DESIGN & ANALYSIS OF DUAL FREQUENCY ANGULAR RING MICROSTRIP ANTENNA WITH PERTURBATION METHOD USING MAPLESOFT

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Submitted By

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CERTIFIFICATE

This is to certify that the report entitled, "**DESIGN & ANALYSIS OF DUAL FREQUENCY ANGULAR RING MICROSTRIP ANTENNA WITH PERTURBATION METHOD USING MAPLESOFT"** submitted by **Mr. KUMAR GAURAV(Roll No.-8518)** in partial fulfillment of the requirements for the award of Master of Engineering Degree in **ELECTRONICS AND CMMUNICATION ENGINEERING** at Delhi College Of Engineering(University Of Delhi) is an authentic work carried out by him under my supervision and guidance. To the best of my/our knowledge, the matter embodied in the thesis has not been submitted to any other University/ Institute for the award of any degree.

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CONTENTS

ABSTRACT

Angular Ring Microstrip Antenna (ARMA) with perturbation at its centre is designed and is studied for its frequency characteristics .A microstrip antenna in its simplest configuration consists of radiating patch on one side of a dielectric substrate (ε_r) ,which has a ground plane on the other side. However they suffer from a drawback that surface waves are formed .These waves reduce antenna efficiency and gain, limit bandwidth, increase end-fire radiation, and limit the applicable frequency antenna.

In 1977 , L. Lewin initiated the concept of analyzing microstrip curve structure with the help of perturbation method .These structure were extensively used in the optical range .However the increasing research and application in the Microwave range led to the frequency of operation to shift into the GHz range. These structures operating in the microwave region are called curve waveguide model. By using second order perturbation radiating power improve up to -60 dB.

The project deals with design and analysis of Angular Microstrip Antenna (ARMA) for making perturbed structure, a curved microstrip bend at any general angle α. Compute perturbation of chamfered α region with respect to α region. Use of second order perturbation & fourth iteration the frequency response of antenna improve. Mathematical software MAPLESOFT was used to solving such equation and MATLAB used for comparing the result. Further a study of the effect of different R/w and angle α on the performance of the antenna is presented and discussed in this thesis.

List of Figure and Table

Chapter 2

Chapter 3

Chapter 4

Chapter 5

ABBREVIATIONS

ARMA: Angular Rectangular Microstrip Antenna.

CMB: Chamfered Microstrip Bend

SOPM: Second Order Perturbation Method

CHAPTER -1

INTRODUCTION

Satellite communication and Wireless communication has been developed rapidly in the past decades and it has already a dramatic impact on human life. In the last few years, the development of wireless local area networks (WLAN) represented one of the principal interests in the information and communication field. Thus, the current trend in commercial and government communication systems has been to develop low cost, minimal weight, low profile antennas that are capable of maintaining high performance over a large spectrum of frequencies. This technological trend has focused much effort into the design of Microstrip (patch) antennas. With a simple geometry, patch antennas offer many advantages not commonly exhibited in other antenna configurations. For example, they are extremely low profile, lightweight, simple and inexpensive to fabricate using modern day printed circuit board technology, compatible with microwave and millimeter-wave integrated circuits (MMIC)[3], and have the ability to conform to planar and non planar surfaces. In addition, once the shape and operating mode of the patch are selected, designs become very versatile in terms of operating frequency, polarization, pattern, and impedance. The variety in design that is possible with Microstrip antenna probably exceeds that of any other type of antenna element.

Using the perturbeted Microstrip Antenna concept in this Dissertation second order perturbated Angular Microstrip antenna is designed with the help of MAPLESOFT one of the most imperial mathematical software which allows to solving complex mathematical differential equation symbolically for radio and microwave application.

1.1 Literature Review

The invention of Microstrip patch antennas has been attributed to several authors, but it was certainly dates in the 1960s with the first works published by Deschamps, Greig and Engleman, and Lewin, among others. After the 1970s research publications started to flow with the appearance of the first design equations. Since then different authors started investigations on Microstrip patch antennas like James Hall and David M. Pozar and there are also some who contributed a lot. Throughout the years, authors have dedicated their investigations to creating new designs or variations to the original antenna that, to some extent; produce either wider bandwidths[2] or multiple-frequency operation in a single element. However, most of these innovations bear disadvantages related to the size, height or overall volume of the single element and the improvement in bandwidth suffer usually from a degradation of the other characteristics. It is the purpose of this thesis to introduce the perturbation techniques produced to improve the narrow bandwidth and low frequency ratio of patch antennas. MAPLESOFT is a symbolic based, partial differential equation simulator solving the current distrbution on 3D and multilayer structures of general shape. Different optimization schemes are available in MATLAB, including Powell optimizer, genetic optimizer, adaptive optimizer, and random optimizer. The variables for optimization defined by MATLAB are controlled by its directions and bounds. In 1995 James Kennedy and Russell Eberhart presented particle-swarm optimization (PSO), an optimizer that models the behavior and intelligence of a swarm of bees, school of fish or flock of birds, and emphasizes both social interaction and nostalgia from the individual's perspective.

1.2 Dissertation Motivation

With bandwidths as low as a few percent, broadband applications using conventional .Microstrip patch designs are limited. Other drawbacks of patch antennas include low efficiency, limited power capacity, spurious feed radiation, poor polarization purity, narrow bandwidth, and manufacturing tolerance problems. Forover two decades, research scientists have developed several methods to increase the bandwidth and low frequency ratio of a patch antenna. Many of these techniques involve adjusting the placement and/or type of element used to feed (or excite) the antenna. Perturbation method is the mathematical method based on Numerical Method technique. Its result is improved by based on iterative process. In satellite communication, antennas with low frequency ratio are very much essential.

1.3 Dissertation Outline

The outline of this thesis is as follows.

Chapter 2: It presents the basic theory of MSAs, including the basic geometries, feeding method and characteristics of the MSA, the advantages and disadvantages of MSAs, impedance matching, shorting techniques, the methods of analysis used for the MSA .

Chapter 3: In this chapter the basics of antenna parameters such as radiation pattern, impedance, VSWR, gain etc. are presented.

Chapter 4: This chapter describes the perturbation method its back ground & how to use in electromagnetic.

Chapter 5: This chapter describes the design of Angular Ring Microstrip patch antenna and simulation result using the MAPLE mathematical simulation software. The simulation results have been presented.

Chapter 6: This chapter contains conclusion and scope of future work.

CHAPTER-2

MICROSTRIP ANTENNA

One of the most exciting developments in antenna and electromagnetic history is the advent of Microstrip antenna (known also as patch antenna). It is probably the most versatile solution to many systems requiring planner radiating element. Microstrip antenna falls into the category of printed antennas: radiating elements that utilize printed circuit manufacturing processes to develop the feed and radiating structure. Of all the printed antennas, including dipole, slots, and tapered slots; Microstrip antenna is by far the most popular and adaptable. This is because of all its salient features: including ease of fabrication, good radiation control, and low cost of production.

The Microstrip antenna is constructed from dielectric substrate and patch metal and that a portion of the metallization layer is responsible for radiation. Microstrip antenna was conceived in the 1950s, and then extensive investigations of the patch antennas followed in the 1970s and resulted in many useful design configurations. Through decades of research, it was identified that the performance and operation of a Microstrip antenna is driven mainly by the geometry of the printed patch and the material characteristics of the substrate onto which the antenna is printed.

A microstrip antenna consists of conducting patch on a ground plane separated by dielectric substrate. This concept was undeveloped until the revolution in electronic circuit miniaturization and large-scale integration in 1970. After that many authors have described the Radiation from the ground plane by a dielectric substrate for different configurations.

The early work of Munson on microstrip antennas for use as a low profile flush mounted antennas on rockets and missiles showed that this was a practical concept for use in many antenna system problems. Various mathematical models were developed for this antenna and its applications were extended to many other fields. The number of papers, articles published in the journals for the last ten years, on these antennas shows the importance gained by them. The micro strip antennas are the present day antenna designer’s choice. Low dielectric constant substrates are generally preferred for maximum radiation. The Communication between humans was first by sound through voice. With the desire for slightly more distance communication came, devices such as drums, then, visual methods such as signal flags and smoke signals were used. These optical communication devices, of course, utilized the light portion of the electromagnetic spectrum. It has been only very recent in human history that the electromagnetic spectrum, outside the visible region, has been employed for communication, through the use of radio. One of humankind's greatest natural resources is the electromagnetic spectrum and the antenna has been instrumental in harnessing this resource. Microstrip antennas are attractive due to their light weight, conformability and low cost. These antennas can be integrated with printed strip-line feed networks and active devices. This is a relatively new area of antenna engineering. The radiation properties of micro strip structures have been known since the mid 1950's.

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

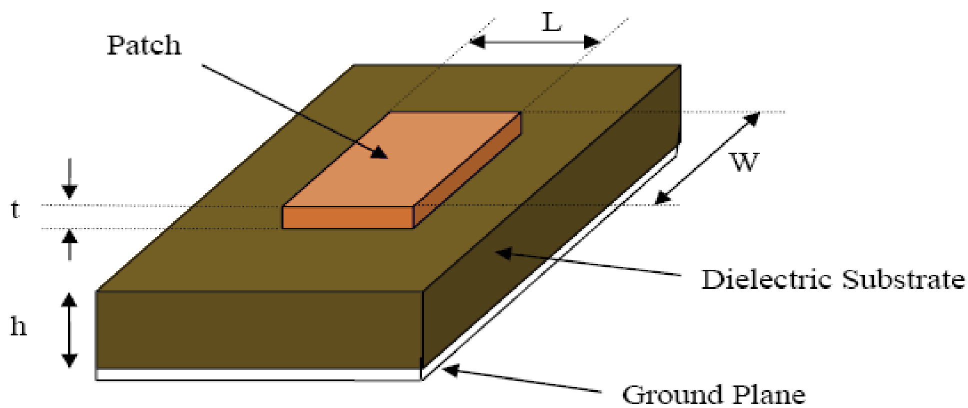


Figure .1 Structure of rectangular Microstrip Patch Antenna

In its most fundamental form, a Microstrip Patch antenna consists of a radiating patch on one side of a dielectric substrate which has a ground plane on the other side as shown in Figure (1). The patch is generally made of conducting material such as copper or gold and can take any possible shape. The radiating patch and the feed lines are usually photo etched on the dielectric substrate.

. A microstrip antenna is characterized by its Length, Width, Input impedance, and Gain and radiation patterns. Various parameters of the microstrip antenna and its design considerations were discussed in the subsequent chapters. The length of the antenna is nearly half wavelength in the dielectric; it is a very critical parameter, which governs the resonant frequency of the antenna. There is no hard and fast rule to find the width of the patch

In order to simplify analysis and performance prediction, the patch is generally square, rectangular, circular, triangular, and elliptical or some other common shape as shown in Figure 2. For a rectangular patch, the length L of the patch is usually $0.3333\lambda_o < L < 0.5\ \lambda_o$, where λ_o *is* the free-space wavelength. The patch is selected to be very thin such that $t << \lambda_o$ (where t is the patch thickness). The height of the dielectric substrate is usually $0.003\ \lambda_o \leq h \leq 0.05\ \lambda_o$. The dielectric constant of the substrate (ε_r) is typically in the range $2.2 \leq \varepsilon_r \leq 12$.

2.1 Different Shape of Microstrip patch Antenna

Conducting patch can take any shape but rectangular and circular configurations are the most commonly used configuration. Other configurations are complex to analyze and require heavy numerical computations.

