna with singly-fed is presented. Depending upon the perturbation the feed location is either on the x or y-axis, or the feed is placed on the diagonal axis of a patch. Note that the feed is always located diagonal to perturbation segments that are appropriately selected to produce two orthogonally degenerate modes in the patch for circularly polarized radiation. Nearly square microstrip antenna is one of the simplest configurations to generate circularly polarized radiation with feed location along the diagonal as shown in Figure 2.10 (a).

![Figure 2.10](image)

Figure 2.10 (a) Diagonally fed nearly square microstrip antenna, and (b) equivalent diagram of nearly square microstrip antenna in the presence of fringing fields

The patch length \((L = L+1)\), and width \((L_2 = L)\) determines the orthogonal resonant frequencies, and are critical parameters in design because of inherent narrow bandwidth of the patch. In practice, the fields are not confined to the patch. A fraction of the fields lie outside the physical dimension of the patch. This is called the fringing field. Considering the effect of the fringing fields, the effective dimensions of diagonally fed nearly square microstrip antenna are different from the physical dimensions. The extension of the dimension of nearly square microstrip antenna due to the fringing field is shown dotted lines in Figure 2.10 (b). For the fundamental \(TM_{10}\) mode, the \(L_i\) should be slightly less than \(\lambda/2\), where \(\lambda\) is the wavelength in the dielectric medium. The
fundamental TM10 mode implies that the field varies one $\lambda/2$ cycle along the length, and no variation along the width of the patch. A nearly square microstrip antenna operating at TM10 mode can be visualized as a transmission line, because the field is uniform along the width and varies sinusoidally along the length. Their equivalent capacitance and the radiation resistance model the fringing fields along the edges and the radiation from the slots. To account for the fringing fields, instead of adding the capacitor at the edges, the dimensions around the periphery of the patch can be extended outwards. This can be explained in terms of the parallel rectangular plates of dimensions $L_1$ and $L_2$, which are separated by a dielectric substrate of thickness $h$. If the fringing fields along the periphery are ignored, then the capacitance of the two parallel plates will be

$$C = \varepsilon_0 \varepsilon_r \frac{L_1 L_2}{h} \tag{2.9}$$

However, due to the fringing capacitance, the effective capacitance $C_e$ of the two parallel plates increases. One of the ways to account for the fringing capacitance is to extend the dimensions of the plate outward, and the value of $C_e$ is calculated from

$$C_e = \varepsilon_0 \varepsilon_r \frac{L_{1e} L_{2e}}{h} \tag{2.10}$$

where, $L_{1e}$ and $L_{2e}$ are the effective dimensions and are equal to:

$$L_{1e} = L_1 + 2\Delta L_1 \tag{2.11}$$

And

$$L_{2e} = L_2 + 2\Delta L_2 \tag{2.12}$$

The $\Delta L_1$ and $\Delta L_2$ are the extensions along the $L_1$ and $L_2$, respectively. The edge extensions are calculated using the empirical formula [41]:

$$\Delta L_1 = 0.412h \frac{\varepsilon_{re1} + 0.300 \, u + 0.264}{\varepsilon_{re1} - 0.258 \, u + 0.813} \tag{2.13}$$

where $u = L_2 / h$

$$\varepsilon_{re1} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + \frac{10}{u}\right)^{-\alpha b} \tag{2.14}$$

where
\[ a = 1 + \frac{1}{49} \ln \left\{ \frac{u^4 + (u/52)^2}{u^4 + 0.432} \right\} + \frac{1}{18.7} \ln \left\{ 1 + \left( \frac{u}{18.1} \right)^3 \right\} \]

\[ b = 0.564 \left( \frac{\varepsilon_r - 0.9}{\varepsilon_r + 0.3} \right)^{0.053} \]

and

\[ \Delta L_2 = 0.412 h \frac{\varepsilon_{re2} + 0.300 \nu + 0.264}{\varepsilon_{re2} - 0.258 \nu + 0.813} \quad (2.15) \]

where \( \nu = L1/h \)

\[ \varepsilon_{re2} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + \frac{10}{\nu} \right)^{-ab} \quad (2.16) \]

where

\[ a = 1 + \frac{1}{49} \ln \left\{ \frac{v^4 + (v/52)^2}{v^4 + 0.432} \right\} + \frac{1}{18.7} \ln \left\{ 1 + \left( \frac{v}{18.1} \right)^3 \right\} \]

\[ b = 0.564 \left( \frac{\varepsilon_r - 0.9}{\varepsilon_r + 0.3} \right)^{0.053} \]

