ADDIS ABABA UNIVERSITY
SCHOOL OF GRADUATE STUDIES

Faculty of Technology
Electrical and Computer Engineering Department

Design of Phase Tunability of Conductor Backed Coplanar Waveguide (CPW) Patch Antenna.

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A Thesis Submitted to the School of Graduate studies of Addis Ababa University, in partial Fulfillment of the Requirements for the Degree of Master of Science in Electrical Engineering (communication engineering).

Advisor: Dr.Ing. Mohmmed Abdo

April, 2008

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FACULTY OF TECHNOLOGY

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Acknowledgements

A number of people have contributed to the successful completion of this research work. Many have given their direct effort to the preparation of this thesis work. I would like to render my gratitude to each one of them.

I would like to express my deepest gratitude to my advisor, Dr. Ing. Mohammed Abdo. His comprehensive and invaluable guidance has added a great deal to the substance of this thesis and taught me an enormous amount during the whole course of the research. In addition, I would like to express my appreciation and lots of thanks to Dr. D.P. Roy for his unreserved cooperation, encouragement and valuable suggestion.

Sincere thanks also go to my brother Ato Desta Arega for all his support and encouragement throughout these challenging years. Finally, I am very thankful to Ato Keros and Ato Sintayhu for their cooperation in using computer lab.

Table of Contents

Chapter One

Chapter Two

Chapter Three

Chapter Four

Chapter Five

Chapter Six

List of Figures

List of Abbreviations

CAD	Computer aided design
CPW	Coplanar wave guide
CPWG	Coplanar wave guide with ground
FCPW	Finite width coplanar waveguide
FDTD	Finite difference time domain
GPS	Global positing system
GUI	Graphic user interface
GP	Half power beam width
MICs	Microwave integrated circuits
MMICs	Monolithic microwave integrated circuits
PC	Personal computer
RF	Radio frequency
SLL	Side lobe level
TEM	Transverse electromagnetic.

Abstract

Traditionally beam steering in phase array antenna system is achieved using tunable phase shifter or switched feed network. Here, a tunable patch antenna element is suggested as a novel alternative to a combination of a radiating element and a phase shifter. A coplanar wave guide patch antenna is chosen due to its suitability to mount external tuning elements, using varactor diode.

Approximate analytical models for coplanar waveguide patch antennas with finite ground plane have been developed in the literature; such approach is limited to finite ground plane coplanar waveguide patch antenna only. Numerical result from approximate made were used as initial value in optimized simulation using math lab and Empire software tool. Coplanar waveguide patch antennas were constructed to re-tune to the desired frequency and tuning element was surface mounted.

The result obtained reveals a significant amount of phase shift with in the desired bandwidth at the center frequency of 1.04GHz. Alternative tuning with ferroelectric elements and distributed tuning is suggested as further research.

CHAPTER ONE

1.1 Introduction

In today's technological world, modern mobile telecommunication systems offer users more and more services every year, from cellular phones, personal digital assistants to global positioning system (GPS) applications. Today information must be accessible to millions of consumers any where in the world regardless of the environments. This causes available frequency bands to be rapidly crowded. Therefore, new frequency bands are allocated in the microwave region. In these frequency bands the atmospheric attenuation is very high. In order to solve the problems, we need to have effective radiation and reception of electromagnetic signals in all wireless applications, which depend on antenna effectiveness in terms of design characteristic. A good design of antenna can relax system requirements and improve system performance.

Effective antennas are impedance matching devices; they match the impedance of the circuit to that of the medium and there by providing an efficient means of radiating the signal into the medium or receiving the signal. The applications mentioned above, especially mobile applications require miniaturizations, imposing strict constraints on the size, gain and cost of the antenna while maintaining efficiency. To achieve the high efficiency and low cost of manufacturing, an attractive alternative is to implement the antenna on the printed circuit board, or ideally on the chip itself that manages the communication, thus lowering manufacturing cost.

Major advances in millimeter wave antennas have been made in recent years, including integrated antenna where active and passive circuits are combined with the radiating elements in one compact unit (monolithic form). A promising approach is to use a coplanar wave guide patch antenna, which effectively uses the surrounding ground plane to achieve better performance [13].

An attempt is made to design tunable antennas to replace the antenna element and phase shifter as a single tunable element and which widely used in phase array antenna, such as reception antennas on cell towers or any other multi user tracking applications. Such approach may be significant in leading to greater miniaturization.

1.2 Objective

Recently, remarkable progress has been achieved in monolithic microwave integrated circuits (MMICs). Hence in this thesis CPW is designed in such a way that it reduces cost and size.

Over all objective of the thesis is that to combine radiating element and a phase shifter as a single tunable patch antenna element. Specifically, a varactor diode is used as a tuning element.

1.3 Overview of the Thesis.

The thesis is organized as: Chapter one an introduction and objective the thesis. Chapter two gives basic concept associated with antenna theory and types. The important terms and fundamental characteristics of antennas are also included.

Chapter three introduces the coplanar wave guide and the advantages of using coplanar waveguide patch antenna and some of printed type transmission lines, their differences and similarity are also presented to give a complete visualization. Lastly, analysis of a coplanar waveguide patch antenna can be used as transmission line modeling. The finite difference time domain method, a method which solves Maxwell's equations as a function of time in discrete time steps at discrete a movie which can be viewed as a function of time. Coplanar patch antennas were designed using analytical results were used as initial values for further optimization using EMPIRE soft ware tool [Appendix A].Chapter four discuses tuning of antenna using varactor diode and compare with previous type of tuning system. In Chapter five, design specifications and simulation results are presented. Finally, chapter six conclusions of the thesis along with recommendations for future work are given.

CHAPTER TWO

Theoretical Development of Antenna and Its Parameters

One of humankind's greatest natural resources is the electromagnetic spectrum and the antenna has been instrumental in harnessing this resource. Its theoretical foundations were formulated based on electromagnetic waves.

2.1 Electromagnetic Wave theory

Time varying and magnetic field interact with each other a time variation of the electric field is always proportional to a spatial variation of the magnetic field, while a time variation of the magnetic field implies a spatial variations of the electric field. Most electromagnetic interactions were discovered in the first half of the nineteenth century by Ampere, Faraday, Gauss and other researches.

In a masterly fashion, James Clerk Maxwell (1831_1879) formulated a compact set of equation that summarizes previously discovered electromagnetic phenomena. More than a century after their introduction, Maxwell's equations still remain valid with several formulations that are continuously up dated to fit the contemporary mathematical notation.

Combining several of Maxwell's equations produces a wave equation: the propagation of electromagnetic waves results directly from electromagnetic interactions [1].

Solving wave equations in the rectangular system of coordinates yields plane waves.

It is useful to review Maxwell's equations, which govern the electric field $\overline{E}$ (volts/meter) and the magnetic field $\overline{H}$ (amperes /meter). Their related variables are the electric displacement $\overline{D} = \varepsilon\overline{E}$ (coulombs/meter$^{2)}$ and the magnetic flux density $\overline{B} = \mu\overline{H}$ (Teslas) respectively, where ε the permittivity of the medium (in vacuum is $\epsilon_{o=8.854\times10^{-12}}$) farad/meter, μ the permeability of the medium (in vacuum is $\mu o=4\pi\times10^{-7}$) henries/meter^2. One tesla is defined as one weber/meter^2=10^4gauss.

Maxwell's equations and their static limits are:

Maxwell's equations	static	Complex Maxwell's equations	
$\nabla \times \overline{E} = -\frac{\partial \overline{B}}{\partial t}$	$=0$	$\nabla \times \overline{E} = -j\omega \overline{B}$	(2.1.1)
$\nabla \times \overline{H} = \overline{J} + \frac{\partial \overline{D}}{\partial t}$	$= \overline{J}$	$\nabla \times \overline{H} = \overline{J} + j\omega \overline{D}$	(2.1.2)
$\nabla \bullet \overline{D} = \rho$	$= \rho$	$\nabla \bullet \overline{D} = \rho$	(2.1.3)
$\nabla \bullet \overline{B} = 0$	$= 0$	$\nabla \bullet \overline{B} = 0$	(2.1.4)

Where $\overline{J}$ is the current density (amperes/meter^2) and the ρ is the charge density (coulombs/meter^3). The dell operator is defined as:

$$\nabla \underline{\underline{\Delta}} \frac{\hat{x}\partial}{\partial x} + \frac{\hat{y}\partial}{\partial y} + \frac{\hat{z}\partial}{\partial z} \qquad (2.1.5)$$

Where $\hat{x}, \hat{y}, and\ \hat{z}$ are unit vectors in Cartesian coordinates.

2.2 Antenna Theory

The field of antennas is vigorous and dynamic, and over the last 50 years antenna technology has been an indispensable partner of the communications revolution. In wireless communication systems, the antenna is one of the critical components. An antenna can be viewed as a device that converts a guided electromagnetic wave on a transmission line to a plane wave propagating in free space. The official IEEE definition of an antenna follows this concept; "that part of a transmitting or receiving system that is designed to radiate or receive electromagnetic waves" [2]. Most antennas are reciprocal devices and behave the same on transmit as on receive mode. Antennas are treated as transmitting or receiving as appropriate for the particular situation. In the receiving mode, antennas act to collect incoming waves and direct them to a common feed point where a transmission line is attached. The antenna, like the eye, is a transformation device converting electromagnetic photons into circuit's currents; but, unlike the eye, the antenna can also convert energy from a circuit into photons radiated into space. Although any circuit can radiate if driven with a signal of high enough frequency, most practical antennas are specially designed to radiate efficiently at a particular frequency. An example of an inefficient antenna is the simple Hertzian dipole antenna, which radiates over wide range of frequencies and is useful for its small size. A more efficient variation of this is half wave dipole, which radiates with high efficiency when the signal wavelength is twice the electrical length of the antenna.

One of the goals of antenna design is to minimize the reactance of the device so that it appears as a resistive load. An "antenna inherent reactance" includes not only the distributed reactance of the active antenna but also the natural reactance due to its location and surroundings (as for example, the capacity relation inherent in the position of the active antenna relative to ground). Reactance diverts energy into the reactive field, which causes unwanted currents that heat the antenna and associated wiring, thereby wasting energy without contributing to the radiated output.

Reactance can be eliminated by operating the antenna at its resonant frequency, when their capacitive and inductive reactances are equal and opposite, resulting in a net zero reactive current. If this is not possible, compensating inductors or capacitors can instead be added to the antenna to cancel its reactance as far as the source is concerned. Once the reactance has been eliminated, what remains is a pure resistance, which is the sum of two parts: the ohmic resistance of the conductors, and the radiation resistance. Power absorbed by the ohmic resistance becomes waste as heat, and that absorbed by the radiation resistance becomes radiated electromagnetic energy. The greater the ratio of radiation resistance to ohmic resistance, the more efficient the antenna.

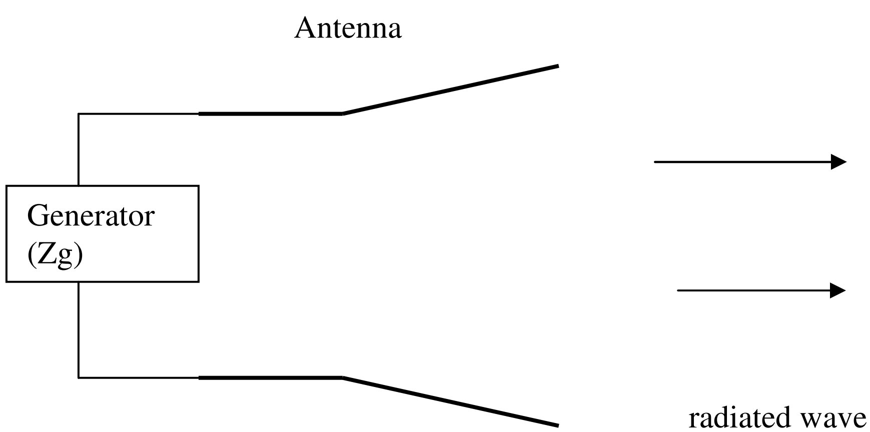


Fig.2.1 .1 Antenna in transmitted mode.

Input impedance is defined as "the impedance presented by an antenna at its terminals or the ratio of the voltage to current at a pair of terminals or the ratio of the appropriate components of the electric to magnetic fields at a point". In figure (2.1.1) black line indicates antenna and the ratio of the voltage to current at these terminals, with no load attached, defines the impedance of the antenna as:

$$Z_A = R_A + jX_A \tag{2.1.6}$$

Where Z_A = antenna impedance (ohms), R_A = antenna resistance (ohms)

X_A =antenna reactance (ohms)

In general the resistive part of (2.1.6) consists of two components; that is

$$R_A = R_r + R_L \qquad (2.1.7)$$

Where

R_r= radiation resistance of the antenna.

R_L= loss resistance of the antenna

Under ideal conditions, energy generated by the source should be totally transferred to the radiation resistance. However, in a practical system there are conduction-dielectric losses due to reflections (mismatch) losses at the interface between the line and the antenna.

2.2.1 Types of Antennas

Many of the major advances in antenna technology, which have been completed in the 1970s through the early 1990s; however, many more issues and challenges are facing us today, especially since the demands for system performances are even greater. Here, we introduce and briefly discuss some forms of the antenna types [1].

2.2.1.1 Wire Antennas

Wire antennas are familiar to the layman because they are seen virtually every where on automobiles, buildings, ships, aircraft, spacecraft, and so on. There are various shapes of wire antennas such as a straight wire (dipole), loop, and helix. They may take the form of a rectangle, square, ellipse, or any other configuration. The circular loop is the most common because of its simplicity in construction.

2.2.1.2 Aperture Antennas

Aperture antennas may be more familiar to the layman today than in the past because of the increasing demand for more sophisticated forms of antennas and the utilization of higher frequencies. Antennas of this type are very useful for aircraft and spacecraft applications, because they can be very conveniently flush-mounted on the skin of the aircraft or spacecraft. In addition, they can be covered with a dielectric material to protect them from hazardous conditions of the environment.

2.2.1.3 Patch antennas

Patch antennas became very popular in the 1970s primarily for space borne applications. Today they are used for government and commercial applications. These antennas consist of a metallic patch on a grounded substrate. The metallic patch can take many different configurations, however, the rectangular and circular patches, are the most popular because of ease of analysis and fabrication, and their attractive radiation characteristics, especially low cross-polarization radiation. Our main focus lies on the printed type of antenna.

2.2.1.4 Array Antennas

Many applications require radiation characteristics that may not be achievable by a single element. It may, however, be possible that an aggregate of radiating element in electrical and geometrical arrangement (an array) will result in the desired radiation characteristics. The arrangement of the array may be such that the radiation from the elements adds up to give a radiation maximum in particular direction or directions, minimum in others or otherwise as desired.

2.2.1.5 Reflector Antennas

The success in the exploration of outer space has resulted in the advancement of antenna theory. Because of the need to communicate over great distances, sophisticated forms of antennas had to be used in order to transmit and receive signals that had to travel millions of miles. A very common antenna form for such an application is a parabolic reflector. Antennas of this type have been built with diameters as large as 305m. Such large dimensions are needed to achieve the high gain required to transmit or receive signals after millions of miles of travel.

2.2.1.6 Lens Antennas

Lenses are primarily used to collimate incident divergent energy to prevent it from spreading in undesired directions. By properly shaping the geometrical configuration and choosing the appropriate material of the lenses, they can transform various forms of divergent energy into plane waves. They can be used in most of the same applications as are the parabolic reflectors, especially at higher frequencies. Their dimensions and weight become exceedingly large at lower frequencies. Lens antennas are classified according to the material from which they are constructed, or according to their geometrical shape.

2:3 Basic Antenna Properties

There are several critical parameters that affect an antennas performance and can be adjusted during the design process. These are: resonant frequencies, impedance, gain, radiation pattern, polarization, efficiency, return loss, reflection coefficient and band width etc. Transmitting antenna may also have a maximum power rating, and receive antennas differ in their nose rejection properties [for detail see Appendix C].

2.3.1 Resonant Frequency

It is related to the electrical length of the antenna. The electrical length is usually the physical length of the wire divided by its velocity factor that is the ratio of the speed of wave propagation in the wire to the speed of light in a vacuum. Typically an antenna is tuned for a specific frequency, and is effective for a range of frequencies usually centered on the resonant frequency. However, the other properties of the antenna (Especially radiation pattern and impedance) change with frequency, so the antennas resonant frequency may merely be close to the center frequency of these other more important properties.

Antennas can be made resonant on harmonic frequencies with lengths that are fractions of the target wave length. Some antenna designs have multiple resonant frequencies, and some are relatively effective over a very broad range of frequencies. The most commonly known type of wide band aerial is the logarithmic or log periodic, but its gain is usually much lower than that of a specific or narrower band aerial.

2.3.2 Gains

Gain as a parameter measures the directionality of a given antenna. An antenna with a low gain emits radiation in all directions equally, where as a high gain antenna will preferentially radiate in particular directions.

Specifically, the gain, directive gain or power gain of an antenna defined as the ratio of the intensity (power per unit surface) radiated by the antenna in a given direction at an arbitrary distance divided by the intensity radiated at the same distance by a hypothetical isotropic antenna.

The gain of an antenna is not added by the antenna is a passive phenomenon power is not added by antenna , but simply redistributed to provided more radiated power in a certain direction than would be antenna. If an antenna has a greater than one gain in some directions, its must have a less than one gain in other directions.

Since energy is conserved by the antenna. An antenna designer must take into account the applications for the antenna when determining the gain. High gain antennas have the advantage of longer range and better signal quality, but must be aimed carefully in a particular direction.

A more common characterization is antenna gain G (f, θ,ϕ), which is the ratio of the power actually transmitted in the direction, θ,ϕ to that which would be radiated if the entire available transmitter power P_T were radiated isotropic ally. That is:

$$G(f,\theta,\phi) \underline{\underline{\Delta}} \eta_R D(f,\theta,\phi) = \frac{P(f,\theta,\phi)}{P_T / 4\pi} \qquad (2.1.8)$$

The ability of antenna to radiate energy in a desired direction is characterized by its antenna directivity, D (f, θ,ϕ), which is the ratio of power actually transmitted in particular direction to that which would be transmitted the power P_{TR} isotropic ally. Therefore, directivity is some times called “directivity over isotropic”

Mathematically is defined as:

$$D(f,\theta,\phi) \underline{\underline{\Delta}} \frac{P(f,\theta,\phi)}{P_T / 4\pi} \qquad (2.1.9)$$

Where P (**f,** θ,ϕ), watt/steradian, integrated over **4π** steradians equals $\mathbf{P_{TR}}$**.**

Therefore, $$\int_{4\pi} D d\Omega = 4\pi \qquad (2.1.10)$$

2.3.3 Bandwidths

The band width of an antenna is the range of frequencies over which it is effective, usually centered on the resonant frequency. The bandwidth of an antenna may be increased by several techniques, including using thick wires, replacing wires with cages to simulate a thicker wire, tapering antenna components and combining multiple antennas into a single assembly and allowing the natural impedance to select the correct antenna, small antennas are usually preferred for convince.

2.3.4 Impedance

As an electromagnetic waves travels through the different parts of antenna of antenna system, it may encounter differences in impedance (E/H, V/I, etc,).

An each interface, depending on the impedance match, some fraction of the wave's energy will reflect back to the source, forming a stand wave in the feed line.

Minimizing impedance differences at each interface (impedance matching) will reduce SWR and maximize power transfer through each part of antenna system.

