DESIGN OF LINEARLY POLARIZED RECTANGULAR MICROSTRIP PATCH ANTENNA USING IE3D/PSO

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CERTIFICATE

This is to certify that the thesis entitled, "**DESIGN OF LINEARLY POLARIZED RECTANGULAR MICROSTRIP PATCH ANTENNA USING IE3D/PSO**" submitted by **Mr. RAHUL RANA** and **Mr. C VISHNU VARDHANA REDDY** in partial fulfillment of the requirements for the award of Bachelor of Technology Degree in **ELECTRONICS AND COMMUNICATION** at the National Institute of Technology, Rourkela (Deemed University) is an authentic work carried out by him under my supervision and guidance.To the best of my/our knowledge, the matter embodied in the thesis has not been submitted to any other University/ Institute for the award of any degree or diploma.

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C Vishnu Vardhana Reddy
Rahul Rana

ABSTRACT

In this project, a novel particle swarm optimization method based on IE3D is used to design an Inset Feed Linearly Polarized Rectangular Microstrip Patch Antenna. The aim of the thesis is to Design and fabricate an inset fed rectangular Microstrip Antenna and study the effect of antenna dimensions Length (L) , Width (W) and substrate parameters relative Dielectric constant (ε_r) , substrate thickness on Radiation parameters of Band width. Low dielectric constant substrates are generally preferred for maximum radiation. The conducting patch can take any shape but rectangular and circular configurations are the most commonly used configuration. Other configurations are complex to analyze and require heavy numerical computations.

The length of the antenna is nearly half wavelength in the dielectric; it is a very critical parameter, which governs the resonant frequency of the antenna. In view of design, selection of the patch width and length are the major parameters along with the feed line depth. Desired Patch antenna design is initially simulated by using IE3D simulator. And Patch antenna is realized as per design requirements.

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CHAPTER 1

INTRODUCTION

Communication between humans was first by sound through voice. With the desire for slightly more distance communication came, devices such as drums, then, visual methods such as signal flags and smoke signals were used. These optical communication devices, of course, utilized the light portion of the electromagnetic spectrum. It has been only very recent in human history that the electromagnetic spectrum, outside the visible region, has been employed for communication, through the use of radio. One of humankind's greatest natural resources is the electromagnetic spectrum and the antenna has been instrumental in harnessing this resource.

1.1 Aim and Objectives

Microstrip patch antenna used to send onboard parameters of article to the ground while under operating conditions. The aim of the thesis is to design and fabricate an inset-fed rectangular Microstrip Patch Antenna and study the effect of antenna dimensions Length (L) , Width (W) and substrate parameters relative Dielectric constant (ε_r), substrate thickness (t) on the Radiation parameters of Bandwidth and Beam-width.

1.2 Overview of Microstrip Antennae

A microstrip antenna consists of conducting patch on a ground plane separated by dielectric substrate. This concept was undeveloped until the revolution in electronic circuit miniaturization and large-scale integration in 1970. After that many authors have described the radiation from the ground plane by a dielectric substrate for different configurations. The early work of Munson on micro strip antennas for use as a low profile flush mounted antennas on rockets and missiles showed that this was a practical concept for use in many antenna system problems. Various mathematical models were developed for this antenna and its applications were extended to many other fields. The number of papers, articles published in the journals for the last ten years, on these antennas shows the importance gained by them. The micro strip antennas are the present day antenna designer's choice.

Low dielectric constant substrates are generally preferred for maximum radiation. The conducting patch can take any shape but rectangular and circular configurations are the most commonly used configuration. Other configurations are complex to analyze and require heavy numerical computations. A microstrip antenna is characterized by its Length, Width, Input impedance, and Gain and radiation patterns. Various parameters of the microstrip antenna and its design considerations were discussed in the subsequent chapters. The length of the antenna is nearly half wavelength in the dielectric; it is a very critical parameter, which governs the resonant frequency of the antenna. There are no hard and fast rules to find the width of the patch.

1.3 Waves on Microstrip

The mechanisms of transmission and radiation in a microstrip can be understood by considering a point current source (Hertz dipole) located on top of the grounded dielectric substrate (fig. 1.1) This source radiates electromagnetic waves. Depending on the direction toward which waves are transmitted, they fall within three distinct categories, each of which exhibits different behaviors.

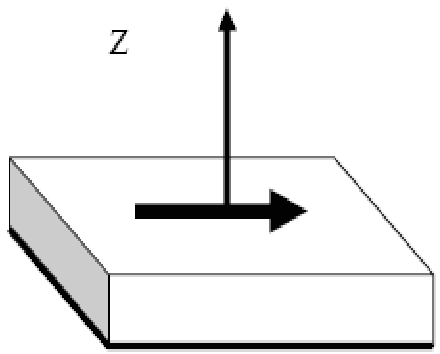


Figure 1.1 Hertz dipole on a microstrip substrate

1.3.1 Surface Waves

The waves transmitted slightly downward, having elevation angles θ between π/2and π - arcsin ($1/\sqrt{\varepsilon_r}$), meet the ground plane, which reflects them, and then meet the dielectric-to-air boundary, which also reflects them (total reflection condition). The magnitude of the field amplitudes builds up for some particular incidence angles that leads to the excitation of a discrete set of surface wave modes; which are similar to the modes in metallic waveguide.

The fields remain mostly trapped within the dielectric, decaying exponentially above the interface (fig1.2). The vector α, pointing upward, indicates the direction of largest attenuation. The wave propagates horizontally along β, with little absorption in good quality dielectric. With two directions of α and β orthogonal to each other, the wave is a non-uniform plane wave. Surface waves spread out in cylindrical fashion around the excitation point, with field amplitudes decreasing with distance (r), say1/*r*, more slowly than space waves. The same guiding mechanism provides propagation within optical fibers.

Surface waves take up some part of the signal's energy, which does not reach the intended user. The signal's amplitude is thus reduced, contributing to an apparent attenuation or a decrease in antenna efficiency. Additionally, surface waves also introduce spurious coupling between different circuit or antenna elements. This effect severely degrades the performance of microstrip filters because the parasitic interaction reduces the isolation in the stop bands.

In large periodic phased arrays, the effect of surface wave coupling becomes particularly obnoxious, and the array can neither transmit nor receive when it is pointed at some particular directions (blind spots). This is due to a resonance phenomenon, when the surface waves excite in synchronism the Floquet modes of the periodic structure. Surface waves reaching the outer boundaries of an open microstrip structure are reflected and diffracted by the edges. The diffracted waves provide an additional contribution to radiation, degrading the antenna pattern by raising the side lobe and the cross polarization levels. Surface wave effects are mostly negative, for circuits and for antennas, so their excitation should be suppressed if possible.

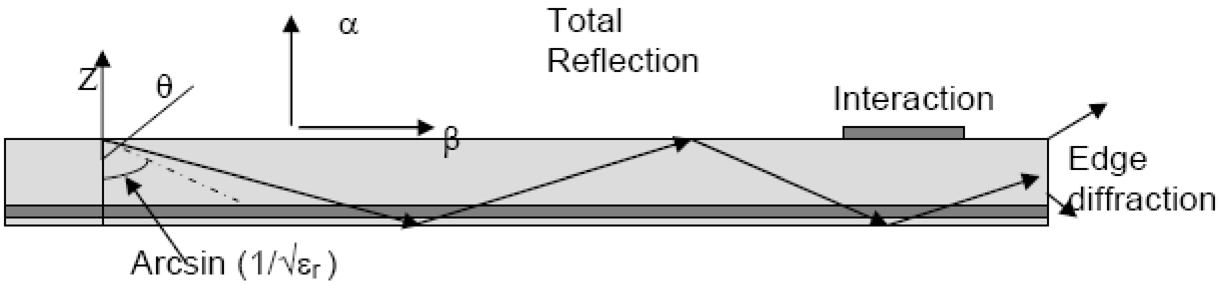


Figure 1.2.Surface waves

1.3.2 **Leaky Waves**

Waves directed more sharply downward, with θ angles between π - arcsin ($1/\sqrt{\varepsilon_r}$) and π, are also reflected by the ground plane but only partially by the dielectric-to-air boundary. They progressively leak from the substrate into the air (Fig 1.3), hence their name laky waves, and eventually contribute to radiation. The leaky waves are also non-uniform plane waves for which the attenuation direction α points downward, which may appear to be rather odd; the amplitude of the waves increases as one moves away from the dielectric surface. This apparent paradox is easily understood by looking at the figure 1.3; actually, the field amplitude increases as one move away from the substrate because the wave radiates from a point where the signal amplitude is larger. Since the structure is finite, this apparent divergent behavior can only exist locally, and the wave vanishes abruptly as one crosses the trajectory of the first ray in the figure.

In more complex structures made with several layers of different dielectrics, leaky waves can be used to increase the apparent antenna size and thus provide a larger gain. This occurs for favorable stacking arrangements and at a particular frequency. Conversely, leaky waves are not excited in some other multilayer structures.

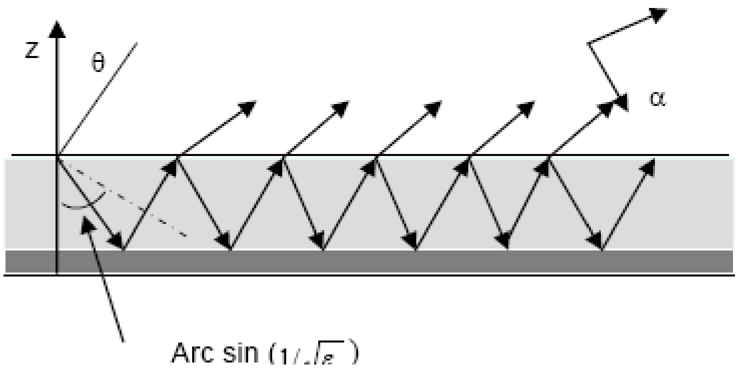


Figure 1.3 Leaky waves

1.3.3 **Guided Waves**

When realizing printed circuits, one locally adds a metal layer on top of the substrate, which modifies the geometry, introducing an additional reflecting boundary. Waves directed into the dielectric located under the upper conductor bounce back and forth on the metal boundaries, which form a parallel plate waveguide. The waves in the metallic guide can only exist for some Particular values of the angle of incidence, forming a discrete set of waveguide modes. The guided waves provide the normal operation of all transmission lines and circuits, in which the electromagnetic fields are mostly concentrated in the volume below the upper conductor. On the other hand, this buildup of electromagnetic energy is not favorable for patch antennas, which behave like resonators with a limited frequency bandwidth.

1.4 Antenna Characteristics

An antenna is a device that is made to efficiently radiate and receive radiated electromagnetic waves. There are several important antenna characteristics that should be considered when choosing an antenna for your application as follows:

- Antenna radiation patterns
- Power Gain
- Directivity
- Polarization

1.5 Organization of the Thesis

An introduction to microstrip antennas was given in Chapter II. Apart from the advantages and disadvantages, the various feeding techniques and models of analysis were listed.

Chapter III deals with the Radiation Parameters and the choice of substrate. The theory of radiation, various parameters and design aspects were discussed. All possible substrates for the design of microstrip antenna with their dielectric constant and permittivity are given.

Chapter IV provides the design and development of microstrip antenna. It provides information about IE3d Software for simulation of Microstrip Antennas, which will be used for cross verification of results for designed antennas. Chapter V gives the Conclusion to this project and suggests the future scope of work.

Chapter 2

MICROSTRIP PATCH ANTENNA

Introduction

Advantages and Disadvantages

Feed Techniques

Method of Analysis

CHAPTER 2

MICROSTRIP PATCH ANTENNA

Microstrip antennas are attractive due to their light weight, conformability and low cost. These antennas can be integrated with printed strip-line feed networks and active devices. This is a relatively new area of antenna engineering. The radiation properties of micro strip structures have been known since the mid 1950's.

The application of this type of antennas started in early 1970's when conformal antennas were required for missiles. Rectangular and circular micro strip resonant patches have been used extensively in a variety of array configurations. A major contributing factor for recent advances of microstrip antennas is the current revolution in electronic circuit miniaturization brought about by developments in large scale integration. As conventional antennas are often bulky and costly part of an electronic system, micro strip antennas based on photolithographic technology are seen as an engineering breakthrough.

2.1 Introduction

In its most fundamental form, a Microstrip Patch antenna consists of a radiating patch on one side of a dielectric substrate which has a ground plane on the other side as shown in Figure 2.1. The patch is generally made of conducting material such as copper or gold and can take any possible shape. The radiating patch and the feed lines are usually photo etched on the dielectric substrate.

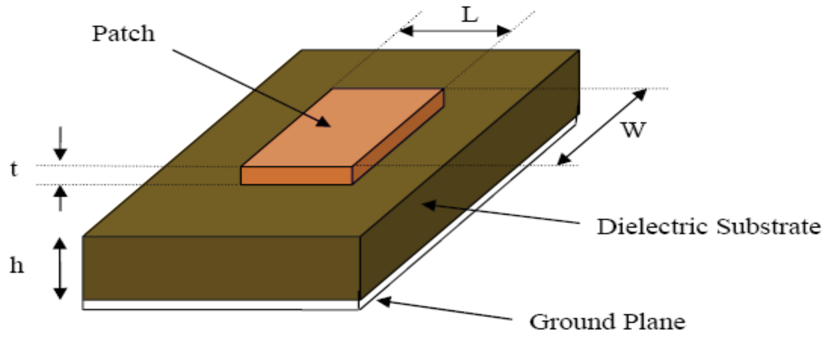


Figure 2.1 Structure of a Microstrip Patch Antenna

In order to simplify analysis and performance prediction, the patch is generally square, rectangular, circular, triangular, and elliptical or some other common shape as shown in Figure 2.2. For a rectangular patch, the length L of the patch is usually $0.3333\lambda_o < L < 0.5\ \lambda_o$, where λ_o is the free-space wavelength. The patch is selected to be very thin such that $t << \lambda_o$ (where *t* is the patch thickness). The height *h* of the dielectric substrate is usually $0.003\ \lambda_o \leq h \leq 0.05\ \lambda_o$. The dielectric constant of the substrate (ε_r) is typically in the range $2.2 \leq \varepsilon_r \leq 12$.

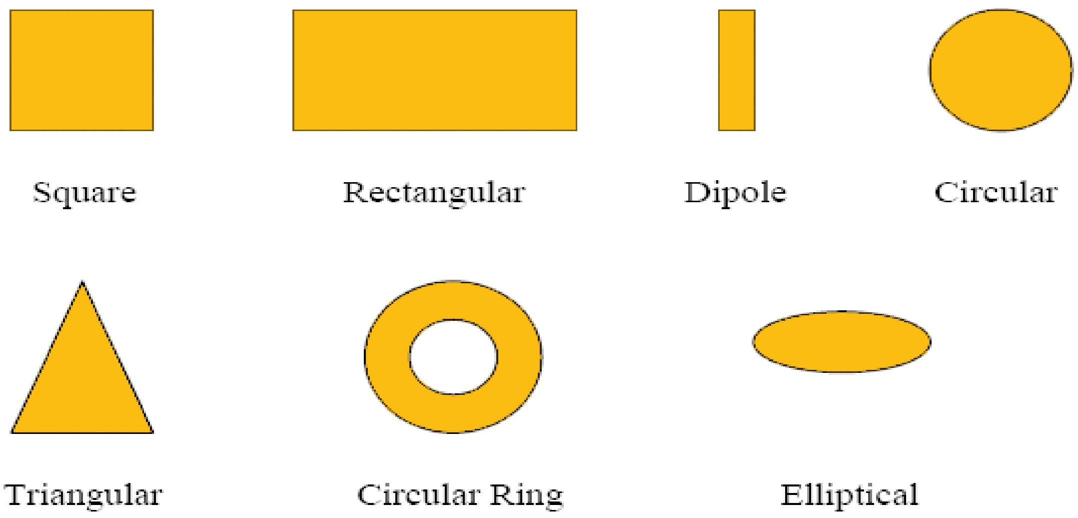


Figure 2.2 Common shapes of microstrip patch elements

Microstrip patch antennas radiate primarily because of the fringing fields between the patch edge and the ground plane. For good antenna performance, a thick dielectric substrate having a low dielectric constant is desirable since this provides better efficiency, larger bandwidth and better radiation [5]. However, such a configuration leads to a larger antenna size. In order to design a compact Microstrip patch antenna, substrates with higher dielectric constants must be used which are less efficient and result in narrower bandwidth. Hence a trade-off must be realized between the antenna dimensions and antenna performance.

