DESIGN OF DEFECTED GROUND STRUCTURE (DGS) FOR MUTUAL COUPLING REDUCTION OF MICROSTRIP ANTENNA ARRAY

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ABSTRACT

Mutual coupling is a well-known effect in multi element array antennas. Generally, mutual coupling is an unwanted phenomenon that distorts the behavior of the radiating elements in an antenna array. Every element in an antenna array affects every other element by radiating over the air or by propagating surface currents through the ground plane. Surface currents can be a bigger problem, especially when antenna elements are closely packed. Microstrip patch antennas are well-known antenna types in different kinds of applications, like mobile and satellite communications. In practice, an array of several antenna elements is used for applications requiring high gain and/or high directivity. However, the performance of such antennas tends to drop due to the strong mutual interaction between the antenna elements. In this project, a two compact T-shaped defected ground structure (DGS) placed back to back are applied to reduce the mutual coupling between elements in a microstrip antenna array design. The proposed compact DGS slots are inserted between the adjacent E-plane coupled elements in the array to limit the propagation of surface waves between the elements of the array. In order to validate the feasibility of the proposed structure, a two-element array with $0.45\lambda_o$ distance between centers of two patches is designed, fabricated, and measured. The measured results show a reduction in mutual coupling of 19 dB obtained between elements at the operation frequency. Radiation patterns have minimal change in the broadside direction but back lobe level is increased because of the existence of the slots in the ground plane.

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CHAPTER 1

INTRODUCTION

1.1 Overview

Antenna is one of the important elements in the RF system for receiving or transmitting signals from and into the air as medium. Without proper design of the antenna, the signal generated by the RF system will not be transmitted and no signal can be detected at the receiver. Antenna design is an active field in communication for future development. Many types of antenna have been designed to suit with most devices. One of the types of antenna is the microstrip patch antenna (MPA). The idea of microstrip antenna was first presented in 1950s but it only got serious attention in the 1970s [1]. Microstrip antennas, in recent years, have been one of the most innovative topics in antenna theory and design. The basic idea of microstrip antenna came from utilizing printed circuit technology not only for the circuit component and transmission lines but also for the radiating elements of an electronic system. In general, the structure of MPA has a conductive strips and patches formed on the top surface of a thin dielectric substrate separating them from a conductive layer on the bottom surface which is the ground for the antenna [2]. A patch is typically wider than a strip and its shape and dimensions are important features of the antenna. Because of their simplicity and compatibility with printed-circuit technology, they are used in wide range of modern microwave applications such as mobile phones and satellite communications [2]. However, the electrical performance of the basic microstrip antenna suffers from a number of serious drawbacks, including very narrow bandwidth, low gain and the excitation of surface waves [3-5].

Over the years, a lot of research have been undertaken to overcome the disadvantages associated with these antennas. Some of the popular established techniques proposed by researchers to enhance the bandwidth of these antennas are increasing the substrate thickness using low dielectric substrate [5]; incorporating various impedance-matching and feeding elements using multi resonators [6], multi slot hole-coupling using planar parasitic elements [7], etc. The second important factor which falls as a disadvantage in the performance of these antennas is the gain that describes the directional property of antennas [8]. In many applications it is necessary to design antennas with very directive characteristics (very high gains) to meet the demands of long distance communication. This can only be accomplished by increasing the electrical size of the antenna [8]. Enlarging the dimensions of single elements often leads to more directive characteristics. Another way to enlarge the dimensions of the antenna, without necessarily increasing the size of the individual elements, is to form an assembly of radiating elements in an electrical and geometrical configuration. This antenna, formed by multi-elements, is referred to as array [8]. By using array in communication systems we enhance the performance of the antenna like increasing gain, directivity scanning the beam of an antenna system, and other functions which are difficult to do with a single element antenna. However, the performance of such antennas tends to drop due to the strong mutual interaction between the antenna elements. Therefore, in the design of microstrip antenna arrays, mutual coupling between radiating elements is an important factor to be considered [9].

1.2 Problem Statement

In many wireless communication systems it is necessary to design antennas with very directive characteristics (high gains) to meet the demands of long distance communication. For these applications, microstrip antenna arrays are preferred due to their well-known attractive features, such as low profile, light weight and low production cost. However, a common disadvantage of microstrip antennas is the reduced radiation efficiency due to generation of surface waves through the substrate layer. In arrays, surface waves have a significant impact on the mutual coupling between array elements. This effect is a potential source of the performance degradation which includes the impedance mismatching, the increased side-lobe level, the deviation of the radiation pattern from the desired one, and the decrease of gain due to the excitation of surface wave. In the application of continuous wave radar, when the transmitting antenna and the receiving antenna are placed closely side by side, the transmitting energy may even blockade the receiver. Therefore, the mutual coupling between elements of antenna arrays is a critical aspect that must be taken into account in the design process, as it can lead to severe degradations in the overall performance.

To suppress surface waves, several studies are conducted including using shorting pins to cancel the capacitive polarization currents of the substrate [10], adding parasitic conducting tape to the middle of two antennas [11, 12], using the dielectric as a band gap structure between elements in the array [13, 14], or in our case using defected ground structures (DGSs) technique which are widely used in microwave circuit and antenna design [15-17]. Many studies have used DGS for filter design, couplers, dividers, and microstrip antennas. Meanwhile, for antenna applications, DGS is mainly applied to the feeding technique of single element microstrip antennas. However, few researches focus on suppressing mutual coupling based on DGS techniques. Therefore, this project presented the design of a two-element Microstrip array with defected ground structure (DGS) to reduce the mutual coupling between the array elements. The proposed DGS is also valid for microstrip antenna arrays with more than two elements.

1.3 Objectives of Project

The objectives of this project are:

I. To design, simulate and fabricate microstrip antenna array operating at a resonant frequency of 2.4 GHz.
II. To study the effect of the mutual coupling between the array elements on the performance of the antenna.
III. To design Microstrip antenna array with defected ground structure to reduce the mutual coupling between them.
IV. To compare the performance of the DGS and conventional antenna.