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

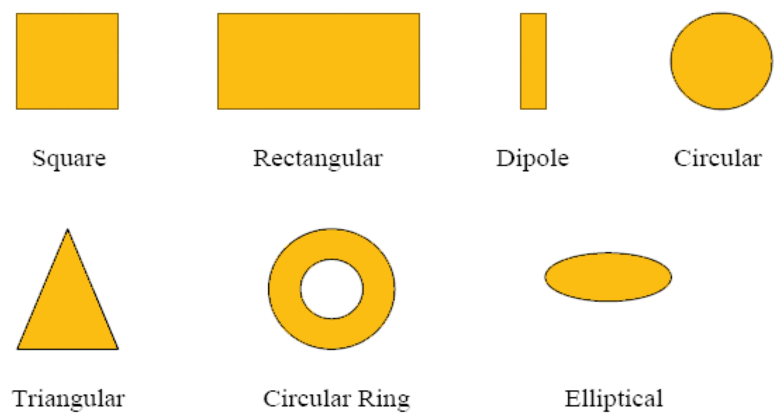


Fig.2 Different Shape of Microstrip Patch Antenna

Dual characteristics are exhibited by rectangular patch having square notch at its centre. Compactness of antenna is also achieved by having square notch at centre as it reduces the size of antenna by 17% of the conventional antenna without slot. But these types of antennas suffer from drawback of excitation of surface waves which leads to lower gain and antenna efficiency. Introduction of a prohibiting the propagation of all the electromagnetic waves of certain band of frequencies. Electromagnetic Band Gap (EBG) structure in the ground plane reduces the excitation of surface waves. The EBG structure consists of a uniformly distributed periodic metallic pattern on one side of a dielectric slab which is capable of eliminating surface wave.

2.2 Advantages and Disadvantages of Microstrip Antenna

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

2.2.1 Advantages

- Light weight and low volume.
- Low profile planar configuration which can be easily made conformal to host surface.
- Low fabrication cost, hence can be manufactured in large
- Supports both, linear as well as circular polarization
- Can be easily integrated with microwave integrated circuits
- Capable of dual and triple frequency operations.
- Mechanically robust when mounted on rigid surfaces.
- They can be easily mounted on missiles, rockets and satellites without major alterations.
- Linear, circular polarizations are possible by simply changing the feed positions.
- Dual frequency antennas can easily be made.

2.2.2 Disadvantages

- Narrow bandwidth
- Low efficiency
- Low Gain
- Extraneous radiation from feeds and junctions

- Poor end fire radiator except tapered slot antennas

- Low power handling capacity.
- Surface wave excitation

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

2.3 Applications of Mcrostrip Antenna

- Satellite Communication
- Doppler and other radars
- Command and Control
- Environmental instrumentation and remote sensing
- Feed elements in complex antennas
- Wireless LANs
-

2.4 Method of Analysis

In common practice, microstrip antennas are evaluated using one of three analysis methods: the transmission line model , the cavity model, and the full-wave model. The transmission line model is the easiest of all, it gives good physical insight. But it is less accurate and more difficult to model coupling effect of antenna. Compared to the transmission line model, the cavity model is more accurate but at the same time more complex and difficult to model coupling effect. In general, when applied properly, the full wave model is very accurate, and very versatile.

2.6.1 Transmission Line Model

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

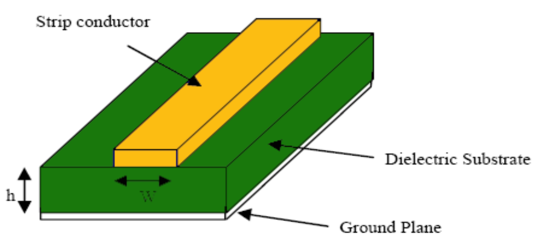


Figure .3 Microstrip Line

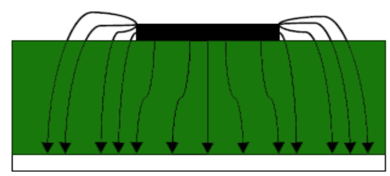


Figure .4 Electric Field Lines

Hence, as seen from Figure-9, most of the electric field lines reside in the substrate and parts of some lines in air. As a result, this transmission line cannot support pure transverse-electric- magnetic (TEM) mode of transmission, since the phase velocities would be different in the air and the substrate. Instead, the dominant mode of propagation would be the quasi-TEM mode. Hence, an effective dielectric constant (ε_{reff}) must be obtained in order to account for the fringing and the wave propagation in the line. The value of ε_{reff} is slightly less than ε_r because the fringing fields around the periphery of the patch are not confined in the dielectric substrate but are also spread in the air as shown in Figure -.8 above. The expression for ε_{reff} is given by Balanis as:

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

ε_r = Dielectric constant of substrate

h = Height of dielectric substrate

W = Width of the patch

Consider Figure -10 below, which shows a rectangular microstrip patch antenna of length L, width W resting on a substrate of height h. The co-ordinate axis is selected such that the length is along the x direction, width is along the y direction and the height is along the z direction.

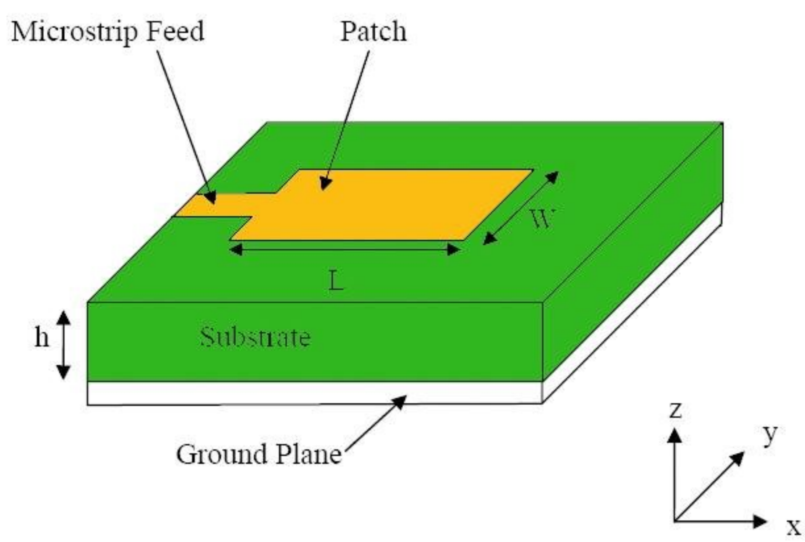
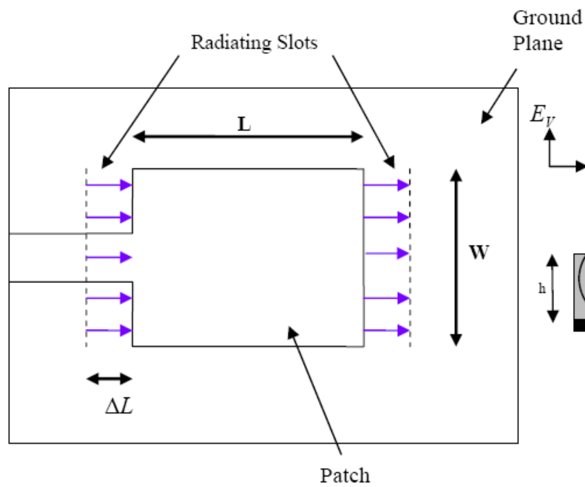


Figure.5 Microstrip Patch Antennas

In order to operate in the fundamental TM_{10} mode, the length of the patch must be slightly less than $\lambda/2$ where λ is the wavelength in the dielectric medium and is equal to $\lambda_o/\sqrt{\varepsilon_{reff}}$

where λ_o *is* the free space wavelength. The $TM10$ mode implies that the field varies one $\lambda/2$ cycle along the length, and there is no variation along the width of the patch. In the Figure -11 shown below, the microstrip patch antenna is represented by two slots, separated by a transmission line of length L and open circuited at both the ends.



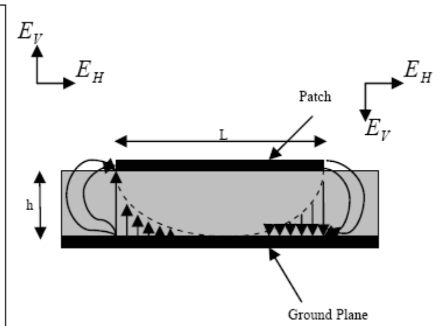


Figure .6 Top View of Antenna ***Figure .7 Side View of Antenna***

Along the width of the patch, the voltage is maximum and current is minimum due to the open ends. The fields at the edges can be resolved into normal and tangential components with respect to the ground plane.

It is seen from Figure -12 that the normal components of the electric field at the two edges along the width are in opposite directions and thus out of phase since the patch is $\lambda/2$ long and hence they cancel each other in the broadside direction. The tangential components (seen in Figure -12), which are in phase, means that the resulting fields combine to give maximum radiated field normal

To the surface of the structure. Hence the edges along the width can be represented as two radiating slots, which are $\lambda/2$ apart and excited in phase and radiating in the half space above the ground plane. The fringing fields along the width can be modeled as radiating slots and electrically the patch of the microstrip antenna looks greater than its physical dimensions. The dimensions of the patch along its length have now been extended on each end by a distance ΔL, which is given empirically by Hammerstad as:

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

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

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

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

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

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

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

Where m and n are modes along L and W respectively.

For efficient radiation, the width W is given by Bahl and Bhartia [15] as:

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

2.6.2 The Cavity Model

Microstrip antenna resembles dielectric loaded cavities, and exhibit higher order resonances. The normalized fields within the dielectric substrate can be found more accurately by treating that region as a cavity bounded by electric conductors (above and below) and by magnetic wall along the perimeter of the patch.

The bases for this assumption are the following points (for height of substrate h<< λ wavelength of wave)

- The fields in the interior region do not vary with z-axis because the substrate is very thin, h<< λ.
- The electric field is z-axis directed only, and the magnetic field has only the transverse components in the region bounded by the patch metallization and the ground plane. This observation provides for the electric walls at the top and bottom.
- The electric current in the patch has no component normal to the edge of the patch metallization, which implies that the tangential component of magnetic field along the edge is negligible, and a magnetic wall can be placed along the periphery.

This approximation model leads to reactive input impedance, and it does not radiate any power. However, the actual fields can be approximated to the generated field of the model and is possible to analyze radiation pattern, input admittance, and resonant frequency

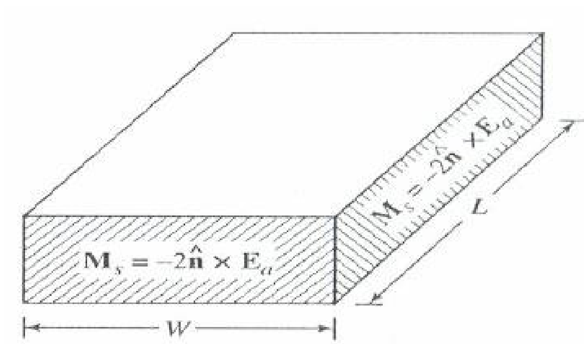


Figure.8 Cavity model of rectangular Microstrip antenna

2.7 Waves on Microstrip Antenna

The mechanisms of transmission and radiation in a microstrip can be understood by considering a point current source (Hertz dipole) located on top of the grounded dielectric substrate (fig.14) This source radiates electromagnetic waves. Depending on the direction toward which waves are transmitted, they fall within three distinct categories, each of which exhibits different behaviors.