2.2 Advantages and Disadvantages

Microstrip patch antennas are increasing in popularity for use in wireless applications due to their low-profile structure. Therefore they are extremely compatible for embedded antennas in handheld wireless devices such as cellular phones, pagers etc... The telemetry and communication antennas on missiles need to be thin and conformal and are often in the form of Microstrip patch antennas. Another area where they have been used successfully is in Satellite communication. Some of their principal advantages discussed by Kumar and Ray [9] are given below:

- Light weight and low volume.
- Low profile planar configuration which can be easily made conformal to host surface.
- Low fabrication cost, hence can be manufactured in large quantities.
- Supports both, linear as well as circular polarization.
- Can be easily integrated with microwave integrated circuits (MICs).
- Capable of dual and triple frequency operations.
- Mechanically robust when mounted on rigid surfaces.

Microstrip patch antennas suffer from more drawbacks as compared to conventional antennas. Some of their major disadvantages discussed by [9] and Garg et al [10] are given below:

- Narrow bandwidth
- Low efficiency
- Low Gain
- Extraneous radiation from feeds and junctions
- Poor end fire radiator except tapered slot antennas
- Low power handling capacity.
- Surface wave excitation

Microstrip patch antennas have a very high antenna quality factor (Q). It represents the losses associated with the antenna where a large Q leads to narrow bandwidth and low efficiency. Q can be reduced by increasing the thickness of the dielectric substrate. But as the thickness increases, an increasing fraction of the total power delivered by the source goes into a surface wave. This surface wave contribution can be counted as an unwanted power loss since it is ultimately scattered at the dielectric bends and causes degradation of the antenna characteristics. Other problems such as lower gain and lower power handling capacity can be overcome by using an array configuration for the elements.

2.3 Feed Techniques

Microstrip patch antennas can be fed by a variety of methods. These methods can be classified into two categories- contacting and non-contacting. In the contacting method, the RF power is fed directly to the radiating patch using a connecting element such as a microstrip line. In the non-contacting scheme, electromagnetic field coupling is done to transfer power between the microstrip line and the radiating patch [5]. The four most popular feed techniques used are the microstrip line, coaxial probe (both contacting schemes), aperture coupling and proximity coupling (both non-contacting schemes).

2.3.1 Microstrip Line Feed

In this type of feed technique, a conducting strip is connected directly to the edge of the Microstrip patch as shown in Figure 2.3. The conducting strip is smaller in width as compared to the patch and this kind of feed arrangement has the advantage that the feed can be etched on the same substrate to provide a planar structure.

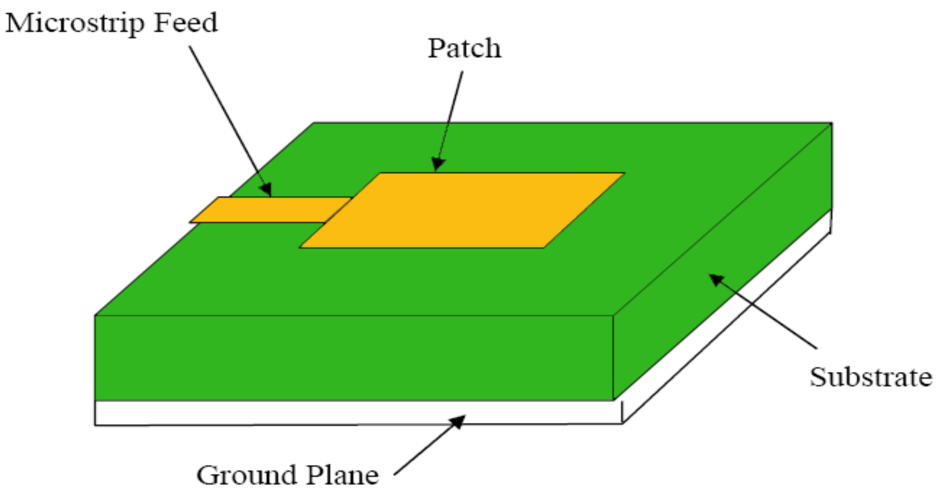


Figure 2.3 Microstrip Line Feed

The purpose of the inset cut in the patch is to match the impedance of the feed line to the patch without the need for any additional matching element. This is achieved by properly controlling the inset position. Hence this is an easy feeding scheme, since it provides ease of fabrication and simplicity in modeling as well as impedance matching. However as the thickness of the dielectric substrate being used, increases, surface waves and spurious feed radiation also increases, which hampers the bandwidth of the antenna [5]. The feed radiation also leads to undesired cross polarized radiation.

2.3.2 Coaxial Feed

The Coaxial feed or probe feed is a very common technique used for feeding Microstrip patch antennas. As seen from Figure 2.4, the inner conductor of the coaxial connector extends through the dielectric and is soldered to the radiating patch, while the outer conductor is connected to the ground plane.

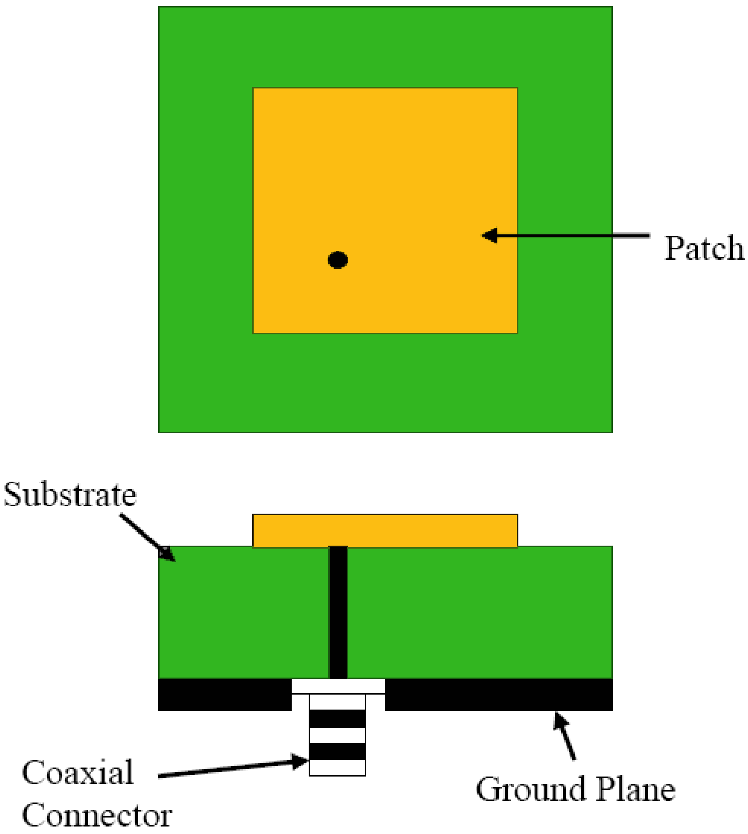


Figure 2.4 Probe fed Rectangular Microstrip Patch Antenna

The main advantage of this type of feeding scheme is that the feed can be placed at any desired location inside the patch in order to match with its input impedance. This feed method is easy to fabricate and has low spurious radiation. However, a major disadvantage is that it provides narrow bandwidth and is difficult to model since a hole has to be drilled in the substrate and the connector protrudes outside the ground plane, thus not making it completely planar for thick substrates ($h > 0.02\lambda_o$). Also, for thicker substrates, the increased probe length makes the input impedance more inductive, leading to matching problems [9]. It is seen above that for a thick dielectric substrate, which provides broad bandwidth, the microstrip line feed and the coaxial feed suffer from numerous disadvantages. The non-contacting feed techniques which have been discussed below, solve these issues.

2.3.3 Aperture Coupled Feed

In this type of feed technique, the radiating patch and the microstrip feed line are separated by the ground plane as shown in Figure 2.5. Coupling between the patch and the feed line is made through a slot or an aperture in the ground plane.

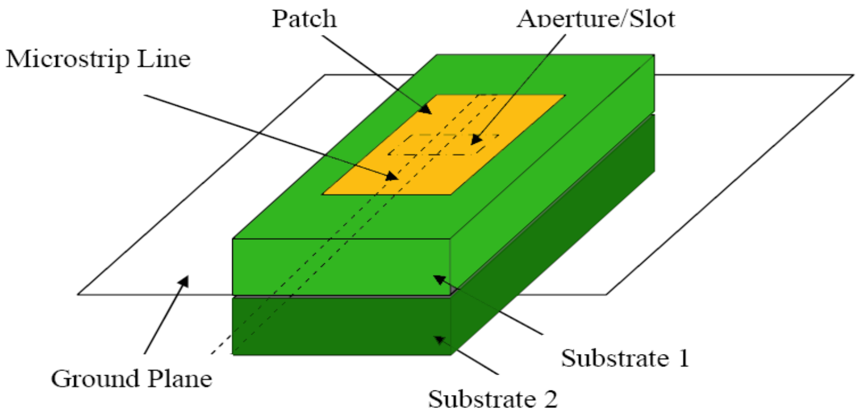


Figure 2.5 Aperture-coupled feed

The coupling aperture is usually centered under the patch, leading to lower cross-polarization due to symmetry of the configuration. The amount of coupling from the feed line to the patch is determined by the shape, size and location of the aperture. Since the ground plane separates the patch and the feed line, spurious radiation is minimized. Generally, a high dielectric material is used for bottom substrate and a thick, low dielectric constant material is used for the top substrate to optimize radiation from the patch [5]. The major disadvantage of this feed technique is that it is difficult to fabricate due to multiple layers, which also increases the antenna thickness. This feeding scheme also provides narrow bandwidth.

2.3.4 Proximity Coupled Feed

This type of feed technique is also called as the electromagnetic coupling scheme. As shown in Figure 2.6, two dielectric substrates are used such that the feed line is between the two substrates and the radiating patch is on top of the upper substrate. The main advantage of this feed technique is that it eliminates spurious feed radiation and provides very high bandwidth (as high as 13%) [5], due to overall increase in the thickness of the microstrip patch antenna. This scheme also provides choices between two different dielectric media, one for the patch and one for the feed line to optimize the individual performances.

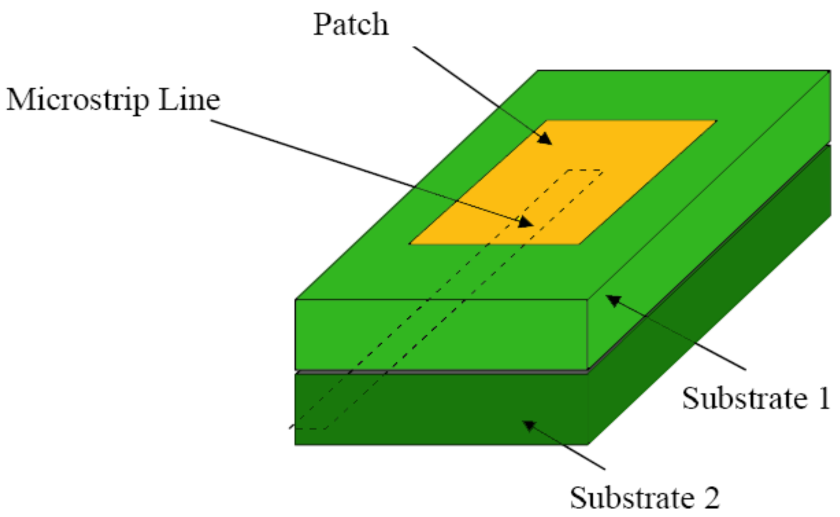


Figure 2.6 Proximity-coupled Feed

Matching can be achieved by controlling the length of the feed line and the width-to-line ratio of the patch. The major disadvantage of this feed scheme is that it is difficult to fabricate because of the two dielectric layers which need proper alignment. Also, there is an increase in the overall thickness of the antenna.

Table 2.1 below summarizes the characteristics of the different feed techniques.

Charact-eristics	Coaxial Probe Feed (Nonpl-anar)	Radiating Edge Coupled (Coplanar)	Nonradiating Edge Coupled (Coplanar)	Gap Cpoupled (Coplanar)	Inset Feed (Coplan-ar)	Proximity Coupled (Planar)	Aperture Coupled (Planar)	CPW Feed (Planar)
Spurious Feed Radiation	More	Less	Less	More	More	More	More	Less
Polarizat-ion Purity	Poor	Good	Poor	Poor	Poor	Poor	Excellent	Good
Fabricati-on Ease	Solder Reqd.	Easy	Easy	Easy	Easy	Alignment Reqd.	Alignment Reqd.	Alignment Reqd.
Reliability	Poor	Better	Better	Better	Better	Good	Good	Good
Impedence Matching	Easy	Poor	Easy	Easy	Easy	Easy	Easy	Easy
BW (at matching)	2-5%	9-12%	2-5%	2-5%	2-5%	13%(30)	21%(33)	3%(39,40)

2.4 Methods of Analysis

The preferred models for the analysis of Microstrip patch antennas are the transmission line model, cavity model, and full wave model [5] (which include primarily integral equations/Moment Method). The transmission line model is the simplest of all and it gives good physical insight but it is less accurate. The cavity model is more accurate and gives good physical insight but is complex in nature. The full wave models are extremely accurate, versatile and can treat single elements, finite and infinite arrays, stacked elements, arbitrary shaped elements and coupling. These give less insight as compared to the two models mentioned above and are far more complex in nature.

2.4.1 Transmission Line Model

This model represents the microstrip antenna by two slots of width *W* and height *h*, separated by a transmission line of length L. The microstrip is essentially a non-homogeneous line of two dielectrics, typically the substrate and air.

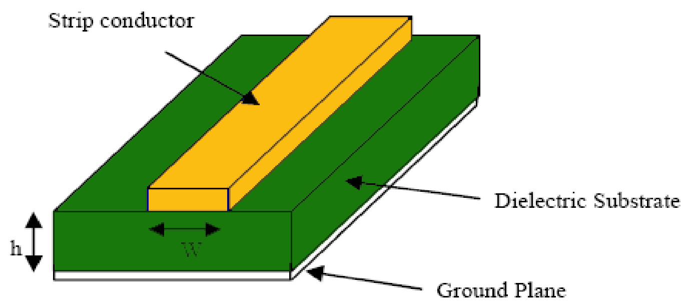


Figure 2.7 Microstrip Line

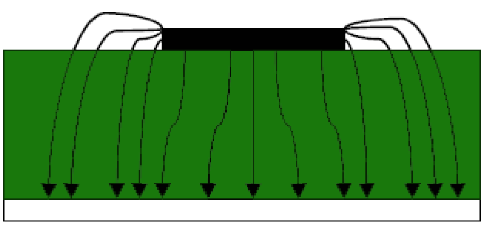


Figure 2.8 Electric Field Lines

Hence, as seen from Figure 2.8, most of the electric field lines reside in the substrate and parts of some lines in air. As a result, this transmission line cannot support pure transverse-electric-magnetic (TEM) mode of transmission, since the phase velocities would be different in the air and the substrate. Instead, the dominant mode of propagation would be the quasi-TEM mode. Hence, an effective dielectric constant (ε_{reff}) must be obtained in order to account for the fringing and the wave propagation in the line. The value of ε_{reff} is slightly less then ε_r because the fringing fields around the periphery of the patch are not confined in the dielectric substrate but are also spread in the air as shown in Figure 3.8 above. The expression for ε_{reff} is given by Balanis [12] as:

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

Where ε_{reff} = Effective dielectric constant

ε_r = Dielectric constant of substrate

h = Height of dielectric substrate

W = Width of the patch

Consider Figure 2.9 below, which shows a rectangular microstrip patch antenna of length *L*, width *W* resting on a substrate of height *h*. The co-ordinate axis is selected such that the length is along the x direction, width is along the *y* direction and the height is along the *z* direction.