The calculated values of \( \Delta L_1 \) or \( \Delta L_2 \) fit the measured resonant frequency within \( \pm 1.6\% \) [42]. It is expected that the use of (2.13) and (2.15) should result in a better fit. The resonant frequency \( f_{10} \) or the effective length \( L_{1e} \) for the dominant TM_{10} mode of a nearly square microstrip antenna is related by the following simple formula [43]

\[ f_{10} = \frac{c}{2L_{1e}\sqrt{\varepsilon_{re1}}} \quad \text{or} \quad L_{1e} = \frac{c}{2f_{10}\sqrt{\varepsilon_{re1}}} \quad (2.17) \]

where \( c \) is the velocity of the light.

Similarly the resonant frequency \( f_{01} \) or the effective length \( L_{2e} \) for the orthogonal TM_{01} mode of a nearly square microstrip antenna is related as
\[ f_{01} = \frac{c}{2L_{2e} \sqrt{\varepsilon_{r2}}} \quad \text{or} \quad L_{2e} = \frac{c}{2f_{01} \sqrt{\varepsilon_{r2}}} \] (2.18)

Using (2.7) and (2.18), the lengths of nearly square microstrip antenna are calculated. From Figure 2.10 (a), the perturbation area of a nearly square microstrip antenna, \( \Delta S = 1L \), and square microstrip antenna area with out perturbation \( S = L^2 \). The empirical design equation for diagonally fed nearly square microstrip antenna is given as [44]:

\[ \left| \frac{\Delta S}{S} \right| = \frac{1}{Q_0} \quad (2.19) \]

where \( Q_0 \) is unloaded quality factor of the microstrip antenna. The design of the patch with perturbation requires \( Q_0 \), which depends on the dimensions, substrate thickness \( h \), and the substrate dielectric constant \( \varepsilon_r \). For the given substrate parameters, the substrate depended quality factor is approximately calculated as [44]:

\[ Q_0(\varepsilon_r) = \frac{Q_0 \sqrt{\varepsilon_r}}{\sqrt{2.55}} \quad (2.20) \]

For better accuracy, \( Q_0 \) should be selected to ensure better radiation efficiency of the microstrip antenna. The amount of perturbation required for circularly polarized radiation is calculated from (2.13). The location of the feed on the diagonal axis is selected for 50\( \Omega \) impedance match or the quarter wave transformer is used for the matching purpose. The sense of circularly polarized radiation changes by switching the feed to the orthogonal diagonal axis. The ratio of the two orthogonal dimensions \( L_1 / L_2 \) of nearly square microstrip antenna should be generally in the range of 1.01 to 1.10 depending upon the substrate parameters. When the patch is fed along the diagonal, then the two resonance modes corresponding to lengths \( L_1 \) and \( L_2 \) are spatially orthogonal. The circularly polarized radiation is obtained at a frequency between the resonance frequencies of these two modes, where the two orthogonal modes have equal magnitudes and are in phase quadrature. In this section, the effect on the input impedance for different values of \( L_2 \) by keeping \( L_1 \) constant has been studied. Nearly square microstrip antenna of length, \( L_1 = 30 \) mm, relative dielectric constant, \( \varepsilon_r = 2.55 \), the substrate height,
\( h = 1.59 \text{ mm} \), and loss tangent, \( \tan \delta = 0.001 \), which is fed along the diagonal at feed location \( F \), the input impedance, voltage standing wave ratio and axial ratio plots for three values of \( L_2 = 0.967 \, L_1 \), \( 0.973 \, L_1 \), and \( 0.983 \, L_1 \) are shown in Figure 2.11 (a) - (c).