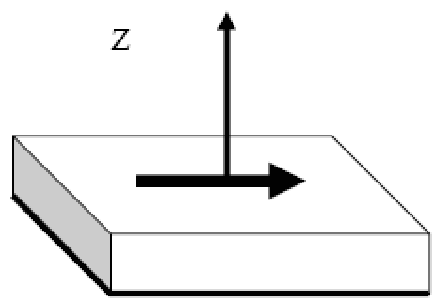


Fig-9 Hertz dipole on Microstrip Substrate

2.7.1 Surface Waves

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

The fields remain mostly trapped within the dielectric, decaying exponentially above the interface (fig -15). The vector α, pointing upward, indicates the direction of largest attenuation. The wave propagates horizontally along β, with little absorption in good quality dielectric. With two directions of α and β orthogonal to each other, the wave is a non-uniform plane wave. Surface waves spread out in cylindrical fashion around the excitation point, with field amplitudes decreasing with distance (r), say $1/r$, more slowly than space waves. The same guiding mechanism provides propagation within optical fibers. Surface waves take up some part of the signal's energy, which does not reach the intended user. The signal's amplitude is thus reduced, contributing to an apparent attenuation or a decrease in antenna efficiency. Additionally, surface waves also introduce spurious coupling between different circuit or antenna elements. This effect severely degrades the performance of microstrip filters because the parasitic interaction reduces the isolation in the stop bands.In large periodic phased arrays, the effect of surface wave coupling becomes particularly obnoxious, and the array can neither transmit nor receive when it is pointed at some particular directions (blind spots). This is due to a resonance phenomenon, when the surface waves excite in synchronism the Floquet modes of the periodic structure. Surface waves reaching the outer boundaries of an open microstrip structure are reflected and diffracted by the edges. The diffracted waves provide an additional contribution to radiation, degrading the antenna pattern by raising the side lobe and the cross polarization levels. Surface wave effects are mostly negative, for circuits and for antennas, so their excitation should be suppressed if possible.

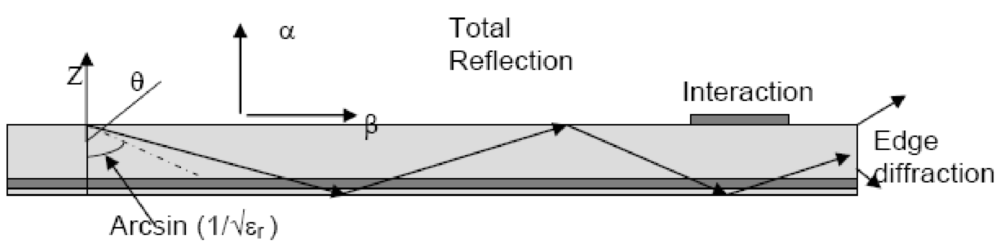


Fig-10 Surface Wave

2.7.2 Leaky Waves

Waves directed more sharply downward, with θ angles between π - arcsin ($1/\sqrt{\varepsilon_r}$) and π, are also reflected by the ground plane but only partially by the dielectric-to-air boundary. They progressively leak from the substrate into the air (Fig 1.3), hence their name laky waves, and eventually contribute to radiation. The leaky waves are also non-uniform plane waves for which the attenuation direction α points downward, which may appear to be rather odd; the amplitude of the waves increases as one moves away from the dielectric surface. This apparent paradox is easily understood by looking at the figure-16; actually, the field amplitude increases as one move away from the substrate because the wave radiates from a point where the signal amplitude is larger. Since the structure is finite, this apparent divergent behavior can only exist locally, and the wave vanishes abruptly as one crosses the trajectory of the first ray in the figure.

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

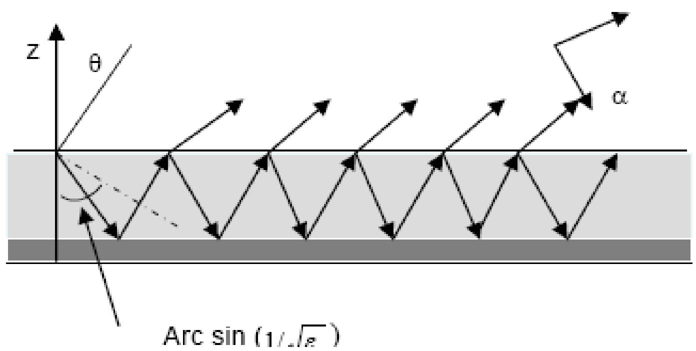


Fig-11 Leaky waves

2.7.3 Guided Waves

When realizing printed circuits, one locally adds a metal layer on top of the substrate, which modifies the geometry, introducing an additional reflecting boundary. Waves directed into the dielectric located under the upper conductor bounce back and forth on the metal boundaries,which form a parallel plate waveguide. The waves in the metallic guide can only exist for some Particular values of the angle of incidence, forming a discrete set of waveguide modes. The guided waves provide the normal operation of all transmission lines and circuits, in which the electromagnetic fields are mostly concentrated in the volume below the upper conductor.

CHAPTER-3

ANTENNA PARAMETERS

To describe the performance of an antenna, some definitions of various parameters are necessary. These parameters are interrelated and not all of them need to be specified for complete description of the antenna performance. Following parameters are being defined as follows:

3.1 Gain and directivity

The gain of an antenna is the radiation intensity in a given direction divided by the radiation intensity that would be obtained if the antenna radiated all of the power delivered equally to all directions. The definition of gain requires the concept of an isotropic radiator; that is, one that radiates the same power in all directions. An isotropic antenna, however, is just a concept, because all practical antennas must have some directional properties. Nevertheless, the isotropic antenna is very important as a reference. It has a gain of unity (g = 1 or G = 0 dB) in all directions, since all of the power delivered to it is radiated equally well in all directions.

Antenna gain Relates the intensity of an antenna in a given direction to the intensity that would be produced by a hypothetical ideal antenna that radiates equally in all directions (isotropically) and has no losses. Since the radiation intensity from a lossless isotropic antenna equals the power into the antenna divided by a solid angle of 4π steradians, we can write the following equation:

$$Gain = 4\pi\left(\frac{\text{Radiation Intensity}}{\text{Antenna Input Power}}\right)$$

$$Gain = 4\pi\left(\frac{\mathrm{U}\,(\theta,\phi)}{\mathrm{Pin}}\right) \qquad \text{Dimensionless Units.}$$

Although the gain of an antenna is directly related to its directivity, the antenna gain is a measure that takes into account the efficiency of the antenna as well as its directional capabilities. In contrast, directivity is defined as a measure that takes into account only the directional properties of the antenna and therefore it is only influenced by the antenna pattern. However, if we assumed an ideal antenna without losses then Antenna Gain will equal directivity as the antenna efficiency factor equals 1 (100% efficiency). In practice, the gain of an antenna is always less than its directivity.

Although the isotopes are a fundamental reference for antenna gain, another commonly used reference is the dipole. In this case the gain of an ideal (lossless) half wavelength dipole is used. Its gain is 1.64 (G = 2.15 dB) relative to an isotropic radiator.The gain of an antenna is usually expressed in decibels (dB). When the gain is referenced to the isotropic radiator, the units are expressed as dBi; but when referenced to the half-wave dipole, the units are expressed as dBd.

3.2 Antenna Polarization

The term polarization has several meanings. In a strict sense, it is the orientation of the electric field vector E at some point in space. If the E-field vector retains its orientation at each point in space, then the polarization is linear; if it rotates as the wave travels in space, then the polarization is circular or elliptical. In most cases, the radiated-wave polarization is linear and either vertical or horizontal. At sufficiently large distances from an antenna, beyond 10 wavelengths, the radiated, far-field wave is a plane wave.

Polarization or plane of polarization of a radio wave can be defined by the direction in which the electric vector E is aligned during the passage of at least one full cycle .polarization refers to the physical orientation of the radiated electromagnetic waves in space .A electromagnetic wave is said to be linearly polarized if they all have the same alignment in space.

In circular polarized antenna ,the plane of polarization rotates in a corkscrew pattern making one complete revolution during each wavelength .A circularly-polarized wave radiates energy the horizontal ,vertical planes as well as every plane in between. If the rotation is clockwise looking in the direction of propagation, the sense is called Right-hand-circular (RHC) .If the rotation is counterclockwise, the sense is called left-hand-circular (LHC) .

3.3 Input impedance

There are the different kinds of impedance relevant to antenna. One is the terminal impedance of the antenna, another is the characteristic impedance of a transmission line, and the third is wave impedance. Terminal impedance[2] is defined as the ratio of voltage to current at the connections of the antenna (the point where the transmission line is connected). The complex form of Ohm's law defines impedance as the ratio of voltage across a device to the current flowing through it. The most efficient coupling of energy between an antenna and its transmission line occurs when the characteristic impedance of the transmission line and the terminal impedance of the antenna are the same and have no reactive component. When this is the case, the antenna is considered to be matched to the line. Matching usually requires that the antenna be designed so that it has a terminal impedance of about 50 ohms or 75 ohms to match the common values of available coaxial cable.

The input impedance of patch antenna is in general complex and it includes resonant and non-resonant part. Both real and imaginary parts of the impedance vary as a function of frequency. Ideally, both the resistance and reactance exhibit symmetry about the resonant frequency. Typically, the feed reactance is very small, compared to the resonant resistance for thin substrates.

If the antenna is matched to the transmission line (ZA=ZO), then the input impedance does not depend on the length of the transmission line.

If the antenna is not matched, the input impedance will vary widely with the length of the transmission line. And if the input impedance isn't well matched to the source impedance, not very much power will be delivered to the antenna. This power ends up being reflected back to the generator, which can be a problem in itself (especially if high power is transmitted). Hence, we see that having a tuned impedance for an antenna is extremely important. In general, the transmission line will transform the impedance of an antenna, making it very difficult to deliver power, unless the antenna is matched to the transmission line. Consider the situation shown in Figure-17 .The impedance is to be measured at the end of a transmission line (with characteristic impedance Z0) and Length L. The end of the transmission line is hooked to an antenna with impedance ZA.

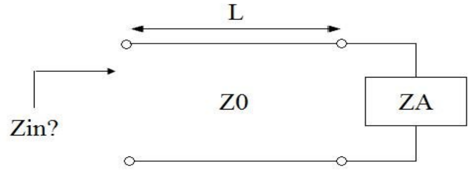


Figure -12 High Frequency Examples for Impedance measurement

It turns out (after studying transmission line theory for a while), that the input impedance Zin is given by:

$$Z_{in} = Z_0 \frac{ZA + jZ_0 \cdot \tan(\frac{2\pi f}{c} L)}{Z_0 + jZA \cdot \tan(\frac{2\pi f}{c} L)}$$

3.4 Voltage Standing Wave Ratio

The standing wave ratio (SWR), also known as the voltage standing wave ratio(VSWR), is not strictly an antenna characteristic, but is used to describe the performance of an antenna when attached to a transmission line. It is a measure of how well the antenna terminal impedance is matched to the characteristic impedance of the transmission line. Specifically, the VSWR is the ratio of the maximum to the minimum RF voltage along the transmission line. The maxima and minima along the lines are caused by partial reinforcement and cancellation of a forward moving RF signal on the transmission line and its reflection from the antenna terminals. If the antenna terminal impedance exhibits no reactive (imaginary) part and the resistive (real) part is equal to the characteristic impedance of the transmission line, then the

antenna and transmission line are said to be matched. It indicates that none of the RF signal sent to the antenna will be reflected at its terminals. There is no standing wave on the transmission line and the VSWR has a value of one. However, if the antenna and transmission line are not matched, then some fraction of the RF signal sent to the antenna is reflected back along the transmission line. This causes a standing wave, characterized by maxima and minima, to exist on the line. In this case, the VSWR has a value greater than one. The VSWR is easily measured with a device and VSWR of 1.5 is considered excellent, while values of 1.5 to 2.0 is considered good, and values higher than 2.0 may be unacceptable.