1.4 The Scope of The project

I. Two element Microstrip Antenna Array with and without DGS in the ground plane was designed.
II. Separate microstrip line feeds was used to feed the two elements of the array to visualize the mutual coupling between the two elements.
III. CST microwave studio was used to simulate the E-plane coupled elements in the array
IV. Antenna with and without DGS has been fabricated to verify the simulation results and tested by using Vector Network Analyzer.
V. The Radiation Pattern of the array is limited to simulations only.

CHAPTER 2

LITERATURE REVIEW

2.1 Introduction to Microstrip Antenna

Microstrip antennas (MSA) received considerable attention in the 1970's, although the first designs and theoretical models appeared in the 1950's [18]. Because of their simplicity and compatibility with printed-circuit technology, they are used in wide range of modern microwave applications such as mobile phones and satellite communications. The MSA are low profile, mechanically robust, inexpensive to manufacture, compatible with MMIC designs and relatively light and compact. They are quite versatile in terms of resonant frequencies, polarization, pattern and impedance. However, microstrip antennas have some drawbacks including low efficiency (due to dielectric and conductor losses), low power, spurious feed radiation (surface waves, strips, etc.), narrow frequency bandwidth, relatively high level of cross polarization radiation. But with technology advancement and extensive research into this area these problems are being gradually overcome. Often microstrip antennas are also referred as patch antennas.

Microstrip Patch antennas, as shown in figure 2.1, consists of a radiating patch on one side of a dielectric substrate which has a ground plane on the other side as shown in Figure 2.1. The patch is generally made of conducting material such as copper or gold and can take any possible shape. The radiating patch and the feed lines are usually photo etched on the dielectric substrate of thickness h, which usually is from $0.003\lambda_0$ to $0.05\lambda_0$ [8].

There are numerous substrates that can be used for the design of microstrip antennas, and their dielectric constants are usually in the range of $2.2 \leq \epsilon_r \leq 12$.

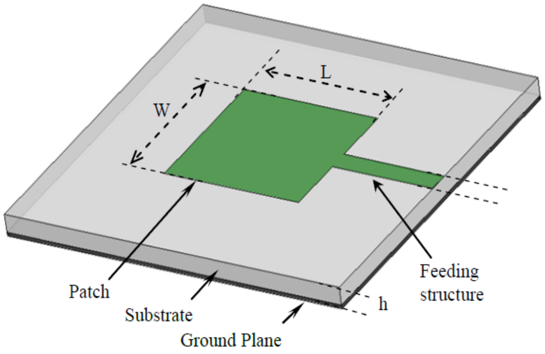


Figure 2.1 Basic microstrip patch antenna [4]

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 [8]. 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.

In order to simplify analysis and performance prediction, the patch is generally square, rectangular, circular, triangular, and elliptical or some other common shape as shown in Figure 2.2. Every shape has its own characteristics but square, rectangular, and circular are the most common configurations because of their easier analysis and fabrication.

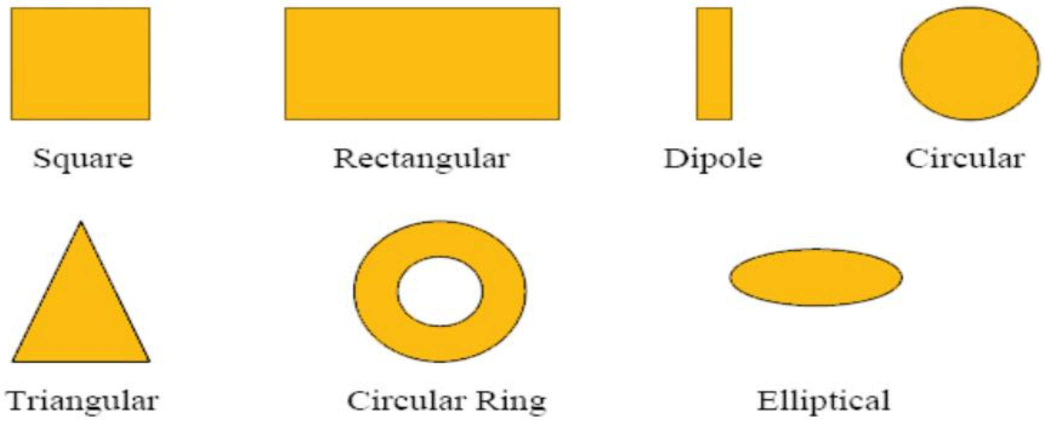


Figure 2.2 Common shapes of microstrip patch elements [19]

2.2 Feed techniques

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

2.2.1 Microstrip Line Feed

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

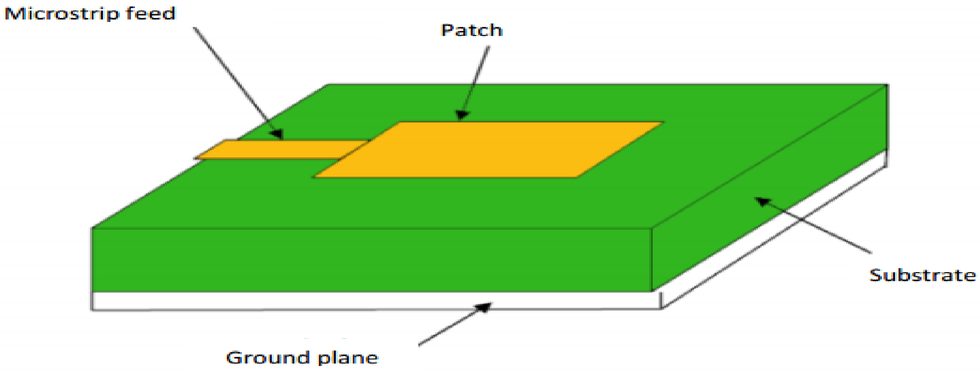


Figure 2.3 Microstrip line feed.

At the edges of the patch, the impedance is generally higher than 50 ohm. To avoid impedance mismatch, sections of quarter-wave transformers can be used to transform from a large input impedance to 50 ohm line, this is shown in figure 2.4.

