DESIGN AND ANALYSIS OF PATCH ANTENNA WITH QUARTER WAVE MULTISECTION TRANSFORMER FOR WLAN/WiMAX APPLICATIONS

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Thesis Submitted By

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Thesis Title

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Date: ____________________ ______________________

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Chapter 1

1 Introduction

Antennas are one of the important elements in RF systems for transmitting or receiving signals through air which acts as a medium. For this, antenna should be properly designed otherwise the signal generated by the RF system will not be transmitted and no signal can be detected at the receiver. Antenna design particularly ‘Microstrip patch’ is an active field of research in communication of small and different shapes of Microstrip patch in view of the miniaturization of devices. Microstrip antenna involves most innovative design area in the antenna engineering due to its low material cost, easy to fabricate, low profile, simple and inexpensive to manufacture using modem printed circuit technology.

1.1 Preface

Impedance matching is also an important factor in RF applications. If properly matched the imaginary part of the input impedance has to be exactly zero, such that there should not be any reactance part of the input impedance.

The quarter-wave transformer is a simple and useful circuit for matching a real load impedance to a transmission line. Also we can extend the single quarter wave section to multisection designs in a methodical manner to provide larger bandwidth if required. For narrow band impedance matching , a single-section transformer may be sufficient.

1.2 Objective of the thesis

We have designed a patch antenna having a feed line made up with quarter wave multisection matching technique. By setting, each section of length λ/4, we have designed a tapered section to match a 220 ohm patch to a 50 ohm source impedance line using binomial transformer matching technique.The passband response (the frequency band where a good impedance match is achieved) of a binomial matching transformer is optimum in the sense that, for a given number of sections, the response is as flat as possible near the design frequency. This type of response, which is also known as maximally flat response.

Here We have also studied the effect of different design parameters like patch length, width, dielectric constant and feed location on performance parameters like Return loss, VSWR, Real and Imaginary impedance, Smith chart, Radiation patterns, Radiation efficiency, gain, directivity and Bandwidth for the patch as well as patch with that matching section.

Chapter 2
Literature Review

2.1. Introduction

In this chapter, We have given an overview of the Microstrip Patch Antenna. We have also discussed it's advantages and disadvantages. And in the end, we have discussed different feed techniques that are used for microstrip antenna.

2.1.1 Microstrip Patch Antenna

A Microstrip patch antenna has a radiating patch which is usually made up of a metal on one side of a dielectric substrate ,and 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. It can take any possible shape required as per application. The radiating patch and the feed lines are etched on the dielectric substrate by means of photolithography.

For a rectangular patch, The length of the patch L is usually in the range of $\lambda o < L < 0.5\ \lambda o$, where λo is the free-space wavelength. The patch thickness has to be very thin in general about $t \ll \lambda o$ (where t is the patch thickness). The height h of the dielectric substrate is usually $0.003\lambda o \leq h \leq 0.05\lambda o$. The dielectric constant of the substrate (ε_r) is typically within the range of 2.2 to 12.

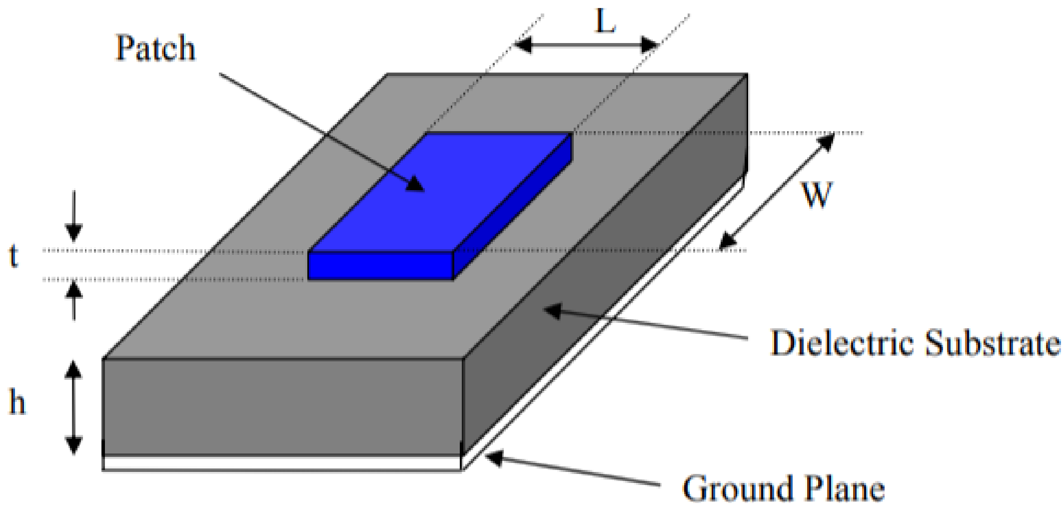


Fig.2.1 Structure of a Microstrip Patch Antenna

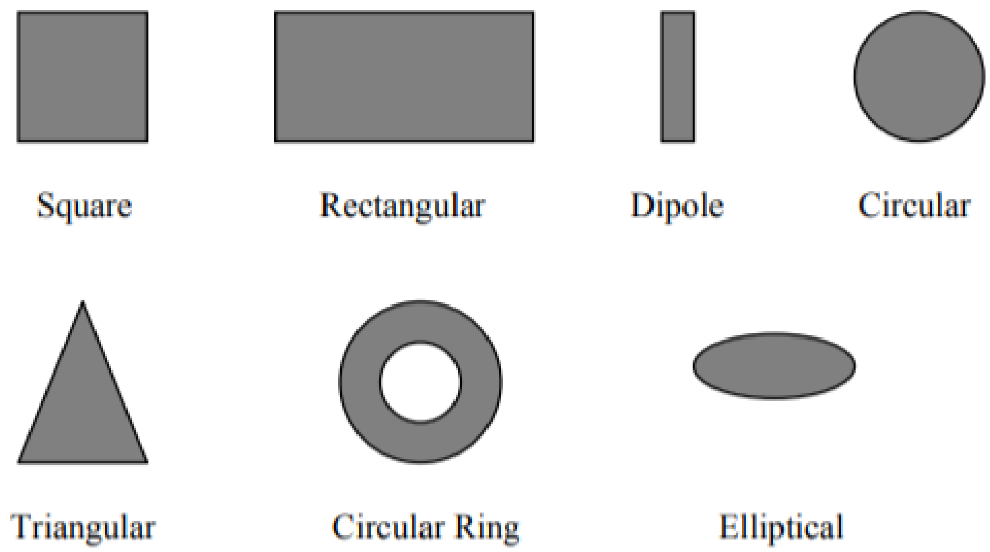


Fig 2.2. Different types of Microstrip Antenna

Microstrip patch antennas generally radiate due to fringing fields between the patch edge and the ground plane . Generally a thick dielectric having low dielectric constant is chosen for good antenna performance since this provides better efficiency, larger bandwidth and better radiation. However, this kind of configuration results a larger antenna size. So, In designing a compact Microstrip patch antenna, dielectrics with higher dielectric constants must be used which are less efficient and also causes narrower bandwidth. Hence we have to compromise between antenna dimensions and antenna performance to get a optimum result.