3.5 Bandwidth

The bandwidth of an antenna is defined as the range of frequency within the performance of the antenna. In other words, characteristics of antenna (gain, radiation pattern, terminal impedance) have acceptable values within the bandwidth limits. For most antennas, gain and radiation pattern do not change as rapidly with frequency as the terminal impedance does. Since the transmission line characteristic impedance hardly changes with frequency, VSWR is a useful, practical way to describe the effects of terminal impedance and to specify an antenna's bandwidth. For broadband antennas, the bandwidth is usually expressed as the ratio of the upper to lower frequencies of acceptable operation. However, for narrowband antennas, the bandwidth is expressed as a percentage of the bandwidth.

3.6 Radiation Efficiency

The radiation efficiency of the patch antenna is affected not only by conductor and dielectric losses, but also by surface-wave excitation - since the dominant TM0 mode of the grounded substrate will be excited by the patch. As the substrate thickness decreases, the effect of the conductor and dielectric losses becomes more severe, limiting the efficiency. On the other hand, as the substrate thickness increases, the surface-wave power increases, thus limiting the efficiency. Surface-

wave excitation is undesirable for other reasons as well, since surface waves contribute to mutual coupling between elements in an array, and also cause undesirable edge diffraction at the edges of the ground plane or substrate, which often contributes to distortions in the pattern and to back radiation. For an air (or foam) substrate there is no surface-wave excitation. In this case, higher efficiency is obtained by making the substrate thicker, to minimize conductor and dielectric losses (making the substrate too thick may lead to difficulty in matching, however, as discussed above). For a substrate with a moderate relative permittivity such as $\varepsilon r = 2.2$, the efficiency will be maximum when the substrate thickness is approximately $\lambda 0 = 0.02$. The radiation efficiency is defined as

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

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

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

Where

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

The dielectric and conductor Q factors are given by

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

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

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

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

where ζ is the conductivity of the metal .The space-wave Q factor is given approximately as

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

where

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

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

With $a2 = -0.16605$, $a4 = 0.00761$, and $c2 = -0.0914153$.
The surface-wave Q factor is related to the space-wave Q factor as

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

^{sw} is the radiation efficiency accounting only for surface-wave Taking into consideration the radiation efficiency of an antenna, we can express a relationship between the antenna's total radiated power and the total power input as:

$$Power\,Radiated = (\text{Antenna Radiation Efficiency})\ (\text{Power Input})$$

In the above formula, antenna radiation efficiency only includes conduction efficiency and dielectric efficiency and does not include reflection efficiency as part of the total efficiency factor. Moreover, the IEEE standards state that "gain does not include losses arising from impedance mismatches and polarization mismatches"

However, a specified frequency is necessary to determine conductor loss. For $h / \lambda 0 < 0.02$, the conductor and dielectric losses dominate, while for $h / \lambda 0 < 0.02$, the surface-wave losses dominate. (If there were no conductor or dielectric losses, the efficiency would approach 100% as the substrate thickness approaches zero.)

3.7 Antenna Efficiency

The total antenna efficiency takes into account all loses in the antenna such as reflections due to mismatch between transmission lines and the antenna, conduction and dielectric loses.

$\varepsilon_{total} = \varepsilon_r \varepsilon_c \varepsilon_d$

Where

ε_{total} is the total efficiency of the antenna.

ε_r is the efficiency due to mismatch losses.

ε_c is the efficiency due to conduction losses.

ε_d is the efficiency due to dielectric losses.

Usually conduction and dielectric efficiency are determined experimentally since they are very difficult to calculate. In fact, they cannot be separated when measured and therefore, it is more helpful to rewrite the equation as:

$$\epsilon_{total} = \epsilon_r \epsilon_{cd} = \left(1 - |\Gamma|^2\right) \epsilon_{cd}$$

Where Γ is the voltage reflection coefficient and, ε_{cd} or $(\varepsilon_c \varepsilon_d)$ is the antenna radiation efficiency which is commonly used to relate the gain and directivity in the antenna .

3.8 Beam width

It is a measure of directivity of an antenna .Antenna beam width is an angular width in degree ,measured on the radiation pattern (major lobe) between points width" between half power point or half power beam width (HPBW) because the power at half power point is just half. Half power beam width is also known as 3db beam width.

3.9 Quality factor

The quality factor is a figure-of-merit that representative of the antenna losses. Typically there are radiation, conduction, dielectric and surface wave losses.

$$\frac{1}{Q_t} = \frac{1}{Q_{rad}} + \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_{sc}}$$

: Total quality factor Q_t

Q_{rad} : Quality factor due to radiation losses

Q_c : Quality factor due to conduction losses

Q_{dm} : Quality factor due to dielectric losses

Q_{sc} : Quality factor due to surface wave

The quality factor, bandwidth and efficiency are antenna figures-of-merit, which are interrelated, and there is no complete freedom to independently optimize each one.

Generally Q is defined in terms of the ratio of the energy stored in the resonator to the energy being lost in one cycle:

$$Q = 2\pi \times \frac{\text{Energy Stored}}{\text{Energy dissipated per cycle}}.$$

The factor of 2π is used to keep this definition of Q consistent (for high values of Q) with the second definition:

$$Q = \frac{f_r}{\Delta f} = \frac{\omega_r}{\Delta\omega},$$

Where f_r is the resonant frequency, Δf is the bandwidth, ω_r is the angular resonant frequency, and $\Delta\omega$ is the angular bandwidth.

The definition of Q in terms of the ratio of the energy stored to the energy dissipated per cycle can be rewritten as:

$$Q = \omega \times \frac{\text{Energy Stored}}{\text{Power Loss}}$$

Physically speaking, Q is 2π times the ratio of the total energy stored divided by the energy lost in a single cycle or equivalently the ratio of the stored energy to the energy dissipated per one radian of the oscillation.

It is a dimensionless parameter that compares the time constant for decay of an oscillating physical system's amplitude to its oscillation period. Equivalently, it compares the frequency at which a system oscillates to the rate at which it dissipates its energy.

Equivalently (for large values of Q), the Q factor is approximately the number of oscillations required for a freely oscillating system's energy to fall off to $1 / e^{2\pi}$, or about 1/535, of its original energy.

The width (bandwidth) of the resonance is given by

$$\Delta f = \frac{f_0}{Q},$$

Where f_0 is the resonant frequency, and Δf, the bandwidth, is the width of the range of frequencies for which the energy is at least half its peak value.

A system with **low quality factor** ($Q < ½$) is said to be over damped

A system with **high quality factor** ($Q > ½$) is said to be under damped

A system with an **intermediate quality factor** ($Q = ½$) is said to be critically damped .

CHAPTER-4

PERTURBATION METHOD

4.1 Perturbation method

Perturbation theory comprises mathematical methods that are used to find an approximate solution to a problem which cannot be solved exactly, by starting from the exact solution of a related problem. Perturbation theory is applicable if the problem at hand can be formulated by adding a "small" term to the mathematical description of the exactly solvable problem. Perturbation theory is a large collection of iterative methods for obtaining approximate solutions to problems involving a small parameter ϵ. These methods are so powerful that sometimes it is actually advisable to introduce a parameter ϵ temporarily into a difficult problem having no small parameter, and then finally to set $\epsilon = 1$ to recover the original problem. This apparently artificial conversion to a perturbation problem may be the only way to make progress. The thematic approach of perturbation theory is to decompose a tough problem into an infinite number of relatively easy ones. Hence, perturbation theory is most useful when the first few steps reveal the important features of the solution and the remaining ones give small corrections.

The three steps of perturbative analysis.

1. Convert the original problem into a perturbation problem by introducing the small parameter ϵ.

2. Assume an expression for the answer in the form of a perturbation series and compute the coefficients of that series.

3. Recover the answer to the original problem by summing the perturbation series for the appropriate value of ϵ.

Step (1) is sometimes ambiguous because there may be many ways to introduce an e. However, it is preferable to introduce e in such a way that the zeroth- order solution (the leading term in the perturbation series) is obtainable as a closed-form analytic expression. Perturbation problems generally take the form of a soluble equation whose solution is altered slightly by a perturbing term Of course; step (1) may be omitted when the original problem already has a small parameter if a perturbation series can be developed in powers of that parameter.

Step (2) is frequently a routine iterative procedure for determining successive coefficients in the perturbation series. A zeroth-order solution consists of finding the leading term in the perturbation series. This involves solving the unperturbed problem, the problem obtained by setting $\epsilon = 0$ in the perturbation problem. A first-order solution consists of finding the first two terms in the perturbation series, and so on. Each of the coefficients in the perturbation series is determined in terms of the previous coefficients by a simple linear equation, even though the original problem was a nonlinear (cubic) equation. Generally it is the existence of a closed-form zeroth-order solution which ensures that the higher-order terms may also be determined as closed-form analytical expressions.

Step (3) may or may not be easy. If the perturbation series converges, its sum is the desired answer. If there are several ways to reduce a problem to a perturbation problem, one chooses the way that is the best compromise between difficulty of calculation of the perturbation series coefficients and rapidity of convergence of the series itself. However, many series converge so slowly that their utility is impaired. Also, we will shortly see that perturbation series are frequently divergent. This is not necessarily bad because many of these divergent perturbation series are asymptotic. In such cases, one obtains a good approximation to the answer when ϵ is very small by summing the first few terms according to the optimal truncation rule. When ϵ is not small, it may still be possible to obtain a good approximation to the answer from a slowly converging or divergent series using the summation methods discussed in next chapter.

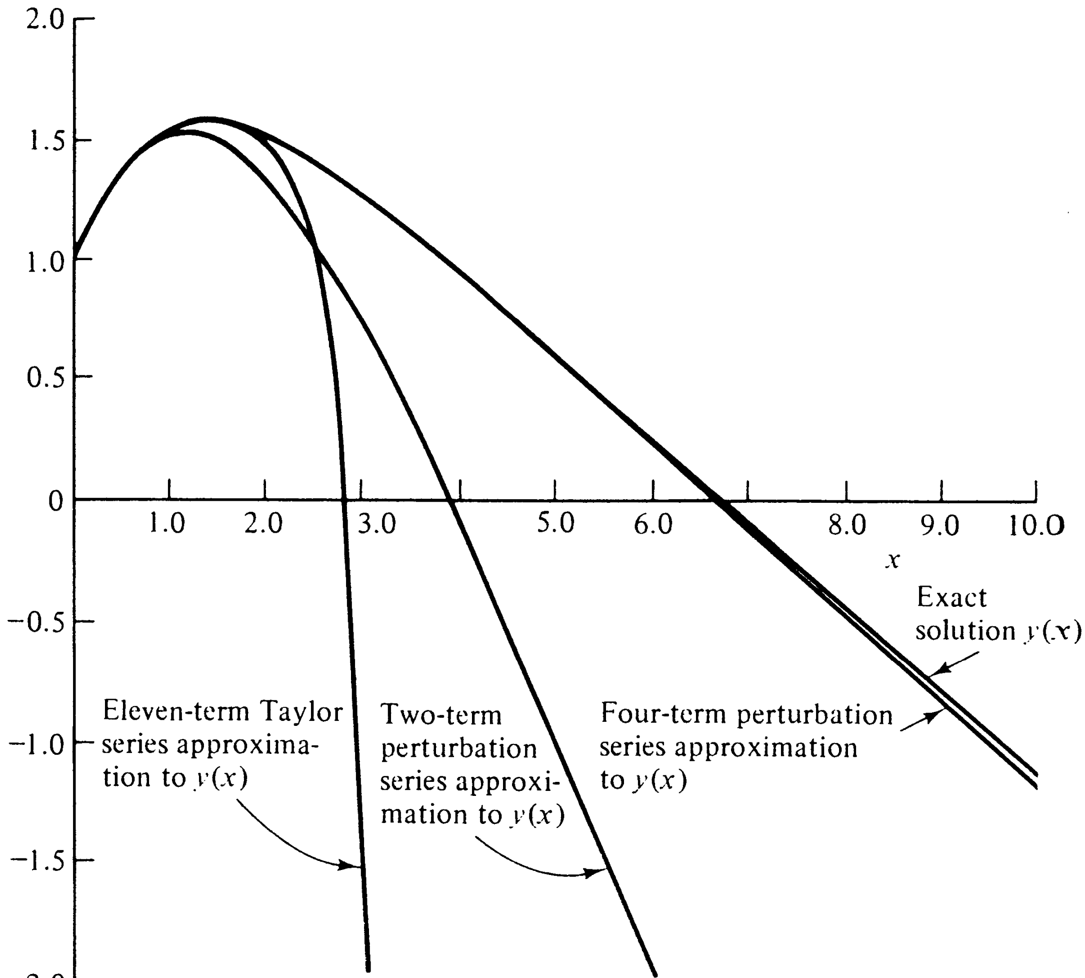


Fig.13 A comparison of Taylor series and perturbation series approximations to the solution of the initial-value problem