2.3.4.1 Impedance Matching

Based on the material and physical dimensions, every antenna has its own characteristic impedance. However, this does not always match the impedance of the transmitters, receivers and transmission lines. This difference produces an undesired reflection of the power when an antenna is feeding. Matching networks are used to match the impedance between the antenna and connected transmission line in order to minimize the power reflection. Factors that should be considered in a matching network design include: complexity, bandwidth, implementation and adjustability. There are different kinds of matching networks and the most common are: L-networks, single-stub tuning (shunt and series), double-stub, quarter-wave transformers, binomial multi-section transformers, Chebyshev multi-section transformers, and tapered lines. Details of how are the performance of these matching networks can be implemented are described by [4]

2.3.4.2 Mutual Impedance

The previous section describes antenna input impedance. It is very important to extend this analysis to include self and mutual impedance of the antenna array elements. When an antenna is in the presence of an obstacle or other element, the current distribution, the field radiation, is altered. As a consequence of this interaction, the input impedance of the antenna varies. The interaction between elements is referred to as driving-point impedance and is the combination of the self-impedance and the mutual impedance between the driven element and the other obstacles or elements. The driving-point impedance depends upon the antenna type, the relative placement of the elements, and the type of feed used to excite the elements.

2.3.4.3 Mutual Coupling

Whether two or more antennas, near each other, are active (transmitting) or passive (receiving), some of the energy of each one will be induced in the others. In general, the amount of energy induced in an antenna array will depend on four factors: the radiation characteristics of each, the relative separation between them, the relative orientation of each, and the scan volume of the array. The effect of this induced energy over a passive element when the antennas in the array are transmitting is called coupling. The coupling effect is the relative change of the driving impedance of each element, and it is usually called mutual impedance driving [3]. For better understanding, we have adopted the following terminology :

Antenna impedance: The impedance looking into a single insolated element. ***Passive driving impedance***: The impedance looking into a single element of an array. With all other elements of the array passively terminated in their normal generator impedance.

Active driving impedance: The impedance looking into a single element of an array with all other elements excited. While *passive driving impedance* has a minor influence in the antenna impedance, we will assume that driving impedance would refer to *active driving impedance*. The impedance of an antenna can be matched to the feed line and radio by adjusting the impedance of antenna of feed line, using the feed line as an impedance transformer.

2.3.5 Radiations pattern

An antenna radiation pattern or antenna pattern is defined as " a mathematical function or a graphical representation of radiation properties of the antenna as a function of space coordinates. In most cases, the radiation pattern is determined in the far field region and is represented as a function of the directional coordinates.Radiation properties includes power flux density, radiation intensity, field strength, directivity phase or polarization." The radiation property of most concern is the two or three dimensional spatial distribution of radiated energy as a function of the observer position along a path or surface of constant radius.

A trace of the received power at a constant radius is called the power pattern. On the other hand, a graph of the spatial variation of the electric (or magnetic) field along a constant radius is called an amplitude field pattern.The radiated power per unit surface is proportional to the squared electrical field of the electromagnetic wave.

The radiation pattern is the locus of points with the same electrical field. In this representation, the reference is usually the best angle of emission. It is also possible to depict the directive gain of the antenna as a function of the direction. Often the gain is given in decibels.

The graphs can be drawn using Cartesian (rectangular) coordinates. The shape of curves can be very different in Cartesian or polar coordinates and with the choice of the limits of the logarithmic scale.

2.3.6 Antenna Efficiency

"Efficiency" is the ratio of power actually radiated to the power put into the antenna terminals. Radiation in an antenna is caused by radiation resistance which can only be measured as part of total resistance including loss resistance. Loss resistance usually results in heat generation rather than radiation, and reduces efficiency. Mathematically, efficiency is calculated as radiation resistance divided by total resistance or the radiation efficiency "η" is the ratio of radiated power to available power as:

$$\eta \underline{\underline{\Delta}} P_{TR} \;/\; P_T \;\leq 1 \qquad (2.1.11)$$

Typically is close to unity for most antennas.

The antenna efficiency e_o is used to take into account losses at the input terminals and within the structure of antenna. Such losses may be due to

1. Reflection because of the mismatch between the transmission line and the antenna.
2. I^2R losses due to conduction and dielectric.

In general, the overall efficiency can be written as:

$$e_o = e_r \, e_c \, e_d \qquad (2.1.12)$$

Where e_o =total efficiency (dimensionless), e_r = reflection (mismatch) efficiency= $(1-|\Gamma|^2)$, e_c = conduction efficiency (dimensionless) and e_d = dielectric efficiency (dimensionless).

Reflections occur as a result of discontinuities, such as an imperfection in an uniform transmission line, or when a transmission line is terminated with other than its characteristic impedance. The reflection coefficient Γ is defined as :

$$\Gamma = \left(\frac{Z_{in} - Z_0}{Z_{in} + Z_0} \right) \tag{2.1.13}$$

Where, Z_{in} is impedance towards the source and Z_o is characteristic impedance of transmission line.

Usually e_c and e_d are very difficult to compute, but they can be determined experimentally.

The magnitude of reflection coefficient Γ always fall in the range [0 1].

2.3.7 VSWR and Return loss

VSWR, which stands for voltage standing wave ratio, also an indicator of reflected waves bouncing back and forth within the transmission line. The SWR can also be defined as the ratio of the maximum amplitude of the electric field strength to its minimum amplitude, i.e. E_{max} / E_{min}. And such, an increase in VSWR corresponds to an increase in dielectric losses. A mismatched antenna reflects some of the incident power back toward the transmitter and since this reflected wave is traveling in the opposite direction as the incident wave, there will be some points along the cable where the two waves are in phase and other points where the waves are out of phase (assuming a sufficiently long cable). If one could attach an RF voltmeter at these two points, the two voltages could be measured and their ratio would be the SWR.

The VSWR can also be defined as interims of reflection coefficient:

$$VSWR = \left| \frac{1+\Gamma}{1-\Gamma} \right| \qquad (2.1.14)$$

The VSWR is always ≥ 1.

Return loss: is a measure of power reflected from imperfections in an electrical or optical communications link. The return loss value describes the reduction in the amplitude of the reflected energy, as compared to the forward energy. In electrical systems, return losses often occur at junctions between transmission line and terminating impedance. It is a measure of the dissimilarity between impedances in metallic transmission lines and loads. For devices that are not perfect transmission lines or purely resistive loads, the return loss value varies with the frequency of the transmitted signal. It will always be a loss, and therefore a negative dB. However one can write -3 dB as simply 3 dB of loss, dropping the negative sign and adding loss. The return loss is simply the amount of power that is "lost" to the load and does not return as a reflection. For example, if a device has 15 dB of return loss, the reflected energy from that device, P_R, is always 15 dB lower than the transmitted energy P_T. Clearly, high return loss is usually desired even though "loss" has negative connotations. Return loss is commonly expressed in decibels.

$$RL(dB) = -20\log[\Gamma] \qquad (2.1.15)$$

2.3.8 Grounds

At frequencies used in antennas, the ground behaves mainly as a dielectric. The conductivity of ground at these frequencies is negligible. When an electromagnetic wave arrives at the surface of an object, two waves are created: one enters the dielectric and the other is reflected.

If the object is a conductor, the transmitted wave is negligible and the reflected wave has almost the same amplitude as the incident one. When the object is a dielectric, the fraction reflected depends (among others things) on the angle of incidence. When the angle of incidence is small (that is, the wave arrives almost perpendicularly) most of the energy traverses the surface and very little is reflected. Most of the electromagnetic waves emitted by an antenna to the ground below the antenna at moderate (say < 60°) angles of incidence enter the earth and are absorbed (lost). But waves emitted to the ground at grazing angles, far from the antenna, are almost totally reflected. At grazing angles, the ground behaves as a mirror. Quality of reflection depends on the nature of the surface. When the irregularities of the surface are smaller, then the wavelength reflection is good.

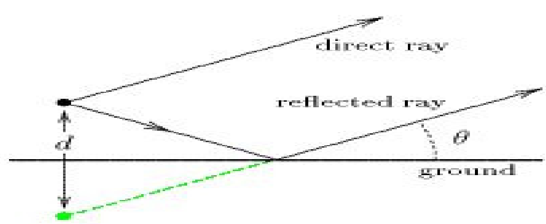


Figure 2.3.8 the wave reflected by earth can be considered as emitted by the image antenna.

2.3.9 Dielectric Substrates

The dielectric substrate is the mechanical back bone of the coplanar waveguide patch. It provides a stable support for the conductor strip and patches that make up connection lines, resonators, and antennas. It ensures that the components that are implemented are properly located and firmly held in place; just as in printed circuits for electronics at low frequencies. The substrate also fulfills an electrical function by concentrating the electromagnetic fields and preventing unwanted radiation in circuits.

The dielectric is an integral part of the connections transmission lines and deposited components: its permittivity and thickness determine the electrical characteristics of the circuit or of the antenna.

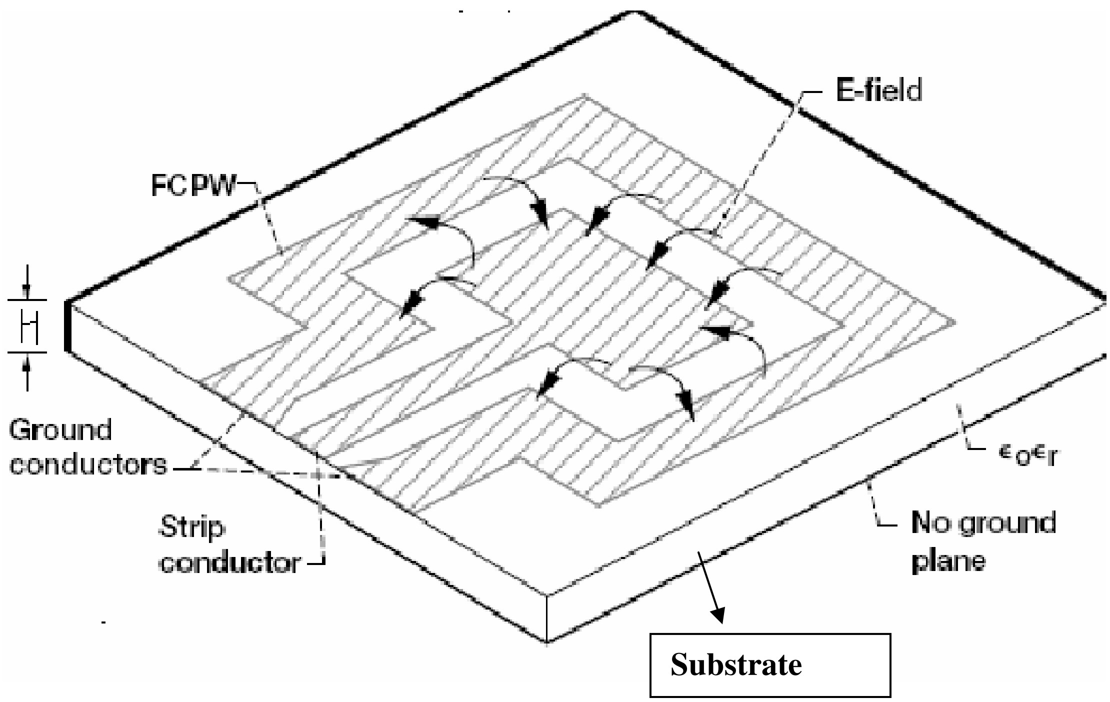


Figure 2.3.9 3d view of CPW patch antenna.

CHAPTER THREE

3.1 coplanar waveguide patch antennas

A coplanar waveguide (CPW) is one type of printed antenna and consists of a strip of thin metallic film deposited on the surface of a dielectric slab with ground electrodes running adjacent and parallel to the strip on the same surface, as shown in Figure 3.1. below.

The coplanar patch antennas share the same advantages of the conventional microstrip antennas, such as: low profile, light weight, easy manufacturing, low cost and easily conform to tight spaces. They also offer many advantages compared to conventional microstrip antennas, these include: higher polarization purity, lower cross talk between adjacent antennas, easy integration to microwave integrated circuits, and broader band width [14, 15].

In addition, a ground on the sample plane as the patch element offers the ability of easy integration of additional external lumped components in the design. More complexity in these antennas performance, such as phase shifting and attenuation can be eliminated through added tunability. Although the band width offered is slightly higher than microstrip antennas, the coplanar waveguide antenna is still considered narrow band antenna, which is suited for limited applications only. Applications where this type of designs might find an inch are GPS, biomedical, bacons, satellite, chip to chip signaling and even a full identification system on a chip.

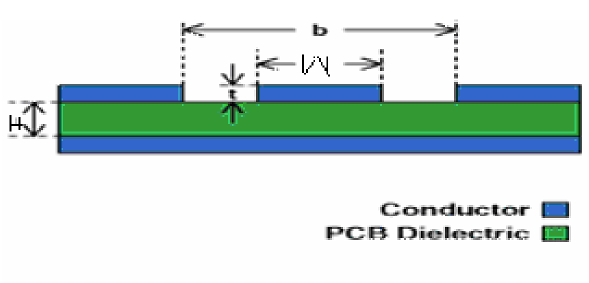


Fig 3.1 coplanar wave guide with backed conductor

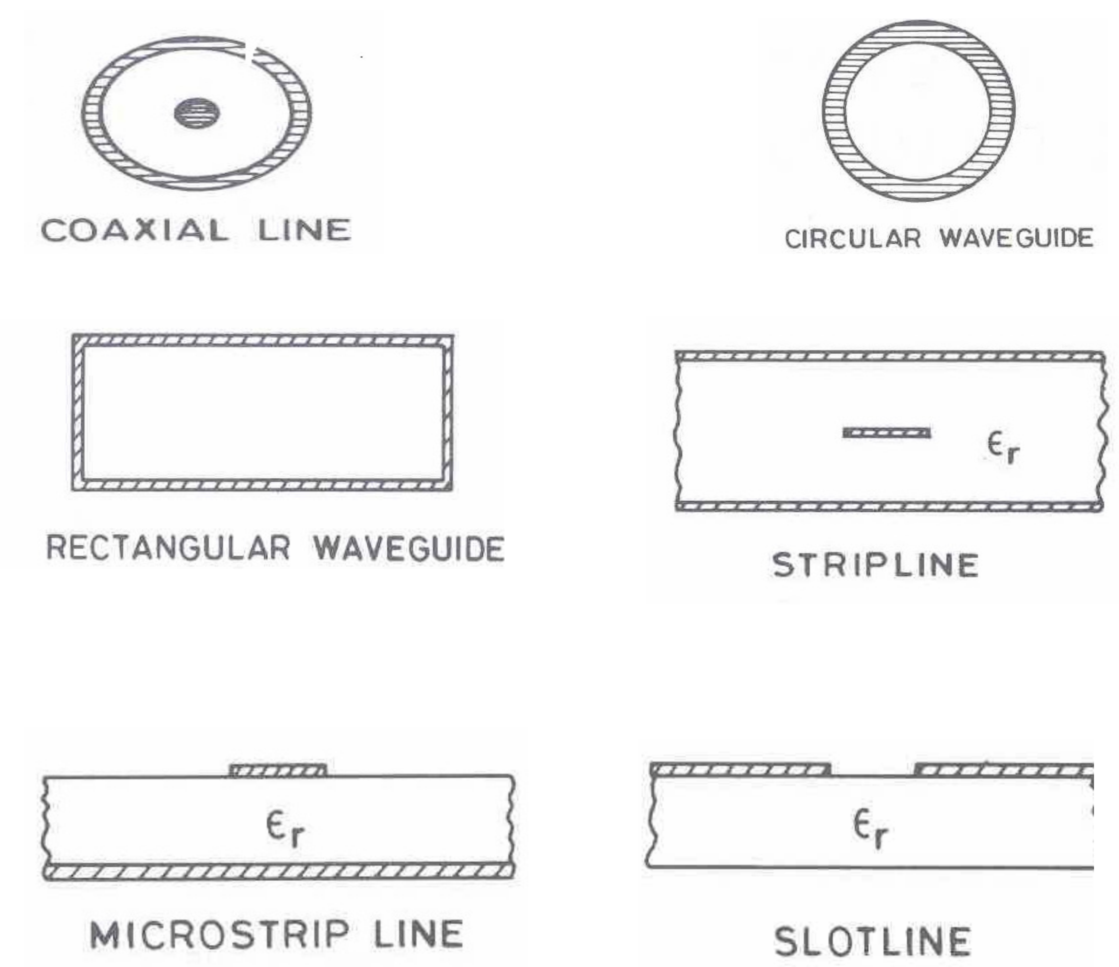
3.2 Finite Coplanar Waveguide

A finite width coplanar waveguide (FCPW) on a dielectric substrate consists of a center strip conductor separated from finite ground planes by narrow slots on either side. The substrate thickness, the strip width, slot width plus the ground plane width are denoted as H, W, and S is (b-W)/2, respectively as shown above in figure 3.1. Thus, finite coplanar waveguide patch antennas are easily adaptable to shunt are easily adaptable to shunt element connections without the need to penetrate the dielectric substrate as in case of microstrip lines [24]. There are some papers on analytical solution of infinite waveguide, which do give expression meaningful for design [14]-[20]. The infinite top ground plane approximation is not valid because the antennas studied in this thesis have the ground plane smaller than the main conductor. The approximation suggested here, is based on the use of three coupled transmission lines. The equations of coupled microstrip lines are readily available from numerous sources and are established [16]-[24]. In conjunction with the above, coupled line approach, transmission line theory is used to model the effects of radiation and thus give a more comprehensive model for the design of coplanar waveguide patch antennas which will allow the insertion of lumped components.

3.3 Planar Transmission line

Microwave integrated circuits offer system engineers prospects of small, batch processed modules radar and communication systems. In the past, radar and communication systems included a variety of non reciprocal magnetic devices. These devices require circularly polarized RF magnetic fields for their operation, and present day microstrip and strip lines do not provide such fields. In addition, the ground plane of these lines, which is located on the opposite side of a dielectric substrate, is not easily accessible devices. Direct dependence of the characteristic impedance on the thickness of the substrate makes it difficult to use a low loss, high dielectric constant material, such as a temperature compensated ceramic.

This is a definite drawback for low frequency operation where size consideration dominates. All these disadvantages may either be overcome or alleviated by a novel integrated circuit transmission line configuration in which all conducting elements, including the ground planes, are on the same side of a dielectric substrate. This is called the coplanar wave guide, a surface strip transmission line. An extensive variety of transmission structures are used at microwave frequencies. These include coaxial line, waveguide, strip line, microstrip line, coplanar waveguide, coplanar strips, slot line, coupled strip lines, coupled microstrip lines and others, see figure below.