2.2 Advantages and Disadvantages

➡They operate at microwave frequencies where traditional antennas are not feasible to be designed.

➡This antenna type has smaller size and hence will provide small size end devices.

➡The microstrip based antennas are easily etched on any PCB and will also provide easy access for troubleshooting during design and development. This is due to the fact that microstrip pattern is visible and accessible from top. Hence they are easy to fabricate and comfortable on curved parts of the device. Hence it is easy to integrate them with MICs or MMICs.

➡As the patch antennas are fed along centreline to symmetry, it minimizes excitation of other undesired modes.

➡The microstrip patches of various shapes e.g. rectangular, square, triangular etc. are easily etched.

➡They have lower fabrication cost and hence they can be mass manufactured.

➡They are capable of supporting multiple frequency bands (dual, triple).

➡They support dual polarization types viz. linear and circular both.

➡They are light in weight.

➡They are robust when mounted on rigid surfaces of the devices.

Following are the disadvantages of Microstrip Antenna:

➡The spurious radiation exists in various microstrip based antennas such as microstrip patch antenna, microstrip slot antenna and printed dipole antenna.

➡It offers low efficiency due to dielectric losses and conductor losses.
➡It offers lower gain.
➡It has higher level of cross polarization radiation.
➡It has lower power handling capability.
➡It has inherently lower impedance bandwidth.
➡The microstrip antenna structure radiates from feeds and other junction points.