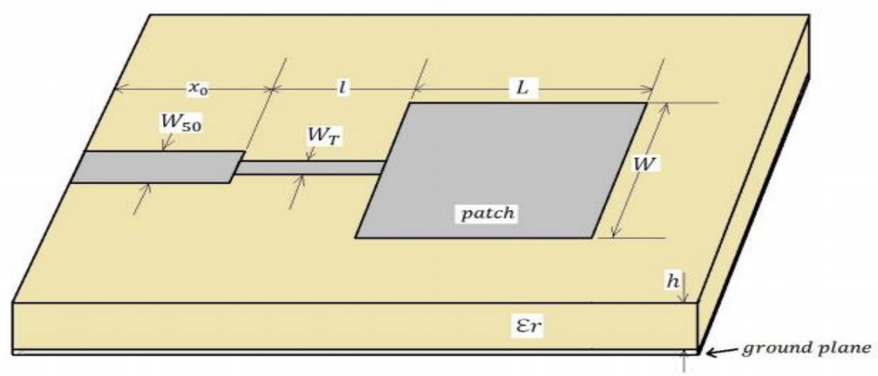


Figure 2.4 Microstrip line feed with quarter wave transformer.

A quarter-wave transformer uses a section of line of characteristic impedance Z_T of $\lambda/4$ long.

$$l = \frac{\lambda}{4} = \frac{\lambda_0}{4\sqrt{\varepsilon_{reff}}} \tag{2.1}$$

To have a matching condition, we want the input impedance of the patch (Z_{in}) equal to the line impedance (Z_0), this can be achieved by using this equation

$$Z_T = \sqrt{Z_0 Z_{in}} \tag{2.2}$$

In microstrip patch antennas, the total input admittance (Y_{in}) is real. Therefore, the resonant input impedance is also real, or

$$Z_{in} = \frac{1}{Y_{in}} = R_{in} \tag{2.3}$$

Another method of matching the antenna impedance is to extend the microstrip line into the patch. Since the input impedance is smaller at points away from the edges (e.g. center of the patch), this is achieved by properly controlling the inset position. Hence this is an easy feeding scheme, since it provides ease of fabrication and simplicity in modeling as well as impedance matching. However as the thickness of the dielectric substrate being used, increases, surface waves and spurious feed radiation also increases, which hampers the bandwidth of the antenna.

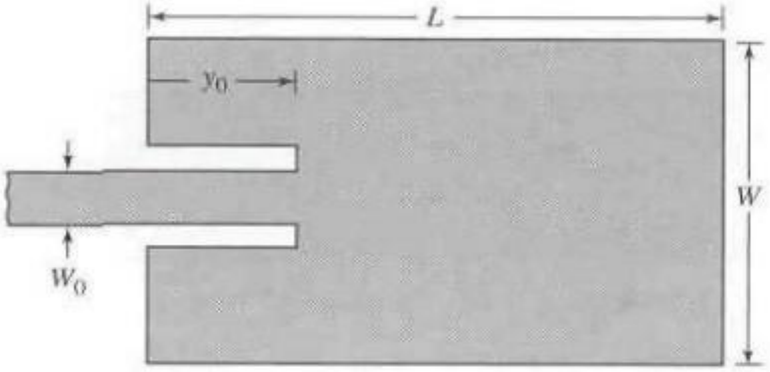


Figure 2.5 recessed microstrip line feed [8].

The impedance of the patch is given by [8]

$$R_{in} = \frac{1}{2(G_1 + G_{12})} \tag{2.4}$$

Where, G1 and G12 are self and mutual conductance’s expressed in section 2.3.1.

The impedance of the patch is also related to the electrical dimensions of the patch and dielectric constant of the substrate, which is given by

$$R_{in} = 90\frac{{\varepsilon_r}^2}{\varepsilon_r - 1}\left(\frac{L}{W}\right)^2 \tag{2.5}$$

The input impedance related to the length of the inset is given by

$$R_{in}(y = y_0) = \frac{1}{2(G_1 + G_{12})} \cos^2\left(\frac{\pi}{L} y_0\right) \tag{2.6}$$

Where y_0 is the inset length from slot at the feeding edge of patch, L is the length of the patch. Therefore,

$$R_{in}(y = y_0) = R_{in}(y = 0) \cos^2\left(\frac{\pi}{L} y_0\right) \tag{2.7}$$

where $R_{in}(y = 0)$ is the input impedance at the leading radiating edge of the patch and $R_{in}(y = y_0)$ is the desired input impedance (50 Ω).

2.2.2 Coaxial Feed

The coaxial feed or probe feed is the most popular technique used for feeding Microstrip patch antennas and is illustrated in Figure 2.6. The coaxial connector is attached to the ground plane and the coaxial center conductor extends through the substrate and is attached to the radiating patch. For coaxial probe the location of the feed is normally located at one third of the distance from the center of the patch to the side.

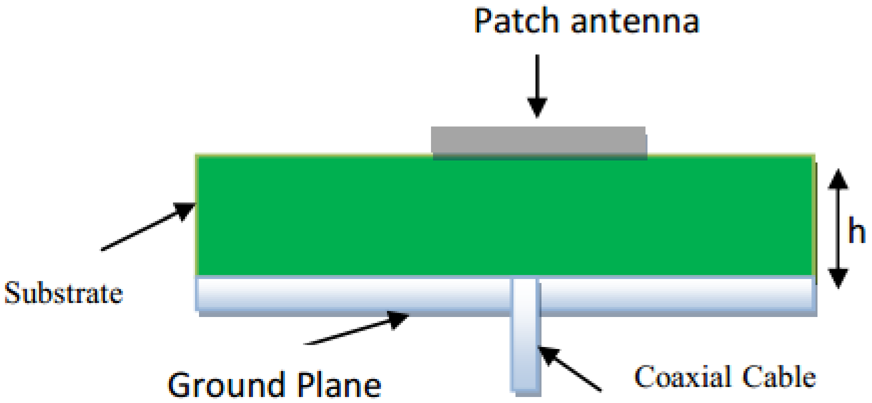


Figure 2.6 Probe fed Rectangular Microstrip Patch Antenna [20].