In singly-fed microstrip antennas, the kink in the impedance plot represents that the two orthogonal modes are excited with equal amplitude and \( 90^\circ \) phase difference. At the kink frequency minimum axial ratio is obtained. Instead of a kink, a small loop or the absence of the loop in the impedance plot yields poor axial ratio at center frequency. The loop in the impedance plot implies that the separation between the two orthogonal modes is large, and hence it is to be reduced to obtain better axial ratio. If there is only a slight bend in the impedance plot without any kink or loop, then the separation between the two modes is to be increased. For the diagonally fed nearly square microstrip antenna, we observed that as the length \( L_2 \) increases, the loop in the impedance plot becomes a kink, and then it disappears as \( L_2 \) increases further. The criticality of the dimension is to be noted. Even though, larger VSWR bandwidth of 4 % is obtained for \( L_2 = 0.967 \, L_1 \) mm due to the loop in the impedance plot, the minimum axial ratio is 4 dB at resonant \( f_0 = 3.01 \text{ GHz} \). For \( L_2 = 0.973 \, L_1 \), the loop becomes kink, hence the VSWR bandwidth is reduced to 3 %, but the minimum axial ratio is improved to 0.5 dB. In this case, the bandwidth for axial ratio \( \leq 3 \text{ dB} \) is 0.8 %. The value of \( L_1/L_2 = 1.027 \) implies separation of 2.7 % between the two resonant lengths, which is slightly greater than the VSWR bandwidth (1.8 %) of the square patch of length \( L \). A lower value of axial ratio could be obtained by fine-tuning the length \( L_3 \) For the feed location shown in Fig. 2.15, left hand circularly polarized radiation is obtained, and if the feed is shifted to the other diagonal, then right hand circularly polarized is obtained.

The axial ratio and VSWR bandwidths are improved by increasing the substrate thickness. When the substrate height is doubled, larger ratio of \( L_1/L_2 = 1.06 \) is taken because bandwidth increases with increase in the substrate thickness. The input impedance and axial ratio variations with frequency are shown in Figure 2.12. For the SMA connector (probe diameter \( d = 1.2 \text{ mm} \)), the kink in the input impedance plot is formed in the inductive region of the Smith chart, which is due to the larger probe...
inductance of increased substrate thickness. The probe inductance is reduced by increasing its diameter to 2.0 mm, which shifts the kink towards the center of the Smith chart as shown in Figure 2.12, thereby improving the VSWR bandwidth and also the axial ratio at the kink. The change in the probe diameter changes the resonance frequency slightly. For the thicker probe, the impedance bandwidth for VSWR ≤ 2 is 7.4 % and the axial ratio bandwidth for AR ≤ 3 dB is 1.7 %.

Figure 2.11  (a) Input impedance (b) VSWR, and (c) axial ratio variation of diagonal fed nearly square microstrip antenna.
The axial ratio bandwidth is improved further when thicker substrate with low dielectric constant is used. In this case, a thicker probe diameter of 4.0 mm is taken to reduce the probe inductance, so that the kink in the impedance plot is within VSWR = 2 circle. VSWR and axial ratio variation with low dielectric constant on thick substrate is shown in Figure 2.13. The bandwidths for VSWR \( \leq 2 \) and axial ratio \( \leq 3 \) dB are 14 % and 2.8 %, respectively. Notice that the bandwidth of the microstrip antenna is limited by its axial ratio and not by its voltage standing wave ratio.

Figure 2.12  (a) Input impedance, and (b) axial ratio variation of singly-fed nearly square microstrip antenna for the substrate height \( h=3.18 \) mm.
2.8 Square Microstrip Antenna with Modified Edge

Instead of using the nearly square microstrip antenna, the edges of the square microstrip antenna are modified by adding stubs or by cutting slots as shown in Figure 2.14. By adding only one stub or by cutting one notch, circularly polarized radiation can also be obtained, but then the configuration is not symmetrical. However as long as the total effective areas of these perturbations are of the same order, the performance of one edge modified is similar to that of two edges modified square microstrip antenna. The approximate value of the total perturbation area is $\Delta S = 2l^2$ for square microstrip antenna with symmetric stubs or notches. The amount of perturbation required for circularly polarized radiation is calculated from (2.14). $\Delta S$ represent the total sum of perturbation segments. The amount of perturbation of individual stub or notch is calculated as $\Delta S_1 = \Delta S_2 = \Delta S/2$.

A square microstrip antenna of length $L = 30$ mm and two square stub of length $l = 0.083 \ L$ with substrate parameters dielectric constant, $\varepsilon_r = 2.55$, substrate thickness or substrate height, $h = 1.59$ mm, and loss tangent, $\tan \delta = 0.001$, yield right hand circularly polarized radiation.
polarized radiation when fed at F. Similarly with two square notch of length \( l = 0.083 \ L \) on the same substrate parameters, yield left hand circularly polarized radiation. For both the configurations the bandwidth for AR \( \leq 3 \) dB and VSWR \( \leq 2 \) are 0.7 % and 3.2 %. In these cases, the values of axial ratio and VSWR bandwidths are similar to that of the nearly square microstrip antenna. We observed that the perturbation area of the stub or the notch is very critical to yield lower axial ratio value, just as in the case of nearly square microstrip antenna, where \( L_1 / L_2 \) ratio is very critical to yield circularly polarized radiation. The advantage of these configurations is that trimming the stub or notch can easily do the fine-tuning.