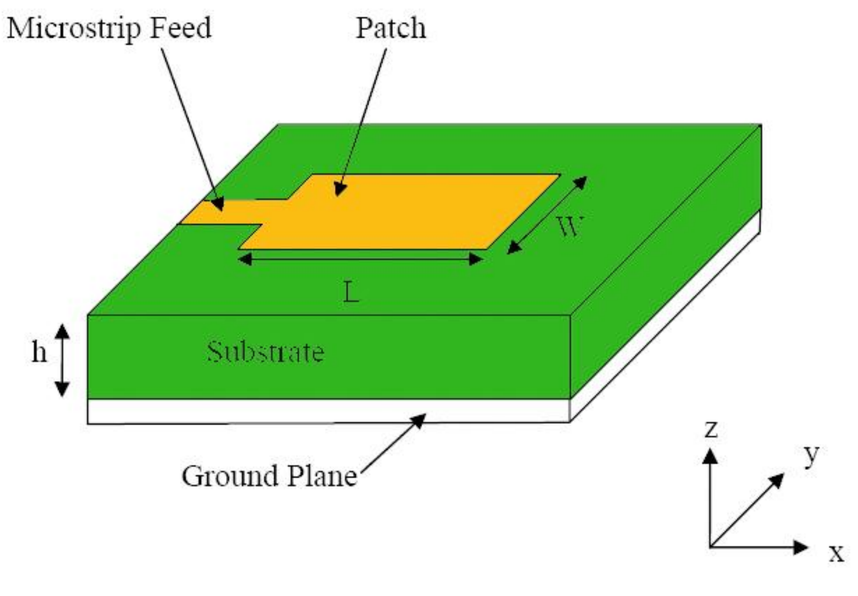


Figure 2.9 Microstrip Patch Antennas

In order to operate in the fundamental TM_{10} mode, the length of the patch must be slightly less than λ/2 where λ is the wavelength in the dielectric medium and is equal to $\lambda_o/\sqrt{\varepsilon_{reff}}$ where λ_o is the free space wavelength. The TM_{10} mode implies that the field varies one λ/2 cycle along the length, and there is no variation along the width of the patch. In the Figure 2.10 shown below, the microstrip patch antenna is represented by two slots, separated by a transmission line of length L and open circuited at both the ends. Along the width of the patch, the voltage is maximum and current is minimum due to the open ends. The fields at the edges can be resolved into normal and tangential components with respect to the ground plane.

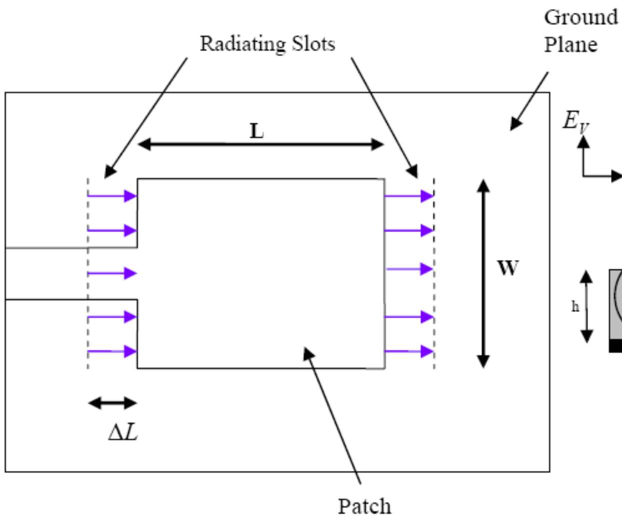


Figure 2.10 Top View of Antenna

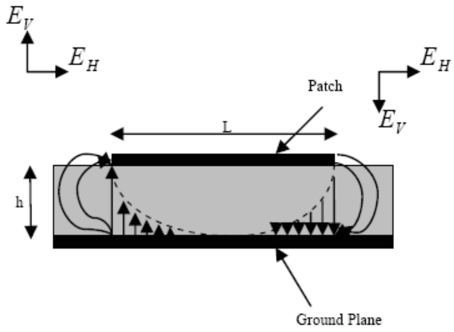


Figure 2.11 Side View of Antenna

It is seen from Figure 2.11 that the normal components of the electric field at the two edges along the width are in opposite directions and thus out of phase since the patch is λ/2 long and hence they cancel each other in the broadside direction. The tangential components (seen in Figure 2.11), which are in phase, means that the resulting fields combine to give maximum radiated field normal to the surface of the structure. Hence the edges along the width can be represented as two radiating slots, which are λ/2 apart and excited in phase and radiating in the half space above the ground plane. The fringing fields along the width can be modeled as radiating slots and electrically the patch of the microstrip antenna looks greater than its physical dimensions. The dimensions of the patch along its length have now been extended on each end by a distance ΔL, which is given empirically by Hammerstad [13] as:

$$\Delta L = 0.412h\frac{(\varepsilon_{reff} + 0.3)\left(\frac{W}{h} + 0.264\right)}{(\varepsilon_{reff} - 0.258)\left(\frac{W}{h} + 0.8\right)}$$

The effective length of the patch L_{eff} now becomes:

$$L_{eff} = L + 2\Delta L$$

For a given resonance frequency f_o, the effective length is given by [9] as:

$$L_{eff} = \frac{c}{2f_o\sqrt{\varepsilon_{reff}}}$$

For a rectangular Microstrip patch antenna, the resonance frequency for any TM_{mn} mode is given by James and Hall [14] as:

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

Where m and n are modes along L and W respectively.
For efficient radiation, the width W is given by Bahl and Bhartia [15] as:

$$W = \frac{c}{2f_o\sqrt{\frac{(\varepsilon_r + 1)}{2}}}$$

3.4.2 Cavity Model

Although the transmission line model discussed in the previous section is easy to use, it has some inherent disadvantages. Specifically, it is useful for patches of rectangular design and it ignores field variations along the radiating edges. These disadvantages can be overcome by using the cavity model. A brief overview of this model is given below.

In this model, the interior region of the dielectric substrate is modeled as a cavity bounded by electric walls on the top and bottom. The basis for this assumption is the following observations for thin substrates (h << λ) [10].

- Since the substrate is thin, the fields in the interior region do not vary much in the *z* direction, i.e. normal to the patch.
- The electric field is *z* directed only, and the magnetic field has only the transverse components H_x and H_y in the region bounded by the patch metallization and the ground plane. This observation provides for the electric walls at the top and the bottom.

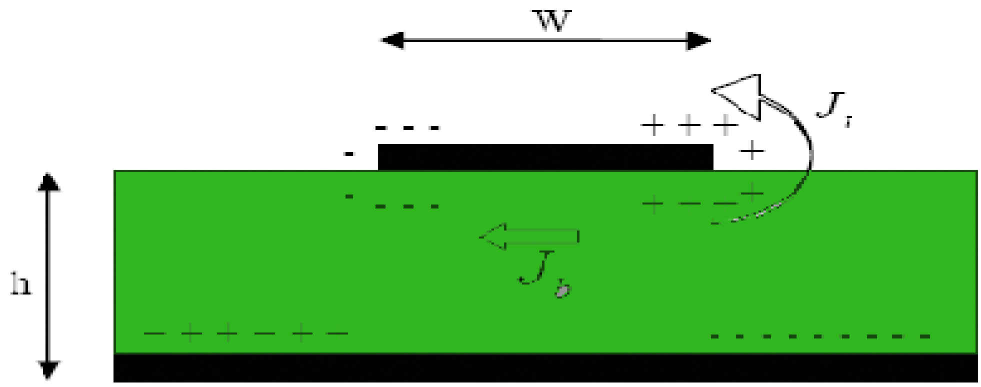


Figure 2.12 Charge distribution and current density creation on the microstrip patch

Consider Figure 2.12 shown above. When the microstrip patch is provided power, a charge distribution is seen on the upper and lower surfaces of the patch and at the bottom of the ground plane. This charge distribution is controlled by two mechanisms-an attractive mechanism and a repulsive mechanism as discussed by Richards [16]. The attractive mechanism is between the opposite charges on the bottom side of the patch and the ground plane, which helps in keeping the charge concentration intact at the bottom of the patch. The repulsive mechanism is between the like charges on the bottom surface of the patch, which causes pushing of some charges from the bottom, to the top of the patch. As a result of this charge movement, currents flow at the top and bottom surface of the patch. The cavity model assumes that the height to width ratio (i.e. height of substrate and width of the patch) is very small and as a result of this the attractive mechanism dominates and causes most of the charge concentration and the current to be below the patch surface. Much less current would flow on the top surface of the patch and as the height to width ratio further decreases, the current on the top surface of the patch would be almost equal to zero, which would not allow the creation of any tangential magnetic field components to the patch edges. Hence, the four sidewalls could be modeled as perfectly magnetic conducting surfaces. This implies that the magnetic fields and the electric field distribution beneath the patch would not be disturbed. However, in practice, a finite width to height ratio would be there and this would not make the tangential magnetic fields to be completely zero, but they being very small, the side walls could be approximated to be perfectly magnetic conducting [5].

Since the walls of the cavity, as well as the material within it are lossless, the cavity would not radiate and its input impedance would be purely reactive. Hence, in order to account for radiation and a loss mechanism, one must introduce a radiation resistance R_R and a loss resistance R_L. A lossy cavity would now represent an antenna and the loss is taken into account by the effective loss tangent δ_{eff} which is given as:

$$\delta_{eff} = 1/Q_T$$

Q_T is the total antenna quality factor and has been expressed by [4] in the form:

$$\frac{1}{Q_T} = \frac{1}{Q_d} + \frac{1}{Q_c} + \frac{1}{Q_r}$$

- Q_d represents the quality factor of the dielectric and is given as :

$$Q_d = \frac{\omega_r W_T}{P_d} = \frac{1}{\tan\delta}$$

where ω_r is the angular resonant frequency.

W_T is the total energy stored in the patch at resonance.

P_d is the dielectric loss.

$\tan\delta$ is the loss tangent of the dielectric.

- Q_c represents the quality factor of the conductor and is given as :

$$Q_c = \frac{\omega_r W_T}{P_c} = \frac{h}{\Delta}$$

where P_c is the conductor loss.

Δ is the skin depth of the conductor.

h is the height of the substrate.

- Q_r represents the quality factor for radiation and is given as:

$$Q_r = \frac{\omega_r W_T}{P_r}$$

where P_r is the power radiated from the patch.

Hence

$$\delta_{eff} = \tan\delta + \frac{\Delta}{h} + \frac{P_r}{\omega_r W_T}$$

Thus, the above equation describes the total effective loss tangent for the microstrip patch antenna.

Chapter 3

RECTANGULAR PATCH ANTENNA

Introduction

Basic Principles of Operation

Resonant Frequency

Radiation Patterns

Radiation Efficiency

Bandwidth

Input Impedence

Improving Perormance

CHAPTER 3

RECTANGULAR PATCH ANTENNA

3.1 Introduction

Microstrip antennas are among the most widely used types of antennas in the microwave frequency range, and they are often used in the millimeter-wave frequency range as well [1, 2, 3]. (Below approximately 1 GHz, the size of a microstrip antenna is usually too large to be practical, and other types of antennas such as wire antennas dominate). Also called patch antennas, microstrip patch antennas consist of a metallic patch of metal that is on top of a grounded dielectric substrate of thickness *h*, with relative permittivity and permeability ε_r and μ_r as shown in Figure 3.1 (usually μ_r=1). The metallic patch may be of various shapes, with rectangular and circular being the most common, as shown in Figure 3.1.

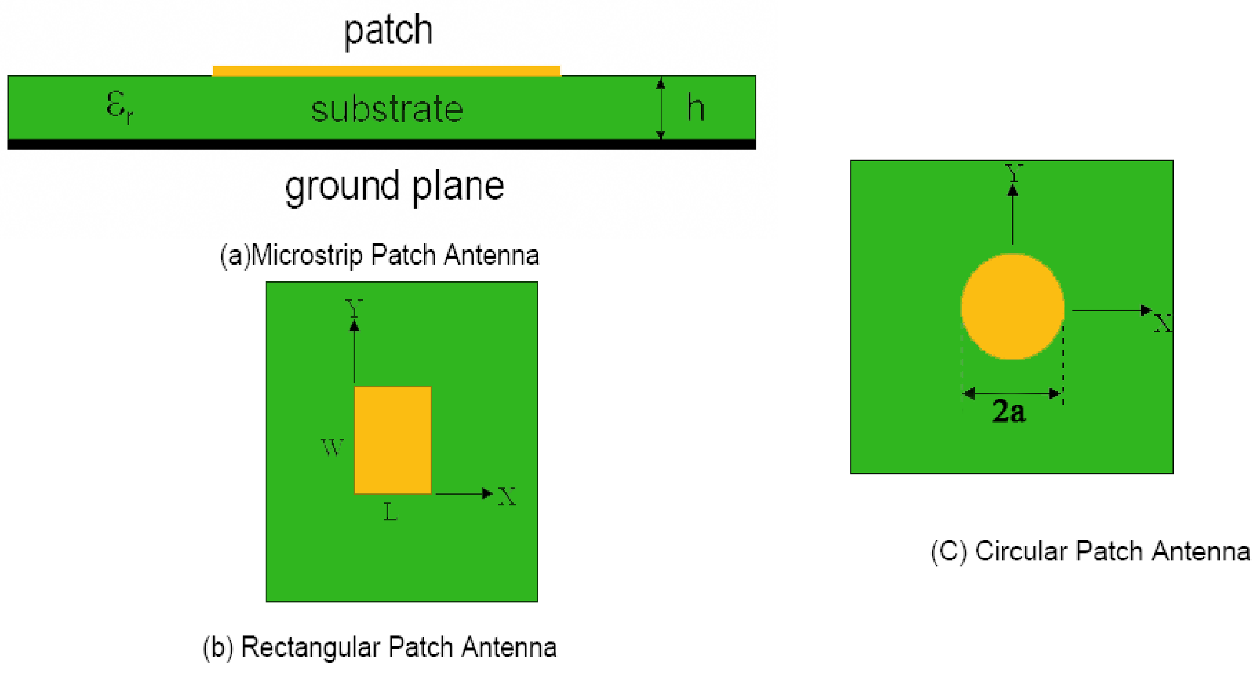


Fig. 3.1/ Rectangular & Circular Patch Antenna

Most of the discussion in this section will be limited to the rectangular patch, although the basic principles are the same for the circular patch. (Many of the CAD formulas presented will apply approximately for the circular patch if the circular patch is modeled as a square patch of the same area.) Various methods may be used to feed the patch, as discussed below. One advantage of the microstrip antenna is that it is usually low profile, in the sense that the substrate is fairly thin. If the substrate is thin enough, the antenna actually becomes "conformal," meaning that the substrate can be bent to conform to a curved surface (e.g., a cylindrical structure). A typical substrate thickness is about 0.02 λ_0. The metallic patch is usually fabricated by a photolithographic etching process or a mechanical milling process, making the construction relatively easy and inexpensive (the cost is mainly that of the substrate material). Other advantages include the fact that the microstrip antenna is usually lightweight (for thin substrates) and durable.

Disadvantages of the microstrip antenna include the fact that it is usually narrowband, with bandwidths of a few percent being typical. Some methods for enhancing bandwidth are discussed later, however. Also, the radiation efficiency of the patch antenna tends to be lower than some other types of antennas, with efficiencies between 70% and 90% being typical.

3.2 Basic Principles of Operation

The metallic patch essentially creates a resonant cavity, where the patch is the top of the cavity, the ground plane is the bottom of the cavity, and the edges of the patch form the sides of the cavity. The edges of the patch act approximately as an open-circuit boundary condition. Hence, the patch acts approximately as a cavity with perfect electric conductor on the top and bottom surfaces, and a perfect "magnetic conductor" on the sides. This point of view is very useful in analyzing the patch antenna, as well as in understanding its behavior. Inside the patch cavity the electric field is essentially *z* directed and independent of the z coordinate. Hence, the patch cavity modes are described by a double index (*m*, *n*). For the (*m*, *n*) cavity mode of the rectangular patch the electric field has the form

$$E_z(x,y) = A_{mn}\cos\left(\frac{m\pi x}{L}\right)\cos\left(\frac{n\pi y}{W}\right)$$

Where L is the patch length and W is the patch width. The patch is usually operated in the (1, 0) mode, so that L is the resonant dimension, and the field is essentially constant in the y direction. The surface current on the bottom of the metal patch is then x directed, and is given by

$$J_{sx}(x) = A_{10}\left(\frac{\pi / L}{j\omega\mu_0\mu_r}\right)$$

For this mode the patch may be regarded as a wide microstrip line of width *W*, having a resonant length *L* that is approximately one-half wavelength in the dielectric. The current is maximum at the centre of the patch, $x = L/2$, while the electric field is maximum at the two "radiating" edges, $x = 0$ and $x = L$. The width *W* is usually chosen to be larger than the length ($W = 1.5\,L$ is typical) to maximize the bandwidth, since the bandwidth is proportional to the width. (The width should be kept less than twice the length, however, to avoid excitation of the (0,2) mode.)