The main advantage of this type of feeding scheme is that the feed can be placed at any desired location inside the patch in order to match with its input impedance. This feed method is easy to fabricate and has low spurious radiation. However, a major disadvantage is that it provides narrow bandwidth and is difficult to model, especially for thick substrates ($h > 0.02\lambda_0$).

Both the microstrip feed line and coaxial probe possess inherent asymmetries which generate higher order modes which produce cross-polarized radiation. To overcome some of these drawbacks, non-contacting aperture coupling feeds have been introduced and discussed below.

2.2.3 Aperture Coupled Feed

In this type of feed technique, the patch ant microstrip feed line are on different sides of the ground plane as shown in Figure 2.7. A slot is cut in the ground plane to couple the electromagnetic to the radiating patch, thus no via connectors needed. This technique is to avoid spurious radiation escapes from the feed line and corrupt the side lobes or polarization of the antenna.

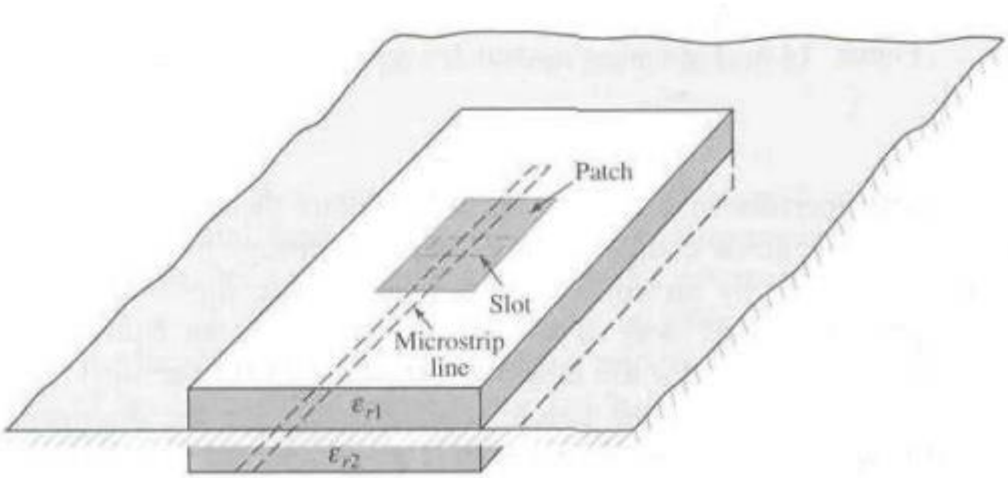


Figure 2.7 Aperture-coupled feed [8].

2.2.4 Proximity Coupled Feed

This configuration of feed technique is also called as the electromagnetic coupling scheme. As shown in Figure 2.8, two dielectric substrates are used such that the patch antenna is on the upper layer substrate and the Microstrip feed line on the lower layer substrate.

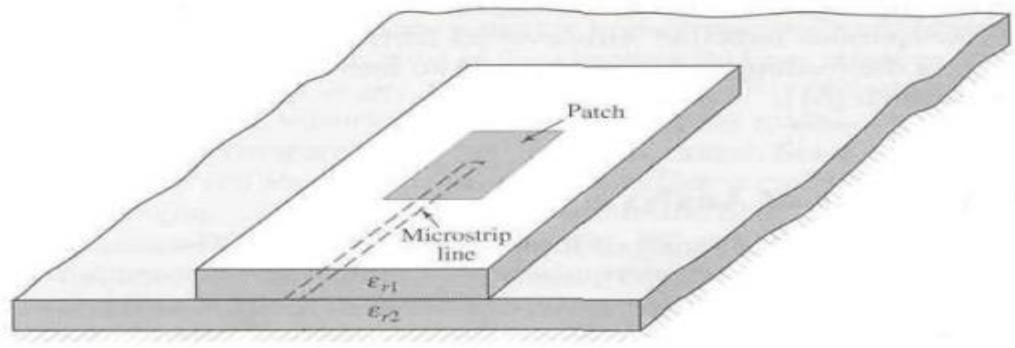


Figure 2.8 Proximity-coupled Feed [8].

The main advantage of this feed technique is that it eliminates spurious feed radiation and provides very high bandwidth due to overall increase in the thickness of the microstrip patch antenna. Matching also can be achieved by controlling the length of the feed line and the width-to-line ratio of the patch.

The major disadvantage of this feed scheme is that it is difficult to fabricate because of the two dielectric layers which need proper alignment. Also, there is an increase in the overall thickness of the antenna.

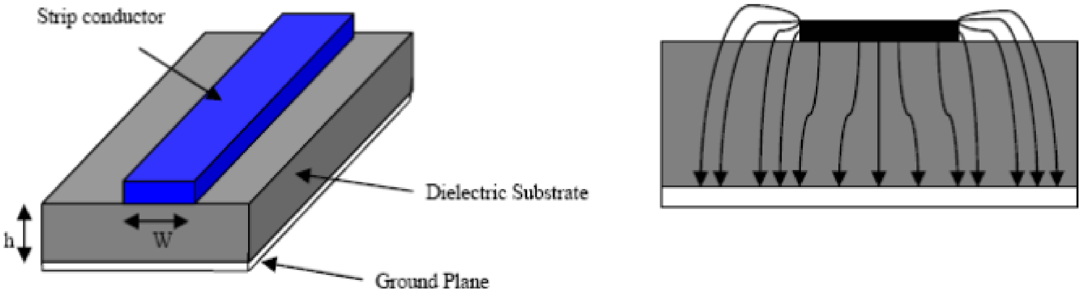
2.3 Methods of Analysis

The preferred models for the analysis of Microstrip patch antennas are the transmission line model, the cavity model, and the full wave model. The transmission line model is the easiest of all, but yields the least accurate results and it lacks the versatility and difficult to model coupling. Compared to the transmission line model, the cavity model is more accurate but is more complex in nature. The full wave models are extremely accurate, versatile and can treat single elements, finite and infinite arrays, stacked elements, arbitrary shaped elements and coupling. However, they usually give less insight as compared to the two models mentioned above and are far more complex in nature.