2.3 Feed Techniques

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

2.3.1 Different types of feeding techniques

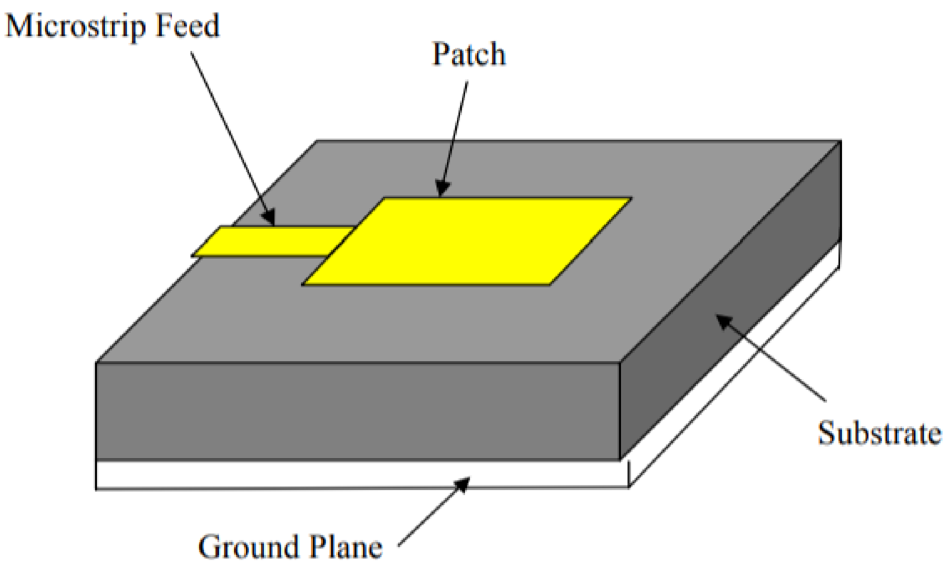


Fig. 2.3. Microstrip Line Feed

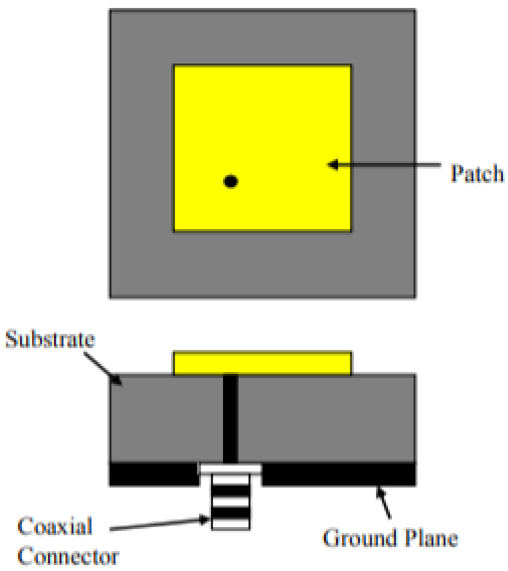


Figure 2.4 Probe fed Rectangular Microstrip Patch Antenna

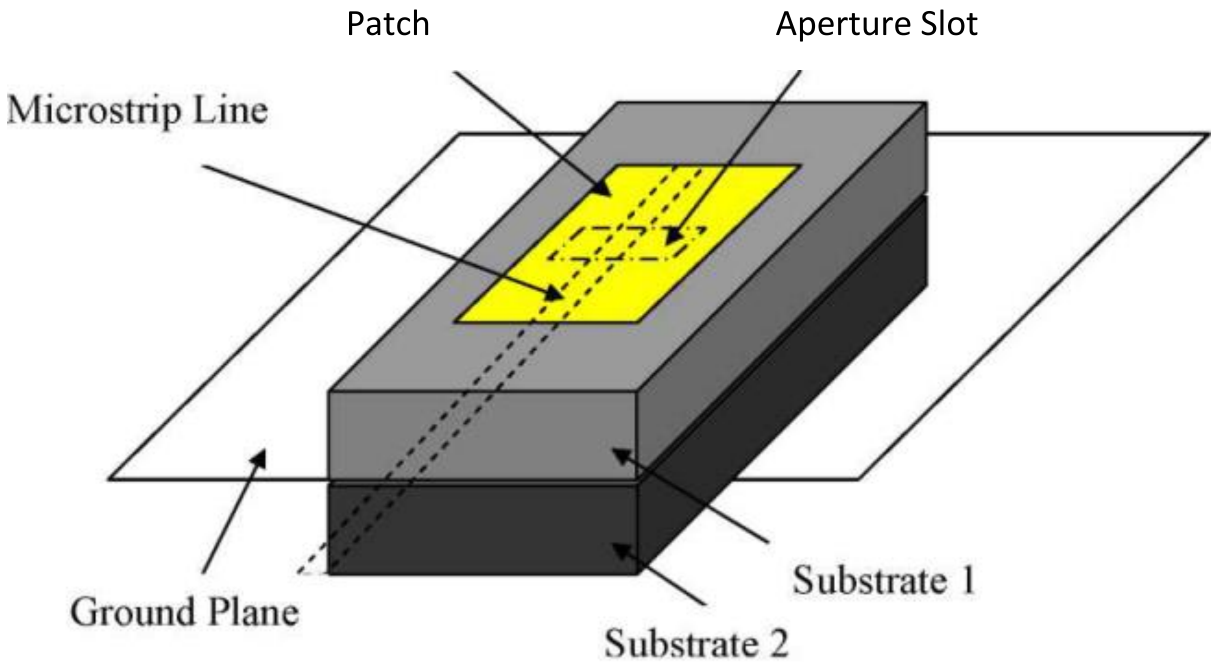


Figure 3.5 Aperture-coupled feed

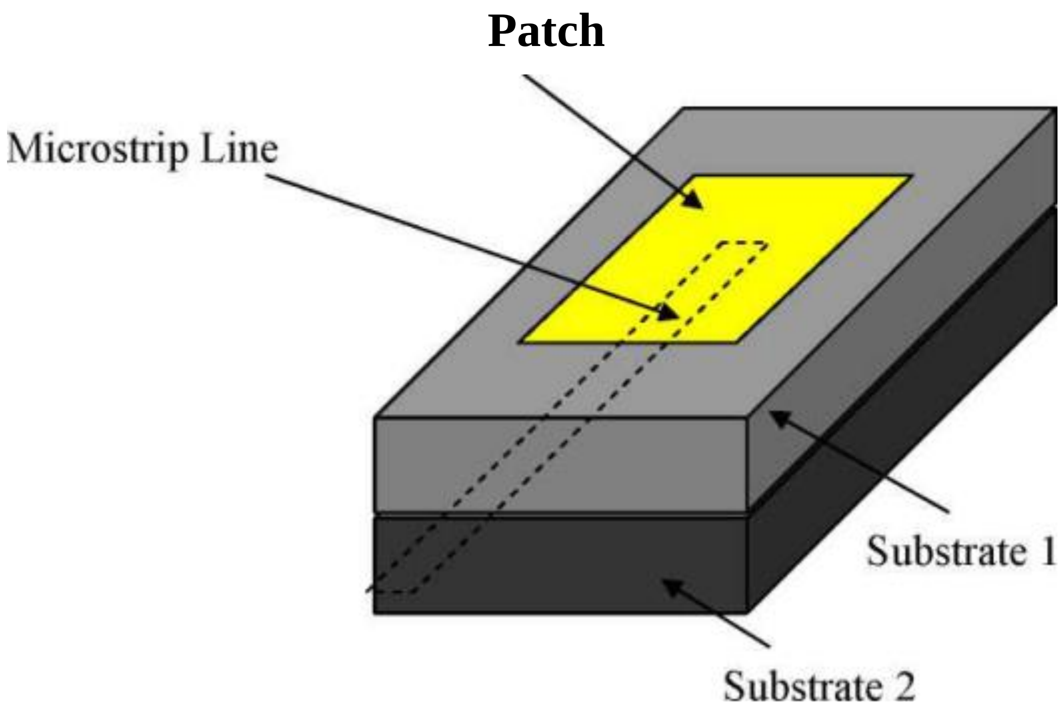


Figure 2.6 Proximity-coupled Feed

Table 3.1 Comparing the different feed techniques

Characteristics	Microstrip Line Feed	Coaxial Feed	Aperture coupled Feed	Proximity coupled Feed
Spurious feed radiation	More	More	Less	Minimum
Reliability	Better	Poor due to soldering	Good	Good
Ease of fabrication	Easy	Soldering and drilling needed	Alignment required	Alignment required
Impedance Matching	Easy	Easy	Easy	Easy
Bandwidth (achieved with impedance matching)	2-5%	2-5%	2-5%	13%

2.5. Antenna Parameters

There are several antenna parameters that are useful to describe the performances of microstrip patch antenna. Some of these are as follows:

2.5.1 Frequency Bandwidth

Frequency bandwidth is the range of frequencies within which the performance of the antenna, with respect to some characteristic, conforms to a specified standard. The bandwidth can be considered to be the range of frequencies, on either side of the centre frequency, where the antenna characteristics are within an acceptable value of those at the centre frequency. In wireless communications, the antenna is required to provide a return loss less than -10dB over its frequency bandwidth.

The frequency bandwidth of an antenna can be expressed as either absolute bandwidth (ABW) or fractional bandwidth (FBW). The f_H and f_L denote the upper edge and the lower edge of the antenna bandwidth, respectively. The ABW is defined as the difference of the two edges and the FBW is designated as the percentage of the frequency difference over the centre frequency, as given in equation 2.1 and 2.2, respectively.

$$ABW = \frac{f_H - f_L}{f_H + f_L} \qquad (2.1)$$

Where, f_H = higher frequency, f_L = lower frequency

$$FBW = 2\frac{f_H - f_L}{f_H + f_L} \qquad (2.2)$$

For broadband antennas, the bandwidth can also be expressed as the ratio of the upper to the lower frequencies, where the antenna performance is acceptable, as shown in equation 2.3

$$BW = \frac{f_H}{f_L} \qquad (2.3)$$

2.5.2 Return Loss

Return loss is a measure of the effectiveness of power delivery from a transmission line to a load such as an antenna. If the power incident on the antenna under test is Pin and the power reflected back to the source is P_{ref}, the degree of mismatch between the incident and reflected power in the travelling waves is given by the ratio P_{in}/P_{ref}. The higher this power ratio is, the better the load and line are matched. Return loss is the negative of the magnitude of the reflection coefficient in dB. Since power is proportional to the square of the voltage, return loss is given by:

$$RL = 10\log_{10}\frac{P_{in}}{P_{ref}}\ dB \qquad (2.4)$$

which is positive quantity if $P_{ref} < P_{in}$

2.4.3 Radiation Pattern

Radiation pattern defines the variation of the power radiated by an antenna as a function of the direction away from the antenna. It is a graphical representation of the radiation properties of the antenna as a function of space coordinates. Also the antenna radiation pattern is a measure of its power or radiation distribution with respect to a particular type of coordinates. We generally consider spherical coordinates as the ideal antenna is supposed to radiate in a spherically symmetrical pattern. Almost in all cases the radiation pattern is determined in the far field.

2.4.4 Antenna Gain

The most important figure of merit that describes the performance of an antenna radiator is the gain. The term antenna gain describes how much power is transmitted in the direction of peak radiation to that of an isotropic source.

The relative gain is the ratio of the power gain in a given direction to the power gain of a reference antenna in its referenced direction. In most of cases the reference antenna is a lossless isotropic source.

When the direction is not specified, the power gain is usually taken in the direction of maximum radiation. A transmitting antenna gain of 3 dB means that the power received far from the antenna will be 3 dB higher than what would be received from a lossless isotropic antenna with the same input power. Similarly, a receiving antenna with a gain of 3 dB in a particular direction would receive 3 dB more power than a lossless isotropic antenna

2.4.5 Group Delay

Group delay is one of the important parameters while discussing the wideband microstrip antenna. Group delay provides the pulse-handling capability. It represents the degree of distortion in the pulse signal. It is a useful measure of time distortion which is usually calculated by differentiating phase with respect to frequency. It evaluates non dispersive behaviour of antenna as a derivative of far field response with respect to frequency. If group delay variation exceeds more than 1 ns, phases are no more linear in far field and phase distortion occurs which can cause a serious problem for wideband applications.

2.4.6 Efficiency

Antenna radiation efficiency is defined as the ratio of power radiated to the input power. It relates the gain and directivity. Radiation efficiency also takes into account conduction and dielectric losses.