At first glance, it might appear that the microstrip antenna will not be an effective radiator when the substrate is electrically thin, since the patch current in (2) will be effectively shorted by the close proximity to the ground plane. If the modal amplitude A_{10} were constant, the strength of the radiated field would in fact be proportional to *h*. However, the Q of the cavity increases as *h* decreases (the radiation Q is inversely proportional to *h*). Hence, the amplitude A_{10} of the modal field at resonance is inversely proportional to *h*. Hence, the strength of the radiated field from a resonant patch is essentially independent of *h*, if losses are ignored. The resonant input resistance will likewise be nearly independent of *h*. This explains why a patch antenna can be an effective radiator even for very thin substrates, although the bandwidth will be small.

3.3 Resonant Frequency

The resonance frequency for the (1, 0) mode is given by

$$f_0 = \frac{c}{2L_e\sqrt{\varepsilon_r}}$$

Where *c* is the speed of light in vacuum. To account for the fringing of the cavity fields at the edges of the patch, the length, the effective length *Le* is chosen as

$L_e = L + 2\Delta L$

The Hammerstad formula for the fringing extension is [1]

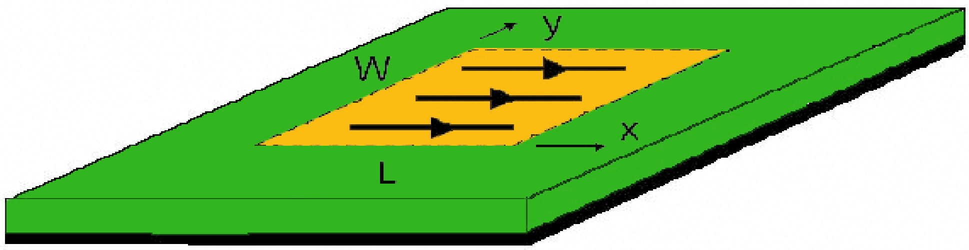
$$\frac{\Delta L}{h} = 0.412\left(\frac{\left(\varepsilon_{eff} + 0.3\right)\left(\frac{W}{h} + 0.264\right)}{\left(\varepsilon_{eff} - 0.258\right)\left(\frac{W}{h} + 0.8\right)}\right)$$

Where

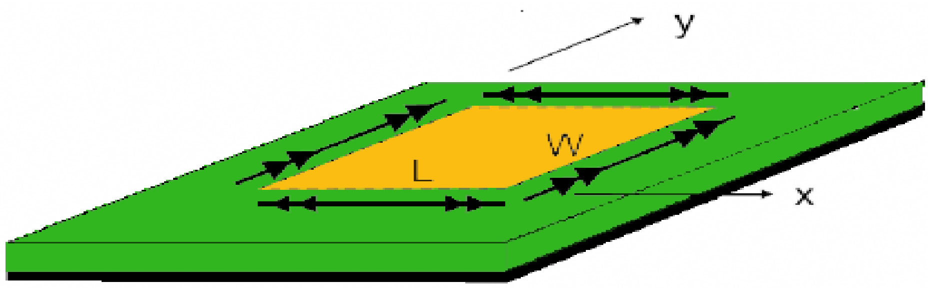
$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + \frac{12h}{W}\right)^{-1/2}$$

3.4 Radiation Patterns

The radiation field of the microstrip antenna may be determined using either an "electric current model" or a "magnetic current model". In the electric current model, the current in (2) is used directly to find the far-field radiation pattern. Figure 3.3a shows the electric current for the (1, 0) patch mode. If the substrate is neglected (replaced by air) for the calculation of the radiation pattern, the pattern may be found directly from image theory. If the substrate is accounted for, and is assumed infinite, the reciprocity method may be used to determine the far-field pattern [5].



(a) Electric Current for (1,0) patch



(b) Magnetic Current for (1,0) patch

Fig. 3.2 / Electric & Magnetic Current Distribution

In the magnetic current model, the equivalence principle is used to replace the patch by a magnetic surface current that flows on the perimeter of the patch. The magnetic surface current is given by

$$\vec{M}_s = -\hat{n} \times \vec{E}$$

Where **E** is the electric field of the cavity mode at the edge of the patch and **n** is the outward pointing unit-normal vector at the patch boundary. Figure 3b shows the magnetic current for the (1, 0) patch mode. The far-field pattern may once again be determined by image theory or reciprocity, depending on whether the substrate is neglected or not [4]. The dominant part of the radiation field comes from the "radiating edges" at $x = 0$ and $x = L$. The two non-radiating edges do not affect the pattern in the principle planes (the E plane at φ = 0 and the H plane at φ = π/2), and have a small effect for other planes.

It can be shown that the electric and magnetic current models yield exactly the same result for the far-field pattern, provided the pattern of each current is calculated in the presence of the substrate at the resonant frequency of the patch cavity mode [5]. If the substrate is neglected, the agreement is only approximate, with the largest difference being near the horizon.

According to the electric current model, accounting for the infinite substrate, the far-field pattern is given by [5]

$$E_i(r,\theta,\phi) = E_i^h(r,\theta,\phi)\left(\frac{\pi WL}{2}\right)\left(\frac{\sin\left(\frac{k_y W}{2}\right)}{\frac{k_y W}{2}}\right)\left(\frac{\cos\left(\frac{k_x L}{2}\right)}{\left(\frac{\pi}{2}\right)^2 - \left(\frac{k_x L}{2}\right)^2}\right)$$

Where

$k_x = k_0 \sin\theta \cos\varphi$

$ky = k0 \sin\theta \sin\varphi$

and E_i^h is the far-field pattern of an infinitesimal (Hertzian) unit-amplitude *x*- directed electric dipole at the centre of the patch. This pattern is given by [5]

$E_\theta^h(r,\theta,\varphi) = E_0 \cos\varphi G(\theta)$

$E_\varphi^h(r,\theta,\varphi) = -E_0 \sin\varphi F(\theta)$

Where

$$E_0 = \left(\frac{-j\omega\mu_0}{4\pi r} \right) e^{-jk_0 r}$$

$$F(\theta) = \frac{2\tan(k_0 h N(\theta))}{\tan(k_0 h N(\theta)) - j\dfrac{N(\theta)}{\mu_r}\sec\theta}$$

$$G(\theta) = \frac{2\tan(k_0 h N(\theta))\cos\theta}{\tan(k_0 h N(\theta)) - j\dfrac{\varepsilon_r}{N(\theta)}\cos\theta}$$

And

$$N(\theta) = \sqrt{n_1^2 - \sin^2\theta}$$

$$n_1^2 = \varepsilon_r \mu_r$$

The radiation patterns (E- and H-plane) for a rectangular patch antenna on an infinite substrate of permittivity ε_r = 2.2 and thickness h/λ_0= 0.02 are shown in Figure 3.4. The patch is resonant with W / L = 1.5. Note that the E-plane pattern is broader than the H-plane pattern. The directivity is approximately 6 dB.

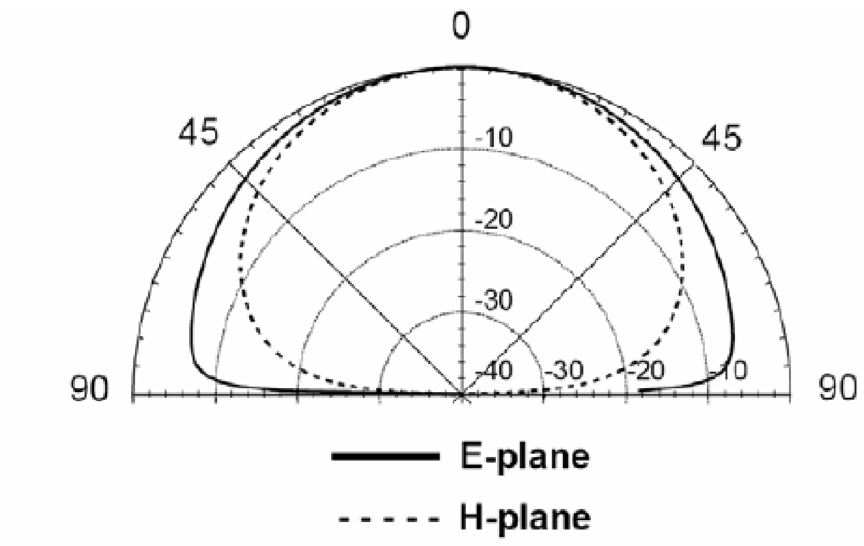


Fig. 3.3 / Radiation Pattern (E & H plane)

3.5 Radiation Efficiency

The radiation efficiency of the patch antenna is affected not only by conductor and dielectric losses, but also by surface-wave excitation - since the dominant TM0 mode of the grounded substrate will be excited by the patch. As the substrate thickness decreases, the effect of the conductor and dielectric losses becomes more severe, limiting the efficiency. On the other hand, as the substrate thickness increases, the surface-wave power increases, thus limiting the efficiency. Surface-wave excitation is undesirable for other reasons as well, since surface waves contribute to mutual coupling between elements in an array, and also cause undesirable edge diffraction at the edges of the ground plane or substrate, which often contributes to distortions in the pattern and to back radiation. For an air (or foam) substrate there is no surface-wave excitation. In this case, higher efficiency is obtained by making the substrate thicker, to minimize conductor and dielectric losses (making the substrate too thick may lead to difficulty in matching, however, as discussed above). For a substrate with a moderate relative permittivity such as ε_r = 2.2, the efficiency will be maximum when the substrate thickness is approximately λ0 = 0.02. The radiation efficiency is defined as

$$e_r = \frac{P_{sp}}{P_{Total}} = \frac{P_{sp}}{P_c + P_d + + P_{sw} + P_{sp}}$$

Where *P sp* is the power radiated into space, and the total input power *P total* is given as the sum of *Pc* - the power dissipated by conductor loss, *Pd*- the power dissipated by dielectric loss, and P_{sw} - the surface-wave power. The efficiency may also be expressed in terms of the corresponding Q factors as

$$e_r = \left(\frac{Q_{sp}}{Q_{Total}} \right)^{-1}$$

Where

$$\frac{1}{Q_{Total}} = \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_{sw}} + \frac{1}{Q_{sp}}$$

The dielectric and conductor Q factors are given by

$$Q_d = \frac{1}{\tan \delta_d}$$

$$Q_c = \frac{1}{2}\eta_0 \mu_r \left(\frac{k_0 h}{R_s} \right)$$

Where $\tan\delta_d$ is the loss tangent of the substrate and *Rs* is the surface resistance of the patch and ground plane metal at radian frequency $\omega = 2\pi f$, given by

$$R_s = \left(\frac{\omega\mu_0}{2\sigma}\right)^{1/2}$$

where σ is the conductivity of the metal.

The space-wave Q factor is given approximately as [6]

$$Q_{sp} = \frac{3}{16}\left(\frac{\varepsilon_r}{pc_1}\right)\left(\frac{L}{W}\right)\left(\frac{1}{h/\lambda_0}\right)$$

Where

$$c_1 = 1 - \frac{1}{n_1^2} + \frac{2/5}{n_1^4}$$

$$p = 1 + \frac{a_2}{10}(k_0W)^2 + \left(a_2^2 + 2a_4\right)\left(\frac{3}{560}\right)(k_0W)^4 + c_2\left(\frac{1}{5}\right)(k_0L)^2$$
$$+ a_2c_2\left(\frac{1}{70}\right)(k_0W)^2(k_0L)^2$$

with $a_2 = -0.16605$, $a4 = 0.00761$, and $c2 = -0.0914153$.

The surface-wave Q factor is related to the space-wave Q factor as

$$Q_{sw} = Q_{sp}\left(\frac{e_r^{sw}}{1 - e_r^{sw}}\right)$$

where e_r^{sw} is the radiation efficiency accounting only for surface-wave loss.

This efficiency may be accurately approximated by using the radiation efficiency of an infinitesimal dipole on the substrate layer [6], giving

$$e_r^{sw} = \frac{1}{1 + (k_0h)\left(\frac{3}{4}\right)(\pi\mu_r)\left(\frac{1}{c_1}\right)\left(1 - \frac{1}{n_1^3}\right)^3}$$

A plot of radiation efficiency for a resonant rectangular patch antenna with $W/L = 1.5$ on a substrate of relative permittivity $\varepsilon_r = 2.2$ or $\varepsilon_r = 10.8$ is shown in Figure 2.5. The conductivity of the copper patch and ground plane is assumed to be $\sigma = 3.0 \times 10^7$ [S/m] and the dielectric loss tangent is taken as $\tan\delta_d = 0.001$. The resonance frequency is 5.0 GHz. (The result is plotted versus normalized (electrical) thickness of the substrate, which does not involve frequency.

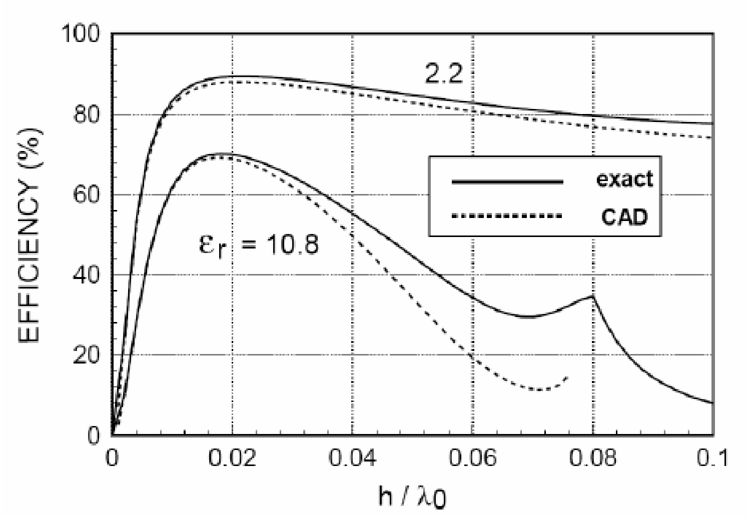


Fig.3.4 / Radiation Efficiency for a rectangular patch Antenna

However, a specified frequency is necessary to determine conductor loss.) For h / λ0 < 0.02, the conductor and dielectric losses dominate, while for h /λ0 < 0.02, the surface-wave losses dominate. (If there were no conductor or dielectric losses, the efficiency would approach 100% as the substrate thickness approaches zero.)

3.6 Bandwidth

The bandwidth increases as the substrate thickness increases (the bandwidth is directly proportional to h if conductor, dielectric, and surface-wave losses are ignored). However, increasing the substrate thickness lowers the Q of the cavity, which increases spurious radiation from the feed, as well as from higher-order modes in the patch cavity. Also, the patch typically becomes difficult to match as the substrate thickness increases beyond a certain point (typically about 0.05 λ0). This is especially true when feeding with a coaxial probe, since a thicker substrate results in a larger probe inductance appearing in series with the patch impedance. However, in recent years considerable effort has been spent to improve the bandwidth of the microstrip antenna, in part by using alternative feeding schemes. The aperture-coupled feed of Figure 2.2c is one scheme that overcomes the problem of probe inductance, at the cost of increased complexity [7].

Lowering the substrate permittivity also increases the bandwidth of the patch antenna. However, this has the disadvantage of making the patch larger. Also, because the $Q$of the patch cavity is lowered, there will usually be increased radiation from higher-order modes, degrading the polarization purity of the radiation.

By using a combination of aperture-coupled feeding and a low-permittivity foam substrate, bandwidths exceeding 25% have been obtained. The use of stacked patches (a parasitic patch located above the primary driven patch) can also be used to increase bandwidth even further, by increasing the effective height of the structure and by creating a double-tuned resonance effect [8].

A CAD formula for the bandwidth (defined by SWR < 2.0) is

$$BW = \frac{1}{\sqrt{2}}\left[\tan\delta_d + \left(\frac{R_s}{\pi\eta_0\mu_r}\right)\left(\frac{1}{h/\lambda_0}\right) + \left(\frac{16}{3}\right)\left(\frac{pc_1}{\varepsilon_r}\right)\left(\frac{h}{\lambda_0}\right)\left(\frac{W}{L}\right)\left(\frac{1}{e_r^{sw}}\right)\right]$$

Where the terms have been defined in the previous section on radiation efficiency. The result should be multiplied by 100 to get percent bandwidth. Note that neglecting conductor and dielectric loss yields a bandwidth that is proportional to the substrate thickness *h*.