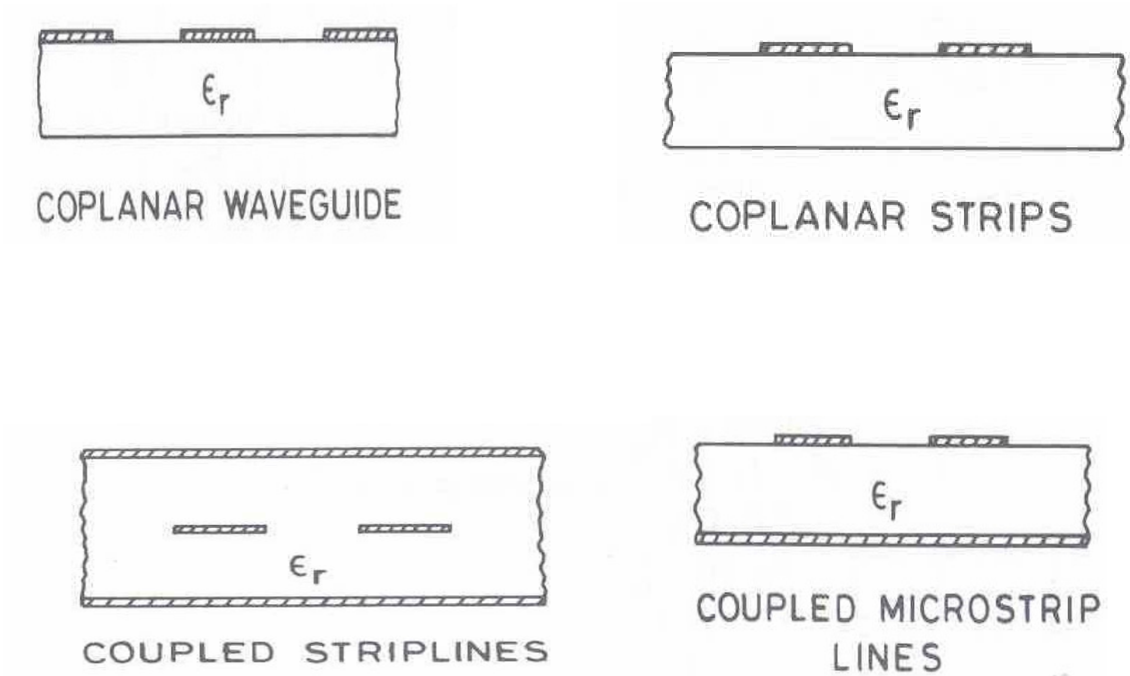


Figure 3.1.1 Transmission structures commonly used in microwave circuits

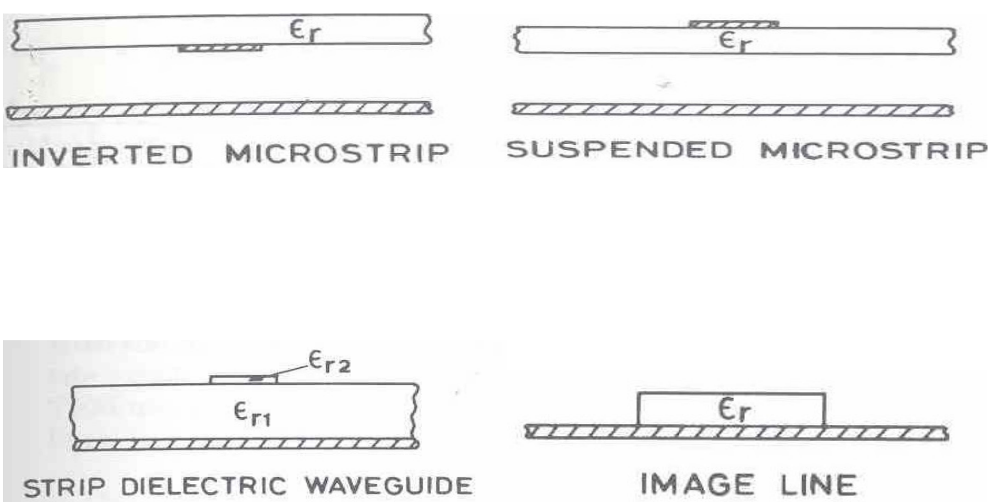


Figure 3.1.2 Other transmission structures for microwave circuits

3.4 Common Types of Printed Circuit Transmission lines.

Microwave signals are transmitted from location to location by waveguide or antennas. The chief difference between the two is that a waveguide confines the electromagnetic field to an area along the path, while an antenna radiates the electromagnetic field into space. A transmission line is sub category of waveguide that uses some physical configuration of metal and /or dielectrics to direct a signal along the path. For analysis, we want to predict two important parameters of a transmission line: Impedance and the electrical length. The most common transmission lines are: strip line, micro strip line, and coplanar waveguide.

3.4.2 Strip Line

Strip (shown in figure below) is one of the most commonly used transmission lines at microwave frequencies. The analysis of strip line is considerably simplified when the thickness "t" of the central strip line is negligible. Strip line was invented by flattening the coaxial line configuration. The presence of top and bottom shields provides good isolation from other signals on the printed circuit board.

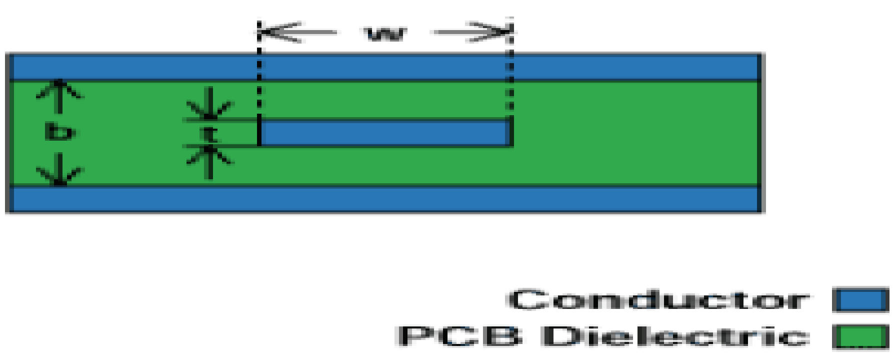


Fig.3.3. Centered strip line

3.4.3 Microstrip Line

The microstrip line simplifies the strip line by removing the upper ground planes. It is probably the most popular planar transmission line because of its ease of fabrication and the ready availability of the signals for probing and circuit connections. Its disadvantage over strip line is that some of the energy transmitted may be coupled into space or adjacent traces.

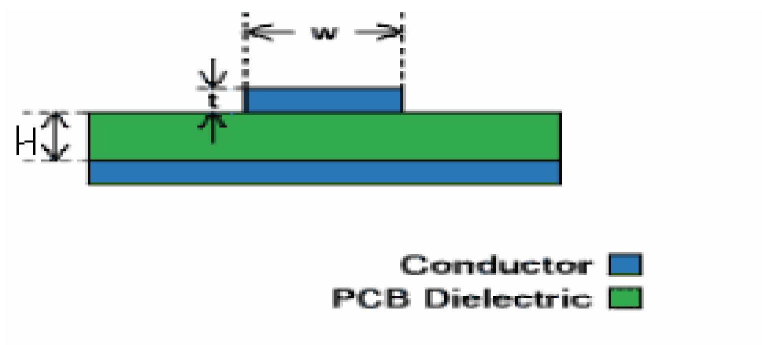


Fig.3.4 microstrip line

3.4.4 Coplanar Waveguide.

All of the transmission lines so far scale the signal conductor’s dimensions to the separation of the ground conductor to set the impedance. Once we choose the dielectric thickness, the other dimensions are determined. Sometimes this can be inconvenient when connecting to other components. When we need transition to a connector or device pin that has a specific pin size, the trace width may turn out to be too wide to fit between pins.

We can taper the dielectric substrate thickness to allow the signal conductor's width to narrow but that is usually impractical. Fortunately, the coplanar waveguide family solves this problem.

Coplanar waveguide (CPW) uses a ground conductor that is coplanar with signal conductor. The structure dimensions are almost independent of substrate thickness and therefore, the impedance is controlled by the signal line width and the ground gap.

What this mean is that we can keep the impedance constant as we taper the signal conductor's width down to meet a pin.

This is perfect for matching to a component pin width changing the substrate thickness. The coplanar waveguide configurations that are most likely to be useful are: coplanar waveguide and coplanar waveguide with ground. Coplanar waveguide concentrates the field in the gap so it will have the best ability to taper in to a pin. However, it may require special attention if you are transitioning from microstrip line into CPW. In this situation, coplanar waveguide with ground (CPWG) may be easier to deal with since you can start with a wide gap (microstrip line) and gradually transition to the coplanar waveguide with ground configuration.

Therefore, we do not have to limit our designs to microstrip lines and strip lines. Transmission lines, such as coaxial line, shielded microstrip line, and the coplanar waveguide family, can help designers solve real world layout problems while maintaining tight control of the waveguide characteristic impedance and physical dimensions [9].

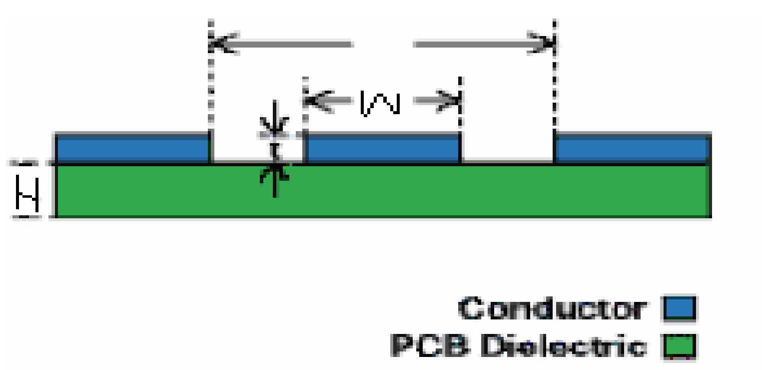


Fig.3.5 coplanar waveguide

3.4.4.1 Waves on CPW

The mechanism of transmission and radiation in CPW can be understood by considering a point current source (Hertz dipole) located on top of the ground dielectric substrate. This source radiates electromagnetic waves.

Depending on the direction toward which waves are transmitted, they fall with in two distinct categories, each of which exhibits quite different behaviors.

3.4.4.1A. Space Waves

Waves transmitted upward, which elevation angles θ between 0 and π/2, move towards free open space above, where they do not find any further interfaces.

These waves thus radiated, with field amplitude decreasing with distance (r) as1/r space waves are essential for all antenna applications, which specifically rely on radiation for their operation. In transmission lines and circuits, on the other hand, they produce spurious leakage and, therefore, are undesirable.

3.4.4.1B. Surface Waves

The waves transmitted slightly down word, having elevation angle, between π/2 and $\pi \text{ arcsin } \left(\frac{1}{\sqrt{\varepsilon_r}}\right)$, meet the ground plane, which reflects them, and then meet the dielectric-to-air boundary, which also reflects them (total reflection condition).

Field amplitudes build up for some particular incident angles, leading to the excitation of surface wave modes in a metallic waveguide. The field remains mostly trapped with in the dielectric, decaying exponentially above the surface waves spread out cylindrical fashion around the excitation point, with field amplitude decreasing with (r), say 1/√r, more slowly than space waves. The same guiding mechanism provides propagation with in optical fibers. The signal's amplitude is thus reduced, contributing to an apparent attenuation or a decrease in antenna efficiency.

3.5 Methods of Analysis

There are many methods of analysis for coplanar waveguides patch antennas. The most popular models are the transmission line, cavity, and full-wave (which include primarily integral equations/moment method). The transmission line model is the easiest of all, it gives good physical insight, but is less accurate and it is more difficult to model coupling. Compared to the transmission line model, the cavity model is more accurate but at the same time more complex. However, it also gives good physical insight and is rather difficult to model coupling, although it has been used successful. In general when applied properly, the full wave models are very accurate, very versatile, and can treat single elements, stacked elements, arbitrary shaped elements, and coupling. However they are the most complex models and usually give less physical insight.

3.5.1Transmission Line Model

It was indicated earlier that the transmission line model is the easiest of all but it yields the least accurate results and it lacks the versatility. However, it does shed some physical insight.

3.5.1.1 Fringing Effect

Because the dimensions of the patch are finite along the length and width. The fields at edges of the path undergo fringing, the fringing fields acts to extend the effective length of the patch thus, the length of a half-wave patch is slightly less than a half wave length in the dielectric substrate material. The amount of fringing is a function of the dimensions of the patch and the height of the substrate. For the principal E-plane (x-y plane) fringing is a function of the ratio of the length of the patch, L to the height, h of the substrate (L/h) and the dielectric constant ϵr of the substrate.

Since for coplanar patch antennas L/h»1, fringing is reduced, however, it must be taken into account because it influences the resonant frequency of the antenna.

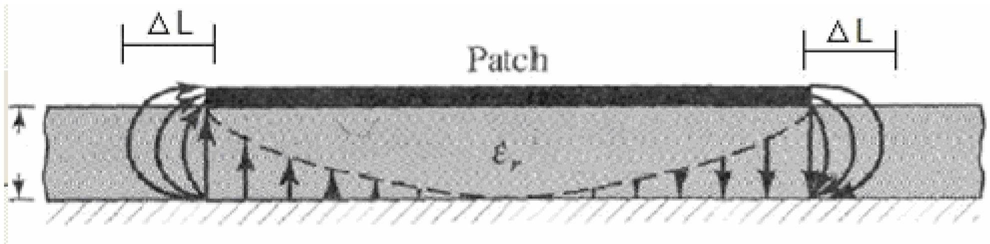


Fig 3.6 Side view of patch antenna with fringing field effect. (Adapted from [1], P. 729)

3:6 Transmission Line Model Analysis for Coplanar waveguide Patch Antenna.

Coplanar patch antennas have a physical structure derived from coplanar waveguide transmission lines. Therefore, a transmission line model is the first and most obvious choice for the analysis and design of coplanar waveguide patch antennas. However, the transmission line model is often regarded as a simplified and some what dated theory. This is true for the original, simple transmission line model, but the accuracy of the improved transmission line model is comparable to that of other more complicated methods. Even mutual coupling between rectangular microstrip antennas can be calculated in a fairly accurate and very efficient way with the transmission line approach.

Any structure supporting wave propagation can be approximated, to a first order, as a transmission line. The coplanar wave guide patch antenna is no exception, in fact, due to the narrow bandwidth of the matching condition; the approximation in terms of equivalent transmission line model of the coplanar waveguide patch antenna is appropriate and very insightful.

The transmission line model treats the antenna as an ideal transmission line terminated in a specific load. The load takes into account the effects of edge capacitance, substrate losses, metal losses, and radiation.

In a relatively thick substrate, the load can also include the effect due to surface waves. This section will derive the basic transmission line model and will not take into account surface waves; a more detailed treatment can be found [25].

3.6.1 Limitations of the sample model

Most of the simple models assume that the electromagnetic fields do not vary across the substrate height. This assumption is only valid when the substrate is thin, as compared with a wavelength, but it is no longer accurate enough for substrate thickness of the order of a tenth of a wave length or longer.

Simple model provide a very valuable first order approximation for analyzing the radiation of thin antennas, except in those particular directions where effects directly related to the presence of the substrate are significant (for instance, surface waves).

Refinements can of course be added afterward to take into account particular effects such as surface waves and to improve the accuracy. There is, however, a limit to what can be done because when several correcting terms that take into account the effects that were neglected by the original approximation are added, simple models eventually tend to become as complex to use as more rigorous interest is lost. More sophisticated approaches, which make an intensive use of the computer to solve Maxwell's equations completely in the presence of boundaries, have been developed.

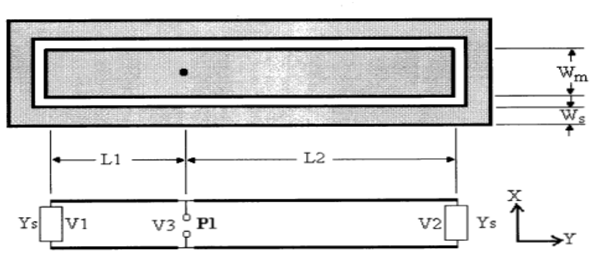


Fig 3.7 Transmission line equivalent to the coplanar waveguide patch antenna.

As illustrated in figure 3.7 a coplanar waveguide patch antenna, in the horizontal direction can be modeled as a pair of transmission lines, with one side terminating at the feed and the other at some load. The reason for the reactance of the load is that the coplanar line does not terminate in a perfect open circuit due to the field not stopping at the edge but extending beyond it, forming stray parasitic, specifically, of capacitive nature. The capacitive nature of the reactance can be included as an extra length of transmission line, which has to be compensated, causing the length of the physical waveguide to be more than a half wavelength. The real part of the load impedance is due to radiation and surface waves, as the transmission line itself does not account for either.

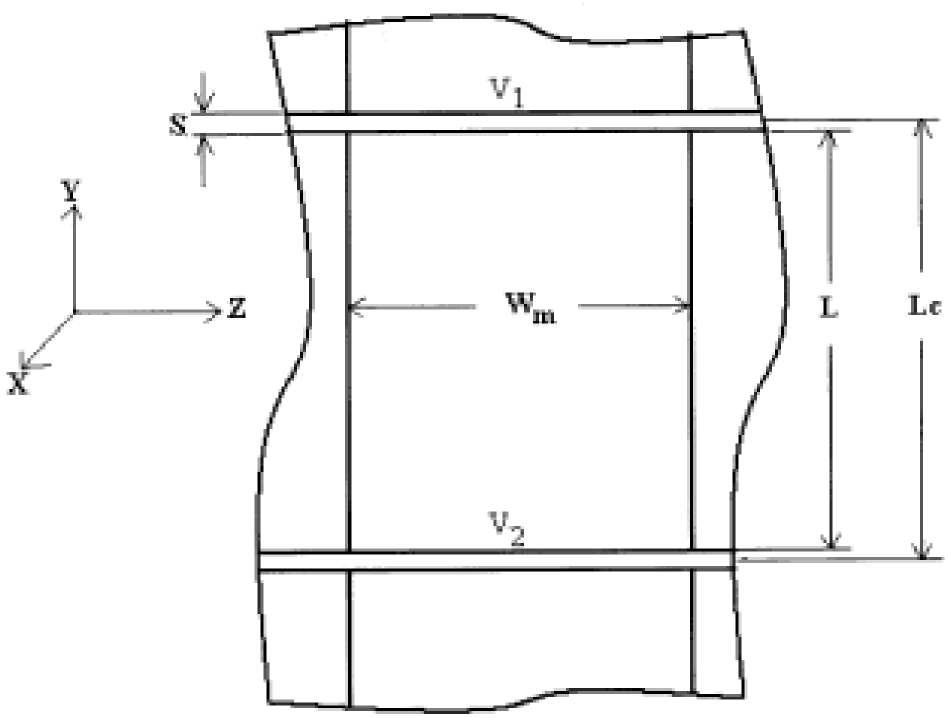


Figure 3.8 Infinite slot models for radiation of a coplanar waveguide patch antenna

A simplified derivation of the load impedance (as given in [11]) models the interaction at edges of the transmission line as that of a pair of infinite slots (figure 3.8). This model is valid for either very long separation of S in which the bottom ground plane interactions dominates or very short separation, in which the top ground plane effects, dominates. Starting with $Y_s = W_s y_s$, where $y_s = g_s + jb_s$ is the admittance per unit length of a uniformly excited slot of infinite length and height H of the substrate, when the ground plane dominates (S>>H), otherwise its S (S<<H). The slots length is the width of the center strip in the actual antenna.

The expressions for the tangential E- field in the slot aperture are given by [11]:

$$E_a = \begin{cases} \frac{V_1}{S}\vec{y}, \frac{L_C - S}{2} \leq y \leq \frac{L_C - S}{2} \\ \frac{V_2}{S}\vec{y}, \frac{L_C - S}{2} \leq -y \leq \frac{L_C + S}{2} \\ 0 .otherwise \end{cases} \tag{3.5.1}$$

Where, L_c is the center distance between the two slots, V_1 and V_2 are the excitation voltages of the two slots and S is the slot width, not necessarily the separation of the top ground plane from the center strip, as indicated above.