2.3.1 Transmission Line Model

This method represents the rectangular microstrip antenna as an array of two radiating slots of width W and height h, separated by a low impedance transmission line of certain length L. Since the dimensions of the patch are finite along the length and the width, the fields at the edges of the patch undergo fringing i.e. the field exists outside the dielectric thus causing a change in the effective dielectric constant. Therefore, in this model, fringing effects must be taken into account.

For a microstrip line shown in figure 2.9(a), typical electric filed lines are shown in figure 2.9(b). This is a nonhomogeneous line of two dielectrics; typically the substrate and air. As can be seen, most of the electric filed lines reside in the substrate and some lines exist in air. As $\frac{w}{h} \gg 1$and $\varepsilon_r \gg 1$, the electric filed lines concentrates mostly in the substrate. Fringing in this case makes the microstrip line look wider electrically compared to its physical dimensions. Since some of the waves travel in the substrate and some in air, an effective dielectric constant ε_{reff} is to account for fringing and wave propagation in the line.



(a) Microstrip line

(b) Electric fiel lines

Figure 2.9 Microstrip line and its field lines [8].

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 Fig. 2.9. The expression for ε_{reff} is given by [8]

$$\varepsilon_{reff} = \left(\frac{\varepsilon_r + 1}{2}\right) + \left(\frac{\varepsilon_r - 1}{2}\right) \left[1 + 12\,\frac{h}{W}\right]^{-\frac{1}{2}} \tag{2.8}$$

where ε_{reff} denotes effective dielectric constant, ε_r stands for dielectric constant of substrate, h represents height of dielectric substrate, and W identifies width of the patch. For dominant TM_{010} mode with no fringing, the length of the patch $L=\lambda/2$, where λ is the wavelength in the dielectric medium and is equal to $\lambda_0/\sqrt{\varepsilon_{reff}}$, where λ_0 is the free space wavelength. The TM_{010} model implies that the field varies one $\lambda/2$ cycle along the length and there is no variation along the width of the patch.

Because of the fringing effects, electrically the patch of the microstrip antenna looks greater than its physical dimensions; this is shown in figure 2.10.

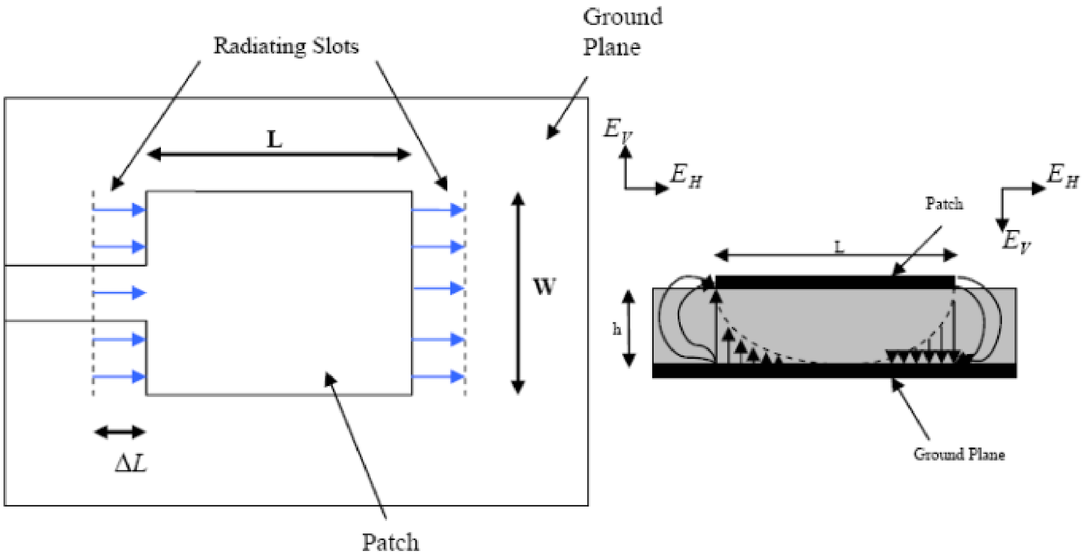


Figure 2.10 Transmission line model for patch antenna [8].

As seen in the figure the dimensions of the patch along its length have been extended on each end by a distance ΔL, which is a function of the effective dielectric constant and the width-to-height ratio (w/h). A practical approximation relation for the normalized extension of the length is given by [8]

$$\Delta L = 0.412h \left[\frac{\left(\varepsilon_{reff} + 0.3\right)\left(\frac{W}{h} + 0.264\right)}{\left(\varepsilon_{reff} - 0.258\right)\left(\frac{W}{h} + 0.8\right)}\right] \tag{2.9}$$

Since the length of the patch has been extended by ΔL on each side, the effective length of the patch is now

$$L_{eff} = L + 2\Delta L \tag{2.10}$$

For a given resonance frequency f_r, the effective length is given by

$$L_{eff} = \frac{c}{2f_r\sqrt{\epsilon_{reff}}} \tag{2.11}$$

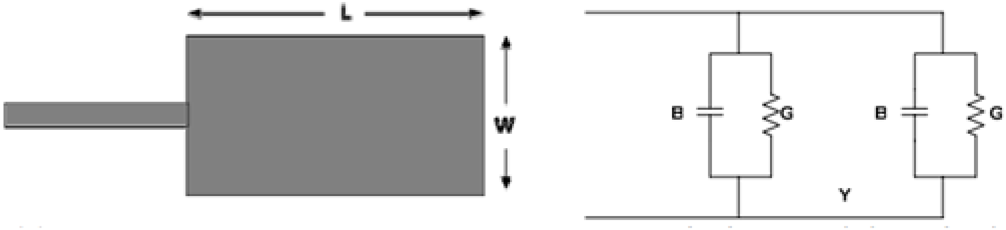
For the dominant TM_{010} mode with no fringing, the resonant frequency f_r of the mictrostip patch antenna is a fucntion of its length. Usually is ginven by the effective length is given by

$$f_r = \frac{1}{2L\sqrt{\epsilon_r}\sqrt{\epsilon_0\mu_0}} = \frac{c}{2L\sqrt{\epsilon_r}} \tag{2.12}$$

Where c is the speed of light in free space. Since Equation (2.12) does not account for fringing, it must be modified to include edge effects and should be computed using

$$f_r = \frac{c}{2L_{eff}\sqrt{\epsilon_{reff}}} \tag{2.13}$$

Figure2.11 shows a typical rectangular patch antenna and its equivalent circuit using transmission-line model.