4.2 Regular and Singular Perturbation

The formal techniques of perturbation theory are a natural generalization of the ideas of local analysis of differential equations. Local analysis involves approximating the solution to a differential equation near the point x = a by developing a series solution about a in powers of a small parameter. Once the leading behavior of the solution near x = a (which we would now refer to as the zeroth-order solution) is known, the remaining coefficients in the series can be computed recursively. The strong analogy between local analysis of differential equations and formal perturbation theory may be used to classify perturbation problems. Recall' that there are two different types of series solutions to differential equations. A series solution about an ordinary point of a differential equation is always a Taylor series having a non vanishing radius of convergence. A series solution about a singular point does not have this form (except in rare cases). Instead, it may either be a convergent series not in Taylor series form or it may be a divergent series. Series solutions about singular points often have the remarkable property of being meaningful near a singular point yet not existing at the singular point. Perturbation series also occur in two varieties. We define a regular perturbation problem as one whose perturbation series is a power series having a non vanishing radius of convergence. A basic feature of all regular perturbation problems is that the exact solution for small. We define singular perturbation problem as one whose perturbation series either does not take the form of a power series or, if it does, the power series has a vanishing radius of convergence. In singular perturbation theory 'there is sometimes no solution to the unperturbed problem; when a solution to the unperturbed problem does exist, its qualitative features are distinctly different from those of the exact solution for arbitrarily small but nonzero ϵ. If there is no such abrupt change in character, then we would have to classify the problem as a regular perturbation problem. When dealing with a singular perturbation problem, one must take care to distinguish between the zeroth-order solution (the leading term in the perturbation series) and the solution of the unperturbed problem, since the latter may not even exist. There is no difference between these two in a regular perturbation theory, but in a singular perturbation theory the zeroth-order solution may depend on ϵ and may exist only for nonzero ϵ.

4.3 Parameter Perturbations

Many physical problems involving the function $u(x,\epsilon)$ can be represented mathematically by the differential equation $L(u, x,\epsilon) = 0$ and the boundary condition $B(u, \epsilon) = 0$, where x is a scalar or vector independent variable and ϵ is a parameter. In general, this problem cannot be solved exactly. However, if there exists an $\epsilon = \epsilon 0$ (e can be scaled so that $\epsilon 0 = 0$) for which the above problem can be solved exactly or more readily, one seeks to find the solution for small ϵ in, say, powers of ϵ; that is

$u(x; \epsilon) = u_0(x) + \epsilon u_1(x) + \epsilon^2 u_2(x) + \ldots\ldots\ldots\ldots\ldots..$

Where u_n is independent of ϵ and $u_0(x)$ is the solution of the problem for $\epsilon = 0$. One then substitutes this expansion into $L(u, x, \epsilon) = 0$ and $B(u, \epsilon) = 0$, expands for small ϵ, and collects coefficients of each power of ϵ. Since these equations must hold for all values of ϵ, each coefficient of ϵ must vanish independently because sequences of ϵ are linearly independent. These usually are simpler equations governing u_n, which can be solved successively.

4.4 Perturbation method in electromagnetic

4.4.1 Cavity Perturbation Theory

Cavity Perturbation Theory describes methods for derivation of perturbation formulae for performance changes of a cavity resonator. These performance changes are assumed to be caused by either introduction of a small foreign object into the cavity or a small deformation of its boundary.

When a resonant cavity is perturbed, i.e. when a foreign object with distinct material properties is introduced into the cavity or when a general shape of the cavity is changed, electromagnetic fields inside the cavity change accordingly. The underlying assumption of cavity perturbation theory is that electromagnetic fields inside the cavity after the change differ by a very small amount from the fields before the change. Then Maxwell's equations for original and perturbed cavities can be used to derive expressions for resulting resonant frequency shifts.

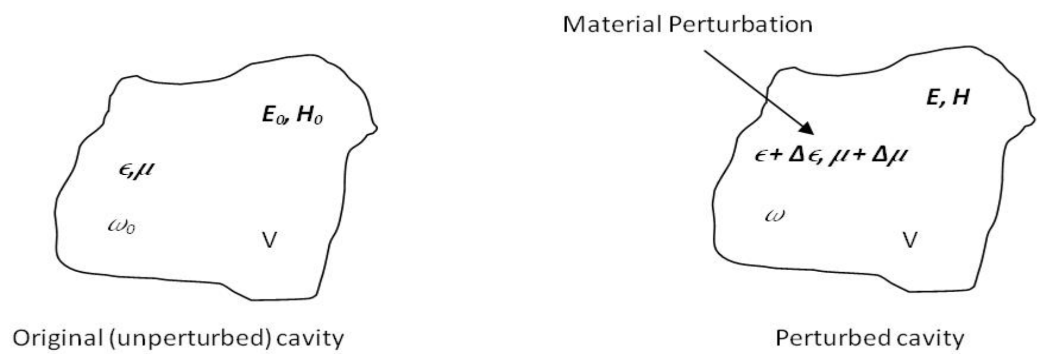


Fig.14 Material perturbation

When a material within a cavity is changed (permittivity and/or permeability), a corresponding change in resonant frequency can be approximated as

$$\frac{\omega - \omega_0}{\omega_0} \approx - \frac{\iiint_V (\Delta\mu |H_0|^2 + \Delta\epsilon |E_0|^2) dv}{\iiint_V (\mu |H_0|^2 + \epsilon |E_0|^2) dv}$$ ……………4.a

where ω is the angular resonant frequency of the perturbed cavity, ω_0 is the resonant frequency of the original cavity, E_0 and H_0 represent original electric and magnetic field respectively, μ and ε are original permeability and permittivity respectively, while Δμ and Δε are changes in original permeability and permittivity introduced by material change.

Expression (4.a) can be rewritten in terms of stored energies as

$$\frac{\omega - \omega_0}{\omega_0} \approx - \frac{1}{W} \iiint_V \left(\frac{\Delta\epsilon}{\epsilon} \cdot \bar{w_e} + \frac{\Delta\mu}{\mu} \cdot \bar{w_m} \right) dv$$ ……………..4.b

where W is the total energy stored in the original cavity and $\bar{w_e}$ and $\bar{w_m}$ are electric and magnetic energy densities respectively.

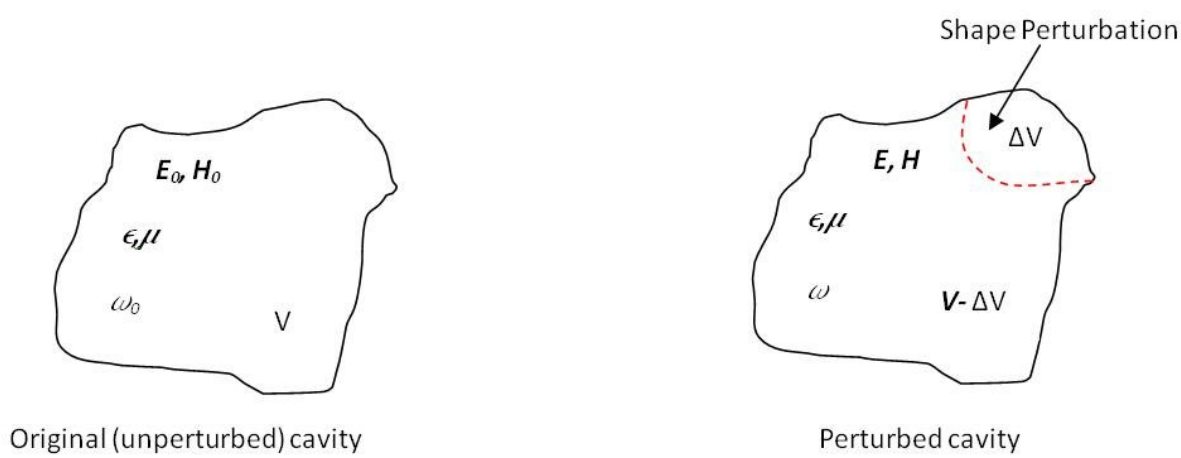


Fig .15 Shape perturbation

When a general shape of a resonant cavity is changed, a corresponding change in resonant frequency can be approximated as

$$\frac{\omega - \omega_0}{\omega_0} \approx \frac{\iiint_{\Delta V} (\mu |H_0|^2 - \epsilon |E_0|^2) dv}{\iiint_V (\mu |H_0|^2 + \epsilon |E_0|^2) dv}$$

………………4.c

Expression (3) for change in resonant frequency can additionally be written in terms of time-average stored energies as

$$\frac{\omega - \omega_0}{\omega_0} \approx \frac{\Delta W_m - \Delta W_e}{W_m + W_e}$$

………………4.d

Where ΔW_m and ΔW_e represent time-average electric and magnetic energies contained in ΔV.

This expression can also be written in terms of energy densities as

$$\frac{\omega - \omega_0}{\omega_0} \approx \frac{(\bar{w_m} - \bar{w_e}) \cdot \Delta V}{W}$$

……………..4.e

CHAPTER-5

DUAL FREQUENCY ANGULAR NOTCH RMA DESIGN

5.1 S-parameter

The S-parameters are members of a family of similar parameters, other examples being: Y-parameters, Z-parameters, H-parameters, T-parameters or ABCD-parameters. They differ from these, in the sense that *S-parameters* do not use open or short circuit conditions to characterize a linear electrical network; instead matched loads are used. These terminations are much easier to use at high signal frequencies than open-circuit and short-circuit terminations. Moreover, the quantities are measured in terms of power.

Many electrical properties of networks of components (inductors, capacitors, resistors) may be expressed using S-parameters, such as gain, return loss, voltage standing wave ratio (VSWR), reflection coefficient and amplifier stability. The term 'scattering' is more common to optical engineering than RF engineering, referring to the effect observed when a plane electromagnetic wave is incident on an obstruction or passes across dissimilar dielectric media. In the context of S-parameters, scattering refers to the way in which the traveling currents and voltages in a transmission line are affected when they meet a discontinuity caused by the insertion of a network into the transmission line. This is equivalent to the wave meeting an impedance differing from the line's characteristic impedance.

Although applicable at any frequency, S-parameters are mostly used for networks operating at radio frequency (RF) and microwave frequencies where signal power and energy considerations are more easily quantified than currents and voltages. S-parameters change with the measurement frequency, so frequency must be specified for any S-parameter measurements stated, in addition to the characteristic impedance or system impedance.