2.4.7 Polarization

Polarization of an antenna in a given direction is defined as “the polarization of the wave transmitted or radiated by the antenna”. When the direction is not defined then the polarization is taken to be in the direction of maximum gain. The polarization of the radiated energy varies with the direction from the center of the antenna, so that different parts of the pattern may have different polarizations. Polarization describes the time varying direction and relative magnitude of the E-field.

At each point on the radiation sphere the polarization is resolved into a pair of orthogonal polarizations, the co-polarization and cross-polarizations. Co-polarization represents the polarization the antenna is intended to radiate while cross-polarization represents the polarization orthogonal to a specified polarization.

2.4.8 Input Impedance

Input impedance is defined as the “the impedance presented by an antenna at its terminals, or ratio of the voltage to current at a pair of terminals, or the ratio of the appropriate components of the electric to magnetic fields at a point.

Chapter 3

3.1. Impedance Matching and Tuning

Preface

This chapter marks techniques of impedance matching techniques to solve real life problems in microwave engineering which is often important in larger design process for a microwave component or system.

3.1 Introduction

The basic idea of impedance matching is illustrated in Figure 3.1, which shows an impedance matching network placed between a load impedance and a transmission line. In order to avoid unnecessary loss of power,The matching network is ideally lossless, and is usually designed in such manner so that the impedance seen looking into the matching network is Z_0. If the impedance matching section is designed properly, the reflections will be eliminated on the transmission line to the left of the matching network, although there will usually be multiple reflections between the matching network and the load, which will include some loss in power transferred and there will be a shift in resonance frequency. This procedure is sometimes referred to as tuning.

- Impedance matching or tuning is important for the following reasons:

Maximum power is delivered when the load is matched to the line (assuming the generator is matched), and power loss in the feed line is minimized.

Impedance matching in a power distribution network (such as an antenna array feed network) may reduce amplitude and phase errors.

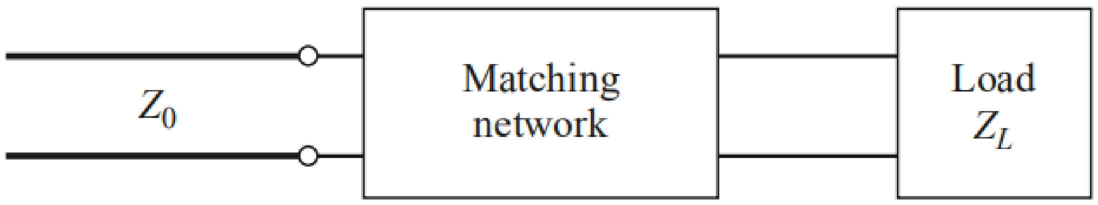


Fig 3.1 A lossless network matching an arbitrary load impedance to a transmission line.

As long as the load impedance, Z_L, has a positive real part, a matching network can always be found. There are various types of options are available, we will discuss the design and performance of several types of practical matching networks. Factors which are important in the selection of a particular matching network may include the following,

Complexity—As with most engineering solutions, the simplest design that satisfies the required specifications is generally preferable. A simpler matching network usually is relatively cheaper, smaller in size , also more reliable, and has losses relatively lesser than a more complex design.

Bandwidth—If a network is matched properly , any type of network can ideally give a perfect match (zero reflection) at a single frequency. But In many types of applications, it is often required to match a load over a range of band of frequencies. There are several ways of doing this, with, of course, a corresponding increase in complexity.

Implementation— Depending on the type of transmission line or waveguide being used, we can select the right type of matching network that is most suitable for the operation. For example, tuning stubs are much easier to implement in waveguides than are multisection quarter-wave transformers.

3.2 THE QUARTER-WAVE TRANSFORMER

The quarter-wave transformer is a simple and useful circuit for matching a real load impedance to a transmission line. An additional feature of the quarter-wave transformer is that it can be extended to multisection designs in a methodical manner to provide broader bandwidth. If only a narrow band impedance match is required, a single-section transformer may suffice. However, as we will see in the next few sections, multisection quarter-wave transformer designs can be synthesized to yield optimum matching characteristics over a desired frequency band.

One drawback of the quarter-wave transformer is that it can only match a real load impedance. A complex load impedance can always be transformed into a real impedance, however, by using an appropriate length of transmission line between the load and the transformer, or an appropriate series or shunt reactive element. These techniques will usually alter the frequency dependence of the load, and this often has the effect of reducing the bandwidth of the match. We will first approach the problem from the impedance viewpoint and then show how this result can also be interpreted in terms of an infinite set of multiple reflections on the matching section.

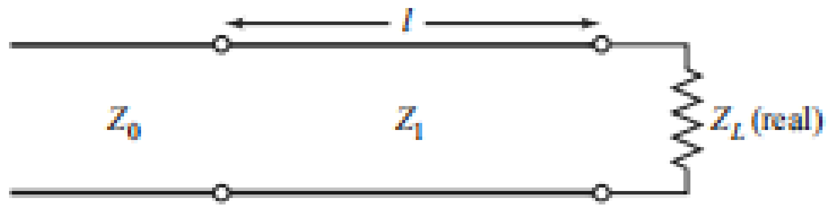


Fig 3.2 A single-section quarter-wave matching transformer.

$l = \lambda_0/4$ at the design frequency f_0

The single-section quarter-wave matching transformer circuit is shown in Figure 3.2, with the characteristic impedance of the matching section given as,

$$Z_1 = \sqrt{Z_0 Z_L} \quad \text{…….(3.1)}$$

At the design frequency, f_0, the electrical length of the matching section need to be $\lambda_0/4$, but for other frequencies the length is not the same, so we cannot get a perfect match at those frequencies.

We will derive an approximate expression for the resulting impedance mismatch versus frequency.

The input impedance seen looking into the matching section is

$$Z_{\text{in}} = Z_1 \frac{Z_L + jZ_1 t}{Z_1 + jZ_L t} \quad \ldots\ldots(3.2)$$

where $t = \tan \beta l = \tan \theta$, and $\beta l = \theta = \pi/2$ at the design frequency f_0.

The resulting reflection coefficient is

$$\Gamma = \frac{Z_{\text{in}} - Z_0}{Z_{\text{in}} + Z_0} = \frac{Z_1(Z_L - Z_0) + jt\left(Z_1^2 - Z_0 Z_L\right)}{Z_1(Z_L + Z_0) + jt\left(Z_1^2 + Z_0 Z_L\right)} \quad \ldots\ldots(3.3)$$

Because $Z_1^2 = Z_0 Z_L$, this reduces to

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0 + j2t\sqrt{Z_0 Z_L}} \quad \ldots\ldots(3.4)$$

The reflection coefficient magnitude is

$$|\Gamma| = \frac{|Z_L - Z_0|}{\left[(Z_L + Z_0)^2 + 4t^2 Z_0 Z_L\right]^{1/2}}$$

$$= \frac{1}{\left\{(Z_L + Z_0)^2/(Z_L - Z_0)^2 + [4t^2 Z_0 Z_L/(Z_L - Z_0)^2]\right\}^{1/2}}$$

$$= \frac{1}{\left\{1 + [4Z_0 Z_L/(Z_L - Z_0)^2] + [4Z_0 Z_L t^2/(Z_L - Z_0)^2]\right\}^{1/2}}$$

$$= \frac{1}{\left\{1 + [4Z_0 Z_L/(Z_L - Z_0)^2]\ \sec^2\theta\right\}^{1/2}} \quad \text{.....(3.5)}$$

since $1 + t^2 = 1 + \tan^2\theta = \sec^2\theta$

If we assume that the operating frequency is near the design frequency f_0, then $l \cong \lambda 0/4$ and $\theta \cong \pi/2$. Then $\sec^2\theta >> 1$, and (3.5) simplifies to

$$|\Gamma| \simeq \frac{|Z_L - Z_0|}{2\sqrt{Z_0 Z_L}} |\cos\theta| \quad \text{......(3.6)}$$

for θ near $\pi/2$.