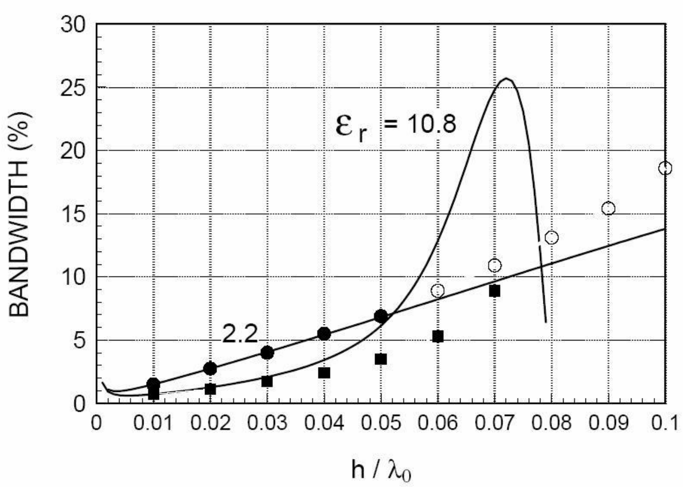


Fig. 3.5/ Calculated & Measured Bandwidth

Figure 2.6 shows calculated and measured bandwidth for the same patch in Figure 2.5. It is seen that bandwidth is improved by using a lower substrate permittivity, and by making the substrate thicker.

3.7 Input Impedence

A variety of approximate models have been proposed for the calculation of input impedance for a probe-fed patch. These include the transmission line method [9], the cavity model [10], and the spectral-domain method [11]. These models usually work well for thin substrates, typically giving reliable results for h / λ0 < 0.02. Commercial simulation tools using FDTD, FEM, or MoM can be used to accurately predict the input impedance for any substrate thickness. The cavity model has the advantage of allowing for a simple physical CAD model of the patch to be developed, as shown in Figure 3.6

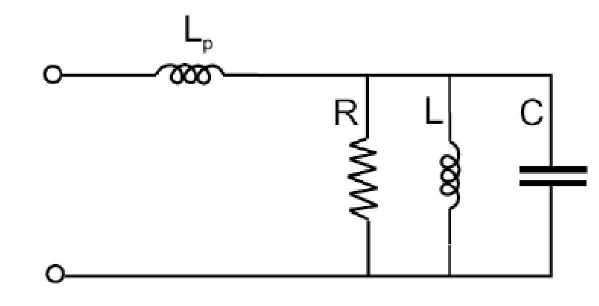


Fig. 3.6 / Equivalent Circuit of Patch Antenna

In this model the patch cavity is modelled as a parallel RLC circuit, while the probe inductance is modelled as a series inductor. The input impedance of this circuit is approximately described by

$$Z_{in} \approx jX_f + \frac{R}{1 + j2Q(f/f_0 - 1)}$$

where f_0 is the resonance frequency, R is the input resistance at the resonance of the RLC circuit (where the input resistance of the patch is maximum), $Q = Q_{total}$ is the quality factor of the patch cavity (20), and $X_f = \omega L_p$ is the feed (probe) reactance of the coaxial probe. A CAD formula for the input resistance R is

$$R = R_{edge} \cos^2\left(\frac{\pi x_0}{L}\right)$$

where the input resistance at the edge is

$$R_{edge} = \frac{\left(\frac{4}{\pi}\right)(\mu_r \eta_0)\left(\frac{L}{W}\right)\left(\frac{h}{\lambda_0}\right)}{\tan\delta_d + \left(\frac{R_s}{\pi\eta_0\mu_r}\right)\left(\frac{1}{h/\lambda_0}\right) + \left(\frac{16}{3}\right)\left(\frac{pc_1}{\varepsilon_r}\right)\left(\frac{h}{\lambda_0}\right)\left(\frac{W}{L}\right)\left(\frac{1}{e_r^{sw}}\right)}$$

A CAD formula for the feed reactance due to the probe is

$$X_f = \frac{\eta_0}{2\pi}\mu_r(k_0 h)\left[-\gamma + \ln\left(\frac{2}{k_0 a\sqrt{\varepsilon_r \mu_r}}\right)\right]$$

where $\gamma = 0.577216$ is Euler's constant.

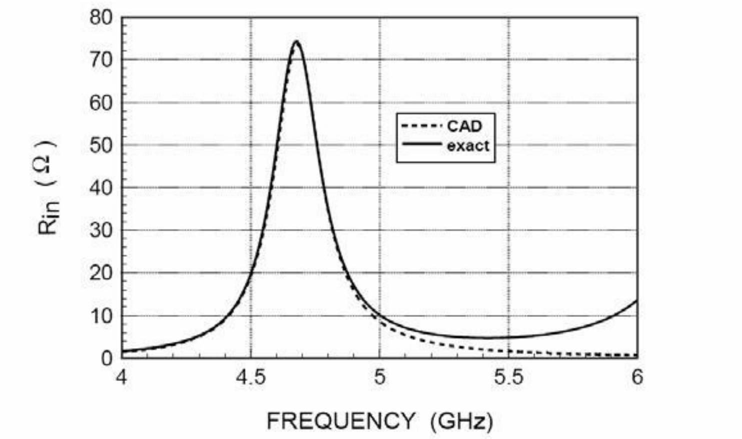


Fig. 3.7(a)

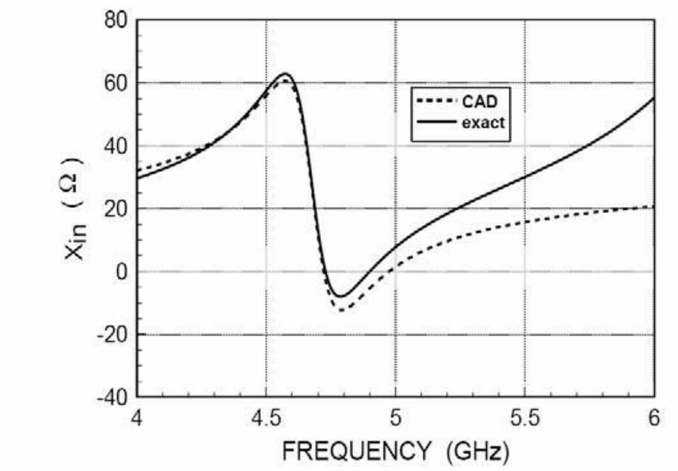


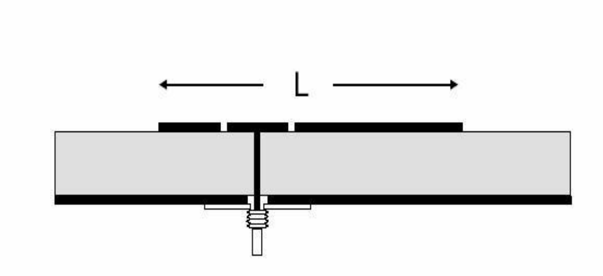
Fig. 3.7 (b)

Fig. 3.7 / Comparison of input Impedances

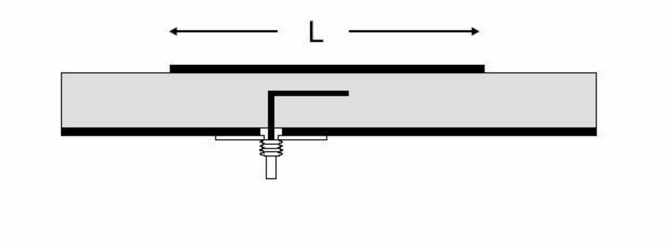
Figure 3.8 shows a comparison of the input impedance obtained from the simple CAD model (30) with that obtained by a more accurate cavity model analysis. At the resonance frequency, the substrate thickness is approximately $0.024\lambda_0$. Near the resonance frequency, the simple CAD model gives results that agree quite well with the cavity model.

3.8 Improving Performance

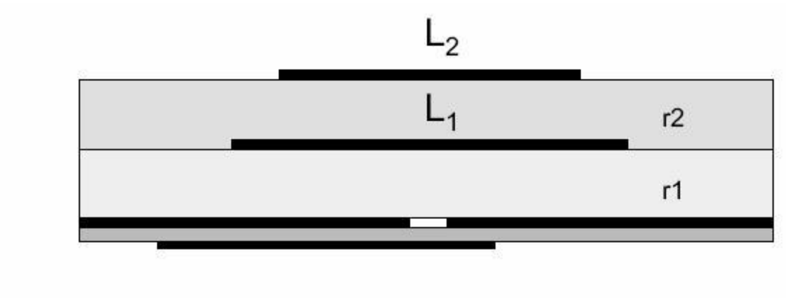
Much research has been devoted to improving the performance characteristics of the microstrip antenna. To improve bandwidth, the use of thick low-permittivity (e.g., foam) substrates can give significant improvement. To overcome the probe inductance associated with thicker substrates, the use of capacitive-coupled feeds such as the top-loaded probe [12] or the L-shaped probe [13] shown in Figure 3.9a and Figure 3.9b may be used. Alternatively, the aperture coupled fed shown in Figure 3.9c may be used, which also has the advantage of eliminating spurious probe radiation. To increase the bandwidth even further, a stacked patch arrangement may be used, in which a parasitic patch is stacked above the driven patch [8]. This may be done using either a probe feed or, to obtain even higher bandwidths, using an aperture-coupled feed (Figure 3.9c).



(a) Top Loading Probe Feeding



(b) L- Probe Feeding



(c) Aperture Coupled Feeding

Fig. 3.8 Feed Types

The bandwidth enhancement is largely due to the existence of a double resonance, and to some extent, to the fact that one of the radiators is further from the ground plane. Bandwidths as large as one octave (2:1 frequency band) have been obtained with such an arrangement. By using a diplexer feed to split the feeding signal into two separate branches, and feeding two aperture-coupled stacked patches with different centre frequencies, bandwidths of 4:1 have been obtained [14]. Parasitic patches may also be placed on the same substrate as the driven patch, surrounding the driven patch. A pair of parasitic patches may be coupled to the radiating edges, the non-radiating edges, or all four edges [15]. This planar arrangement saves vertical height and allows for easier fabrication, allows the substrate area occupied by the antenna is larger, and there may be more variation of the radiation pattern across the frequency band since the current distribution on the different patches changes with frequency. Broad banding may also be achieved through the use of slots cut into the patch, as in the "U-slot"patch design [16]. This has the advantage of not requiring multiple layers or increasing the size of the patch as with parasitic elements.

Another variation of the microstrip antenna that has been introduced recently is the "reduced surface wave" microstrip antenna shown in Figure 3.10 [17].

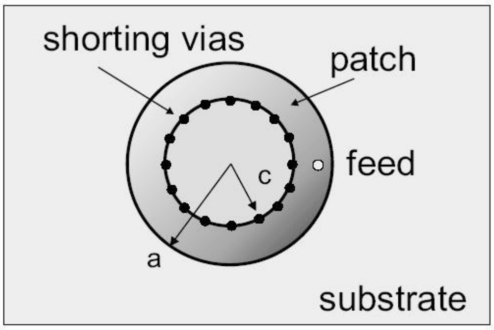


Fig. 3.9/ Reduced surface wave microstrip antenna

This design is a variation of a circular patch, with an inner ring of vias that creates short-circuit inner boundary. By properly selecting the outer radius, the patch excites very little surface-wave field, and also only a small amount of lateral (horizontally propagating) radiation. The inner short-circuit boundary is used to adjust the dimensions of the patch cavity (between the inner and outer boundaries) to make the patch resonant. The reduced surface-wave and lateral radiation result in less edge diffraction from the edges of the supporting ground plane, giving smoother patterns in the front-side region and less radiation in the backside region. Also, there is less mutual coupling between pairs of such antennas, especially as the separation increases.

3.9 Linear Polarization

Antenna Polarization is a very important parameter when choosing and installing an antenna. It helps to have a good grasp of all the aspects of this subject. 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 radiates wholly in one plane containing the direction of propagation. In a circular polarized antenna, the plane of polarization rotates in a circle making one complete revolution during one period of the wave. An antenna is said to be vertically polarized (linear) when its electric field is perpendicular to the Earth's surface.

A circular polarized wave radiates energy in both the horizontal and vertical planes and all planes in between. The difference, if any, between the maximum and the minimum peaks as the antenna is rotated through all angles, is called the axial ratio or ellipticity and is usually specified in decibels (dB). If the axial ratio is near 0 dB, the antenna is said to be circular polarized. If the axial ratio is greater than 1-2 dB, the polarization is often referred to as elliptical.

3.9.1 Important Considerations

Polarization is an important design consideration. The polarization of each antenna in a system should be properly aligned. Maximum signal strength between stations occurs when both stations are using identical polarization.

On line-of-sight (LOS) paths, it is most important that the polarization of the antennas at both ends of the path use the same polarization. In a linearly polarized system, a misalignment of polarization of 45 degrees will degrade the signal up to 3 dB and if misaligned 90 degrees the attenuation can be 20 dB or more. Likewise, in a circular polarized system, both antennas must have the same sense. If not, an additional loss of 20 dB or more will be incurred.

Linearly polarized antennas will work with circularly polarized antennas and vice versa. However, there will be up to a 3 dB loss in signal strength. In weak signal situations, this loss of signal may impair communications. Cross polarization is another consideration. It happens when unwanted radiation is present from a polarization which is different from the polarization in which the antenna was intended to radiate. For example, a vertical antenna may radiate some horizontal polarization and vice versa. However, this is seldom a problem unless there is noise or strong signals nearby.

3.9.2 Typical Applications

Vertical polarization is most often used when it is desired to radiate a radio signal in all directions such as widely distributed mobile units. Vertical polarization also works well in the suburbs or out in the country, especially where hills are present. As a result, nowadays most two-way Earth to Earth communications in the frequency range above 30 MHz use vertical polarization.

Horizontal polarization is used to broadcast television in the USA. Some say that horizontal polarization was originally chosen because there was an advantage to not have TV reception interfered with by vertically polarized stations such as mobile radio. Also, man made radio noise is predominantly vertically polarized and the use of horizontal polarization would provide some discrimination against interference from noise.

3.9.3 Other Considerations

If your antenna is to be located on an existing tower or building with other antennas in the vicinity, try to separate the antennas as far as possible from each other. In the UHF range, increasing separation even a few extra feet may significantly improve performance from problems such as desensitization.

When setting up your own exclusive communications link, it may be wise to first test the link with vertical and then horizontal polarization to see which yields the best performance (if any). If there are any reflections in the area, especially from structures or towers, one polarization may outperform the other.

On another note, when radio waves strike a smooth reflective surface, they may incur a 180 degree phase shift, a phenomenon known as specular or mirror image reflection. The reflected signal may then destructively or constructively affect the direct LOS signal. Circular polarization has been used to an advantage in these situations since the reflected wave would have a different sense than the direct wave and block the fading from these reflections.

3.9.4 Diversity Reception

Even if the polarizations are matched, other factors may affect the strength of the signal. The most common are long and short term fading. Long term fading results from changes in the weather (such as barometric pressure or precipitation) or when a mobile station moves behind hills or buildings. Short term fading is often referred to as "multipath" fading since it results from reflected signals interfering with the LOS signal.

Some of this fading phenomenon can be decreased by the use of diversity reception. This type of system usually employs dual antennas and receivers with some kind of "voting" system to choose the busiest signal. However, for best results, the antennas should be at least 20 wavelengths apart so that the signals are no longer correlated. This would be 20-25 feet at 880 MHz, quite a structural problem. Nowadays we are inundated with mobile radios and cellular telephones. The polarization on handheld units is often random depending on how they are held by the user. This has led to new studies which have found that polarization diversity can be an advantage. The most important break through in this area is that the antennas at the base station do not have to be separated physically as described above. They can be collocated as long as they are orthogonal and well isolated from each other. Only time will tell if these systems are truly cost effective.

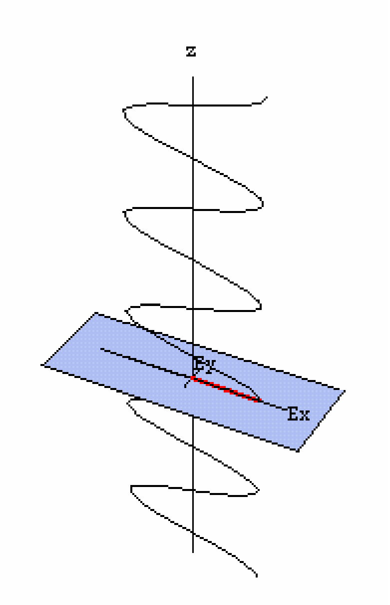


Fig 3.10 Linear Polarization

Chapter 4

MICROSTRIP PATCH ANTENNA AND RESULTS

CHAPTER 4

MICROSTRIP PATCH ANTENNA DESIGN AND RESULTS

In this chapter, the procedure for designing a rectangular microstrip patch antenna is explained. Next, a compact rectangular microstrip patch antenna is designed for use in cellular phones. Finally, the results obtained from the simulations are demonstrated.