(a) Rectangular Patch (b) Transmission Model equivalent

Figure 2.11 Rectangular microstrip patch and its equivalent circuit transmission model [8].

Each radiating slots corresponding to the edges of the patch is represented by parallel equivalent admittance, Y, which is given by

$$Y = G + JB \tag{2.14}$$

where Gand Brepresents the conductance and the susceptance produced by the slot or the radiating edge of the microstrip patch. These conductance and susceptance are typically expressed as [8],

$$G = \frac{W}{120\lambda_0}\left[1 - \frac{1}{24}(k_0h)^2\right] \qquad \frac{h}{\lambda_0} < \frac{1}{10} \tag{2.15}$$

$$B = \frac{W}{120\lambda_0}[1 - 0.6336ln(k_0h)] \qquad \frac{h}{\lambda_0} < \frac{1}{10} \tag{2.16}$$

where W is the width of the microstrip patch and h is the height of the substrate, λ_0 is the free-space wavelength of the propagating electromagnetic wave. Thus the resonant input impedance can be given as,

$$R_{in}(y = 0) = \frac{1}{2(G_1 + G_{12})} \tag{2.17}$$

Where G_{12} is the mutual conductance between the slots which are representative of patch edges and is given by,

$$G_{12} = \frac{1}{120\pi^2} \int_0^{\pi} \left[\frac{\sin\left(\frac{k_0 W}{2} \cos\theta\right)}{\cos\theta} \right]^2 J_0(k_0 L \sin\theta) \sin^3\theta d\theta \qquad (2.18)$$

where J_0 is the Bessel function of the first kind and order zero.

2.3.2 Cavity Model

Cavity model is more complex than transmission line model and provides more accurate results. In this model, the microstrip antennas is based on the assumption that the region between the microstrip patch and ground plane is a resonance cavity bounded by ceiling and floor of electric conductors and magnetic walls along the edge of the conductor as shown in figure (2.12) [8], [19].

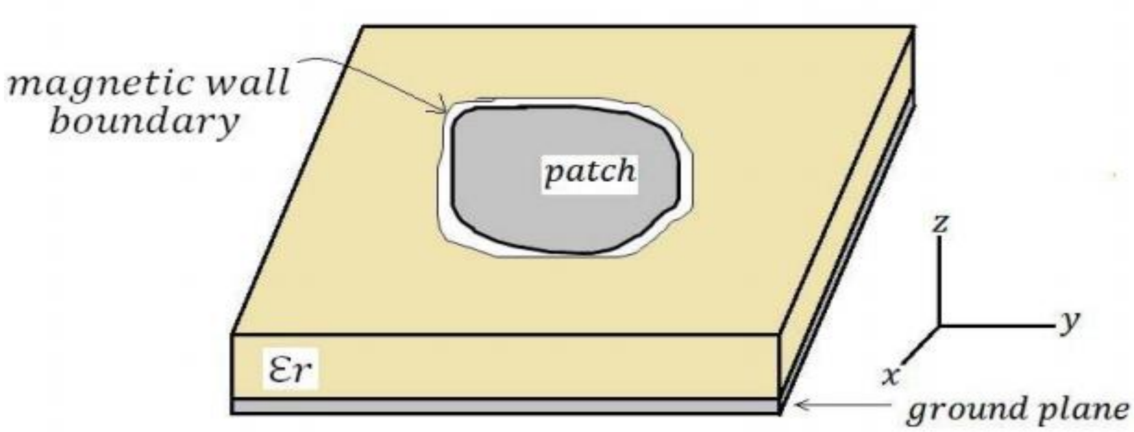


Figure 2.12 Magnetic wall model of a microstrip patch antenna

Since thin substrates are used, the field inside the cavity is assumed to be uniform along the thickness of the substrate [8]. In cavity model, the analysis is made simple, by expressing the electromagnetic fields within the patch substrate, as a summation of the various resonant modes of the two dimensional radiator (i.e., the patch in this case). Since the normal substrates that are used to produce the microstrip patch antennas are thin, the usual assumption is that the field inside the cavity is uniform along the thickness of the substrate [19].

The cavity model makes the following assumptions [21]

- There are only three field components in the region enclosed by the cavity: E component in the z axis (E_z) and two components of ؛ along the x and y axis (H_x , H_y).
- Because h (height of the substrate) is very thin (h<<λ), field in the interior region do not vary with z-coordinates for all frequencies.
- The electric current in the microstrip patch has no component normal to the edge of the patch at any point.

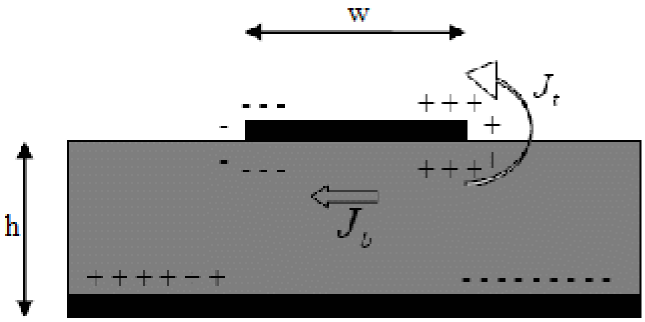


Figure 2.13 Charge distribution and current density creation on microstrip patch [8]

When the microstrip antenna is connected to a microwave source, a charge distribution is seen on the upper and lower surfaces of the patch and at the bottom of the ground plane. The charge distribution is controlled by two mechanisms; attractive and repulsive. The attractive force is between the opposite charges on the patch and on the ground plane, it creates a current density inside the dielectric J_b at the bottom of the patch.

The repulsive mechanism is between the like charges on the bottom surface of the patch, which causes pushing of some charges from the bottom, to the top of the patch. As a result of this charge movement, currents flow at the top and bottom surface of the patch. The cavity model assumes that the height to width ratio (i.e. height of substrate and width of the patch) is very small and as a result of this the attractive mechanism dominates and causes most of the charge concentration and the current to be below the patch surface.