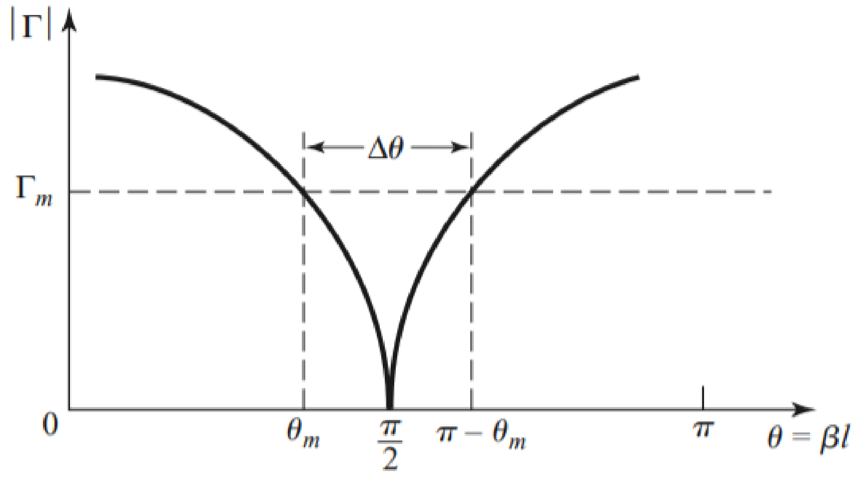


Fig 3.3 Approximate behaviour of the reflection coefficient magnitude for a single-section quarter-wave transformer operating near its design frequency.

This result gives the approximate mismatch of the quarter-wave transformer near the design frequency, as sketched in Figure 3.3.

If we set a maximum value, Γ_m, for an acceptable reflection coefficient magnitude, then the bandwidth of the matching transformer can be defined as,

$$\Delta\theta = 2\left(\frac{\pi}{2} - \theta_m\right) \qquad \text{.......(3.7)}$$

since the response of (3.5) is symmetric about $\theta = \pi/2$, and $\Gamma = \Gamma_m$ at $\theta = \theta_m$ and at $\theta = \pi - \theta_m$. Equating Γ_m to the exact expression for the reflection coefficient magnitude in (3.5) allows us to solve for θ_m:

$$\frac{1}{\Gamma_m^2} = 1 + \left(\frac{2\sqrt{Z_0 Z_L}}{Z_L - Z_0}\sec\theta_m\right)^2 \qquad \text{......(3.8)}$$

or

$$\cos\theta_m = \frac{\Gamma_m}{\sqrt{1-\Gamma_m^2}} \frac{2\sqrt{Z_0 Z_L}}{|Z_L - Z_0|} \quad \text{...... (3.9)}$$

If we assume TEM lines, then

$$\theta = \beta\ell = \frac{2\pi f}{v_p}\frac{v_p}{4f_0} = \frac{\pi f}{2f_0} \quad \text{........(3.10)}$$

and so the frequency of the lower band edge at $\theta = \theta_m$ is

$$f_m = \frac{2\theta_m f_0}{\pi} \quad \text{........(3.11)}$$

and the fractional bandwidth is, using (3.10),

$$\begin{aligned}\frac{\Delta f}{f_0} &= \frac{2(f_0 - f_m)}{f_0} = 2 - \frac{2f_m}{f_0} = 2 - \frac{4\theta_m}{\pi} \\ &= 2 - \frac{4}{\pi}\cos^{-1}\left[\frac{\Gamma_m}{\sqrt{1-\Gamma_m^2}}\frac{2\sqrt{Z_0 Z_L}}{|Z_L - Z_0|}\right]\end{aligned} \quad \text{.......(3.12)}$$

Fractional bandwidth is usually expressed as a percentage, $100\Delta f/f_0\%$. Note that the bandwidth of the transformer increases as Z_L becomes closer to Z_0 (a less mismatched load).

The above results are strictly valid only for TEM lines. When non-TEM lines (such as waveguides) are used, the propagation constant is no longer a linear function of frequency, and the wave impedance will be frequency dependent.

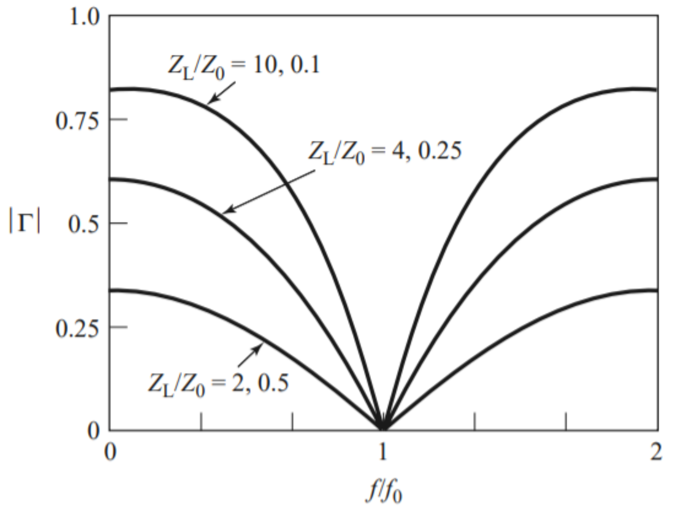


Fig 3.4 Reflection coefficient magnitude versus frequency for a single-section quarter wave matching transformer with various load mismatches.

These factors serve to complicate the general behaviour of quarter-wave transformers for non-TEM lines, but in practice the bandwidth of the transformer is often small enough that these complications do not substantially affect the result. Another factor ignored in the above analysis is the effect of reactance associated with discontinuities when there is a step change in the dimensions of a transmission line. This can often be compensated by making a small adjustment in the length of the matching section.

Figure 3.4 shows a plot of the reflection coefficient magnitude versus normalized frequency for various mismatched loads. It is to be noted the trend of increased bandwidth for smaller load mismatches.

3.3 THE THEORY OF SMALL REFLECTIONS

The quarter-wave transformer provides a simple means of matching any real load impedance to any transmission line impedance. For applications requiring more bandwidth than a single quarter-wave section can provide, multisection transformers can be used. The design of such transformers is the subject of the next two sections, but prior to that material we need to derive some approximate results for the total reflection coefficient caused by the partial reflections from several small discontinuities. This topic is generally referred to as the theory of small reflections.