4.1 Design Specifications

The three essential parameters for the design of a rectangular Microstrip Patch Antenna:

- Frequency of operation (f_o): The resonant frequency of the antenna must be selected appropriately. The Mobile Communication Systems uses the frequency range from 2100-5600 MHz. Hence the antenna designed must be able to operate in this frequency range. The resonant frequency selected for my design is 2.4 GHz.
- Dielectric constant of the substrate (ε_r): The dielectric material selected for our design is RT Duroid which has a dielectric constant of 2.45. A substrate with a high dielectric constant has been selected since it reduces the dimensions of the antenna.
- Height of dielectric substrate (h): For the microstrip patch antenna to be used in cellular phones, it is essential that the antenna is not bulky. Hence, the height of the dielectric substrate is selected as 1.58 mm.

Hence, the essential parameters for the design are:

- f_o = 2.4 GHz
- ε_r = 2.45
- h = 1.58 mm

4.2 Design Procedure (PSO/IE3D)

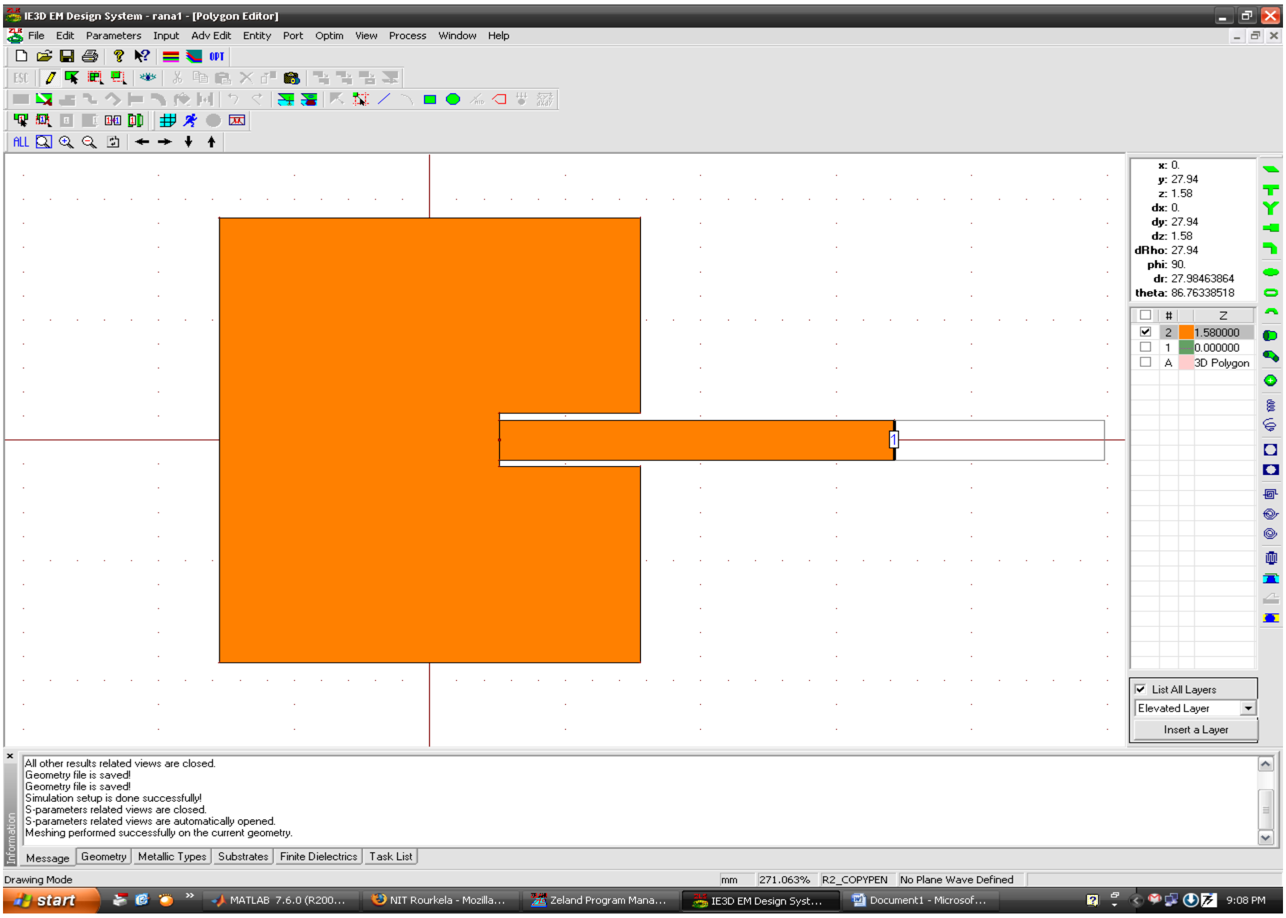


Figure 4.1: Microstrip patch antenna designed using IE3D

The transmission line model described in chapter 2 will be used to design the antenna.

Step 1: Calculation of the Width (*W*):

The width of the Microstrip patch antenna is given as:

$$W = \frac{c}{2 f_o \sqrt{\frac{(\varepsilon_r + 1)}{2}}}$$

Substituting c = 3.00e+008 m/s, ε_r = 2.45 and f_o = 2.4 GHz, we get:

W = 0.0475 m = 47.5 mm

Step 2: Calculation of Effective dielectric constant ($\varepsilon reff$):

The effective dielectric constant is:

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

Substituting ε_r = 2.45, W = 47.5 mm and h = 1.58 mm we get:

$\varepsilon reff$ = **2.3368**

Step 3: Calculation of the Effective length (L_{eff}**):**

The effective length is:

`

$$L_{eff} = \frac{c}{2f_o\sqrt{\varepsilon_{reff}}}$$

Substituting ε_{reff} = 2.3368, c = 3.00e+008 m/s and f_o = 2.4 GHz we get:

L_{eff}**= 0.0406 m = 40.625 mm**

Step 4: Calculation of the length extension (ΔL**):**

The length extension is:

$$\Delta L = 0.412h\frac{\left(\varepsilon_{reff} + 0.3\right)\left(\frac{W}{h} + 0.264\right)}{\left(\varepsilon_{reff} - 0.258\right)\left(\frac{W}{h} + 0.8\right)}$$

Substituting ε_{reff} = 2.3668, W = 47.5 mm and h = 1.58 mm we get:

ΔL **= 0.81 mm**

Step 5: Calculation of actual length of patch (L**):**

The actual length is obtained by:

$$L = L_{eff} - 2\Delta L$$

Substituting L_{eff}= 40.625 mm and ΔL = 0.81 mm we get:

L **= 39 m = 39 mm**

Step 6: Calculation of the ground plane dimensions (L_g and W_g):

The transmission line model is applicable to infinite ground planes only. However, for practical considerations, it is essential to have a finite ground plane. It has been shown by [9] that similar results for finite and infinite ground plane can be obtained if the size of the ground plane is greater than the patch dimensions by approximately six times the substrate thickness all around the periphery. Hence, for this design, the ground plane dimensions would be given as:

$$L_g = 6h + L = 6(1.5) + 39 = 48 \text{ mm}$$

$$W_g = 6h + W = 6(1.5) + 47.5 = 56.5 \text{ mm}$$

Step 7: Determination of Inset feed depth (y_0):

An inset-fed type feed is to be used in this design. As shown in Figure 4.1, the feed depth is given by y_0. The feed point must be located at that point on the patch, where the input impedance is 50 ohms for the resonant frequency. Hence, a trial and error method is used to locate the feed point. In this case we use PSO to obtain the optimum feed depth, where the return loss (R.L) is most negative (i.e. the least value). According to [5] there exists a point along the length of the patch which gives the minimum return loss.

$$R_{in}(y = y_0) = R_{in}(y = 0)\cos^4(\pi * y_0/L)$$

Where, $R_{in}(y=0) = 0.5 * (G_1 \pm G_{12})$

$$Z_c = \begin{cases} \dfrac{60}{\sqrt{\epsilon_{reff}}}\ln\left[\dfrac{8h}{W_0} + \dfrac{W_0}{4h}\right], & \dfrac{W_0}{h} \le 1 \\[2ex] \dfrac{120\pi}{\sqrt{\epsilon_{reff}}\left[\dfrac{W_0}{h} + 1.393 + 0.667\ln\left(\dfrac{W_0}{h} + 1.444\right)\right]}, & \dfrac{W_0}{h} > 1 \end{cases}$$

Where,

$$G_1 = \begin{cases} \dfrac{1}{90}\left(\dfrac{W}{\lambda_0}\right)^2 & W \ll \lambda_0 \\[2ex] \dfrac{1}{120}\left(\dfrac{W}{\lambda_0}\right) & W \gg \lambda_0 \end{cases}$$

And,

$$G_{12} = \frac{1}{120\pi^2} \int_0^{\pi} \left[\frac{\sin\left(\frac{k_0 W}{2} \cos\theta\right)}{\cos\theta} \right]^2 J_0(k_0 L \sin\theta) \sin^3\theta \, d\theta$$

Using the first equation (assuming that Z_C in the second equation is 50 Ω) where R_{in} ($y=y_0$) = 50 Ω we get:

$$\mathbf{y_0 = 13\ mm}$$

4.3 Simulation Setup and Results

The software used to model and simulate the Microstrip patch antenna is Zeland Inc's IE3D. IE3D is a full-wave electromagnetic simulator based on the method of moments. It analyzes 3D and multilayer structures of general shapes. It has been widely used in the design of MICs, RFICs, patch antennas, wire antennas, and other RF/wireless antennas. It can be used to calculate and plot the S_{11} parameters, VSWR, current distributions as well as the radiation patterns.

4.3.1 Return Loss

The inset feed used is designed to have an inset depth of 13.2mm, feed-line width of 5.6mm and feed path length of 37mm. A frequency range of 2.2-3.5 GHz is selected and 151frequency points are selected over this range to obtain accurate results.

The center frequency is selected as the one at which the return loss is minimum. As described in chapter 2, the bandwidth can be calculated from the return loss (RL) plot. The bandwidth of the antenna can be said to be those range of frequencies over which the RL is greater than -9.5 dB (-9.5 dB corresponds to a VSWR of 2 which is an acceptable figure). Using PSO, the optimum feed depth is found to be at Y_o = 13.2mm where a RL of -27 dB is obtained. The bandwidth of the antenna for this feed point location is calculated (as shown below in Figure 4.4) to be 23.28 MHz and a center frequency of 1.9120 GHz is obtained which is very close to the desired design frequency of 1.9 GHz.

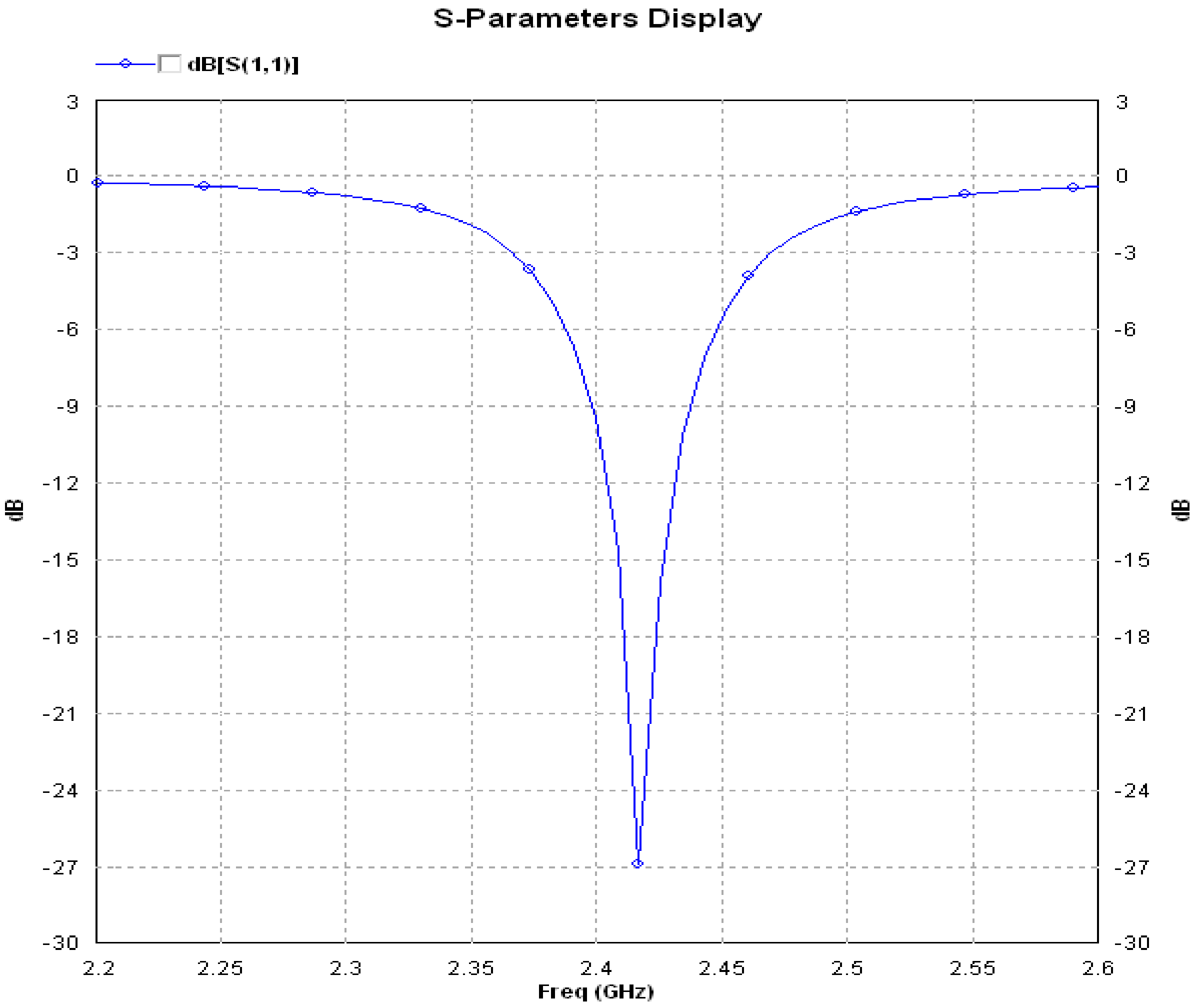


Figure 4.2: S-parameter plot for Return loss v/s frequency

The **Smith Chart**, invented by Phillip H. Smith (1905-1987), is a graphical aid or nomogram specializing in radio frequency (RF) engineering to assist in solving problems with transmission lines and matching circuits. The Smith Chart is plotted on the complex reflection coefficient plane in two dimensions and is scaled in normalized impedance (the most common), normalized admittance or both, using different colours to distinguish between them. These are often known as the Z, Y and YZ Smith Charts respectively. Normalized scaling allows the Smith Chart to be used for problems involving any characteristic impedance or system impedance, although by far the most commonly used is 50 ohms. As such, figures 4.3 and 4.4 show that the input impedance of the port was matched with the normalized Z_C value of 50 Ω at the frequency 2.408 GHz, which is near the operating frequency of 2.4 GHz.