Much less current would flow on the top surface of the patch and as the height to width ratio further decreases, the current on the top surface of the patch would be almost equal to zero, which would not allow the creation of any tangential magnetic field components to the patch edges. Hence, this would allow the four side walls to be modeled as perfect magnetic conducting surfaces which ideally would not disturb the magnetic field and in turns the electric field distribution beneath the patch". This good approximation to the cavity model leads us to deal with the side walls as perfect magnetic conducting walls. We have mentioned before that the field inside the cavity has three field components E_z, H_X and H_Y; the wave equation (2.19) can be re-written as equation (2.20)

$$\nabla \times \nabla \times \vec{E} - k^2\vec{E} = -j\omega\mu_0\vec{J} \tag{2.19}$$

$$\nabla^2 E_Z - k^2 E_Z = j\omega\mu_0\hat{Z} \bullet \vec{J} \tag{2.20}$$

Where $k^2 = \omega^2\mu_0\varepsilon_0\varepsilon_r$ is the wave number.

$\overrightarrow{J =}$ Electric current density fed by the feed line to the patch.

$\hat{Z}$ Is the unit vector normal to the plane of the patch.

In addition we have on the top and the bottom conductors:

$$\hat{n} \times \vec{E} = 0 \tag{2.21}$$

And on the walls:

$$\hat{n} \times \vec{H} = 0 \tag{2.22}$$

2.3.3 Full wave Model

Full wave models are very versatile and can provide very accurate results. Method of Moments [22], the Finite Difference Time Domain method [23], the Finite element method (FEM) [24] are all belong to this category, they are suitable for volumetric configurations. The Finite Element Method (FEM) is more popular amongst these methods and in this method the region of interest is divided into any number of finite surfaces or volume elements depending upon the planar or volumetric structures to be analyzed. These discretized units, generally referred to as finite elements, are well defined geometrical shapes, such as triangular elements for planar configurations and tetrahedral and prismatic elements for three dimensional configurations [25].

2.4 Microstrip Antenna Arrays

In many wireless communication systems it is necessary to design antennas with very directive characteristics (high gains) to meet the demands of long distance communication that may not be achievable by a single element antenna. The radiation from the single element is often very wide in pattern with large beam angles. This is not good for point to point communications, which requires antennas that are more directive in nature i.e. Radar applications. Also, a single radiating element often generates radiation patterns with unacceptable bandwidth, efficiency, and gain parameters. All these and more make the utilization of a single element antenna not recommendable. Therefore, the implementation of antennas in array configuration overcomes these drawbacks [8].

Antenna arrays are basically a collection of radiating elements, geometrically arranged in a specific manner, to generate the required radiation pattern. Each antenna in the array is known as an element, and it can be anything from simple dipole antenna, monopole antennas, horn antennas, or as in this case microstrip patches.

Arrays are categorized into uniform and non-uniform. Uniform arrays are the simplest one-dimensional array antenna, where the signal inputted in each identical element consists of identical amplitude and equal differential phase distribution. This class of array has the narrowest main-lobe and considerable amount of side-lobes. On the other hand, non-uniform array antenna with unequal amplitude distribution yields a more controlled side-lobe level.

The phased array is a special type of antenna array, where the spatial distributions of the radiated fields are electronically scanned to enhance the desired signal, by introducing differential phase (and/or magnitude) in the input signal of the radiating elements. Phased-array antennas have been developed mainly for radar applications but are being used more now for space-based communications applications because of their advantages in scanning, re-configurability, weight, and power [26]. Also, because of the development in the integration technology of small microwave circuits, has led the deployment of these antennas in ground, ship, air and space communication.

2.4.1 Feed network

The elements of antenna array can be fed by a single line called the series-feed network or by multiple lines called corporate-feed network [2].

In series-fed linear array, the radiating elements are connected in series along a uniform transmission line as shown in figure 2.14. There are two types of series feed configuration, which are the traveling-wave series-fed array and the resonant series-fed array. In the case of traveling-wave series-fed array, the impedance of the transmission line and the radiating elements are matched. In this type of array, the element spacing along the transmission line controls the phases of the elements. The elements must be placed one wavelength to achieve broadside radiation. Spacing of less than a wavelength will result in an off-broadside beam. The direction of main beam will shift as frequency changes. Since the wave on the transmission line is a traveling wave, energy decays along the feed line toward the last element. A matched load at the end can absorb the remaining energy or the remaining energy will be reflected to the in input in phase for the broadside beam.

For the resonant series-fed array, the impedance at the junction of the feed line and the radiators are not matched. But the antenna elements are properly spaced. The element spacing can be one wavelength or multiple integrals of one wavelength.

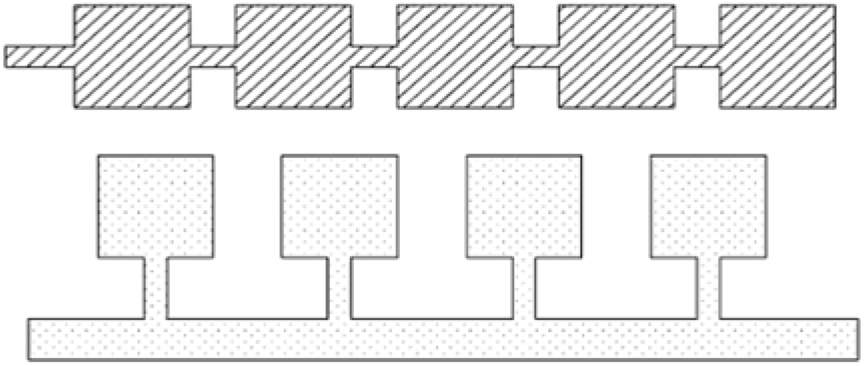


Figure 2.14 Series-fed array (4-element) [27].

The second type of array feeding network is parallel feed which is also called the corporate feed. This feeding method is to feed all elements in parallel by a power division transmission line. The corporate feed transmission line splits into N transmission lines at regular intervals. This classical arrangement uses T-junctions that split the signal using impedance matching techniques. This is shown in figure 2.15.