The Fourier transform of (3.5.1) [see appendix B]

3.6.2 Coupled Line

Coupled microstrip lines are used in a variety of circuit functions, namely directional couplers, filters, impedance matching networks and delay lines. There are numerous papers dealing with the analysis, design and applications of coupled microstrip lines. Both quasi-static and full-wave analyses have been reported [9]. Here we are analyzing coplanar waveguide (lines).The properties of coupled lines are determined by the self and mutual inductances and capacitance associated with the two lines. Under the quasi-TEM approximation, the self inductance can be expressed in terms of self capacitance by using simple relation. It is also found that for most of the practical circuits using symmetric coupled microstrip lines, the mutual inductance and the mutual capacitance are inter related and it is not necessary to determine the mutual inductance separately.

Therefore, only capacitance parameters are evaluated for coplanar waveguide. These capacitances can be expressed in terms of even and odd mode values for the two modes of propagation. The line parameters of the finite ground plane coplanar waveguide can be obtained by modeling the coplanar waveguide as the center of three coupled microstrip lines. Since the two sides are predominately grounded, the odd mode could be considered as the dominant mode, as a first order approximation. The break up of total line capacitance in to a parallel plate and two fringing capacitances, one for each side of the strip, is also shown. The fringing capacitances for the even mode can be obtained from the fringing capacitance for uncoupled microstrip lines. The odd mode fringing capacitances are determined with help of an equivalent geometry for coupled strip lines and coplanar strips. Using these fringing capacitances the total even and odd mode capacitances may be shown as below geometry.

3.6.2.1 Quasi-TEM Approximation

When the longitudinal components of the fields for the dominant mode of a coplanar wave guide patch remain very much smaller than the transverse components, they may be neglected. This a balanced mode, meaning that the electric field in the transverse cross section is symmetrically oriented from the center strip to the two ground planes.

In this case, the dominant mode then behaves like a TEM mode, and the TEM transmission line theory is applicable for the coplanar waveguide patch as well. This is called the quasi-TEM approximation and it is valid over most of the operating frequency ranges of coplanar waveguide patch.

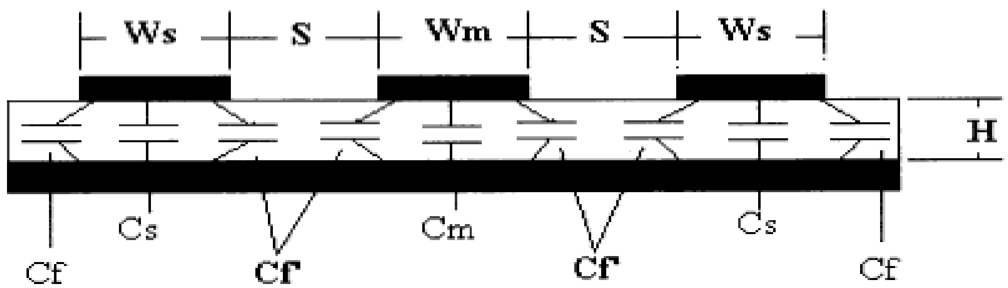


Fig.3.9a Even mode representation of a four-wire coupled transmission line and its equivalent capacitance network

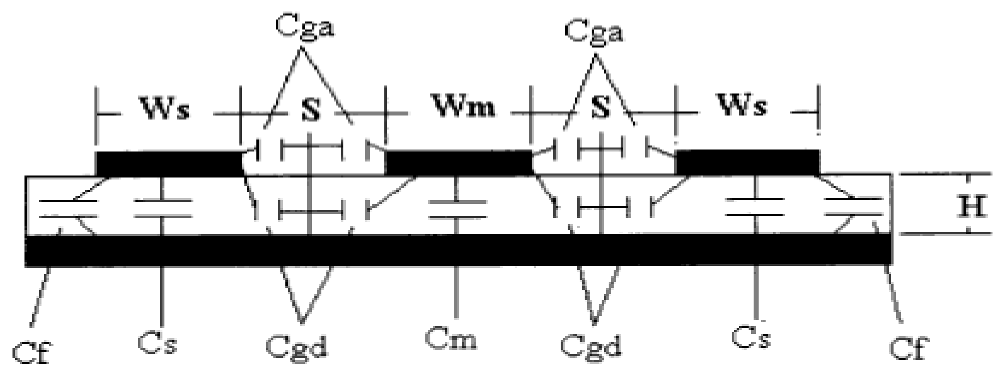


Fig.3.9b Odd mode representation of a four-wire coupled transmission line and its equivalent capacitance network.

Where, $\mathbf{C_m}$= capacitance in the middle strip.

$\mathbf{C_s}$=capacitance in the side strip.

$\mathbf{C_f}$=edge capacitance.

$\mathbf{C_{f'}}$= fringe capacitance.

$\mathbf{C_{ga}}$=capacitance associated with slot in air.

$\mathbf{C_{gd}}$=capacitance in the substrate.

The procedures developed for two coupled lines [3] are extended to include the three line model shown in figure 3.9. The odd and even mode capacitance for center and side strips can be defined as shown in figure 3.9a and 3.9b.

$$C_{me} = 2C_{f'} + C_m \qquad (3.2.5a)$$

$$C_{se} = C_f + C_{f'} + C_S \qquad (3.2.5b)$$

$$C_{mo} = C_m + 2(C_{gd} + C_{ga}) \qquad (3.2.5c)$$

$$C_{so} = C_S + C_f + C_{ga} + C_{gd} \qquad (3.2.5d)$$

Where, **C_{me}**, **C_{mo}**, **C_{se}**, **C_{so}** are the even and odd middle (center) and side strip capacitors respectively, the rest of the parameters are cross coupling capacitances. These capacitances can be described in terms of physical parameters as [8]:

$$C_m = \varepsilon_o \varepsilon_r \frac{W_m}{H} \qquad (3.2.6a)$$

Where Wm is width of patch, H is substrate thickness, **Єr** is dielectric constant of the substrate, and $Є_O$ is dielectric constant of free space.

$$C_s = \varepsilon_o \varepsilon_r \frac{W_S}{H} \qquad (3.2.6b)$$

Where Ws is width of upper conductor and Cs side capacitance.

$$2C_f = \frac{\sqrt{\varepsilon_{re}}}{c} Z_O - C_{m,s} \qquad (3.2.7)$$

Where, **Cm**, **s** are the capacitance of either the middle or the side strip as defined in (3.2.7) for the associated capacitance. And C_f is the fringing capacitance of a coplanar waveguide of width **Ws/H**, impedance **Zo** and the effective dielectric constant **Єre.**

$$C_{f'} = \frac{C_f}{1 + A\left(\frac{H}{S}\right)\tanh\left(12\frac{S}{H}\right)}\sqrt{\frac{\varepsilon_r}{\varepsilon_{re}}} \qquad (3.2.8)$$

Where,

$$A = \exp\left(-0.1\exp 2.33 - 2.53\frac{W}{H}\right)$$

Where **S** is slot width, **H** is the thickness of dielectric substrate and **W** is found by:

$$W = \frac{W_s + W_m}{2} \qquad (3.2.9)$$

Where Ws is the width of upper conductor and Wm is the patch width.

Cga is the capacitance term in the odd mode for the fringing field across the gap, in the air region. It is obtained from geometry of coplanar line and is given by [3]

$$Cga = \varepsilon_o \frac{K(k')}{K(k)}, \quad \text{Where } k' = \sqrt{1-k^2} \qquad (3.2.10)$$

Where k =tanh (πw/2h) and K represents a complete elliptic function of the first kind with k' its complementary function given by

$$K(k') = K'(k)$$

The ratio of the complete elliptic function K (k) and its complement K' (k) is given by [8]

$$\frac{K'(k)}{K(k)} = \begin{cases} \left[\frac{1}{\pi}\ln\left(2\frac{1+\sqrt{k'}}{1-\sqrt{k'}}\right)\right]^{-1}, \text{ for } 0 \le k \le 0.7 \\ \left[\frac{1}{\pi}\ln\left(2\frac{1+\sqrt{k}}{1-\sqrt{k}}\right)\right]^{-1}, \text{ for } 0.7 \le k \le 1 \end{cases} \tag{3.2.11}$$

Cgd represents the capacitance in the odd mode for the fringing field across the gap, in dielectric region.

It is evaluated by modifying the corresponding capacitance expression for coupled strip lines and is given by [3]:

$$C_{gd} = \frac{\varepsilon_o \varepsilon_r}{\pi}\ln\left\{\coth\left(\frac{\pi S}{4H}\right)\right\} + 0.65 C_f \left[\frac{0.02H}{S}\sqrt{\varepsilon_r} + 1 - \varepsilon_r^{-2}\right] \tag{3.2.12}$$

The characteristic impedances and effective dielectric constants for the two modes can be obtained from the capacitance values using the following relations [3]

$$\varepsilon_{re}^{i} = \frac{C_i}{C_i^a} \tag{3.2.13}$$

And

$$Z_{oi} = \left[c\sqrt{C_i C_i^a}\right]^{-1} \tag{3.2.14}$$

Where i stand for the even or odd mode and $\mathbf{C^a}$ *denotes the capacitance with air as dielectric, c is speed of light* (3×10^8 *m/s*).

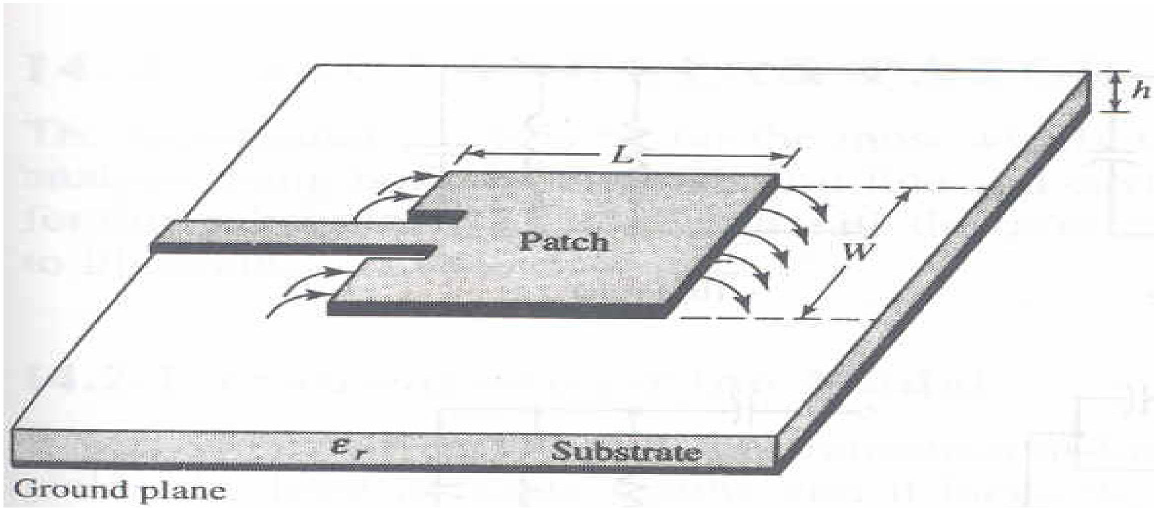
Here, ϵr_e and Zo_i are the line parameters, and $\mathbf{C_i^a}$ is the equivalent capacitance when $\boldsymbol{\epsilon_{re}=1}$.

Calculations indicate that the even mode capacitance is accurate to with in 3 percent for **Єr≥1. T**he accuracy for odd mode capacitance improves for higher order values of Єr**.**

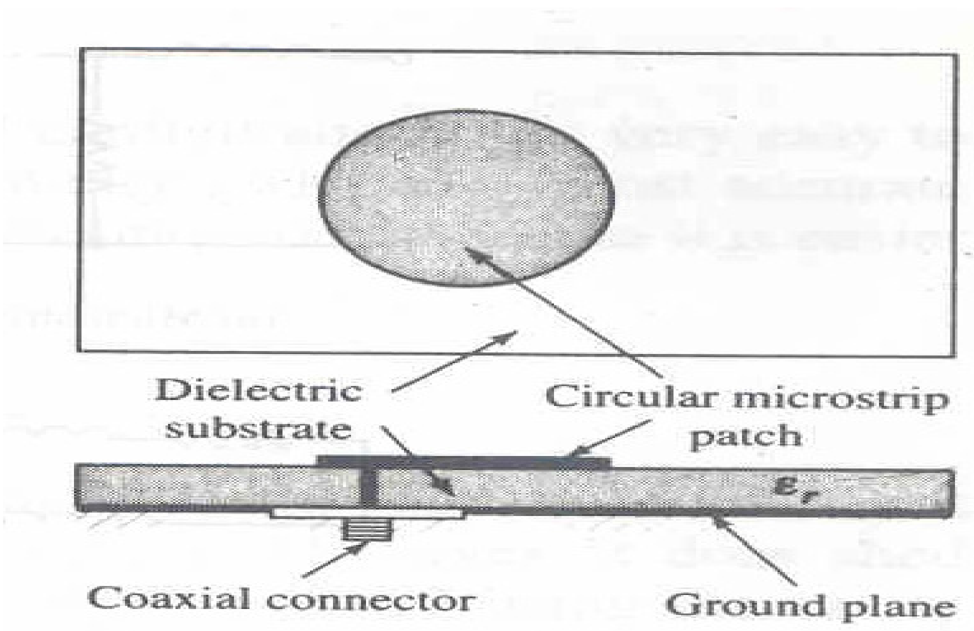
3.6.3 Feeding techniques

There are many configurations that can be used to feed CPW patch antennas. The four most popular are: the microstrip line, coaxial probe, aperture coupling and proximity coupling. The microstrip feed line is also a conducting strip, usually of much smaller width compared to the patch. The microstrip line is easy to fabricate, simple to match by controlling the inset position and rather simple to model. However as the substrate thickness increases surface waves and spurious feed radiation increase, which for practical designs limit the bandwidth. The direct coaxial probe feed is simple to implement by extending the center conductor of the connector attach to the ground plane up to the patch. Impedance can be adjusted by proper placement of the probe feed.

As the probe distance from the patch edge, y, is increased, the input resistance is reduced by the factor $\mathbf{\cos^2 (\pi y/L)}$**.** A disadvantage of the probe feed is that it introduces an inductance that prevents the patch from being resonant, if t is 0.1λ or greater. Also, probe radiation can be a source of cross polarization.



a) microstrip line feed



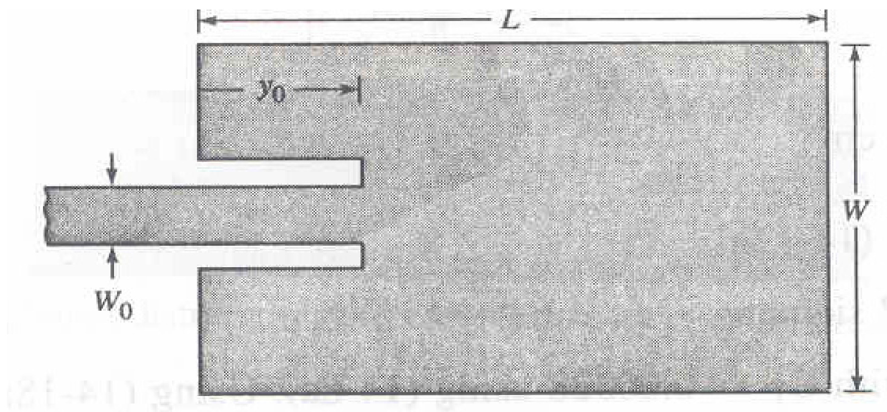
(b) Probe feed

Figure 3.10 Typical feeds for patch antennas

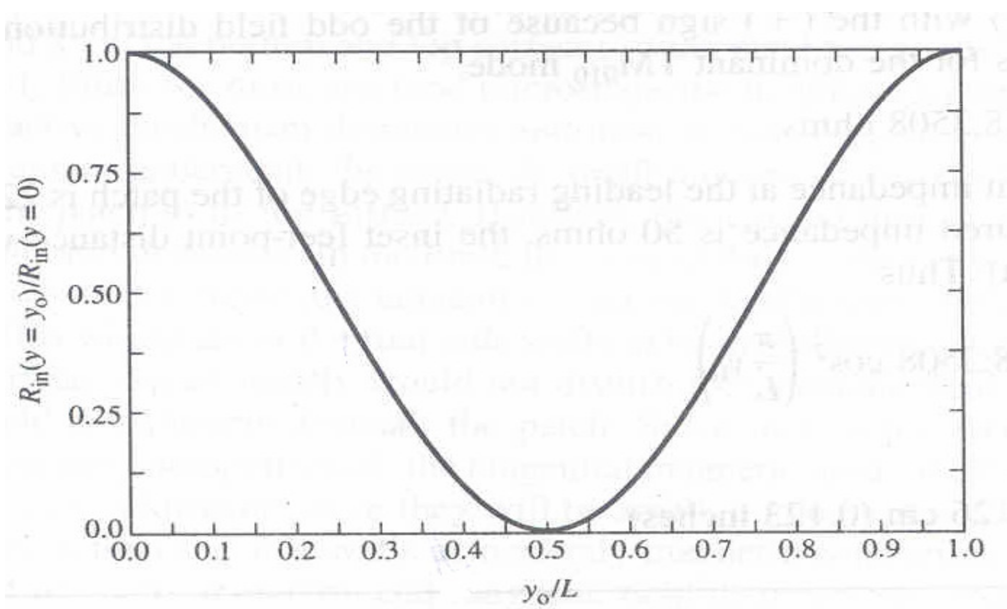
The feed point impedance is found from [27]:

R(y) is the input impedance at point y along the antenna.

$$R(y) = Z_{Odd} \cos^2\left(\frac{\pi y}{L}\right) \qquad (3.2.15)$$



a) Recessed microstrip line feed



b) Normalized input resistance [1]

Figure 3.11 Recessed microstrip-line feed and variation of normalized input resistance.

A plot of the normalized value of is shown in figure 3.11(b). The values obtained agree fairly agree well with experimental data. However, the inset feed introduces a physical notch, which in turn introduces a junction capacitance. The physical notch and its corresponding junction capacitance influence slightly the resonance frequency, which typically may vary by about 1%. Figure 3.11(b) that the maximum value occurs at the edge of the slot ($y_o = 0$) where the voltage is maximum and the current is minimum. The minimum value (zero) occurs at the center of the patch (y_o=L/2) where the voltage is zero and current is maximum. As the inset feed-point moves from the edge towards the center of the patch the resonant input impedance decreases monotonically and reaches zero at the center. When the value of the inset feed point approaches the center of the patch (y_o=L/2), the $\cos^2$ (πy_o/L) function varies very rapidly; therefore the input resistance also changes rapidly with the position of feed pint. To maintain very accurate values, a close tolerance must be preserved.

3.6.3.1 Feed point location

The rectangular patch antenna can be represented by a transmission line model of two narrow radiating apertures separated by a transmission line with a length and width equal to that of the patch antenna and with characteristic impedance Z01 (and corresponding characteristic admittance Y01), as shown in Figure 1.C.

The radiating apertures on the edges of the patch antenna are each represented by an admittance YL1 and YL2, where YL1 = G1 + jB1 and YL2 = G2 + jB2. Since the radiating apertures are identical, the aperture conductance and susceptances are equal (G1 = G2, B1 = B2), and consequentially, the aperture admittances are equal (YL1 = YL2 = YL). The input signal feed point is located at a point along the length of the transmission line that is chosen to minimize the antenna reactance and maximize the input impedance match (50Ω for a typical microstrip feed).