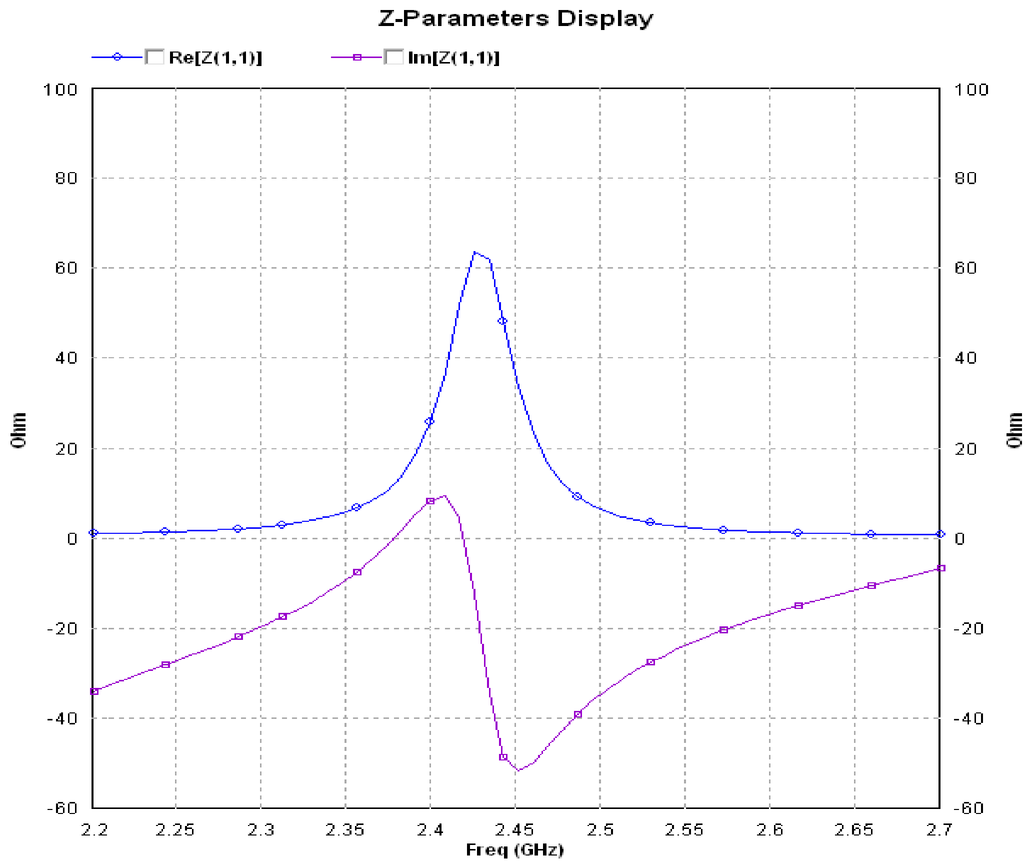


Figure 4.3: Z-parameter plot for Input impedance (Z_C)

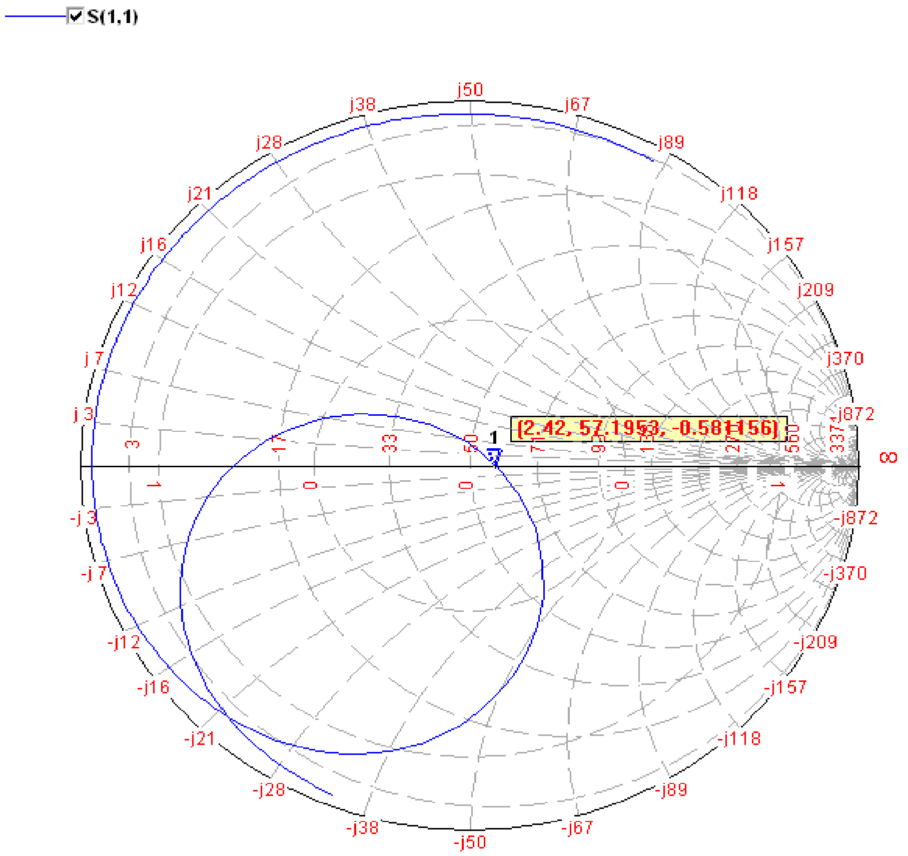


Figure 4.4: Smith Chart Display

4.3.2 Radiation Pattern plots

Since a Microstrip patch antenna radiates normal to its patch surface, the elevation pattern for $\varphi = 0$ and $\varphi = 90$ degrees would be important. Figure 4.5 below shows the gain of the antenna at 2.4 GHz for $\varphi = 0$ and $\varphi = 90$ degrees.

The maximum gain is obtained in the broadside direction and this is measured to be 1.87 dBi for both, $\varphi = 0$ and $\varphi = 90$ degrees.

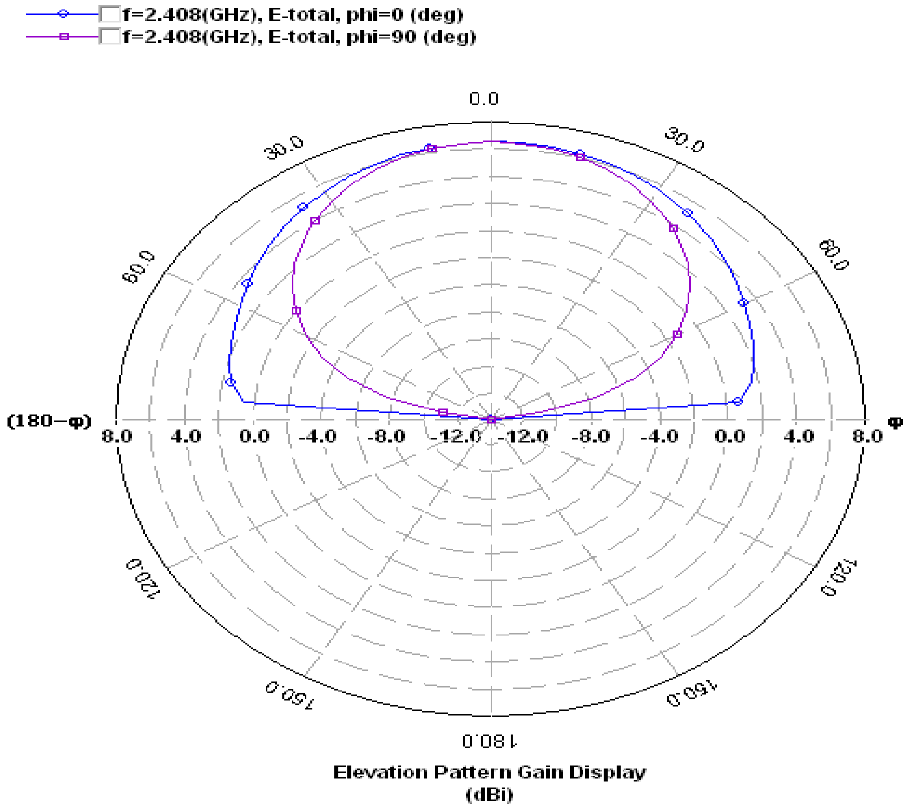


Figure 4.5: Elevation Pattern for $\varphi = 0$ and $\varphi = 90$ degrees

Mesh generation is the practice of generating a polygonal or polyhedral mesh that approximates a geometric domain to the highest possible degree of accuracy. The term "grid generation" is often used interchangeably. Typical uses are for rendering to a computer screen or for physical simulation such as finite element analysis or computational fluid dynamics. The triangulated zones in the mesh shown in figure 4.6 indicate the points in the grid where the current distributed is concentrated.

S-parameters are calculated from the average current distribution of the cross section, and thus the exact current distribution is not required to be precise.

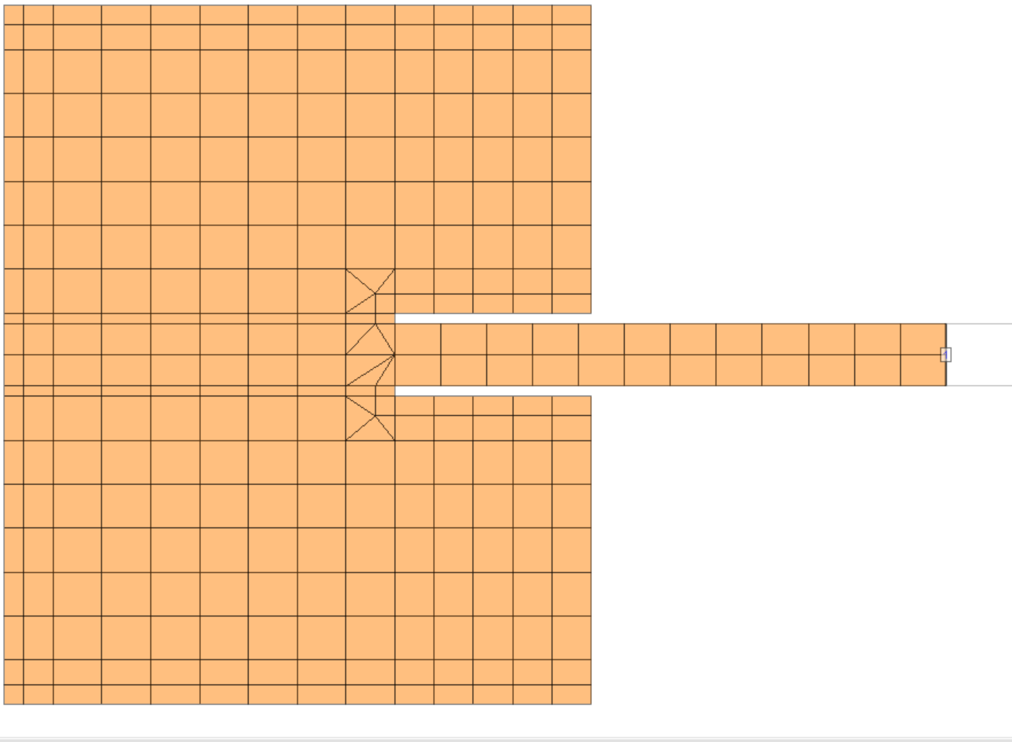


Figure 4.6: IE3D Mesh pattern of the patch antenna

4.3.3 Current Distribution

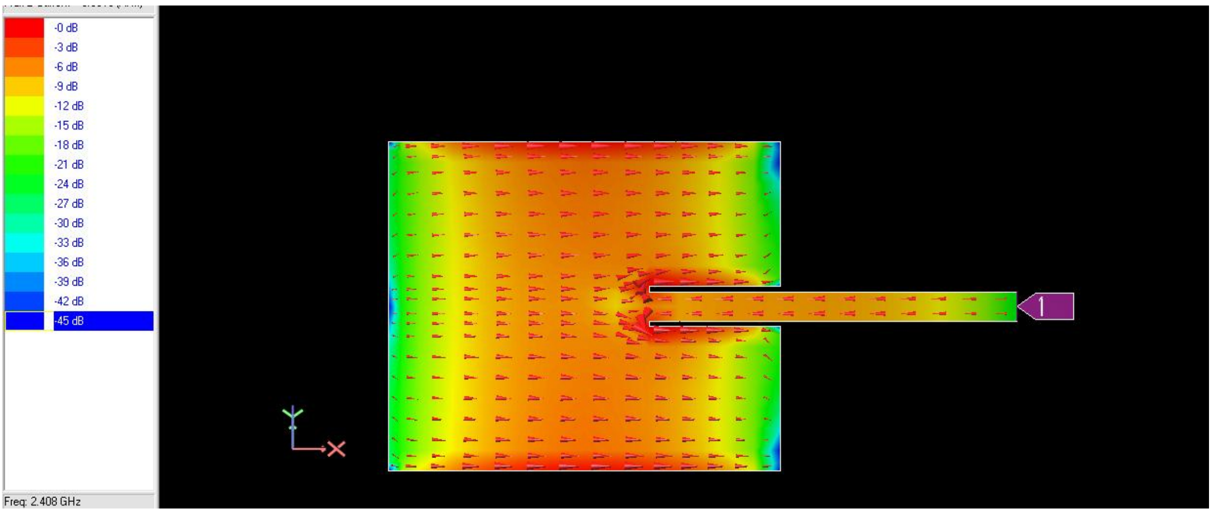


Figure 4.7: 3D Current distribution Plot

The 3D current distribution plot gives the relationship between the co-polarization (desired) and cross-polarization (undesired) components. Moreover it gives a clear picture as to the nature of polarization of the fields propagating through the patch antenna. Figure 4.7 clearly shows that the patch antenna is ***linearly polarized***.

4.3.4 Other Parameters

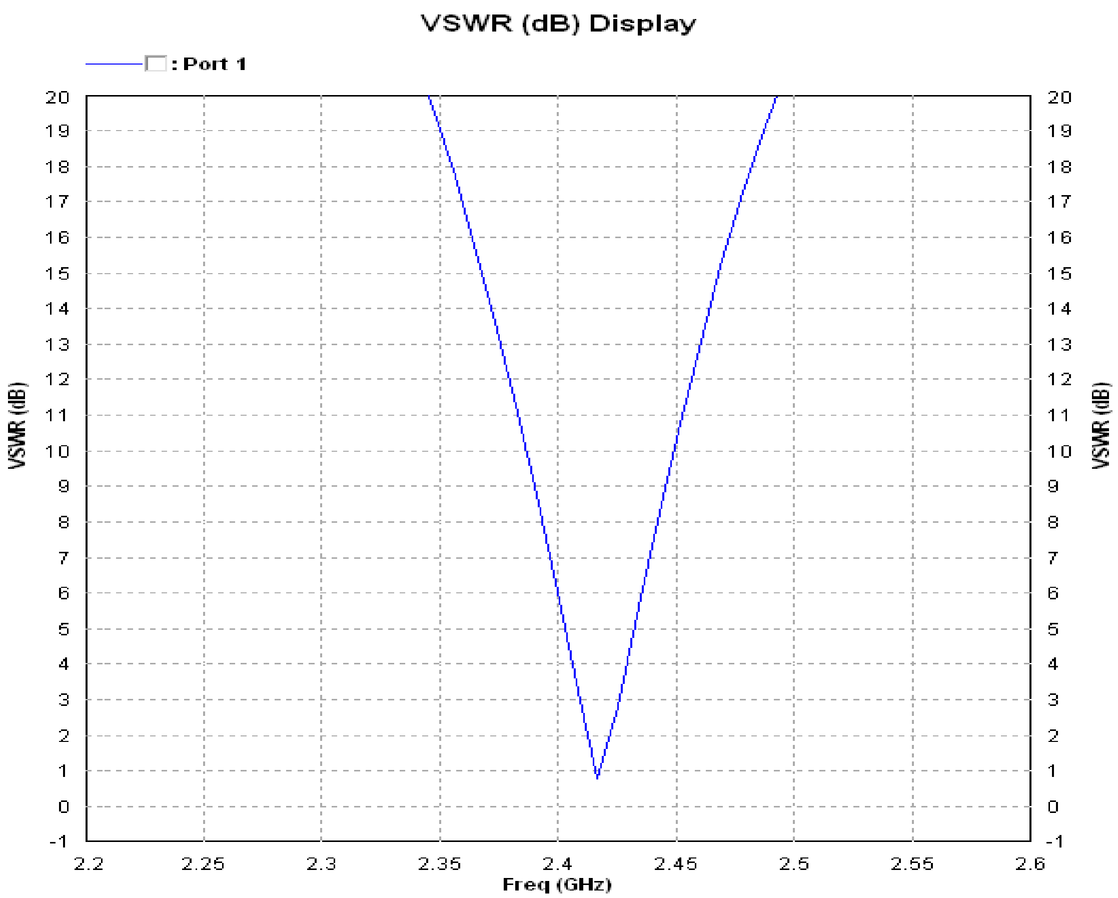


Figure 4.8: VSWR v/s frequency plot

The most common case for measuring and examining VSWR is when installing and tuning transmitting antennas. When a transmitter is connected to an antenna by a feed line, the impedance of the antenna and feed line must match exactly for maximum energy transfer from the feed line to the antenna to be possible. When an antenna and feed line do not have matching impedances, some of the electrical energy cannot be transferred from the feed line to the antenna. Energy not transferred to the antenna is reflected back towards the transmitter. It is the interaction of these reflected waves with forward waves which causes standing wave patterns.