2-Port S-Parameters

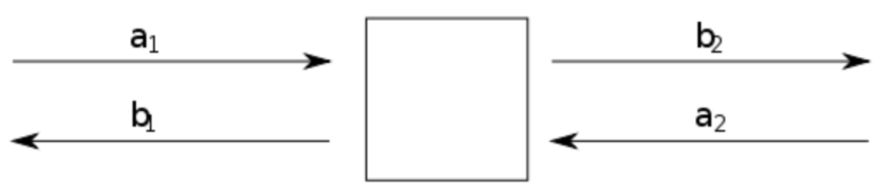


Fig.16 Model of S-Parameter

The S-parameter matrix for the 2-port network is probably the most commonly used and serves as the basic building block for generating the higher order matrices for larger networks. In this case the relationship between the reflected, incident power waves and the S-parameter matrix is given by:

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

Expanding the matrices into equations gives:

$$b_1 = S_{11}a_1 + S_{12}a_2$$

$$b_2 = S_{21}a_1 + S_{22}a_2$$

Each equation gives the relationship between the reflected and incident power waves at each of the network ports, 1 and 2, in terms of the network's individual S-parameters, S_{11}, S_{12}, $S_{21}$and S_{22}. If one considers an incident power wave at port 1 (a_1) there may result from it waves exiting from either port 1 itself (b_1) or port 2 (b_2). However if, according to the definition of S-parameters, port 2 is terminated in a load identical to the system impedance (Z_0) then, by the maximum power transfer theorem, b_2 will be totally absorbed making a_2 equal to zero. Therefore

$$a_1 = \frac{V_1 + Z_0 I_1}{2\sqrt{Z_0}} \quad a_2 = \frac{V_2 - Z_0 I_2}{2\sqrt{Z_0}}$$

$$b_1 = \frac{V_1 - Z_0 I_1}{2\sqrt{Z_0}} \quad b_2 = \frac{V_2 + Z_0 I_2}{2\sqrt{Z_0}}$$

$$a_1 = \frac{V_1 + Z_0 I_1}{2\sqrt{Z_0}} = \frac{1}{\sqrt{Z_0}} V_{1+} \qquad a_2 = \frac{V_2 - Z_0 I_2}{2\sqrt{Z_0}} = \frac{1}{\sqrt{Z_0}} V_{2-}$$

$$b_1 = \frac{V_1 - Z_0 I_1}{2\sqrt{Z_0}} = \frac{1}{\sqrt{Z_0}} V_{1-} \qquad b_2 = \frac{V_2 + Z_0 I_2}{2\sqrt{Z_0}} = \frac{1}{\sqrt{Z_0}} V_{2+}$$

Each 2-port S-parameter has the following generic descriptions:

S_{11} is the input port voltage reflection coefficient
S_{12} is the reverse voltage gain
S_{21} is the forward voltage gain

S_{22} is the output port voltage reflection coefficient

S-Parameter properties of 2-port networks

An amplifier operating under linear (small signal) conditions is a good example of a non-reciprocal network and a matched attenuator is an example of a reciprocal network. In the following cases we will assume that the input and output connections are to ports 1 and 2 respectively which is the most common convention. The nominal system impedance, frequency and any other factors which may influence the device, such as temperature, must also be specified.

Input return loss

Input return loss (RL_{in}) is a scalar measure of how close the actual input impedance of the network is to the nominal system impedance value and, expressed in logarithmic magnitude, is given by

$$RL_{\mathrm{in}} = |20\log_{10}|S_{11}||\,\mathrm{dB}.$$

By definition, return loss is a positive scalar quantity implying the 2 pairs of magnitude (|) symbols. The linear part $|S_{11}|$ is equivalent to the reflected voltage magnitude divided by the incident voltage magnitude.

5.2 Calculation of S-parameter of Angular Ring Microstrip Antenna

The microstrip lines are modeled by equivalent ideal magnetic wall wave-guides for which the electromagnetic field solutions are known . The field solutions in the microstrip ring segment are derived by utilizing a perturbation analysis of a modified (magnetic wall) curved waveguide model. Other techniques have been formulated to evaluate the fields inside curved metallic waveguides. The perturbation solution for the fields in the equivalent curved waveguide model developed here is readily adaptable to the mode-matching procedure and is used to calculate the properties of the curved microstrip bend discontinuities. The frequency-dependent reflection and transmission coefficients of curved microstrip

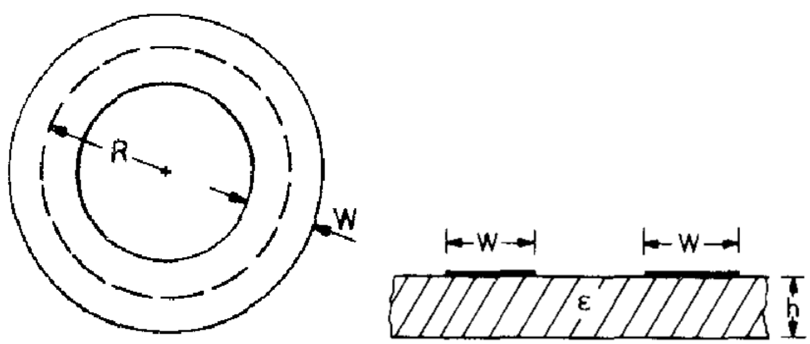


Fig.17 Annular Ring Microstrip Antenna

The magnetic wall curved waveguide models for the open- and closed-ring microstrip resonators are shown in Fig. 1. The model is characterized by its effective dimensions and the medium permittivity which are determined from the solution of the corresponding microstriplie problem, and the inclusion of the effect of curvature on the model. The model assumes that the substrate height h is small (h<< A, the wavelength) and, hence, the fields are constant along the z-direction.

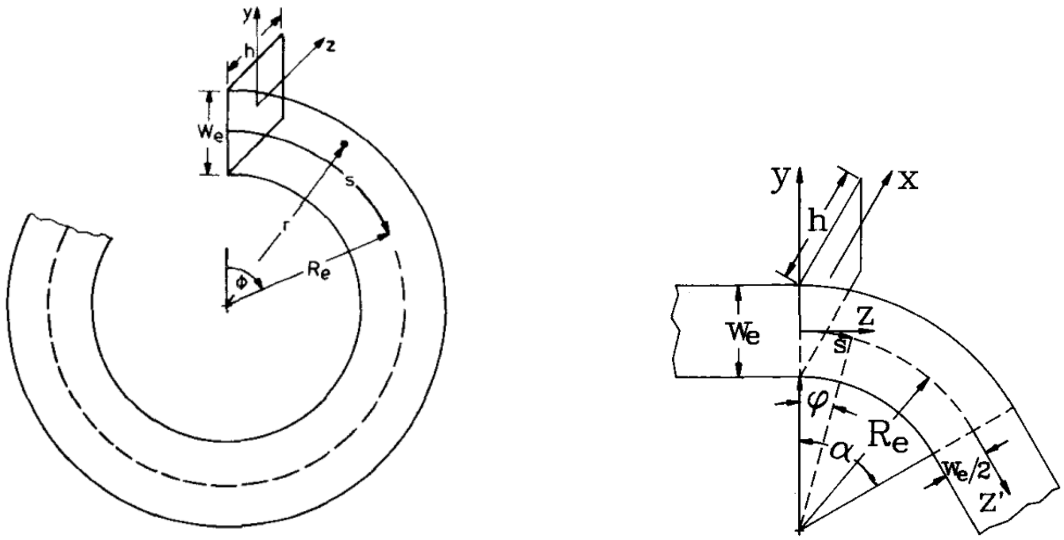


Fig.18 Angular Microstrip Antenna

The corresponding magnetic wall waveguide model for the microstrip lines, shown in Fig. 2, is characterized by its frequency-dependent effective width we and effective permittivity E, . Since the curved microstrip bend contains no discontinuities in width and permittivity (i.e., has constant width w and permittivity ϵ), a modified (magnetic wall) curved waveguide model is used here for the microstrip ring segment where the effective width and the effective permittivity are identical to the effective quantities for the microstrip lines Hence, no discontinuities by which non existing modes could be excited are introduced by the modified curved waveguide model. The model effective radius is given by

$$R_e = \frac{R}{2} + \frac{1}{2}\sqrt{R^2 + (w_e - w)w_e}\,.$$

The waveguide models assume that the substrate height h is small compared with the wavelength; hence, the mode and $TE_{n,0}$ modes with transversal components E_x and H_y, exist inside the waveguides. The application of the mode-matching method given in a complete set of transversal field solutions with orthogonal properties in all three regions. A complete set of solutions for the transversal field in regions I and 11 (i.e., in the straight waveguides) was derived in reference planes at z = 0 and z' = 0, is given by

$$E_x^{\mathrm{I}} = -\sum_{n=0}^{\infty} \sqrt{Z_n}\left(a_n^{\mathrm{I}} e^{-j\beta_n z} + b_n^{\mathrm{I}} e^{j\beta_n z}\right) \cdot \sqrt{\frac{\delta_n}{w_e h}} \cos\left[\frac{n\pi}{w_e}\left(y - \frac{w_e}{2}\right)\right]$$

----------------------------5.a

$$H_y^{\mathrm{I}} = -\sum_{n=0}^{\infty} \sqrt{Y_n}\left(a_n^{\mathrm{I}} e^{-j\beta_n z} - b_n^{\mathrm{I}} e^{j\beta_n z}\right) \cdot \sqrt{\frac{\delta_n}{w_e h}} \cos\left[\frac{n\pi}{w_e}\left(y - \frac{w_e}{2}\right)\right]$$

----------------------------5.b

$$E_x^{\mathrm{II}} = -\sum_{n=0}^{\infty} \sqrt{Z_n}\left(a_n^{\mathrm{II}} e^{j\beta_n z'} + b_n^{\mathrm{II}} e^{-j\beta_n z'}\right) \cdot \sqrt{\frac{\delta_n}{w_e h}} \cos\left[\frac{n\pi}{w_e}\left(y - \frac{w_e}{2}\right)\right]$$

-----------------------------5.c

$$H_y^{\mathrm{II}} = \sum_{n=0}^{\infty} \sqrt{Y_n}\left(a_n^{\mathrm{II}} e^{j\beta_n z'} - b_n^{\mathrm{II}} e^{-j\beta_n z'}\right) \cdot \sqrt{\frac{\delta_n}{w_e h}} \cos\left[\frac{n\pi}{w_e}\left(y - \frac{w_e}{w}\right)\right]$$

--------------------------- -5.d

And

$$\beta_n^2 = k_e^2 - \frac{n^2\pi^2}{w_e^2}, \qquad k_e^2 = \mu_0 \epsilon_e \omega^2.$$

Here β_n is the phase constant, $Z_n = 1/ Y_n = \omega\mu_0 /\beta$, is are the characteristic wave impedance, and a_n , b_n in the region the wave equation and boundary conditions for $E_n = E_{x,n}$ in the curved orthogonal coordinate system as characterized by $u_1 = x$, $u_2 = y$, and $u_3 = s = R_{e\varphi}$ with corresponding metric coefficients $h_1 = h_2 = 1$and $h_3 = 1 + y/R_e$ are given as

$$\left(1+\frac{y}{R_e}\right)^2\frac{\partial^2 E_n}{\partial y^2}+\frac{1}{R_e}\left(1+\frac{y}{R_e}\right)\frac{\partial E_n}{\partial y}+\frac{\partial^2 E_n}{\partial s^2}$$
$$+k_e^2\left(1+\frac{y}{R_e}\right)^2 E_n=0$$

------------------------- 5.e

And

$$\frac{\partial E_n}{\partial y}=0 \qquad \text{at} \qquad y=\pm\frac{w_e}{2}.$$

The solutions for E_n are expressed as

$$E_n=\left(1+\frac{y}{R_e}\right)\psi_n(y)e^{-j\tilde{\beta}_n s}.$$

These form a complete set of eigenfunctions which are orthogonal with respect to the weighting function $(1 + y/R_e)^{-1}$

$$\int_{-w_e/2}^{w_e/2} E_m E_n\left(1+\frac{y}{R_e}\right)^{-1} dy=0 \qquad \text{for } m\neq n.$$