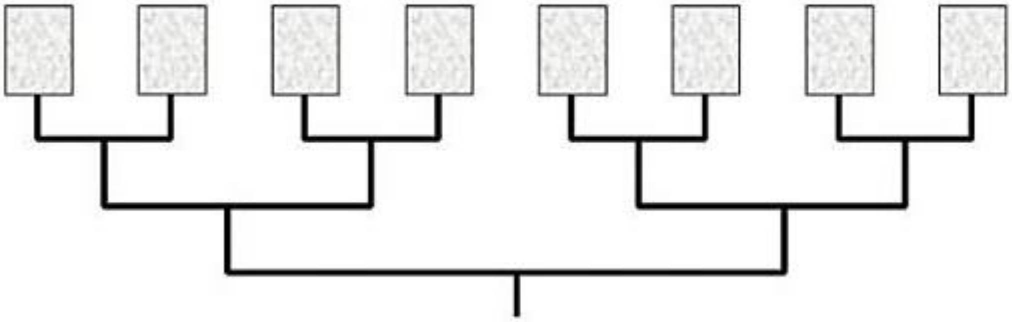


Figure 2.15 Corporate-fed array (8-element) [27].

2.4.2 Quarter-wave transformer

In order to match the input port to each element of an array, appropriate feeding circuits should be employed. The quarter-wave transformer is used extensively. The equation which relates the input impedance, to the load impedance and the characteristic impedance is

$$Z_0 = \sqrt{Z_{input} Z_{load}} \tag{2.23}$$

Using this method, smooth transition of power can be achieved between two transmission lines with different impedances [28]. But since the matching section is quarter wave long only for the design frequency, λ/4 transformers are associated with low impedance bandwidth. Figure 2.16, shows microstrip patch antenna with quarter-wavelength transmission line. The impedance of microstrip patch antennas ranges from 150 to 300 Ω, the purpose of using λ/4 transformers is to match the impedance of the patch with 50Ω line, because most of microwave applications are designed with an input impedance of 50Ω.

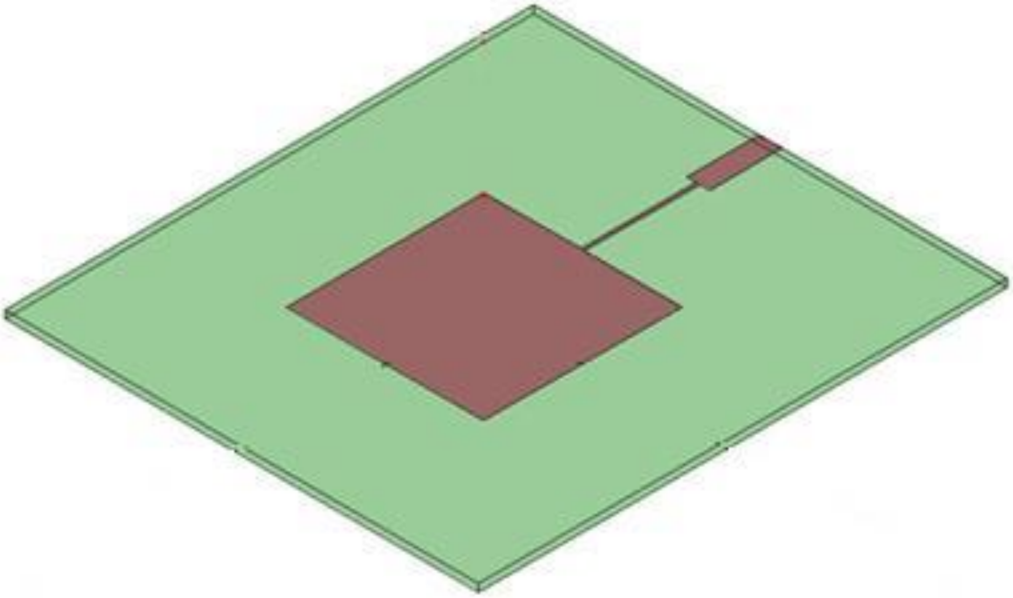


Figure 2.16 MPA with quarter-wavelength transmission line.

2.4.3 Array network power dividers

The design process of the antenna array requires the dimensions of associated array feeder. Array feeder is a power divider network that provides required transmission, reflection and isolation properties. Various types of microstrip power divider circuits have been developed by the researchers. The most popular ones being the quadrature hybrid, annular or ring, Wilkinson and T-junction power divider. By definition, a -3dB power divider is ideally a passive lossless reciprocal three port device that divides power equally in magnitude and phase.

Figure 2.17 shows a three port lossless and reciprocal power divider with all matched ports.

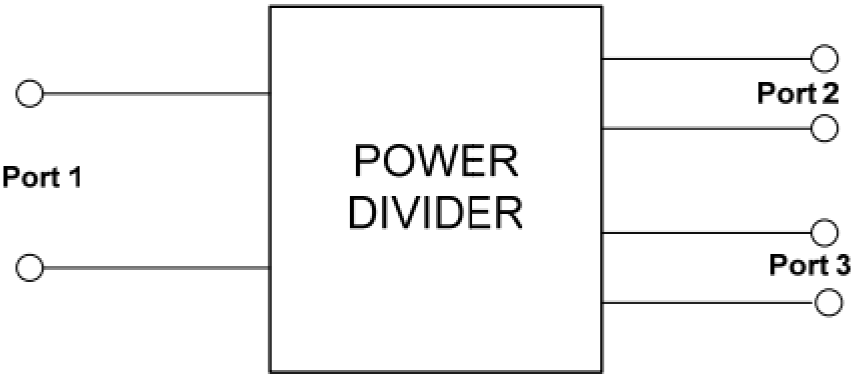


Figure 2.17 Block diagram power divider.

Three-port network power dividers with one input and two outputs have a scattering matrix with the following nine independent elements

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix} \tag{2.24}$$

For reciprocal networks the [S] matrix is symmetric and ($S_{ij} = S_{ji}$). Ideally, to avoid any loss of power, the network would be lossless and matched at all ports. When all ports are matched ($S_{ii} = 0$) and the reciprocal matrix reduces to 6.

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