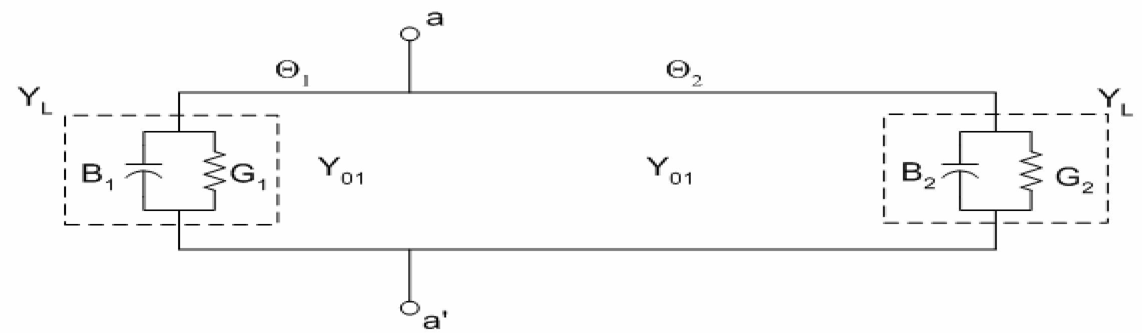


Figure 3.12 transmission line model of rectangular patch antenna

According to transmission line theory,

According to transmission line theory,

$$Y_{aa'} = Y_{01}\frac{Y_L + jY_{01}\tan(\theta_1)}{Y_{01} + jY_L\tan(\theta_1)} + Y_{01}\frac{Y_L + jY_{OL}\tan(\theta_2)}{Y_{01} + jY_L\tan(\theta_2)} \qquad (3.2.16)$$

Where,

$$\theta_1 = \beta.l_1,\ \theta_2 = \beta.l_2\ ,\ \beta = \frac{2\pi}{\lambda_o}$$

Where Yaa' represents the input admittance at the signal feed point. The length of the patch antenna, corresponding to a transmission line of electrical length θ1 + θ2, is set to approximately λo/2 so that the electric field polarization at each radiating aperture is opposite to the other. This maximizes broadside radiation above the patch and results in a negligible susceptance B1, B2 in the load admittance, YL, such that YL ≈ GL.

If θ1 is set to 45°, or λo/8, and θ2 is set to 135°, or 3λ/8, then the reactive component of Yaa' disappears and the formula for Yaa' reduces to,

$$Y_{aa'} = 4.\frac{Y^2{}_{01}.G_L}{Y^2{}_{01} + G^2{}_L} \qquad (3.2.17)$$

The reciprocal of Yaa' yields a feed point impedance of,

$$Z_{aa'} = \frac{1}{4}[\frac{1}{G_L} + Z_{O1}.G_L] \qquad (3.2.18)$$

Since the feed point impedance should equal 50Ω and both GL and Z01 include the transmission line width, W, as a variable, the width that yields the best impedance match can be solved for by setting Zaa' = 50Ω and solving for W in the following way:

From basic transmission line theory,

$$Z_{01} = \sqrt{\frac{L}{C}} \qquad (3.2.19)$$

Where

$$C = \varepsilon . F(g), \quad L = \frac{\mu}{F(g)}$$

and from Schwarz-Christoffel transformations applied to a standard microstrip transmission line geometry,

$$F(g) = \frac{2.w}{h} + \frac{2}{\pi} + \frac{2}{\pi}.\ln(1 + \frac{\pi .w}{h}) \qquad (3.2.20)$$

So Z01 simplified as a function of h and W results in,

$$Z_{01} = \frac{\sqrt{\frac{\mu}{\varepsilon}}}{\frac{2.w}{h} + \frac{2}{\pi} + \frac{2}{\pi}.\ln(1 + \frac{\pi .w}{h})} \qquad (3.2.21)$$

Next, from aperture antenna theory derived fully in [1], the aperture conductance GL is approximated by means of a Fourier transform as,

$$G_L = \frac{w}{120.\lambda_o}[1 - \frac{1}{24}(\beta_o .h)^2] \qquad (3.2.22)$$

Where, $\beta_O = w\sqrt{\mu\varepsilon}$

Characteristic impedance and effective dielectric constant can be found

From [3],

$$Z_O\sqrt{\varepsilon_{re}} = 60\ln\left(\frac{8H}{W} - 0.358 + \frac{1}{0.931(\frac{H}{W}) + 0.736}\right) \qquad (3.2.23)$$

And

$$\varepsilon_{re} = 1 + 0.5(\varepsilon_r - 1)\left[\left(1 + 10\frac{H}{W}\right) - 0.555 + 1\right] \qquad (3.2.24)$$

From equations (3.5.9) – (3.5.18), of which the result of (3.5.17) and (3.5.19) gives the line parameters 0f CPW patch antenna which can be used with microstrip design equation to yield the length and feed point location for a particular antenna design.

3.6.4 Losses in CPW

When quasi-static approximation is valid, one can use Wheeler's incremental inductance formula for evaluating ohmic losses, and the expression for attenuation in CPW becomes. The total loss in CPW as [8]:

$$\alpha_T = \alpha_C + \alpha_d \qquad (3.2.25)$$

Where αc is loss due to conductor, αd is loss due to dielectric.

$$\alpha_c^{cp} = 0.023\frac{R_s}{Z_{OCP}}\left[\frac{\partial Z_{ocp}^a}{\partial S} - \frac{\partial Z_{ocp}^a}{\partial w_m} - \frac{\partial Z_{ocp}^a}{\partial t}\right], db/m \qquad (3.2.26)$$

Where superscript, CP denotes coplanar waveguide. After subsisting the expression for various partial derivatives in equation (3.5.25) the final expression for conductor loss becomes

$$\alpha_c^{cp} = 4.88\times10^{-4} R_s \varepsilon_{re} Z_{oc} \frac{p^c}{S\pi}(1+\frac{w_m}{S}).\left\{\frac{\frac{1.25}{\pi}\ln\frac{4\pi S}{t}+1+\frac{1.25\,t}{\pi w_m}}{[2+\frac{w_m}{S}-1.25\,t(1+\ln\frac{4\pi S}{t})]}\right\} \quad (3.2.27)$$

Where P^c is given by

$$P^C = \begin{cases} \dfrac{1}{(1-K)\sqrt{K}}, for\, 0.7 \le K \le 1.0 \\ \\ \dfrac{K}{(1-K')(K')}\left[\dfrac{K(k)}{K(k')}\right], for\, 0 \le K \le 0.7 \end{cases} \quad (3.2.28)$$

The expression for the attenuation constants due to dielectric loss in CPW is the same as that for microstrip lines and can be written as [8]:

$$\alpha_d = 27.3\frac{\varepsilon_r}{\sqrt{\varepsilon_{re}}}.\frac{\varepsilon_{re}-1}{\varepsilon_r-1}.\frac{\tan\delta}{\lambda_o}, db/m \quad (3.2.29)$$

Where Rs is surface resistivity of the conductors, λo is free space wave length and tanδ is the loss tangent of the dielectric.

Chapter-four

Tunability

4:1 Introduction

Normally, tunability in antennas implies frequency scanning over a certain frequency band. Here, tunability will be regarded as a change in the phase of the input impedance as a function of frequency. The overall goal is to combine radiating elements and a phase shifter as a single tunable patch element.

The advantage of external tunable components is in their ease of mounting. Also, most devices are not suitable for integration of control devices which are commonly manufactured on Si, GaAs, or Ge. Since the devices are small in size and sealed in their own packages they often do not interfere with the radiation of the antenna or coupling in the system. Therefore, they do not need to be simulated electromagnetically as part of the antenna system. This offers the designers a system approach where each component could be treated separately thereby saving time and accuracy in numerical simulations. However, the disadvantage of the added external devices is their size and cost. External devices need extra room and often have separate power requirements, which could be different than from the other part of the system. In addition, there is extra cost of packaging and interconnects from such devices to the antenna elements.

Internal devices, due to their high proximity to the antenna itself, sometimes need to be taken into account electromagnetically, with in simulation, thus making them cumbersome to use. As cumbersome as they are to simulate, they do offer simple, low cost alternatives to the external modules. Even though no closed form models exist for most cases, approximate circuit models are very good and can be used for practical design. In the design presented here, a varactor diode is used as the tuning device. A very commonly used device in the microwave and RF arena, the varactor diode functions as a voltage tuned capacitor.

4.2 Mounting of Components on CPW

Since CPW is an open structure, it is possible to mount into the circuit a wide variety of active or passive lumped miniaturized elements, such as resistor, capacitor, inductors, diodes, transistors, dielectric resonators, and is connected to the conducting strips that make up the circuit. However, unlike microstrip, CPW does not need hole drilling through the substrate to connect to the ground, which causes additional kind of loss in the case of microstrip.

4.2.1 Lumped Elements in microwave circuits

Passive elements in conventional microwave circuits are mostly distributed and employ sections of transmission lines. This is because of the large sizes of discrete lumped elements (resistor, inductors, and capacitors) used in electronics circuits at low frequencies. However, if the sizes of lumped elements can be reduced to values much smaller than the wavelength, we should also be acceptable at microwave frequencies, and are designed on the basis of this consideration.

Lumped elements are especially suitable for monolithic MICs and for broadband hybrid circuits with minimal size requirements, such as large ratio impedance transformers.

Computer-aided design of circuits using lamped elements requires a complete and accurate characterization of thin film lumped elements at microwave frequencies. This necessitates the development of comprehensive mathematical models which take into account the presence of ground planes, proximity effect, fringing fields, parasitic, etc.

Design of resistors, inductors, and capacitors at microwave frequency may be arrived at by considering them as very short sections (much smaller than the wave length) of TEM transmission lines.

4.3 Models for Microwave Semiconductor Devices

Recent advances in semiconductor device technology have enabled microwave engineers to use solid state devices extensively for various applications. Devices like PIN diodes, varactor diodes, Schottky-barrier diodes, and point-contact diodes are now universally used in circuits such as mixers, harmonic multipliers, parametric amplifiers, phase shift, and detectors.

4.4 Varactor Diode

A varactor diode is a variable-reactance, non-linear circuit element, the variable reactance is provided by junction capacitance, which varies as a function of applied voltage . Varactor diodes find many applications such as modulators, harmonic generators, parametric amplifiers, up converters, and **tuning elements**.

The semi conductor capacitor is a semiconductor diode (a unidirectional element) whose junction capacitance can be varied by varying an externally applied bias voltage.

Reverse bias voltage causes the depletion zone, thus insulating the two sides of the junction from each other. When the junction is used as a variable capacitor, an external potential is applied. The junction assumes the characteristics of a parallel plate capacitor. The p and n zones represent the plates of the capacitor and the depletion zone represents the dielectric between the plates (see fig 4.4.1). When the potential is applied across the junction in direction which would not normally cause conduction, the carriers will be pulled a way from the depletion zone to a distance that is a direct function of the applied potential. The higher the potential, the further the carriers are pulled away from the depletion zone.

Therefore, the higher the potential the further the parallel plates are pulled apart and the lower the capacitance and as the bias voltage changes so does the depletion region and thereby changing the capacitance in a very controlled way.

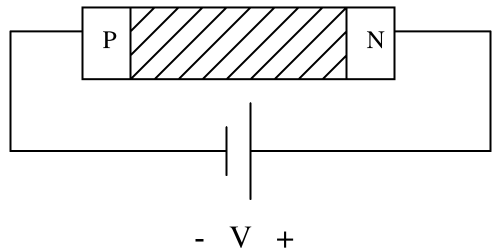


Figure 4.4.1 p-n junction as a function of reverse biased voltage (varactor diode).

In the case of an abrupt junction, abrupt transition from the P- zone, to the depletion zone, to n-zone, such as the case for alloyed junctions. The active elements of varactor diode are the p-n junction (or the metal-semiconductor junction in the case Schottky-barriers varactor diodes). The junction capacitance, C_j, is given by:

$$C_j = \frac{Cjo}{(1 - \frac{V}{\phi})^{\gamma}} \qquad (4.1)$$

Where Cjo is the zero-bias junction capacitance, V is the applied bias (assumed negative when reverse biased), Φ is the contact potential, and the constant γ is the capacitance-voltage slope exponent.

The series resistance is given by;

$$Rs = V/I \qquad (4.\ 2)$$

Where,

$$I = Is\left[\exp\frac{qV}{nkT} - 1\right] \qquad (4.3)$$

Where q is the electronics charge, k is the Boltzmann's constant, T is absolute temperature, and I_s, the reverse saturation current, and Rs, series resistance. The ideality factor, n is about two for point-contact diodes and about 1.06 for Schottky-barrier diodes.

4.5 Implementation of Varactor Diode

The varactor diode is mounted on the top surface of the antenna in certain location, where the electric field is maximum and thus controls the resonant frequency of the antenna. The diode acts as a matching element when operated over a large range and thereby significantly changing the resonance frequency of the antenna.

Using the transmission line method introduced in section 3.5 a varactor diode is added as a tuning element in a planar waveguide patch antenna. The diode, with an appropriate DC offset, acts as a capacitor applied on the line, the location in the electric model being directly related to the position in the physical structure.

Therefore, the effect can be predicted very accurately, and optimization can be made utilizing ordinary circuit simulators instead of field simulators, provided that the effects of the component's package are negligible. The circuit model of a tuning element has also to include the packaging effects. Therefore, only a circuit model would not give an accurate representation but it would give a good intuitive understanding. In addition, just adding a capacitor does not induce great change because the phase shift due to a capacitor is a constant 90 degrees.

In order to provide a variable phase shift, a more complex circuit is needed. Here the circuit conducted is parallel tank circuit.

The reasoning for using a parallel tank as oppose to a series tank circuit is that at resonance the impedance is very high and therefore should not disturb greatly the original operation of the antenna. When the lump circuit moves out of resonance its reactance becomes either inductive or capacitive and thereby changing the characteristic of the antenna by appearing to either shorten its length or increase it.

This apparent change of length also changes the impedance match of the antenna to the load impedance, for small changes in capacitance; the input impedance of the antenna does not change much but its phase changes more significantly.

A condition of resonance will be experienced in a tank circuit (Figure below) when the reactances of the capacitor and inductor are equal to each other. Because inductive reactance increases with increasing frequency and capacitive reactance decreases with increasing frequency, there will only be one frequency where these two reactances will be equal.

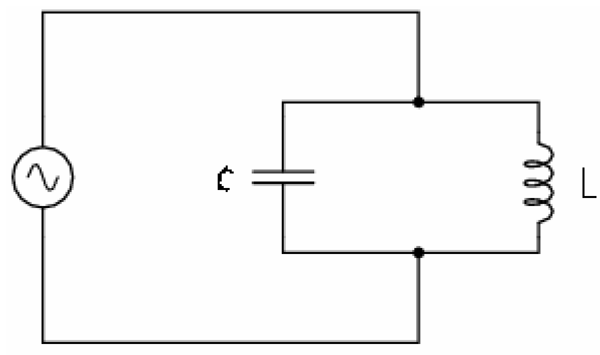


Figure 4.4.2 simple parallel (tank circuit).

$$Z_{parallel} = \frac{1}{\frac{1}{Z_L} + \frac{1}{Z_C}} \qquad (4.4)$$

Now, we use the parallel impedance formula to see what happens to total Zt.

We can't divide any number by zero and arrive at a meaningful result, but we can say that the result approaches a value of *infinity* as the two parallel impedances get closer to each other. What this means in practical terms is that, the total impedance of a tank circuit is infinite (behaving as an *open circuit*) at resonance.

Chapter five

Design specifications and Simulation Results.

In this chapter, the design specification and the results obtained from the simulation are demonstrated. The following procedures were followed to construct the antennas; first the size of the antenna was estimated using analytical approximations. These results served as initial values for further optimization using Empire software tool. Once the desired optimization limits are achieved, the obtained result for the antenna geometry was drawn on a CAD drawing program.

5.1 Design Specifications

The main parameters in the design of coplanar waveguide patch antenna are:

Frequency of operation (f_o): the resonant frequency selected for my design is1.o4GHz.

Dielectric constant of the substrate (ε_r): the dielectric material selected for my design is silicon which has a dielectric constant of 7.85, neither small nor large so as to get reasonable size of CPW patch.

Height of dielectric substrate (H): for the CPW patch antenna is not bulky. Hence, the height of the dielectric substrate is selected as 2.4mm.

The slot width (S): mostly taken as 1/16 inch [9].

The transmission line model described in chapter 3 will be used to design CPW patch and also use microstrip equations for finding patch dimension [1]

$$W = \frac{c}{2f_o\sqrt{\frac{\varepsilon_r + 1}{2}}} \qquad (5.1.1)$$

Substituting c=3e8 m/s, ε_r= 7.85 and f_o=1.04GHz in to equations and get the result.

Effective dielectric constant (ε_{reff}): equation (3.2.17) gives the effective dielectric constant as

$$\varepsilon_{reff} = 1 + 0.5(\varepsilon_r - 1)\left[(1 + 10\frac{H}{W}) - 0.55 + 1\right] \qquad (5.1.17)$$

Similarly, length of the patch can be found as [1]

$$L = \frac{c}{2f_o\sqrt{\varepsilon_{reff}}} \qquad (5.1.2)$$

The expression for extended length, ΔL due to fringing in CPW in the same as that of microstrip lines [1]. i.e

$$\frac{\Delta L}{H} = 0.412\frac{(\varepsilon_{re} + 0.3)(\frac{w_m}{H} + 0.264)}{(\varepsilon_{re} - 0.258)(\frac{w_m}{H} + 0.8)} \qquad (5.1.3)$$

Where H is height substrate, W_m is width of the patch and effective dielectric constant.
In this case, Єre is given by (3.2.21)
Since the length of the patch has been extended by ΔL on each side, the effective length of the patch is now

$$L_{eff} = L + 2\Delta L \qquad (5.1.4)$$

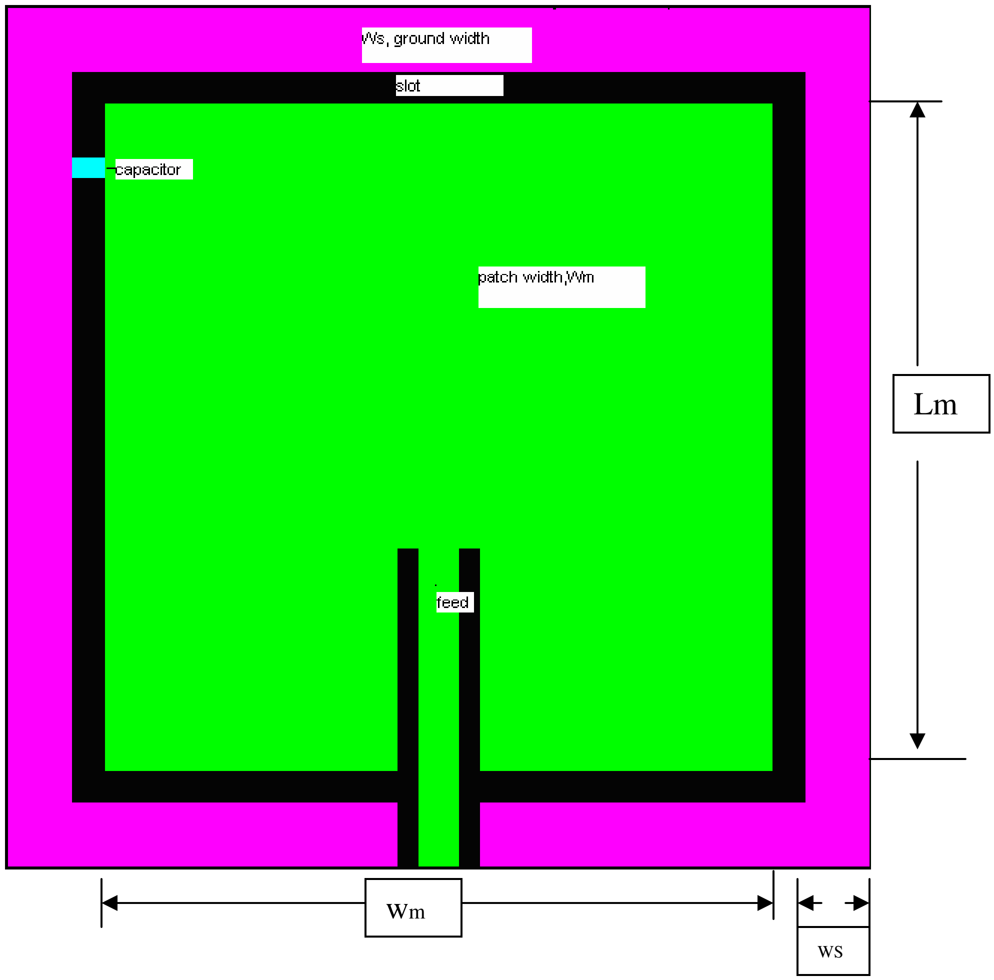


Figure 5.2.1 Top view of CPW patch antenna.