3.3.1 Single-Section Transformer

We will derive an approximate expression for the overall reflection coefficient, Γ, for the single-section matching transformer shown in Figure 3.5. The partial reflection and transmission coefficients are,

$$\Gamma_1 = \frac{Z_2 - Z_1}{Z_2 + Z_1} \qquad \text{........(3.13)}$$

$$\Gamma_2 = -\Gamma_1 \qquad \text{........(3.14)}$$

$$\Gamma_3 = \frac{Z_L - Z_2}{Z_L + Z_2} \qquad \text{........(3.15)}$$

$$T_{21} = 1 + \Gamma_1 = \frac{2Z_2}{Z_1 + Z_2} \quad \text{……..(3.16)}$$

$$T_{12} = 1 + \Gamma_2 = \frac{2Z_1}{Z_1 + Z_2} \quad \text{……..(3.17)}$$

We can compute the total reflection, Γ, seen by the feed line using either the impedance method, or the multiple reflection method,

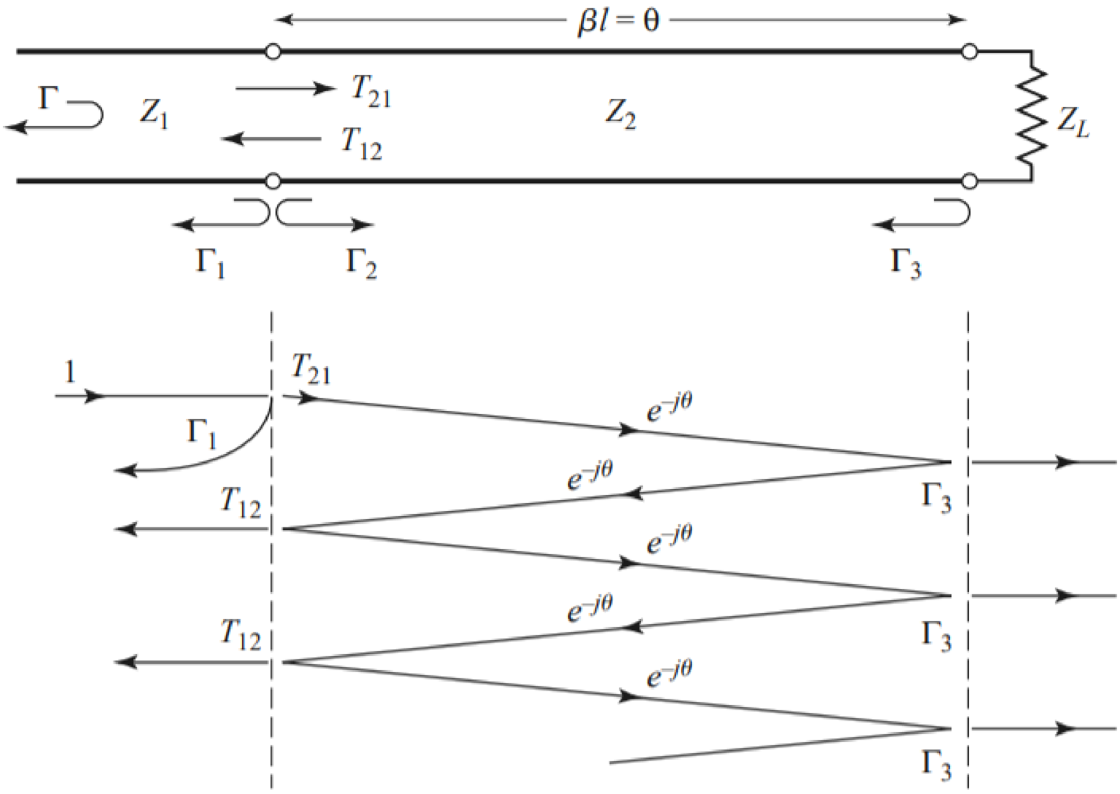


Fig 3.5 Partial reflections and transmissions on a single-section matching transformer.

For our present purpose the latter technique is preferred, so we express the total reflection as an infinite sum of partial reflections and transmissions as follows:

$$\Gamma = \Gamma_1 + T_{12}T_{21}\Gamma_3 e^{-2j\theta} + T_{12}T_{21}\Gamma_3^2\Gamma_2 e^{-4j\theta} + \cdots$$

$$= \Gamma_1 + T_{12}T_{21}\Gamma_3 e^{-2j\theta} \sum_{n=0}^{\infty} \Gamma_2^n \Gamma_3^n e^{-2jn\theta} \qquad \text{........(3.18)}$$

The summation of the geometric series,

$$\sum_{n=0}^{\infty} x^n = \frac{1}{1-x} \qquad \text{for } |x| < 1$$

allows us to express (3.18) in closed form as,

$$\Gamma = \Gamma_1 + \frac{T_{12}T_{21}\Gamma_3 e^{-2j\theta}}{1 - \Gamma_2\Gamma_3 e^{-2j\theta}} \qquad \text{......... (3.19)}$$

From (3.17), (3.18), and (3.19), we use $\Gamma_2 = -\Gamma_1$, $T_{21} = 1 + \Gamma_1$, and $T_{12} = 1 - \Gamma_1$ in (3.19) to give,

$$\Gamma = \frac{\Gamma_1 + \Gamma_3 e^{-2j\theta}}{1 + \Gamma_1\Gamma_3 e^{-2j\theta}} \qquad \text{.......... (3.20)}$$

If the discontinuities between the impedances Z1, Z2 and Z2, ZL are small, then $|\Gamma_1 \Gamma_3| = 1$, so we can approximate (3.20) as ,

$$\Gamma \simeq \Gamma_1 + \Gamma_3 e^{-2j\theta}$$
………. (3.21)

This result expresses the intuitive idea that the total reflection is dominated by the reflection from the initial discontinuity between Z_1 and Z_2 (Γ_1), and the first reflection from the discontinuity between Z_2 and Z_L ($\Gamma_3 e^{-2j\theta}$). The $e^{-2j\theta}$ term accounts for the phase delay when the incident wave travels up and down the line.

3.3.2 Multisection Transformer

Considering the multisection transformer shown in Figure 3.6, which consists of N equal-length (commensurate) sections of transmission lines. We will derive an approximate expression for the total reflection coefficient Γ .