--------------5.f

The orthogonality property for the functions $\psi_n(y)$ then immediately follows from

$$\int_{-w_e/2}^{w_e/2}\psi_m(y)\psi_n(y)\left(1+\frac{y}{R_e}\right)dy=0 \qquad \text{for } m\neq n.$$

---------------5.g

The total field strength Ex with unknown coefficients c_n is given by

$$E_x=\sum_{n=0}^{\infty}c_n E_n=\left(1+\frac{y}{R_e}\right)\sum_{n=0}^{\infty}c_n\psi_n(y)e^{-j\tilde{\beta}_n s}$$

The magnetic field is readily found in terms of E_x from Maxwell's equations. For the transverse magnetic field component H_y we get

$$H_y = -\frac{1}{j\omega\mu_0}\left(1+\frac{y}{R_e}\right)^{-1}\frac{\partial E_x}{\partial s}$$

$$= \frac{1}{\mu_0\omega}\sum_{n=0}^{\infty} c_n\tilde{\beta}_n\psi_n(y)e^{-j\tilde{\beta}_n s}.$$

--------------5.h

A perturbation solution for the electromagnetic field can be found by expanding E_n and β_n along s in a power series in the effective radius of curvature R_e of the curved waveguide as shown in for a curved waveguide with electric walls:

$$E_n = e^{-j\tilde{\beta}_n s}\left(\phi_{0,n} + \frac{\phi_{1,n}}{R_e} + \frac{\phi_{2,n}}{R_e^2} + \cdots\right)$$

$$= \left(1+\frac{y}{R_e}\right)\psi_n(y)e^{-j\tilde{\beta}_n s}$$

$$\tilde{\beta}_n^2 = \beta_n^2\left(1+\frac{B_{1,n}}{R_e} + \frac{B_{2,n}}{R_e^2} + \cdots\right)$$

The quantities φ_n and β_n are the solutions for a straight waveguide (R_e tend infinity) and are given by

$$\phi_{0,n} = \cos\left[\frac{n\pi}{w_e}\left(y-\frac{w_e}{2}\right)\right]$$

$$\beta_n^2 = k_e^2 - \frac{n^2\pi^2}{w_e^2}, \qquad k_e^2 = \mu_0 \epsilon_e \omega^2$$

The degree of accuracy and the complexity of the expressions obviously depend on the number of terms used in the expansions. For a second-order solution the expansion functions $\varphi 1,n$ and $\varphi_{2,n}$ as well as the phase constant β_n are given as

$$\phi_{1,n} = \begin{cases} k_e^2 y\left(\frac{w_e^2}{4} - \frac{y^2}{3}\right) & \text{for } n = 0 \\ \frac{w_e^2}{2n^2\pi^2}\left\{\frac{d\phi_{0,n}}{dy}\left[\beta_n^2\left(y^2 - \frac{w_e^2}{4}\right) - \frac{k_e^2 w_e^2}{n^2\pi^2}\right] - k_e^2 y \phi_{0,n}\right\} & \text{for } n > 0 \end{cases}$$

$$\phi_{2,n} = \begin{cases} \frac{k_e^2 y^2}{6}\left[2y^2 - w_e^2 + \frac{k_e^2}{60}\left(6w_e^4 - 15y^2 w_e^2 + 8y^4\right)\right] & \text{for } n = 0 \\ p_n(y)\cdot\phi_{0,n} + q_n(y)\frac{d\phi_{0,n}}{dy} & \text{for } n > 0 \end{cases}$$

$$p_n(y) = \frac{w_e^2 y^2}{8n^2\pi^2}\left[k_e^2\left(\frac{7k_e^2 w_e^2}{n^2\pi^2} - 4\right) + \beta_n^4\left(\frac{w_e^2}{2} - y^2\right)\right]$$

$$q_n(y) = \frac{y w_e^4}{48n^4\pi^4}\left[12k_e^2\left(\frac{7k_e^2 w_e^2}{n^2\pi^2} - 4\right) + \beta_n^2\left(w_e^2 - 4y^2\right)\left(9k_e^2 - 4\beta_n^2\right)\right]$$

$$\bar{\beta}_n^2 = \begin{cases} k_e^2\left[1 - \frac{w_e^2}{12R_e^2}\left(1 - \frac{2}{5}k_e^2 w_e^2\right)\right] & \text{for } n = 0 \\ k_e^2\left(1 - \frac{n^2\pi^2}{k_e^2 w_e^2}\right) + \frac{\pi^2}{6R_e^2}\left[n^2 + \frac{12 - n^2\pi^2}{2n^2\pi^4}k_e^2 w_e^2 - \frac{21 + n^2\pi^2}{2n^4\pi^6}k_e^4 w_e^4\right] & \text{for } n > 0. \end{cases}$$

A form of the field solution in region which is suitable for the mode-matching method can be constructed by superimposing the field solutions obtained by alternately placing a magnetic wall at $s = 0$ (E^a_x, H^a_y) and $s = R_e\alpha$ (E^b_x, H^b_y)

$$E_x^a = \left(1 + \frac{y}{R_e}\right) \sum_{n=0}^{\infty} c_n^a \psi_n \cos\left(\tilde{\beta}_n s\right)$$

$$H_y^a = \frac{1}{j\omega\mu_0} \sum_{n=0}^{\infty} c_n^a \tilde{\beta}_n \psi_n \sin\left(\tilde{\beta}_n s\right)$$

This leads to

$$c_n^a = \frac{j\omega\mu_0}{\tilde{\beta}_n I_n \sin\left(\tilde{\beta}_n R_e \alpha\right)} \sum_{p=0}^{\infty} \sqrt{Y_p}\left(a_p^{\mathrm{II}} - b_p^{\mathrm{II}}\right) \sqrt{\frac{\delta_p}{w_e h}}\, K(p,n)$$ --------------------5.i

$$c_n^b = \frac{j\omega\mu_0}{\tilde{\beta}_n I_n \sin\left(\tilde{\beta}_n R_e \alpha\right)} \sum_{p=0}^{\infty} \sqrt{Y_p}\left(a_p^{\mathrm{I}} - b_p^{\mathrm{I}}\right) \sqrt{\frac{\delta_p}{w_e h}}\, K(p,n)$$ --------------------5.j

With

$$I_n = \int_{-w_e/2}^{w_e/2} \psi_n^2(y) \left(1 + \frac{y}{R_e}\right) dy$$ --------------------5.k

And

$$K(p,n) = \int_{-w_e/2}^{w_e/2} \phi_{0,p}(y) \psi_n(y) \left(1 + \frac{y}{R_e}\right) dy$$ --------------------5.l

af

$$\left(a_n^{\mathrm{I}} + b_n^{\mathrm{I}}\right)\sqrt{\frac{Z_n w_e}{\delta_n h}}$$

$$= -\sum_{m=0}^{\infty} \left[c_m^a + c_m^b \cos\left(\tilde{\beta}_m R_e \alpha\right)\right] K(n,m) \qquad \text{---------------------5.m}$$

$$\left(a_n^{\mathrm{II}} + b_n^{\mathrm{II}}\right)\sqrt{\frac{Z_n w_e}{\delta_n h}}$$

$$= -\sum_{m=0}^{\infty} \left[c_m^a \cos\left(\tilde{\beta}_m R_e \alpha\right) + c_m^b\right] K(n,m) \qquad \text{---------------------5.n}$$

The coefficients c_n^a and c_n^b can then be eliminated by inserting (5.i) and (5.j) into (5.m) and (5.n). Thus, an infinite set of linear equations for the wave amplitudes a_1, b_1 and a_2, b_2" is obtained. In order to obtain numerical results, this infinite set of equations is truncated to 2M + 2 equations, where M is the highest-order mode to be considered. The scattering parameters S_{11} for an incident TEM mode can be found by setting a = 1 for n = 0 and j = 1 or 2, with all other 2)'s set equal to 0. Then the scattering parameters $S_{ij} = b_0$ can be determined with standard routines for solving a set of linear equations.

5.3 Results

The transmission characteristics of typical curved microstrip bends have been computed, and good convergence with increasing number of higher order modes has been found. As seen in previous fig, a minimum of three higher order modes must usually be considered in order to obtain a negligible truncation error. In the results for the curved microstrip bends presented in this paper, seven higher order modes were taken into account to ensure a negligible truncation error. Fig.21 show the magnitudes of the scattering parameters of curved microstrip bends with $\alpha = 55^0$ and $R / w = 2$, all normalized with respect to the microstrip impedance.

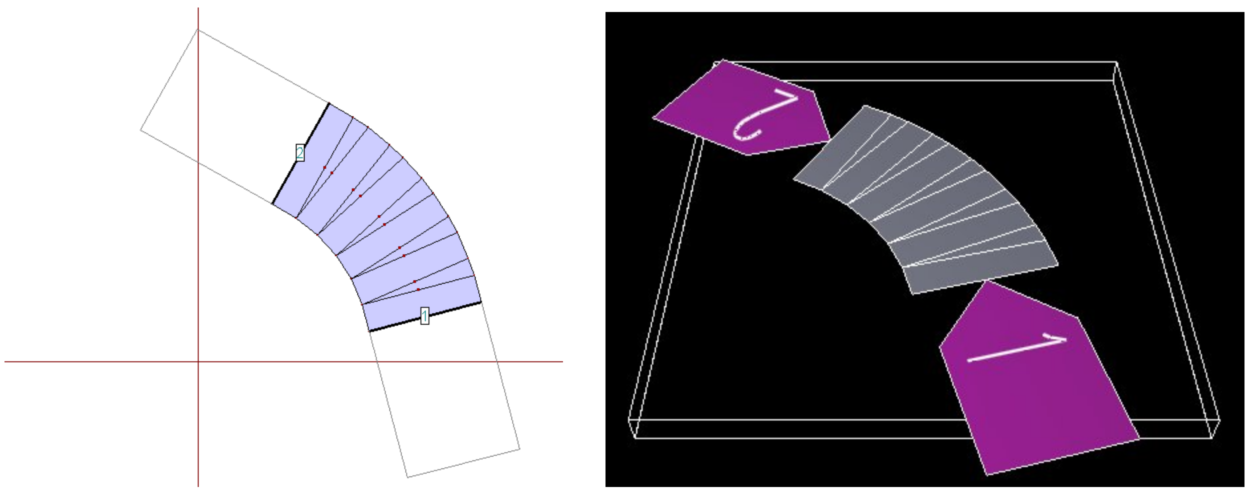


Fig.19 Angular ring structure with edge feed

It should be noted that the accuracy of these second- order perturbation solutions depends on the curvature Re (or R/w) of the curved waveguide model higher order terms may need to be included, especially when the inner radius of the microstrip goes to zero . However, the first-order perturbation solutions for the microstrip ring resonator are very close to the exact solutions computed for a wide range of parameters; therefore a corresponding superior accuracy of the second-order perturbation solutions is expected here including the results . In addition, for moderate microstrip curvature , the results based on the waveguide model presented in this thesis are in good agreement with IE3D data for microstrip curved bends on thin GaAs substrates .