![Figure 2.14 Diagonal fed square microstrip antenna with (a) two stubs, and (b) two notches along its opposite edges.](image)

2.9 **Square Microstrip Antenna with Modified Corner**

Square microstrip antenna with modifying corners is also used to produce two orthogonally degenerate modes for circularly polarized radiation. Small isosceles right angle triangular patches or small square patches are removed from the diagonally opposite corners of the square patch shown in Figure 2.15. Chopping of two diagonally opposite corners makes the resonance frequency of the mode along this diagonal to be higher than that for the mode along the un-chopped diagonal. The patch is fed along the central axis so that the orthogonal modes are generated. Instead of chopping the corners, small square patches could be added at the corners as shown in Figure 2.15 to obtain
circularly polarized radiation. For these configurations the perturbation area is calculated using the following relation:

\[
\frac{\Delta S}{S} = \frac{1}{2Q_0}
\]  

(2.21)

In all these configurations, only one corner could be modified to yield circularly polarized radiation. Since two corners modified configurations are symmetrical, the details are given for only these cases. The perturbation area of corners chopped square microstrip antenna is \(\Delta S = l^2\). When isosceles right angle triangular patches of side length \(l = 0.12L\) are removed from the two corners, left hand circularly polarized radiation is obtained for the feed at F. The axial ratio and the VSWR bandwidths are 1 % and 3 % for \(\text{AR} \leq 3\ \text{dB}\) and \(\text{VSWR} \leq 2\), respectively.

![Figure 2.15](image)

Figure 2.15  square microstrip antennas with modified diagonally opposite corners (a) small isosceles right angle triangles removed (b) small squares removed, and (c) small squares added.

The perturbation area of square microstrip antenna with small squares added or removed from the corner is \(\Delta S = 2l^2\). When a small square of length \(l = 0.083L\) is removed from the two corners of square the perturbation area, left hand circularly polarized radiation is obtained for the feed at F. When a small squares of length \(l = 0.067L\) is added at the two corners of the square the perturbation area as shown in Figure 2.20 (c), right hand circularly polarized radiation is obtained for the feed at F. In the letter case, the resonance frequency and axial ratio bandwidth are slightly smaller than that of the corner chopped cases, because of its larger patch area.
2.11 Square Microstrip Antenna with a Diagonal Slot

A square microstrip antenna with a rectangular diagonal slot and the feed along its central axis is shown in Figure 2.16. The difference in the resonance frequencies of the orthogonal modes is caused by the rectangular slot, which makes the path lengths of the two diagonals unequal. In this case, the ratio of length and width of the slot governs the circularly polarized radiation characteristics of the antenna for given substrate specifications. The perturbation area of rectangular slot is $\Delta S = l \times w$. Square microstrip antenna of length $L = 30$ mm with rectangular slot of length $l = 0.25 \ L$ and width $w = 0.07 \ L$, fed at $F$, left hand circularly polarized radiation is obtained. The axial ratio and VSWR bandwidths are 1 % and 3 %, respectively. These theoretical results follow the same trend as observed in the measured results [45].

![Figure 2.16 Square MSA with a diagonal slot](image)

In the entire singly-fed modified square microstrip antennas, the ratio of the two orthogonal modes is very critical to yield circularly polarized radiation with minimum axial ratio at the resonance frequency. Due to the design limitations and fabrication error (tolerance in the patch dimensions and the substrate parameters), it is possible that the results may not be optimum and fine-tuning of the dimensions is required. Whether to increase or decrease the dimensions could be determined by looking at the input impedance plot. By fine-tuning the dimensions, lower axial ratio can be obtained with larger axial ratio bandwidth. In this regard, stub configurations are more suitable as fine tuning can be easily done.
2.12 Matlab based free software for antenna simulation

Matlab based free software for antenna design and simulation is developed by Sergey N Makarow. It is not fully developed like other commercial software such as ADS or IE3D. It is not much more user friendly and while using it one has to write many codes to analyze the required antenna. I have used this software. I have written some codes of my own, designed a microstrip patch antenna for 2.5 GHz. and simulated. The codes are not included in this report. The simulated results are given in following figure 2.17 to 2.19.

![Figure 2.17](image1.png)  
**microstrip patch antenna for 2.5 GHz**

![Figure 2.18](image2.png)  
**S\_{11} Versus Frequency plot**

![Figure 2.19](image3.png)  
**S\_{11} Versus Frequency plot**