Figure 5.2.1 shows one possible configuration used for the tuning element. The element can be placed on many locations along the line as well as multiple elements can be used. The planar transmission line configuration allows ease of element placement. Since the element can remain in the same plane provided the spacing between the center conductor and the ground plane is small enough to accommodate the element placement.

5.2 Simulation Setup and Results

The software used to model and simulate the CPW patch antenna are math lab and Empire software. Empire software analyzes 3D and multilayer structures of general shapes. It has been widely used in the design of MICs, MMICs, patch antennas, wire antennas, and other RF/wireless antennas. It can be used to calculate and plot the S_{11} parameters, phase tuning, current distributions as well as the radiation patterns.

An evaluation version of the software was used to obtain the results for the thesis.

For simplicity, the length and the width of the patch and the ground plane have been rounded off to the following values: patch length, patch width and conductor width (Lm=65.7mm, Wm= 65.9mm and Ws=6.4mm) respectively.

Figure 5.2.2 is the graphical solution for feed point impedance-matching by width variation, by utilizing MATHLAB, to optimize for the desired frequency to yield a well-match rectangular patch antenna, by setting 'Zaa' equal to 50Ω and solving for the antenna width, W. See math lab. Code for 'variable width patch antenna 'm', the code is written to perform this task. Note that for the following commonly-used formula for the effective dielectric constant, ε_{reff}, of the CPW patch antenna was substituted for relative dielectric constant $\varepsilon_{r,}$ in the equation for Zo1.

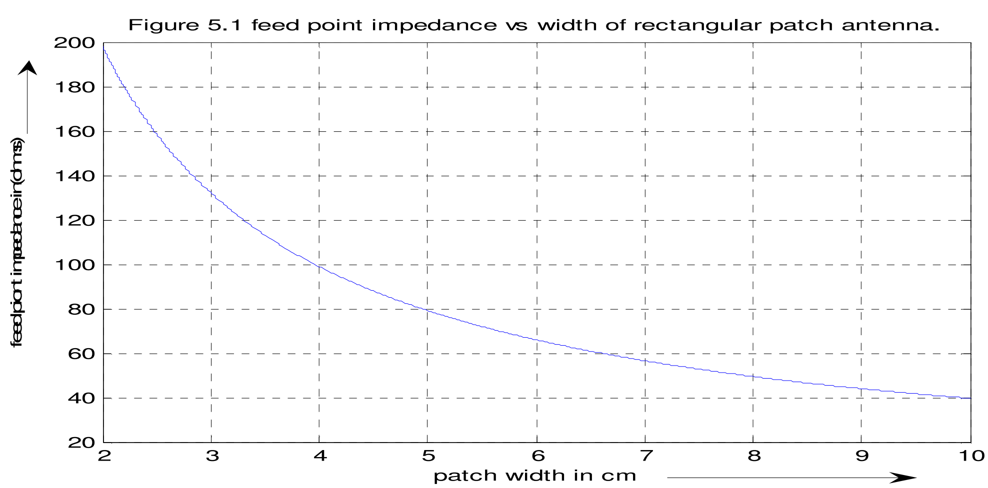


Fig 5.2.2 graphical solution for finding feed point.

Figure 5.2.3 is the graphical representation of equation (4.1), the equation can be used when V<0.2V. The positive voltage, Φ, is the contact potential of the pin junction. As the pin junction becomes more reversed biased (V<0), the depletion capacitance decreases. However, when the varactor junction becomes slightly forward biased, the capacitance increases rapidly. This is shown in Figure 5.3.2 (math lab code in appendix C). The varactor diode is mounted on a tuned antenna in the location determined. The location of the diode was optimized for maximum phase shift by changing the placement location on the patch geometry and the location with the biggest phase change response was selected using Empire software.

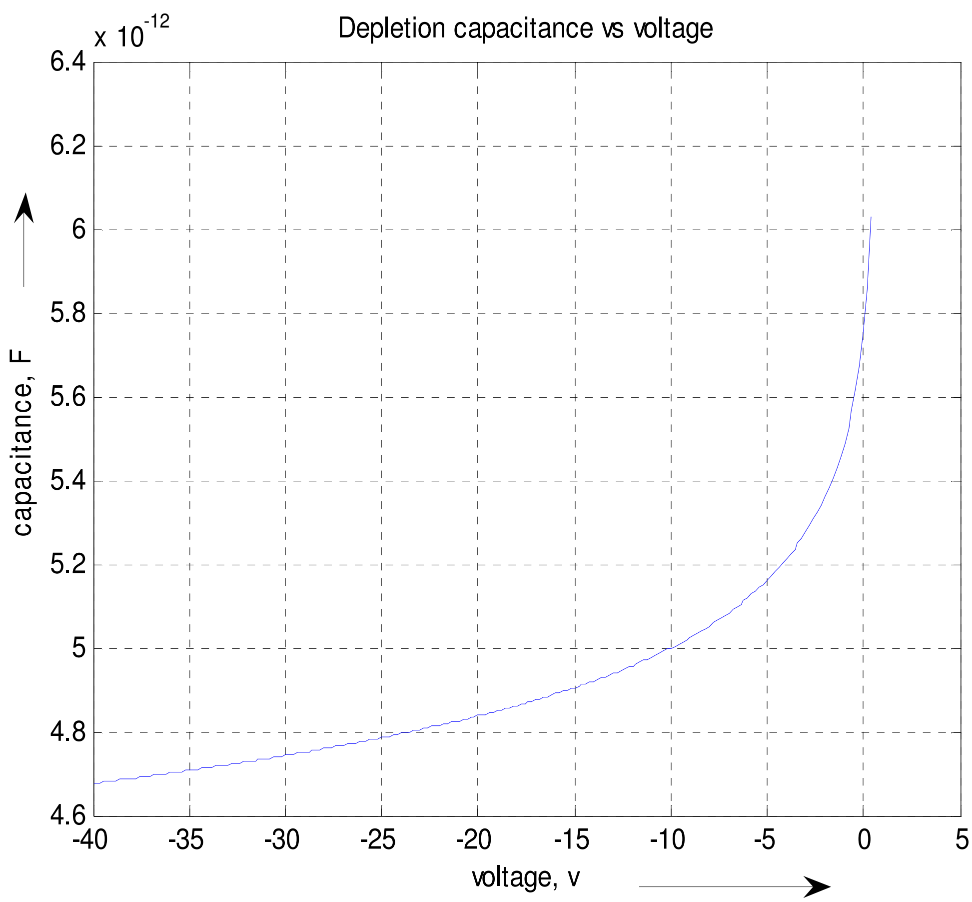


Fig 5.2.3 depletion capacitance of a p-n junction of varactor diode

Figure 5.2.4 depicts the magnitude of scattering coefficient S_{11} from a conductor backed coplanar waveguide patch antenna with one tuning varactor diode. The result indicates that resonance frequency around 1.05GHz and for better antenna performance the return loss less than -9.5, which fulfill the Federal Communication Commission (FCC) ,

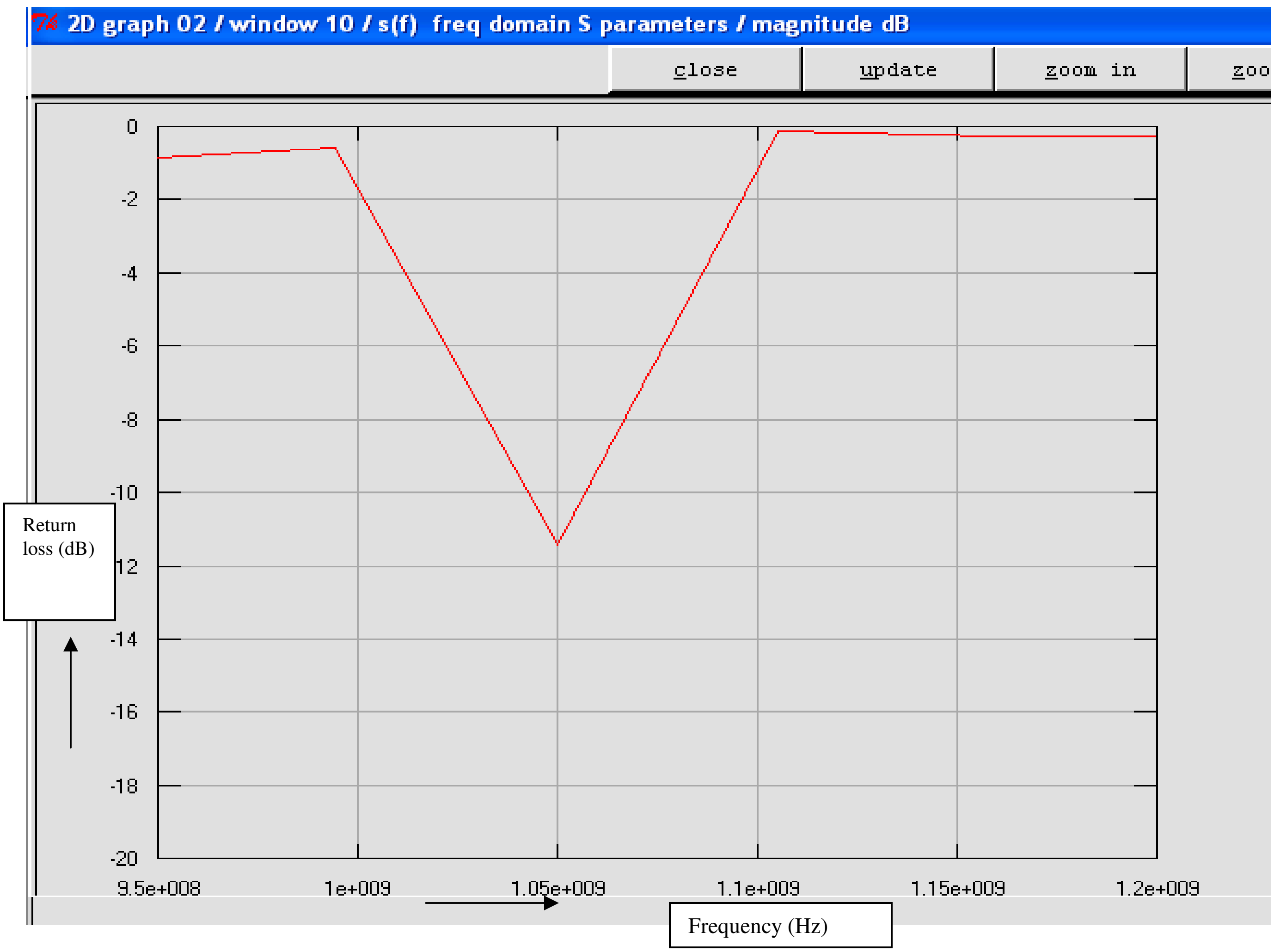


Fig 5.3.4 Return loss for CPW patch antenna

Figure 5.3.5 depicts the phase of scattering coefficient S_{11} from a conductor backed coplanar waveguide patch antenna with one tuning varactor diode. The result tells us a significant amount of phase shift with the given band width at around resonance frequency 1.05GHz. The resonance frequency is shifted from 1.04GHz to 1.05GHz due to the notch effect of microstrip feeder.

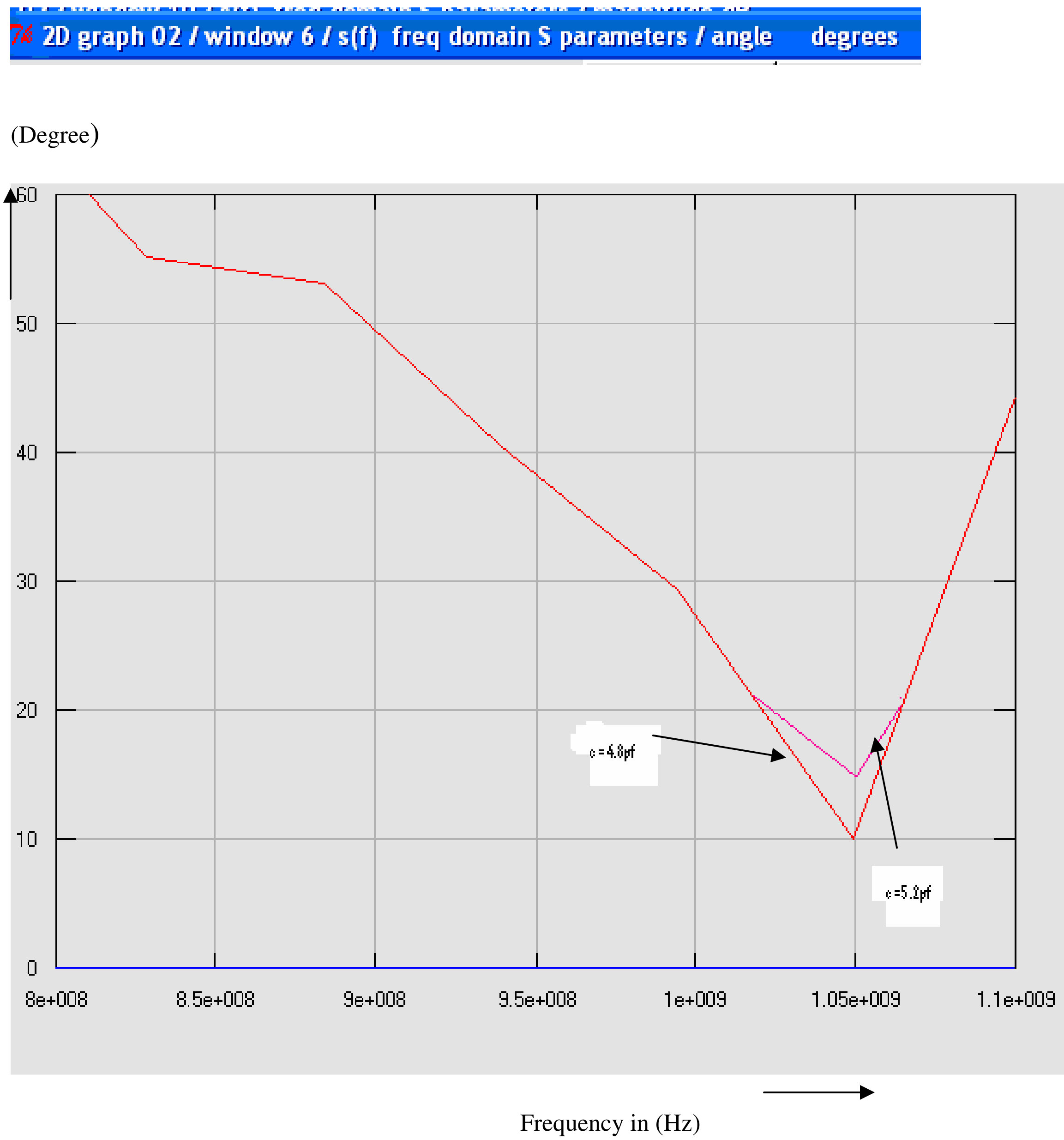


Fig 5.2.5 phase vs. frequency response of CPW patch antenna.

The input impedance curve tells us the magnitude and phase angle of the input impedance of antenna at the specified frequency. It also indicates matching of antenna, because it intersects the unit circle of the smith chart. The curve cut near the center line of smith chart indicates that the matching is become perfect.

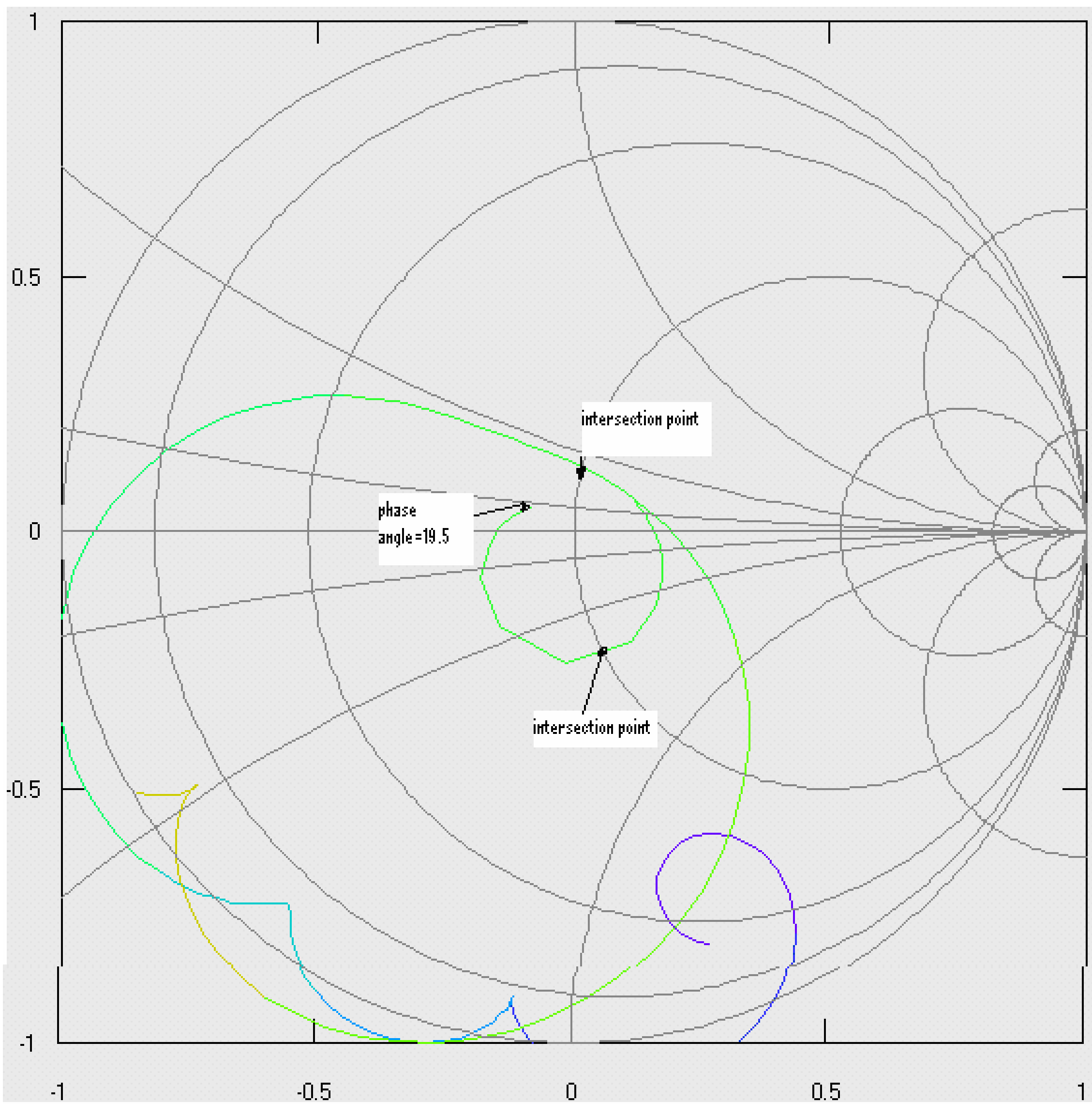


Fig 5.2.6 Input impedance curve.