Matching the impedance of the antenna to the impedance of the feed line is typically done using an antenna tuner. The tuner can be installed between the transmitter and the feed line, or between the feed line and the antenna. Both installation methods will allow the transmitter to operate at a low VSWR. Ideally, VSWR must lie in the range of 1-2 which is achieved in figure 4.8 for the frequency 2.408 GHz, near the operating frequency value.

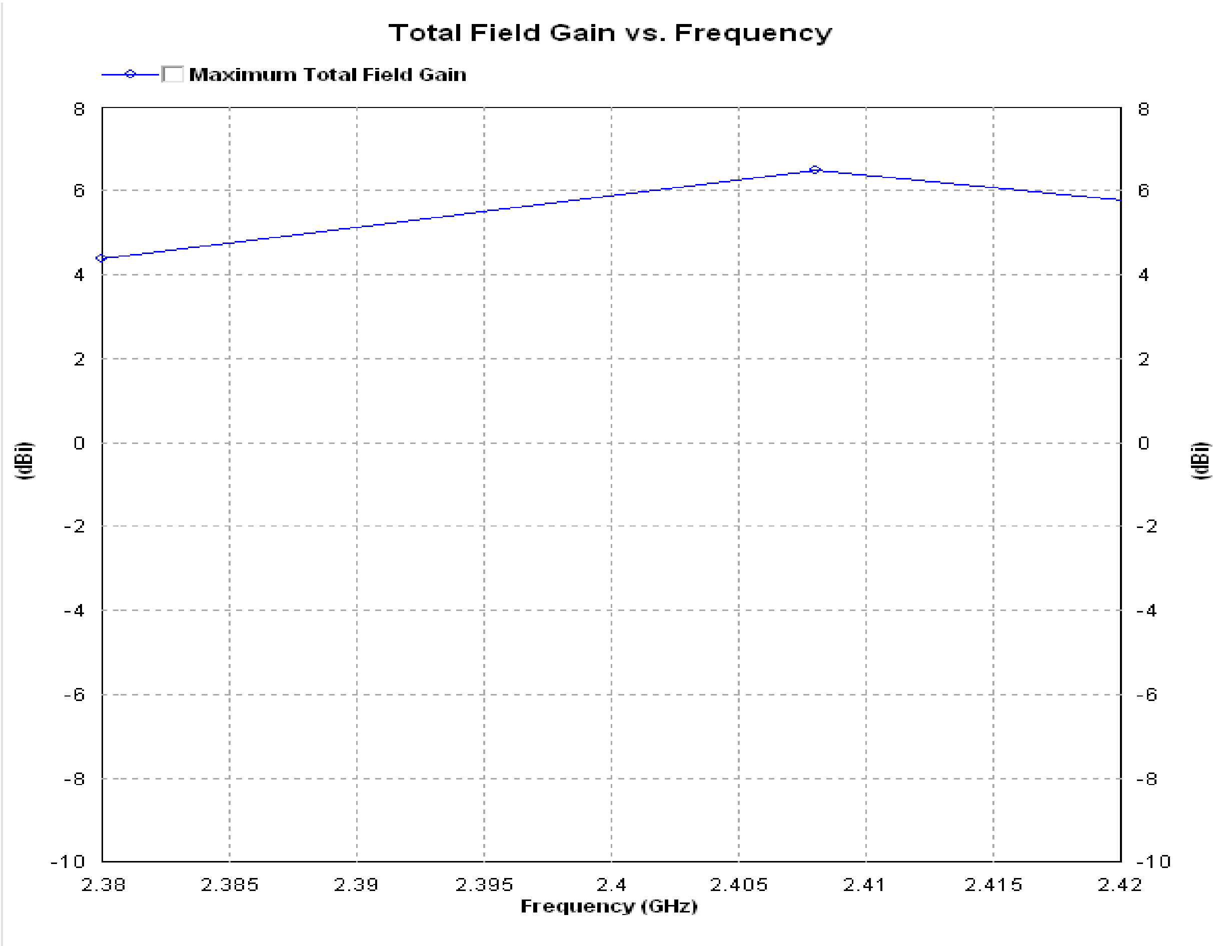


Figure 4.9: Gain v/s frequency plot

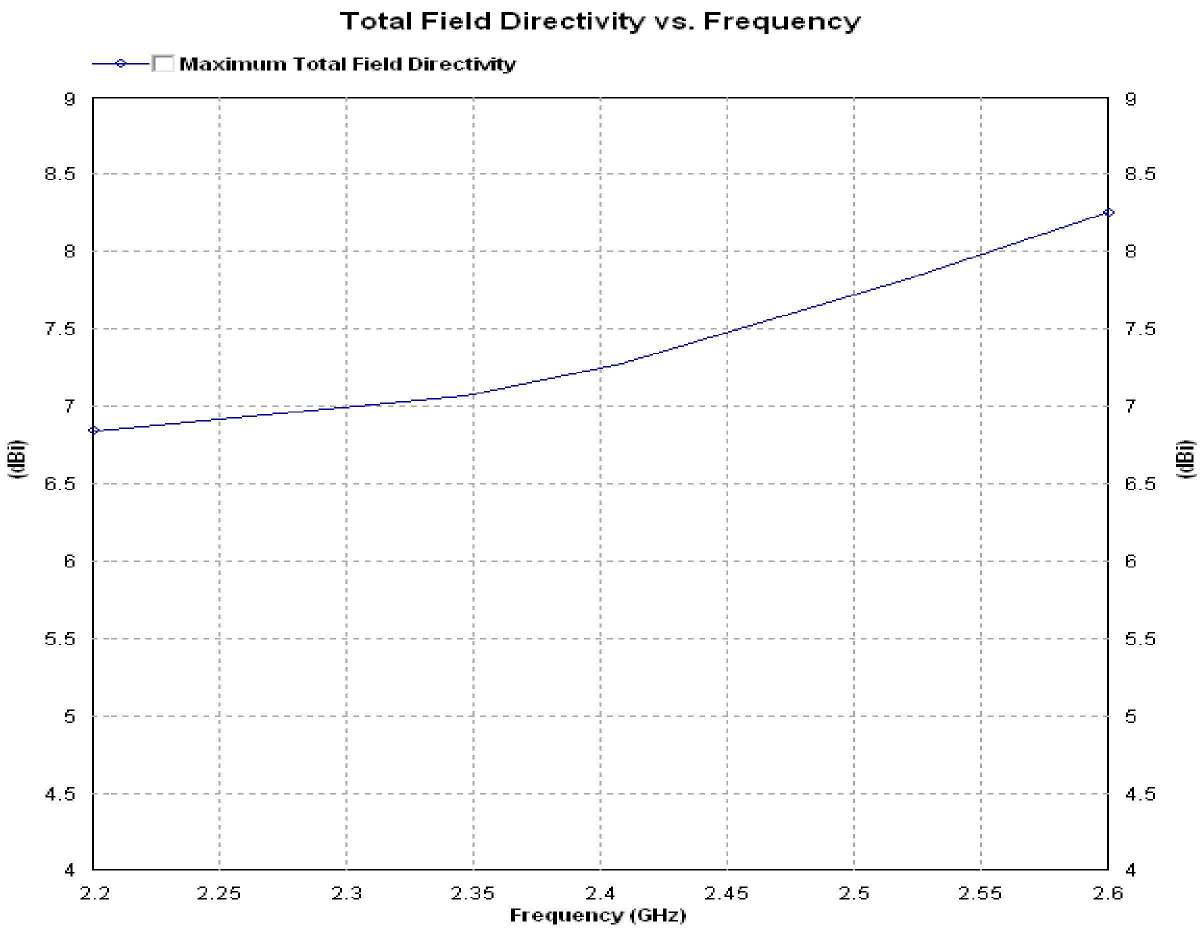


Figure 4.10: Directivity v/s frequency plot

4.4 PSO/IE3D Linkage

Design method by combining particle swarm optimization with IE3D is used to obtain the antenna parameters. IE3D is a full-wave, method-of-moments-based electromagnetic simulator solving the current distribution on 3D and multilayer structures of general shape. Different optimization schemes are available in IE3D, including Powell optimizer, genetic optimizer, and random optimizer. The variables for optimization defined by IE3D are controlled by its directions and bounds. However, the variables can only be connected with another by a fixed rate. More complicated relations between variables cannot be set in IE3D. Optimization with complicated variations, such as the optimization of the width and the horizon position for a vertical slot on a patch simultaneous, may cause a overlap problem in IE3D. In this article, design method by combining particle swarm optimization with IE3D is used to obtain the parameters of the linearly polarized Microstrip patch antenna. Particle swarm optimization is a robust stochastic evolutionary computation technique based on the movement and intelligence of swarm, which is very easy to understand and implement. It can be manipulated according to the following equations in

$$v_{id} = w \times v_{id} + c_1 \times \mathrm{rand}() \times (p_{id} - x_{id}) + c_2 \times \mathrm{rand}() \times (p_{gd} - x_{id}) \tag{1}$$

$$x_{id} = x_{id} + v_{id} \tag{2}$$

where x_{id} is the position of the ith particle along the dth dimension, v_{id} is the velocity of the ith particle along the dth dimension, p_{id} is the best position of the ith particle along the dth dimension, p_{gd} is the global best particle position along the dth dimension, c_1 and c_2 are two positive constants, rand() and Rand() are two random functions in the range (0, 1) and w is the inertia weight. PSO algorithm uses the fitness evaluation to represent how well a solution satisfies the design parameters. Each parameter used to evaluate the fitness is referred to as a fitness factor. The fitness factors must together quantify the result. A common means to do this is called the method of weighted aggregation (MWA). By combining PSO with IE3d, an efficient method for antenna design and optimization is presented. The flow chart of a typical PSO/IE3D is shown in Figure 4.11. The variables for optimization defined by IE3D are saved in a *.sim* file, and the simulated results of return loss are saved in a *.sp* file. By changing the variables saved in the *.sim* file using the PSO program, an optimization for complicated structure can be performed. Fitness function value is obtained by calculating the simulated results saved in the *.sp* file. A *.bat* file is used to call the PSO program and IE3D. Powerful fitness functions can be used in the PSO/IE3D method, which makes it more robust than the optimizers available in IE3D.

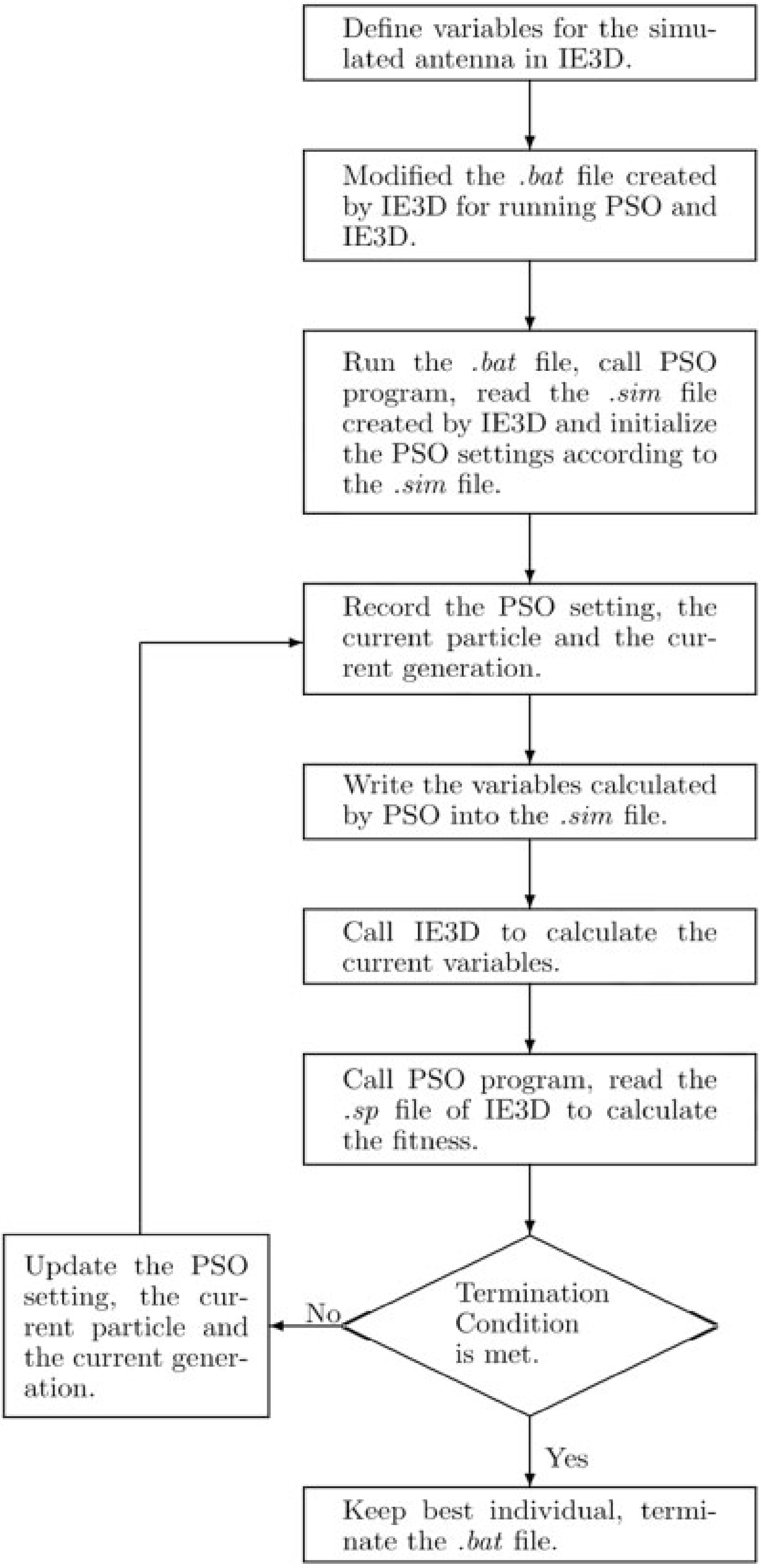


Figure 4.11: Flow chart illustrating the steps of a PSO/IE3D algorithm

4.4.1 Development Model

We use to pre-emptive parameters to check the veracity of the PSO matlab code, before we begin to optimize the parameters of the Microstrip patch antenna.

Case I: Resonant frequency:

We calculate the resonant frequency a Microstrip antenna, using its parameters like width (W), length (L), permittivity of the substrate (ε_r) and height (h) of the substrate.

We apply Particle Swarm Optimization technique for optimization of ΔL. The optimized ΔL is used for calculating the resonant frequency of rectangular Microstrip patch antenna.

Case II: Feed point calculation

The input impedance of rectangular Microstrip patch antenna is a vital parameter in deciding the amount of input power delivered to the antenna, thus, reducing the coupling effect of the RF signal to the nearly circuits. The calculation of an exact 50 ohms input impedance of a rectangular Microstrip patch antenna becomes extremely difficult when the antenna size is drastically small. In this paper, an attempt has been made to exploit the capability of PSO technique to calculate the input impedance by searching the feed point position. The feed point is calculated by using below equation which is optimized by PSO technique to get accurate results.

$$Rin\ (\ y = y\ 0\ ) = Rin\ (\ y = 0\ )\ \cos^4(\pi * y_0/L)$$

Using these two cases, we get an approximated optimization values as:

L = 39.4 mm

W = 46.9 mm

y_0 = 13.2 mm

We use the above values to verify the final optimization results generated by the PSO/IE3D matlab code.

Chapter 5

CONCLUSION AND FUTURE PROSPECTS

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CONCLUSION AND FUTURE PROSPECTS

The optimization of the Microstrip Patch is partially realized which concludes that the PSO code was functioning correctly. The future scope of work revolves around increasing the efficiency and decreasing the run time of the PSO code by using a distributive computing platform. Realization of results by the modified PSO would be concluded with the fabrication of the patch of the Microstrip Patch Antenna. The investigation has been limited mostly to theoretical study due to lack of distributive computing platform. Detailed experimental studies can be taken up at a later stage to find out a design procedure for balanced amplifying antennas.

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