Partial reflection coefficients can be defined at each junction, as follows:

$$\Gamma_0 = \frac{Z_1 - Z_0}{Z_1 + Z_0}$$
………… (3.22)

$$\Gamma_n = \frac{Z_{n+1} - Z_n}{Z_{n+1} + Z_n}$$
………… (3.23)

$$\Gamma_N = \frac{Z_L - Z_N}{Z_L + Z_N}$$

…………. (3.24)

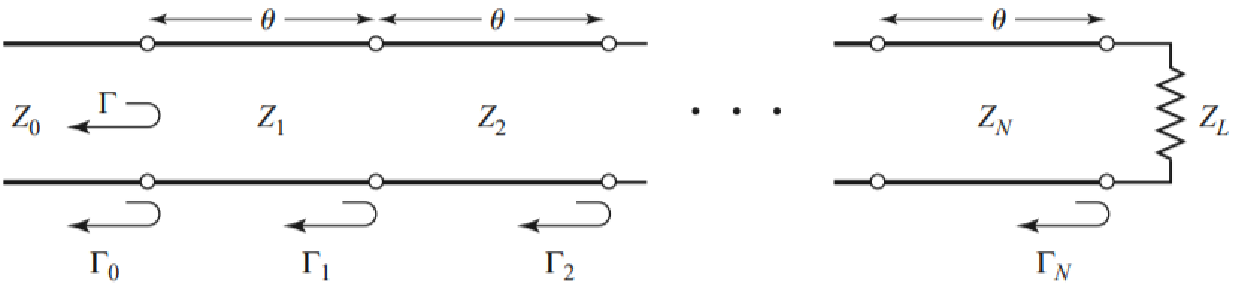


Fig 3.6 Partial reflection coefficients for a multisection matching transformer.

We also assume that all Zn increase or decrease monotonically across the transformer and that Z_L is real. This implies that all Γ_n will be real and of the same sign ($\Gamma_n > 0$ if $Z_L > Z_0$; $\Gamma_n < 0$ if $Z_L < Z_0$). Using the results of the previous section allows us to approximate the overall reflection coefficient as

$$\Gamma(\theta) = \Gamma_0 + \Gamma_1 e^{-2j\theta} + \Gamma_2 e^{-4j\theta} + \cdots + \Gamma_N e^{-2jN\theta}$$

…………… (3.25)

Further assume that the transformer can be made symmetrical, so that $\Gamma_0 = \Gamma_N$, $\Gamma_1 = \Gamma_{N-1}$, $\Gamma_2 = \Gamma_{N-2}$, and so on. (Note that this does not imply that the Zn are symmetrical.) Then (3.25) can be written as,

$$\Gamma(\theta) = e^{-jN\theta}\left\{\Gamma_0[e^{jN\theta} + e^{-jN\theta}] + \Gamma_1[e^{j(N-2)\theta} + e^{-j(N-2)\theta}] + \cdots\right\}$$

…………… (3.26)

If N is odd, the last term is $\Gamma_{(N-1)/2}(e^{j\theta} + e^{-j\theta})$, while if N is even, the last term is $\Gamma_{N/2}$. Equation (3.26) is seen to be of the form of a finite Fourier cosine series in θ, which can be written as

$$\Gamma(\theta) = 2e^{-jN\theta}\left[\Gamma_0 \cos N\theta + \Gamma_1 \cos(N-2)\theta + \cdots + \Gamma_n \cos(N-2n)\theta + \cdots + \frac{1}{2}\Gamma_{N/2}\right] \quad \text{for N even} \qquad \ldots\ldots\ldots (3.27)$$

$$\Gamma(\theta) = 2e^{-jN\theta}[\Gamma_0 \cos N\theta + \Gamma_1 \cos(N-2)\theta + \cdots + \Gamma_n \cos(N-2n)\theta + \cdots + \Gamma_{(N-1)/2} \cos\theta] \quad \text{for N odd} \qquad \ldots\ldots\ldots\ldots (3.28)$$

The importance of these results lies in the fact that we can synthesize any desired reflection coefficient response as a function of frequency (θ) by properly choosing the Γn and using enough sections (N).

3.3.3 BINOMIAL MULTISECTION MATCHING TRANSFORMER

The passband response (the frequency band where a good impedance match is achieved) of a binomial matching transformer is optimum in the sense that, for a given number of sections, the response is as flat as possible near the design frequency. This type of response, which is also known as maximally flat, is determined for an N-section transformer by setting the first N − 1 derivatives of $|\Gamma(\theta)|$ to zero at the centre frequency, f_0. Such a response can be obtained with a reflection coefficient of the following form:

$$\Gamma(\theta) = A(1 + e^{-2j\theta})^N \quad \text{.......... (3.29)}$$

Then the reflection coefficient magnitude is

$$\begin{aligned}|\Gamma(\theta)| &= |A||e^{-j\theta}|^N|e^{j\theta} + e^{-j\theta}|^N \\ &= 2^N|A||\cos\theta|^N\end{aligned} \quad \text{......... (3.30)}$$

Note that $|\Gamma(\theta)| = 0$ for $\theta = \pi/2$, and that $d^n|\Gamma(\theta)|/d\theta^n = 0$ at $\theta = \pi/2$ for $n = 1,2,...,N-1$. ($\theta = \pi/2$ corresponds to the center frequency, f_0, for which $l = \lambda/4$ and $\theta = \beta l = \pi/2$.)

We can determine the constant A by letting $f \rightarrow 0$. Then $\theta = \beta l = 0$, and (3.30) reduces to

$$\Gamma(0) = 2^N A = \frac{Z_L - Z_0}{Z_L + Z_0} \quad \text{.......... (3.31)}$$

since for f = 0 all sections are of zero electrical length. The constant A can then be written as ,

$$A = 2^{-N}\frac{Z_L - Z_0}{Z_L + Z_0} \quad \text{.......... (3.32)}$$

Next we expand $\Gamma(\theta)$ in (3.31) according to the binomial expansion,

$$\Gamma(\theta) = A(1 + e^{-2j\theta})^N = A\sum_{n=0}^{N} C_n^N e^{-2jn\theta} \quad \text{........ (3.33)}$$

Where

$$C_n^N = \frac{N!}{(N-n)!n!}$$

are the binomial coefficients. Note that $C_N^n = C_{N-n}^N$, $C_0^N = 1$, and $C_N^n = N = C_{N-1}^N$. The key step is now to equate the desired passband response, given by (3.33), to the actual response as given (approximately) by (3.26)

$$\Gamma(\theta) = A\sum_{n=0}^{N} C_n^N e^{-2jn\theta} = \Gamma_0 + \Gamma_1 e^{-2j\theta} + \Gamma_2 e^{-4j\theta} + \cdots + \Gamma_N e^{-2jN\theta}$$

…….. (3.34)

This shows that the Γ_n must be chosen as,

$$\Gamma_n = AC_n^N$$

………. (3.35)

where A is given by (3.32) and C_n^N is a binomial coefficient.

At this point, the characteristic impedances, Zn, can be found via (3.22), (3.24) , (3.25), but a simpler solution can be obtained using the following approximation . Because we assumed that the Γ_n are small, we can write,

$$\Gamma_n = \frac{Z_{n+1} - Z_n}{Z_{n+1} + Z_n} \simeq \frac{1}{2} \ln \frac{Z_{n+1}}{Z_n}$$

……… (3.36)

Since ln x ≈ 2(x − 1)/ (x + 1) for x close to unity, Then, using (3.36) and (3.34) gives,

$$\ln \frac{Z_{n+1}}{Z_n} \simeq 2\Gamma_n = 2AC_n^N = 2(2^{-N})\frac{Z_L - Z_0}{Z_L + Z_0}C_n^N \simeq 2^{-N}C_n^N \ln \frac{Z_L}{Z_0}$$

…………. (3.37)

which can be used to find Z_{n+1}, starting with n = 0. This technique has the advantage of ensuring self-consistency, in that Z_{N+1} computed from (3.37) will be equal to Z_L, as it should. Exact design results, including the effect of multiple reflections in each section, can be found by using the transmission line equations for each section and numerically solving for the characteristic impedances.