Chapter Six

6.1Conclusion and Recommendation

In this thesis, a novel application of a tunable antenna is suggested as an attractive alternative for a combination of a radiating element and a phase shifter. This concept has been implemented on a conductor backed coplanar waveguide patch antenna due to its stability to surface mount external tuning elements. Due to limitation on software, tuning approach was based on varactor diode, but the chosen antenna geometry is very suitable for ferroelectric as well as distributed tuning along the perimeter of the path.

Here, approximate analytical models for the coplanar waveguide patch antenna with finite ground plane, were developed. These models are effectively used to determine approximate dimensions of the patch and the feed location for the given size of the ground and the slot width. Using initial results based on these approximate methods in the commercially available empire software, faster convergence of optimal results have been achieved. Another thing is that even though both antenna exhibit narrow bandwidths, a conductor backed coplanar waveguide patch antenna wider bandwidth response for equal dimensions of the patch.

One of the challenging situations encountered in the design of patch antennas is to obtain the exact location of the feed point (port) and to get coupling in the better coupling in the case of multi-layer configuration. But I tried to logical approximation using math lab software and the located point where better performance is obtained.

Another difficulty in design of CPW patch antenna is to use tuning element directly, due to Empire soft does no an access to use it. Hence I used the equation of varactor diode to get capacitor values (using math lab) and by using the value of the capacitor from the graph and put it in to the designed patch antenna as shown Figure 5.2.1.

Very difficult to integrate radiating element and phase shifter in monolithic form, in traditional beam steering phase array antenna phase shifter. But it is effective for both lower and higher phase shifter. So the design presented is limited to narrow band, i.e traditional method is better in this aspect. Otherwise, the result obtained tells us the variation of the phase with in certain band (narrow bandwidth) is equally acceptable for the low phase shifter.

This type of approach eliminates the need of complex systems for the purpose of tenability, at least at the hard ware level, and thereby allows inexpensive implementation of massive antenna arrays for low power operations, due to the capacitive change being a small signal effect. For higher power applications, distributed tuning seems to be a very attractive option to achieve even higher phase shifter. Further research is to be evaluated for distribute tuning along the slot using ferroelectric and magnetic materials.

Appendix A

General Overview of EMPIRE^{TM} Software

EMPIRE^{TM} is an establish and versatile electromagnetic field simulator based on up on the Finite-Difference Time- Domain Method (FDTD), At the IMST GmbH, this powerful method has been applied to build an electromagnetic field simulator for the analysis of packages, Interconnects, Radiators, waveguide elements (EMPIRE^{TM}) and EMC problems. The exact knowledge of electromagnetic fields and the propagation of waves is a prerequisite for the design of radio frequency (RF) elements. In former times, RF design tasks were based on measurements and simple models, leading to time and cost consuming design procedures. Today, affordable computer hard ware have and soft ware have established and simulations programs can accurately predict the electro-magnetic behaviour of new products so, field simulators have become a new standard for RF calculations. Unlike the Finite element method, which has long been established in computing electromagnetic fields, the FDTD method had its break through in the late 80's when extensive research and development lead to a number of improvements in applying advanced boundary conditions, e.g., free space or waveguides, and, therefore, reducing significantly the area of simulation.

Today, its applicability covers the whole area of three-dimensional (3D) field simulations for RF designer. Originally intended to analyze planar passive structures, EMPIRE^{TM} simulator continuously extended to capture more and more high frequency applications. It has proven its functionality in numerous comparisons with measurements and was successfully applied in many different microwave components designs.

The method is based on the most efficient algorithm known which is very accurate and simplification is hardly necessary since it solves the discredited Maxwell's equations directly.

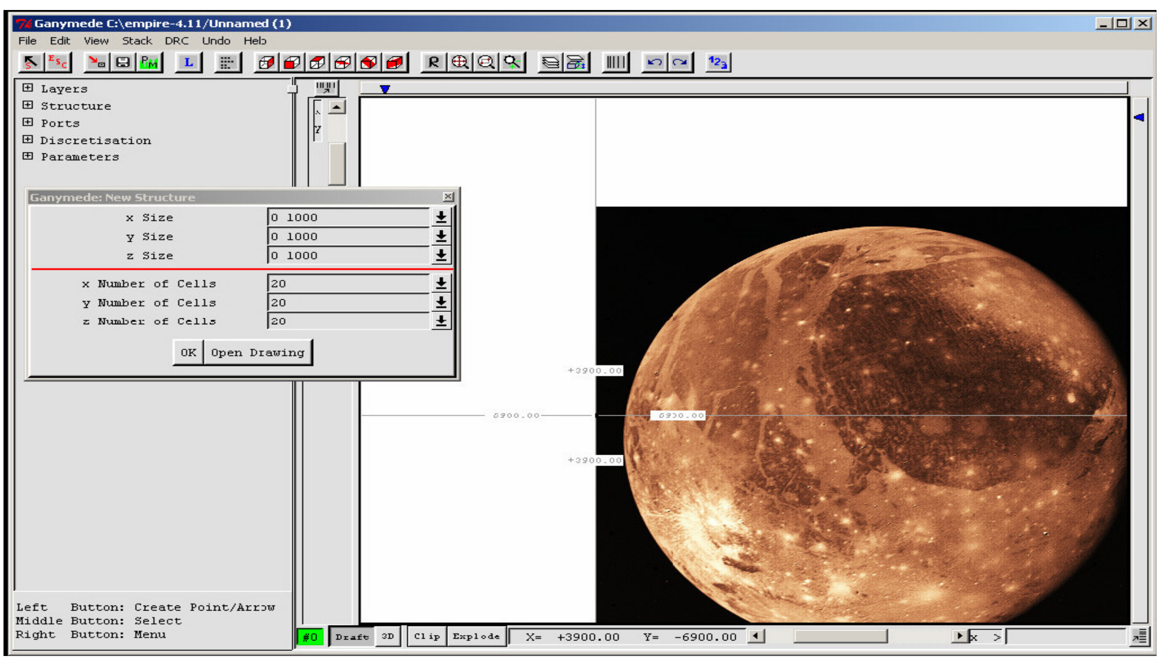
The EMPIRE™ Graphical User Interface Ganymede editor works on all supported platforms, like windows NT/2000/XP, Suse and Red Hat linux. Graphics date can be easily exchanged with other CAD programs for both 2D layout data (DXF 12,Gerber, GDS11, …) as well as 3D structures in STL standard which supported by nearly all CAD tools. Multiple port scattering parameters, radiation patterns, field plots etc are generated for a user defined frequency range with in only one simulation process. Monitoring and animation can give physical insight into the electromagnetic phenomena while accurate results are obtained with little effort.

1.1 setting prefences

It is assumed that the Ganymede icon is installed on the desktop of the PC or the EMPIRE™ software is installed correctly on a workstation, e.g. the command empire is known.

Step 1: double click the Ganymede icon

If the installation set-up was done successfully, the Ganymede window as displayed in figure A.1.1 Empire to start.



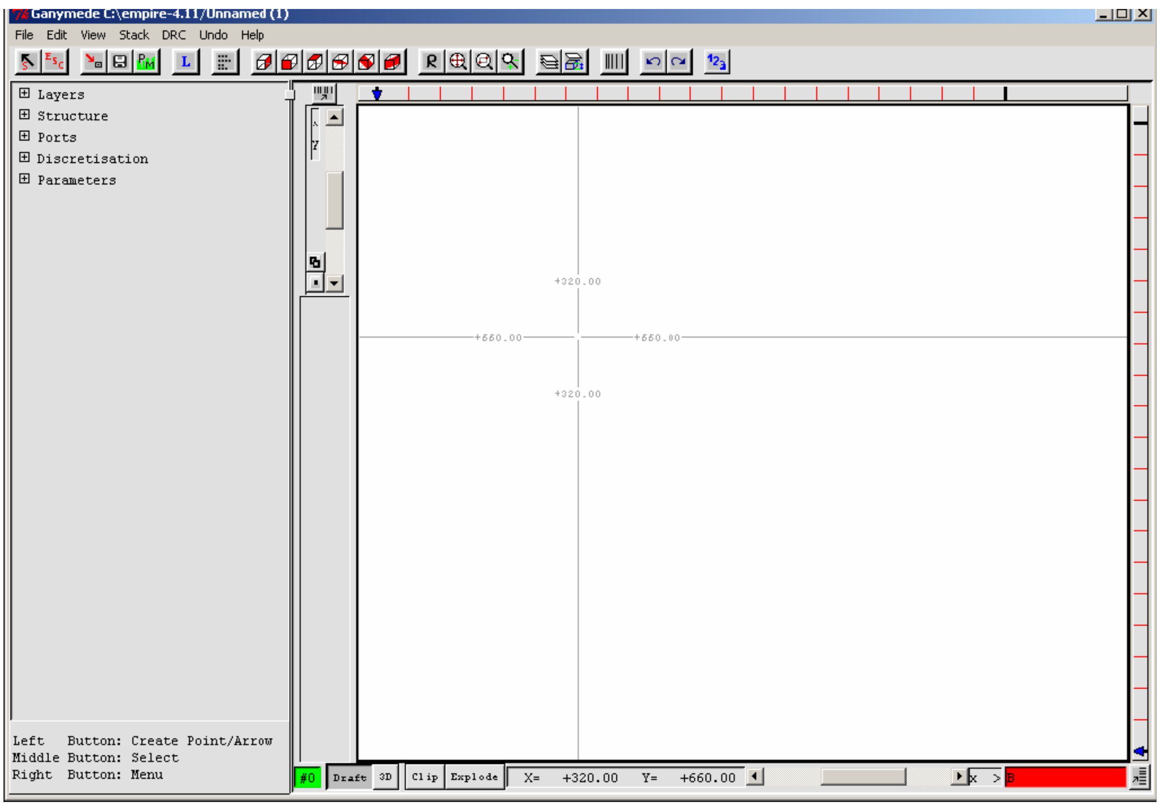
Step 2: creating the drawing area

Enter the size of structure: X0, Y0, Z0 and X1, Y1, Z1

Enter the basic discretization:

Press ok

As can be seen in the discretization bars, bold lines mark the entered boundaries. The blue triangles mark the origin, black lines are fixed, red lines are un-fixed grid lines. The cross-hair cursor displays the current co-ordinate values, as shown figure A.1.2 after entering drawing area



Save Settings: file save as

Create the project directory C:\empire-data\tutorial using and select the new folder.

1.2 Creating and defining layers

Each object that will be created needs a layer and its property has to be defined.

1. Press Layers Create Layer on the left list.
2. Press sign of the newly created layer metal200

Then other properties like: material property, metal property etc can be defined.

Once we know how to create layer and define its properties the next thing is discussing how to create objects like boxes, discs, cylinders, curved objects and slotted objects.

Note: In using this software after writing text or numbers we have to press the enter button on the keyboard. If not we will get the default value.

1.3 Defining Structures

(a) Box: For example if we want to create our working area, which is a box, we do as follows.

Step1: double click the Ganymede editor

Step2: Enter the x, y, z size and the respective number of cells, say,

x size →0 100000

y size →0 200000

z size →0 70000

and the corresponding number of cells, say,

x number of cells → 100

y number of cells → 200

z number of cells → 70

Step 3: Press the OK button

Step 4: Press the Layers (a default layer named A is created)

Step 5: Optionally change layer's color or name

Step 6: Press metal 200 Object Definition General Objects

Material object → rel.permitivity 1

→conductivity 0

→ priority 100

→ Tangent delta none

→ Frequency for tangent delta in Hz 1e9

Step 7: Press the OK button

Step 8: Press Structure create box

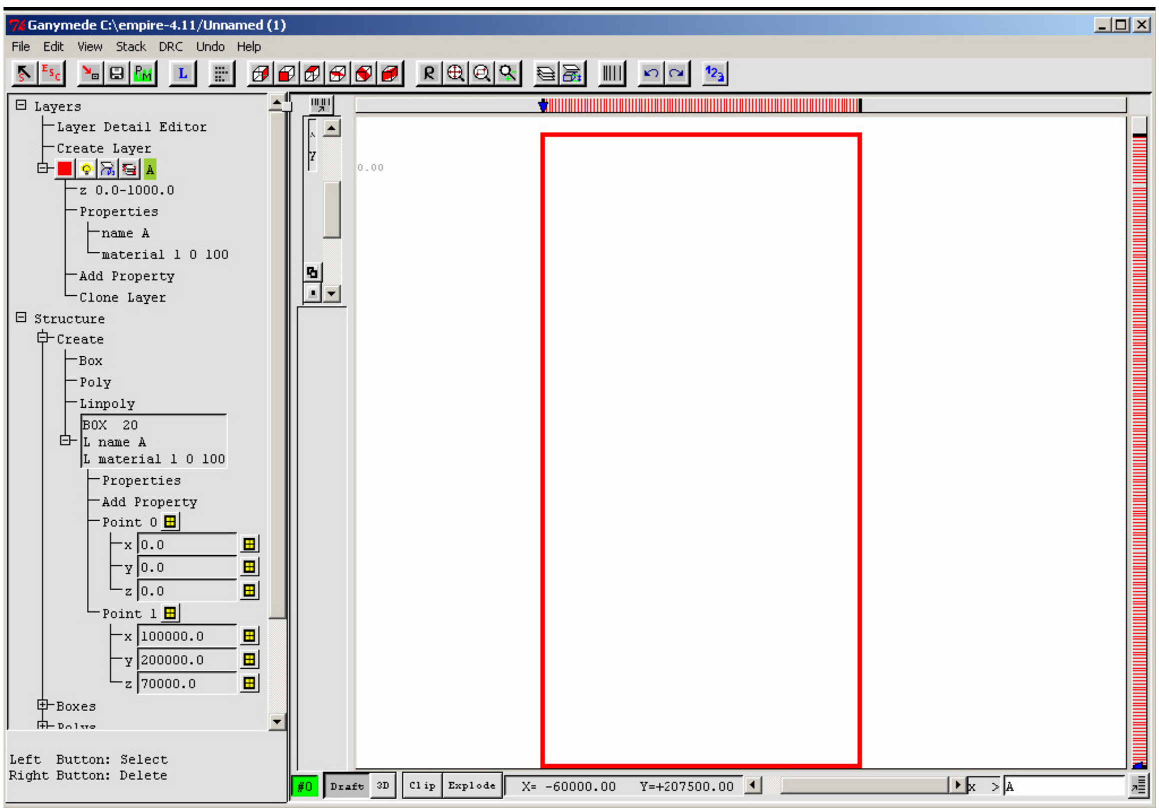
Point 0: x 0 and Point1: x 0; x 100000

y 0; y 200000

z 0; z 70000

Then we get the following figure.

Fig A.1.3 2D view



If we click 3D at bottom, you will see 3D as shown below

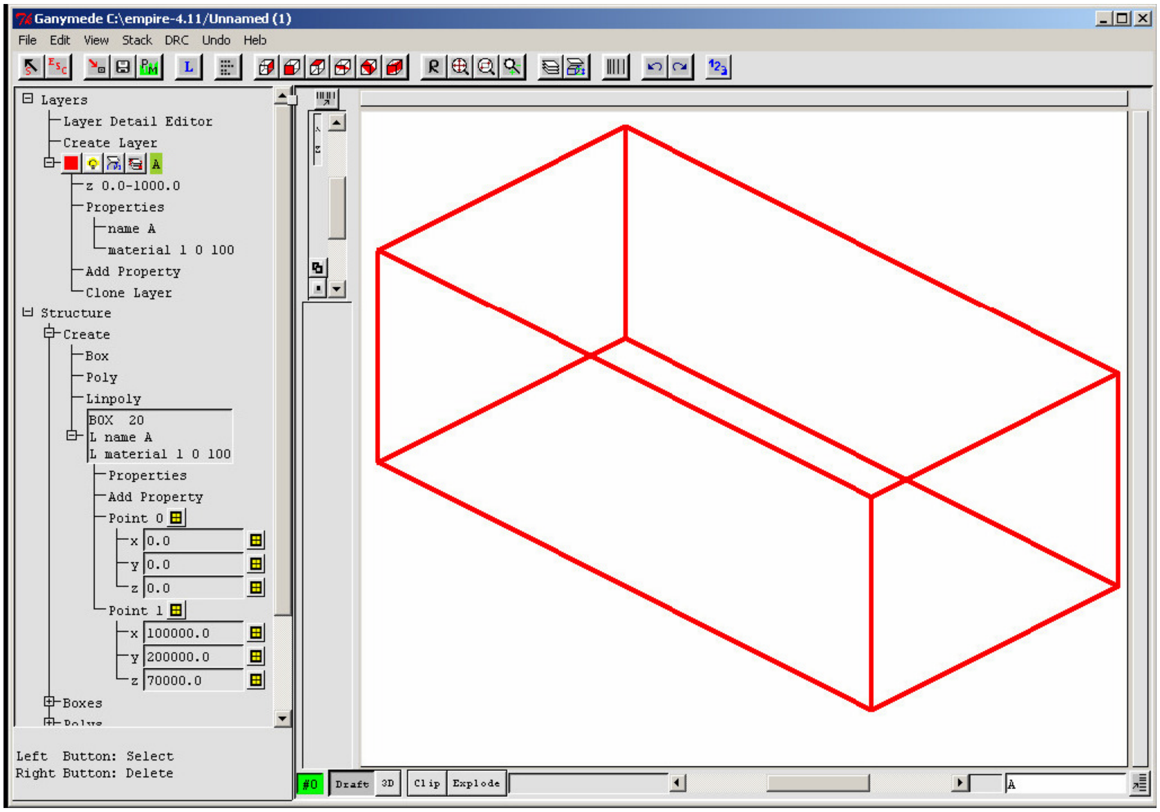


Figure A.1.4 3D view

The next thing is descretization.

Step 1: Press the Top View button

Step 2: Press the left top square button with diagonal arrow

Step 3: Press Empty Descretization that is appeared at the left margin

Step 4: Select the port using middle mouse button and press the left top corner square button with diagonal arrow

Step 5: Press the Propose Discretization Local button

Step 6: Along the width of mobile phone fix by pressing on the top descretization bar using Alt + left click

Step 7: Press on the empty gray space along the width

Step 8: Write frequency and dielectric constant on the key board and press the Enter key

Automatically you will see discretization

The descretized 2D and 3D views are as follows. We have to note that during the 3D view the surrounding layer and the casing layers should be turned off and later turned on for simulation.

Simulation Procedure

Step 1: Press the Export & Save button close

Step 2: Press the Start Simulation button

Step 3 : Enter center frequency 1.04GHz , Bandwdth30MHz, start 500MHz,stop 2.8GHz

Step4: Press End Criteria→Resonance Estimation Switch ON →
Order of the mAvg system 40

Step 5: Press cleanup setup→cleanup selector (select all)

Step 6: Press Processing Start→complete simulation→yes

Now the simulator starts to run.