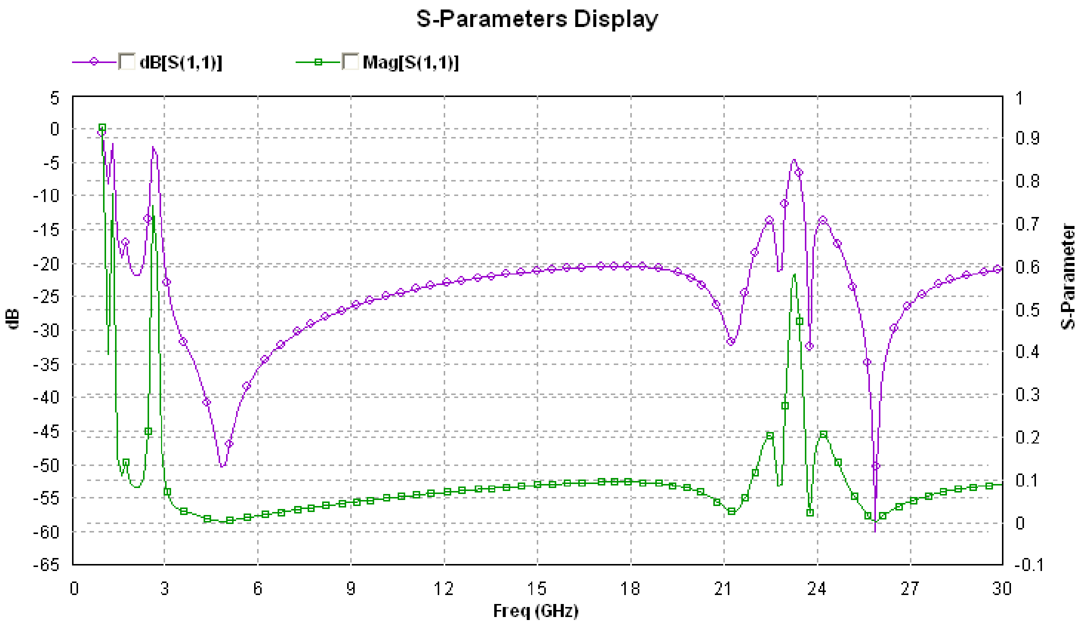


Fig .20 Magnitude of the reflection coefficient of frequency for a curved microstrip bend with a = 55" and R / w = 2, a chamfered acute-angle bend w = 1.2 mm, h = 0.635 mm, and ϵ = 9.8 . (IE3D)

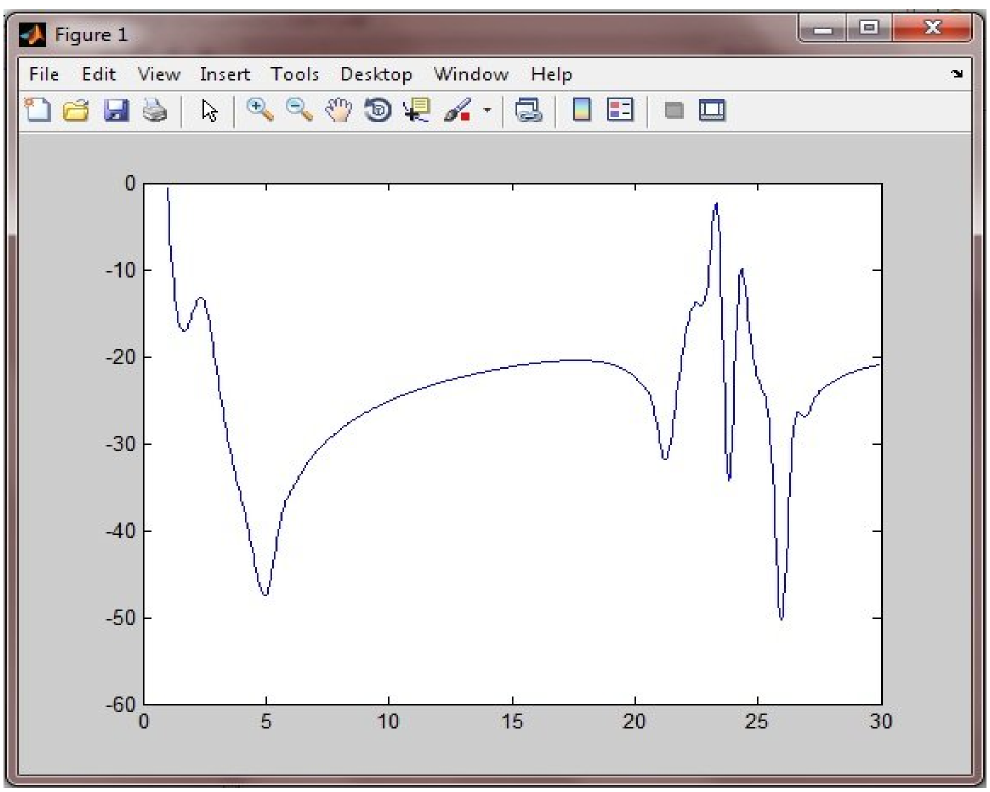


Fig :21 Magnitude of the reflection coefficient of frequency for a curved microstrip bend with a = 55" and R / w = 2, a chamfered acute-angle bend w = 1.2 mm, h = 0.635 mm, and ϵ = 9.8 . (second order perturbed)

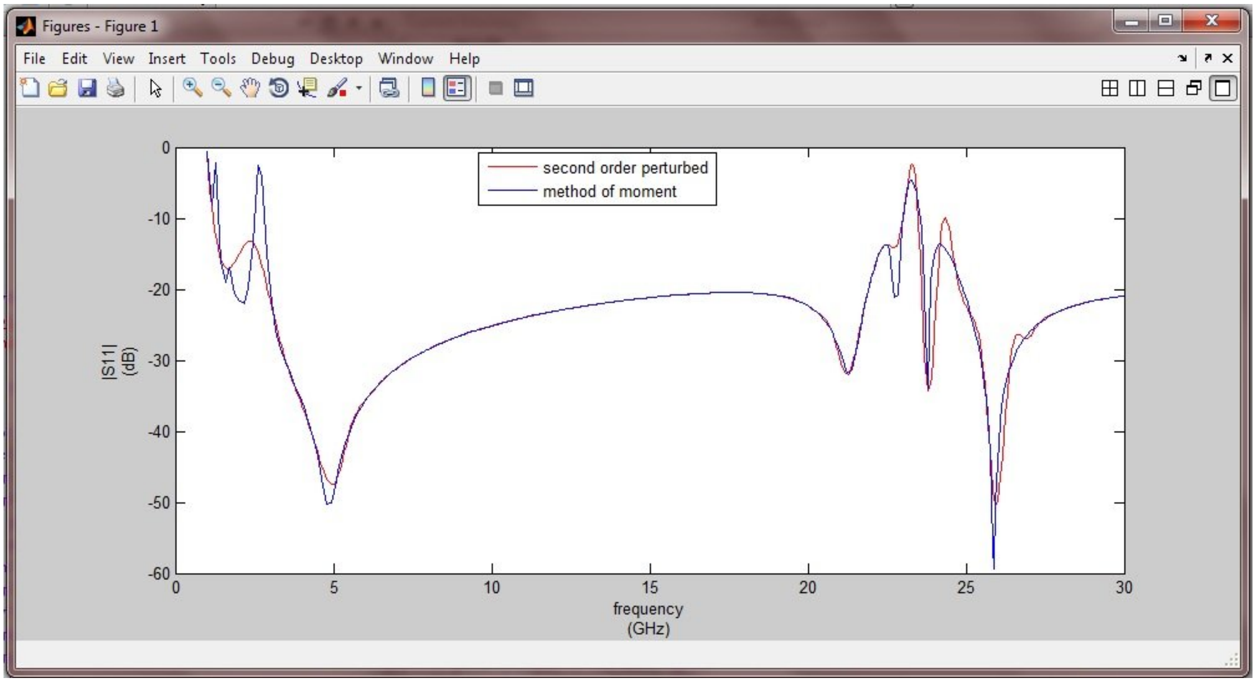


Fig .22 comparison the result of perturbation method &method of moment

CHAPTER- 5

CONCLUSION AND FUTURE PROSPECT

5.4 Conclusion

A method for calculating the frequency-dependent scattering parameters of curved microstrip bends has been described and computational results have been presented. The results have been compared with those obtained for the acute-angle and the chamfered acute-angle bends and show an improvement in the transmission properties. The results for the scattering parameters of the curved microstrip bend converge very fast with increasing number of higher order modes considered and have been shown to be consistent for large curvatures and bends with small angle .

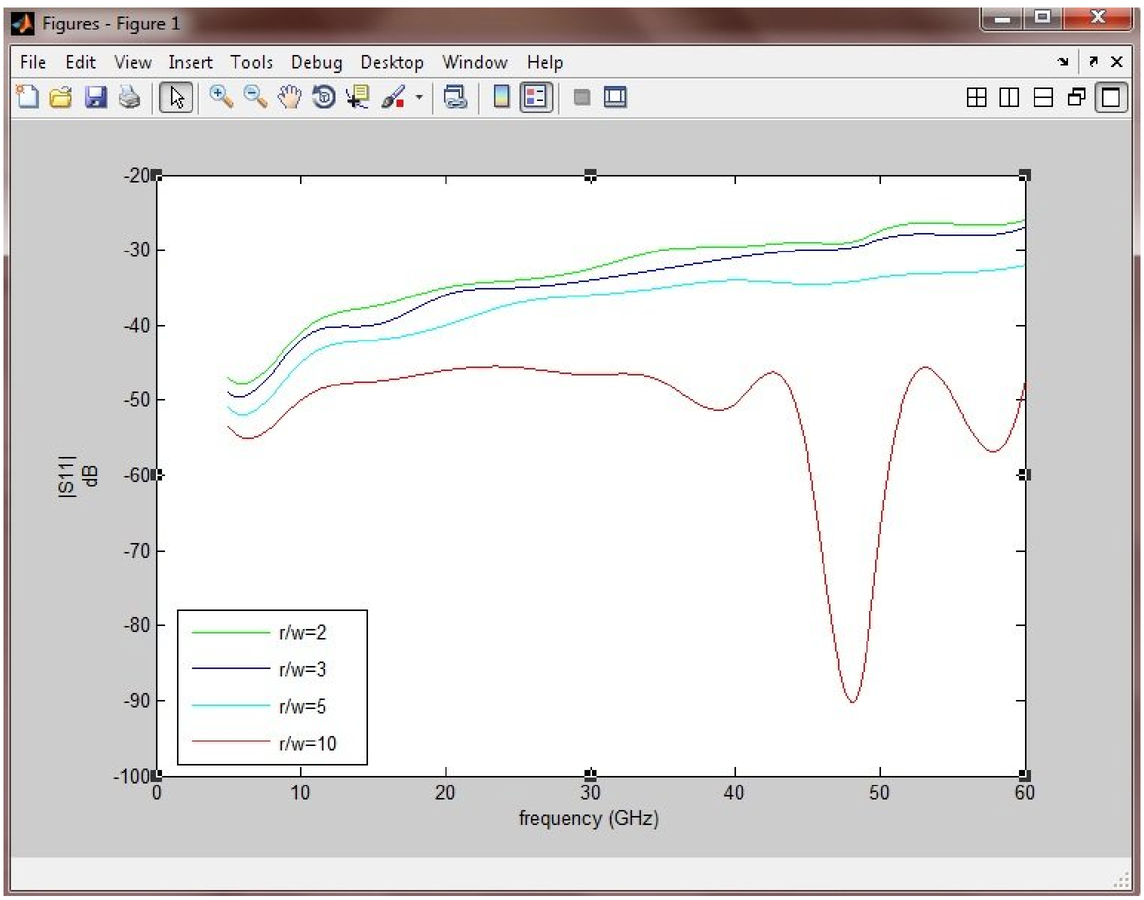


Fig:.23 Magnitude of the reflection coefficient as a function of frequency for a curved microstrip bend with Q = 90”, w = 73 µm, h = 100 µm, ϵ = 12.9 (Z = 50Ω), and various radii R.

5.2 Future Prospect

- Improve the order of perturbation.
- Changing the shape of the patch from angular to annular circular or elliptical.
- Studying the effect of change in the number of angle
- Properties of the antenna.
- Varying the probe positions.
- Trying different types of antenna feeding techniques.

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