Step 7: Press the Visualization setup 2

NOTE: for more detail read EMPIRE™ manual available E-mail: empire.support@imst.de.

Appendix B

The Fourier transforms 0f (3.5.1)

$$\overline{E_a} = \int_{-\infty}^{\infty} E_a e^{jk_y y} dy \qquad \text{(B.1)}$$

Where, K_y is the y component of the propagation constant K. Since the Fourier transform, Ea has only a y component, so as Ea, leading to

$$\overline{E_y} = jS \frac{\sin(k_y S/2)}{k_y S/2}\left(V_1 e^{k_y L_C/2} + V_2 e^{-jk_y L_C/2}\right) \qquad \text{(B.2)}$$

Substitution of (3.5.3) into the expression for the complex power [5]

$$P = \frac{k}{4\pi\eta} \int_{-k}^{k} |E_Y|^2 \frac{dk_y}{\sqrt{k_y^2 - k^2}} \qquad \text{(B.3)}$$

And

$$q = \frac{k}{4\pi\eta}\left(\int_{k}^{+\infty} |k_y|^2 \frac{dk_y}{\sqrt{k_y^2 - k^2}} + \int_{-\infty}^{-k} |E_y|^2 \frac{dk_y}{k_y^2 - k^2} \right) \qquad \text{(B.4)}$$

And η is the wave impedance in the substrate. The real and imaginary parts of the power can be related to the circuit of figure 3.8 to form the following equations:

$$p + jq = 0.5 y_s \left(|V_1|^2 + |V_2|^2\right) + y_m R_e (V_1 V_2) \qquad \text{(B.5)}$$

Setting $V_2 = 0$, we get the following integral equations

$$g_s = \frac{k}{\pi\eta}\int_0^k \frac{\sin 2(k_y S/2)}{(k_y S/2)^2}\frac{dk_y}{\sqrt{k^2 - k_y^2}} \quad \text{(B.6)}$$

And

$$b_s = \frac{k}{\pi\eta}\int_k^\infty \frac{\sin 2(k_y S/2)}{(k_y S/2)^2}\frac{dk_y}{\sqrt{k^2 - k_y^2}} \quad \text{(B.7)}$$

The above set equations could be expressed as the double integrals of the Bessel function of the kind Jo (v) and second kind Yo (v) of order zero as follows:

$$g_s = \frac{k}{\eta s^2}\int_0^{ks}\int_0^u J_0(v)dvdu \quad \text{(B.8)}$$

And

$$b_s = -\frac{k}{\eta s^2}\int_0^{ks}\int_0^u Y_0(v)dvdu \quad \text{(B.9)}$$

The double integrals can be expanded into series form and after, omitting terms of fourth order and higher, will simplify into:

$$g_s \approx \frac{k}{2\eta}\left(1 - \frac{s^2}{24}\right) \quad \text{(B.10)}$$

And

$$b_s \approx \frac{k}{\pi\eta}\left\{\left(\ln\left(\frac{s}{2}\right) + 0.577216 - \frac{3}{2}\right)\left(1 - \frac{s^2}{24}\right) + \frac{s^2}{288}\right\} \quad \text{(B.11)}$$

Where, s = kS.

Green's Function of a Hertzan Dipole in a Grounded substrate

Following the derivation of [28]:

The expression for the electric field, in terms of the Green's functions and current distribution is

$$E_x(x, y) = \iint G_{xx} J_x(x_s, y_s) dy_s dx_s + \iint G_{xy} J_y(x_s, y_s) dy_s dx_s \qquad \text{(B.12)}$$

And

$$E_y(x, y) = \iint G_{yx} J_x(x_s, y_s) dy_s dx_s + \iint G_{yy} J_y(x_s, y_s) dy_s dx_s \qquad \text{(B.13)}$$

Where $G_{xx}, G_{xy}, G_{yx} and, G_{yy}$ are the spatial-domain dyadic Green's functions which can be obtained by the inverse Fourier transform of the spectral-domain Green's functions as follows:

$$G_{ij}(x, y; x_s, y_s) = \frac{1}{r_o} \iint \tilde{G}_{ij}(k_x, k_y) e^{-jk_x(x, x_s)} e^{-jk_y(y, y_s)} dk_x dk_y \qquad \text{(B.14)}$$

Where i and j are either x or y and $\tilde{G}_{ij}$ is the spectral domain Green's function given as follows

$$\tilde{G}_{xx}(k_x, k_y) = \frac{1}{\varepsilon_2}\left[\frac{k_o^{\,2} n_o^{\,2} - k_x^{\,2}}{D_e(k)} A_1(k) - \frac{k_x^{\,2}}{D_e(k) D_m(k)} B_1(k)\right] \qquad \text{(B.15)}$$

And

$$\tilde{G}_{yy}(k_x, k_y) = \frac{1}{\varepsilon_2}\left[\frac{k_o^{\,2} n_2^{\,2} - k_y^{\,2}}{D_e(k)} A_1(k) - \frac{k_y^{\,2}}{D_e(k) D_m(k)} B_1(k)\right] \qquad \text{(B.16)}$$

And

$$\tilde{G}_{xy}(k_x,k_y)=\frac{-k_xk_y}{\varepsilon_2}\left[\frac{A_1(k)}{D_e(k)}+\frac{B_1(k)}{D_e(k)D_m(k)}\right]=\tilde{G}_{yx}(k_x,k_y) \qquad \text{(B.17)}$$

Where

$$D_e(k)=q_2\left(q\sinh q_1b+\frac{q_1}{\mu_1}\cosh q_{1b}\right)\cosh q_2(h-b)+$$
$$q_2\left(\frac{q_2}{\mu_2}\sinh q_1b+\frac{\mu_2qq_1}{\mu_1q_2}\cosh q_1b\right)\sinh q_2(h-b) \qquad \text{(B.18)}$$

$$D_m(k)=\frac{q_2n_2^{\;2}}{\mu_2}\left(q_1\eta_1\sinh q_1b+qn_1^{\;2}\cosh q_1b\right)\cosh q_2(h-b)+$$
$$\left(q_2^{\;2}n_1^{\;2}\cosh q_1b+\frac{n_2^{\;4}\mu_1qq_1}{\mu^2{}_2}\sinh q_2b\right)\sinh q_2(h-b) \qquad \text{(B.19)}$$

And

$$A_1(k)=\left[\frac{q_1}{\mu_1}\sinh q_1b\cosh q_2(h-z_s)+\frac{\mu_2qq_1}{\mu_1q_2}\cosh q_1b\sinh q_2(h-z_s)\right]\sinh q_2(z_s-b)+$$
$$\left[\left(q\sinh q_2(h-z_s)+\frac{q_2}{\mu_2}\cosh q_2(h-z_s)\right)\sinh q_1b\cosh q_2(z_s-b)\right]\sinh q_1b\cosh q_2(z_s-b)$$

$$B_1(k)=q_2\Big\{\left(n_2^{\;2}-n_1^{\;2}\right)\Big[\left(q\sinh q_2(h-z_s)+\frac{q_2}{\mu_2}\cosh q_2(h-z_s)\right)\left(\frac{qn_2^{\;2}}{\mu_2}\cosh q_2(h-z_s)+\right.$$
$$\left.q_2\sinh q_2(h-z_s)\right)\sinh q_1b\cosh q_1b-\left(1-n_2^{\;2}\right)\left[\frac{q_1}{\mu_1}\cosh q_1b\sinh q_2(z_s-b)+\frac{q_2}{\mu_2}\sinh q_1b\cosh q_2(z_s-b)\right]$$
$$\left[q_2n_1^{\;2}\cosh q_1b\sinh q_2(z_s-b)+\frac{\mu_1q_1n_2^{\;2}}{\mu_2}\sinh q_1b\cosh q_2(z_s-b)\right]\Big]\Big\}$$

(B.20)

Where,

$$k = \sqrt{k_x^{\ 2} - k_y^{\ 2}} \qquad \text{(B.21a)}$$

$$q = \sqrt{k^2 - k_o^{\ 2}} \qquad \text{(B.21b)}$$

$$q_1 = \sqrt{k^2 - k^2{}_1} \qquad \text{(B.21c)}$$

$$q_2 = \sqrt{k^2 - k^2{}_2} \qquad \text{(B.21d)}$$

$$n_1 = \sqrt{\mu_1 \varepsilon_1} \qquad \text{(B.21e)}$$

$$n_2 = \sqrt{\eta_2 \varepsilon_2} \qquad \text{(B.21f)}$$

$$k_o = \omega\sqrt{\mu_o \varepsilon_o} \qquad \text{(B.21g)}$$

$$k_1 = k_o n_1 \qquad \text{(B.21h)}$$

$$k_2 = k_o n_2 \qquad \text{(B.21i)}$$

Appendix C

C. 1 parameter definitions

There are several parameters, which are commonly used to describe the performance criteria of an antenna or an array of antenna elements. The antenna engineer's goal is to realize the design that meets those specifications. These parameters such as: gain, input impedance, bandwidth, beam width, etc...Will be defined in this section.

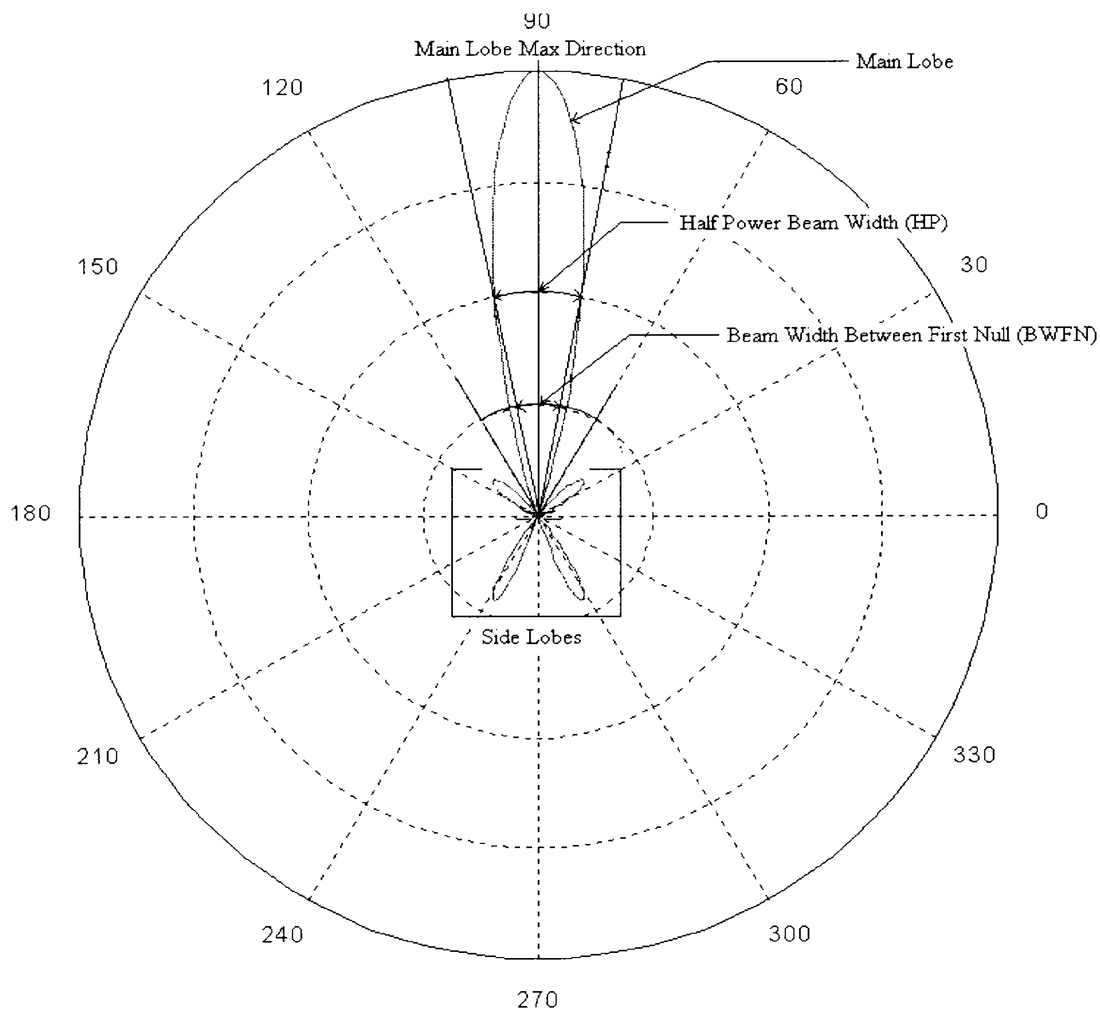


Figure c.1.1 A typical antenna radiation pattern

Figure C.1.1 shows a typical far field radiation pattern from an antenna. It has all the components that could be used to describe the beam. The radiation pattern is typically plotted or rectangular coordinates, the angle represent the angular position, at a given radius, which usually in the far field, around a fixed point of in the antenna; the r-axis represents the intensity measured. In rectangular coordinates, the x-axis is usually the angle and the y-axis is the intensity. The graphs are usually in the logarithmic units of decibels

The pattern displayed in the polar graph may consist of a number of lobes. The lobe with the larger radius is considered as the main lobe.

The lobes directly in back of it are called the back lobes and the ones to the side are called side lobes. The radiation pattern may be represented by a number of different methods, one of which is a cross section, as represented by figure c.1.1, another is 3-D plot, and many other exist. The coordinates and, convention has it, is given by F(θ,ϕ) or P(θ,ϕ) for power.

A couple of other definitions are associated with the radiation patterns they will be defined next. The side lobe level is the amount of power concentrated at the side lobe compared to main lobe and is defined as

$$SLL_m = 20Log_{10}\left|\frac{F(SLL)}{F(MAX)}\right| \qquad (C.1.1)$$

Where, F(SLL) is the maximum side lobe level and F(MAX) is the maximum level at the main lobe.

The width of the main lobe is usually indicated in a quantity called half power beam width (HP) which is defined in a plane containing the direction of the maximum of a beam, the angle between the two directions in which the radiation intensity is one-half(-3dB) of its maximum value.

An auxiliary quantity often used to evaluate directivity and also gain is called the radiation intensity.

$$U(\theta,\phi)=\frac{1}{2}\mathrm{Re}\{\vec{E}\times\vec{H}^*\}\cdot .r^2\hat{r}=U_m|F(\theta,\phi)|^2 \qquad \text{(c.1.2)}$$

Where U (θ,ϕ) is the radiation intensity, $\vec{E}$ is the electric field vector and $\vec{H}^*$ is the conjugate of the magnetic field vector. U_m is the maximum radiation intensity.

The average radiation intensity can be defined as the integral of the radiation intensity over solid angleΩ, which is also related to the total power p as follows

$$U_{avg}=\frac{1}{4\pi}\oiint_S U(\theta,\phi)d\Omega=\frac{P}{4\pi} \qquad \text{(C.1.3)}$$

Directivity is now defined as the proportion of radiation intensity in a given direction to the average radiation intensity

$$D(\theta,\phi)=\frac{U(\theta,\phi)}{U_{avg}}=\frac{4\pi}{\Omega_A}|F(\theta,\phi)|^2 \qquad \text{(C.1.4)}$$

And

$$\Omega_A=\iint|F(\theta,\phi)|^2d\Omega \qquad \text{(C.1.5)}$$

Maximum directivity is then given as

$$D=\frac{U_m}{U_{avg}}=\frac{4\pi}{\Omega_A} \qquad \text{(C.1.6)}$$

A quantity similar to directivity but conveys a real practical meaning is called the power gain of antenna.

The power gain is a measure of how efficient the antenna is as a radiator, as it is a ratio of the radiation intensity to the input power

$$G(\theta,\phi)=\frac{4\pi U(\theta,\phi)}{P_m}=\frac{e_r U(\theta,\phi)}{U_{avg}} \qquad \text{(C.1.7)}$$

Where e_r is the efficiency or the proportion of power radiated to the input power.

Another related quantity is the maximum gain, which is defined similar to the maximum directivity:

$$G = e_r D \qquad \text{(C.1.8)}$$

Where G is maximum gain and D is the maximum directivity.

A quantity, which is related directly to the special behavior of the antenna radiation and its orientation, is the antenna polarization state. There are four main polarization states, others are just a super position of those four, a visual description is shown in figure C.1.1.

An antenna, just like any other device can be translated into an electric network. For the purpose of this section, the antenna will be viewed as a one port network. If one looks into that port, which shall be designated as port 1, there will be an impedance which will consists of a real and imaginary parts $Z = R + jX$, where the reactance, x, conveys the power lost due to ohmic losses and radiation losses. In an antenna design the desire is to maximize the radiative power and minimize the stored and ohmic losses.

The impedance of an antenna can be viewed in another, more intuitive way. The one port network can be described in terms of its scattering parameters matrix. Power going into 1 can either continue into the antenna or reflect back. If power reflects back, that power is not radiated. The power that goes into the port either gets absorbed or radiated but it is not possible to differentiate the absorbed and radiated power from the scattering coefficient alone. These statements could be represented mathematically as:

$$P_{in} - P_{reflected} = P_{radiated} + P_{ohmic} \qquad \text{(C.1.9)}$$

In an antenna, the reflected power is to minimized while the radiative absorption is to be maximized, but again, ohmic absorption is to be minimized.

MATLAB Code

Variable_Width_Patch_Antenna.m

```
clear all; close all;
%Define Constants
er = 4.8; %Relative dielectric constant of substrate
mu0 = 4*pi*10^-7;
e0 = 8.854*10^-12;
f = 1.5*10^9; %Center frequency of antenna (Hz)
h = 0.00159 %Thickness of substrate (m)
lambda0 = 1/ (sqrt(mu0*er*e0)*f); %Wavelength (m)
%Vary Width from 4cm to 20cm
W = 0.04:0.00001:0.2;
 %Relative Dielectric Constant (from Empirically-Determined Formula [1])
ereff = (er + 1)/2 + ((er - 1)/2).*(1 + 12.*(h./W)).^-0.5;
%Characteristic Impedance of Transmission Line
Fg = 2.*W./h + 2/pi + (2./pi).*log(1 + pi.*W./h);
Z01 = sqrt(mu0./(ereff.*e0))./Fg;
%Conductance of Rectangular Aperture
GL = (W./(120.*lambda0)).*(1 -
(1/24).*(2.*pi.*f.*sqrt(mu0.*e0).*h).^2);
%Feed point Impedance of Patch Antenna
Zaa = 0.25.*((1./GL) + Z01.^2.*GL);
plot(W.*100,Zaa)
xlabel('Width (cm)')
ylabel('Feed point Impedance (Ohms)')
Title('Feed point Impedance vs. Width for 1.5GHz Rectangular Patch
Antenna')
```

MATHLAB CODE

Depletion capacitance of varactor diode

```
Cj1=4.5e-12; Vs1= -10;
Cj2=6.5e-12; Vs2=-3;
Vc=0.65
Num=Cj1/CJ2;
```

```
Den=(Vc-Vs2)/(Vc-Vs1);
M==log10(Num)/log10(Den);
Cjo=Cj1*(1-(Vs1/Vc))^M;
Vs=-40:0.2:40;
K= length(Vs);
For i=1:k
   Cj(i)=Cjo/(1-(Vs(i)/Vc))^M;
end
plot  (Vs, Cj, ‘b’)
x label ( ‘voltage’)
y label (‘capacitance, F’)
Title (‘depletion capacitance Vs. voltage’)
Axis ([-30, 2, e-12,14e-12])
```

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