The bandwidth of the binomial transformer can be evaluated as follows. As in Section 3.3.3, let Γ_m be the maximum value of reflection coefficient that can be tolerated over the passband. Then from (3.30)

$$\Gamma_m = 2^N |A| \cos^N \theta_m$$

……… (3.38)

where $\theta_m < \pi/2$, is the lower edge of the passband, as shown in Figure 3.3. Thus,

$$\theta_m = \cos^{-1}\left[\frac{1}{2}\left(\frac{\Gamma_m}{|A|}\right)^{1/N}\right] \qquad \text{……….. (3.39)}$$

and using (3.12) gives the fractional bandwidth as ,

$$\frac{\Delta f}{f_0} = \frac{2(f_0 - f_m)}{f_0} = 2 - \frac{4\theta_m}{\pi}$$

$$= 2 - \frac{4}{\pi}\cos^{-1}\left[\frac{1}{2}\left(\frac{\Gamma_m}{|A|}\right)^{1/N}\right] \qquad \text{………… (3.40)}$$

3.4. CHEBYSHEV MULTISECTION MATCHING TRANSFORMERS

In comparison with the binomial transformer, the multisection Chebyshev matching transformer optimizes the bandwidth of the antenna at the expense of passband ripple. Compromising on the flatness of the passband response leads to a bandwidth that is substantially better than that of the binomial transformer for a given number of sections. The Chebyshev transformer is designed by equating Γ(θ) to a Chebyshev polynomial, which defines the optimum characteristics of this type of transformer.

3.4.1 Chebyshev Polynomials

The nth-order Chebyshev polynomial is a polynomial of degree n, denoted by $T_n(x)$. The first four Chebyshev polynomials are:

$$T_1(x) = x$$

$$T_2(x) = 2x^2 - 1$$

$$T_3(x) = 4x^3 - 3x$$

$$T_4(x) = 8x^4 - 8x^2 + 1$$

Higher order polynomials can be found using the following recurrence formula:

$$T_n(x) = 2xT_{n-1}(x) - T_{n-2}(x)$$

If we plot the first polynomials in figure 3.7 , we can observe the following properties ,

For $-1 \leq x \leq 1$, $|Tn(x)| \leq 1$. In this range the Chebyshev polynomials oscillate between ±1. This is the equal-ripple property, and this region will be mapped to the passband of the matching transformer

For $|x| > 1$, $|Tn(x)| > 1$. This region will map to the frequency range outside the passband.

For $|x| > 1$, the $|Tn(x)|$ increases faster with x as n increases.

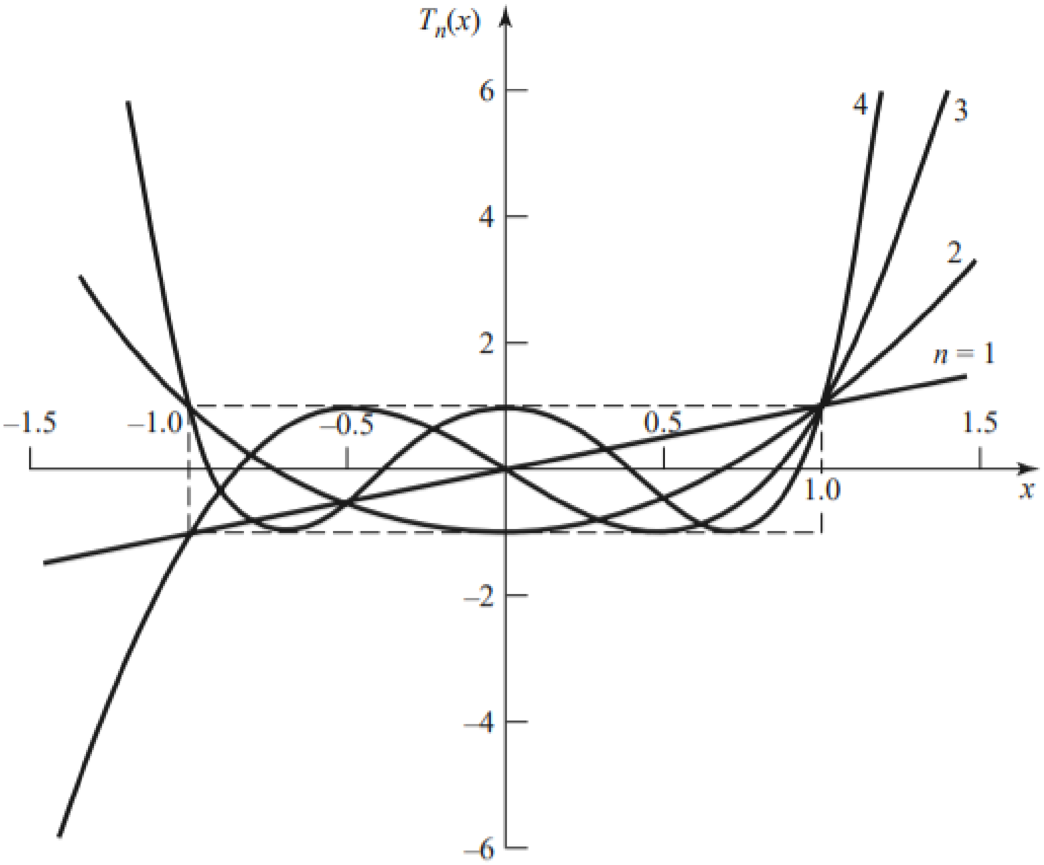


Fig 3.7 The first four Chebyshev polynomials, $T_n(x)$.

Now let $x = \cos\theta$ for $|x| < 1$. Then it can be shown that the Chebyshev polynomials can be expressed as,

$$T_n(\cos\theta) = \cos n\theta$$

or more generally as

$$T_n(x) = \cos(n \cos^{-1} x) \quad \text{for } |x| < 1, \quad \ldots\ldots (3.41)$$

$$T_n(x) = \cosh(n \cosh^{-1} x) \quad \text{for } x > 1. \quad \ldots\ldots (3.42)$$

We desire equal ripple for the passband response of the transformer, so it is necessary to map θ_m to $x = 1$ and $\pi - \theta_m$ to $x = -1$, where θ_m and

$\pi - \theta_m$ are the lower and upper edges of the passband, respectively, as shown in Figure 3.7. This can be accomplished by replacing cos θ in (3.41) with cos θ/cos θ_m :

$$T_n\left(\frac{\cos\theta}{\cos\theta_m}\right) = T_n(\sec\theta_m\cos\theta) = \cos n\left[\cos^{-1}\left(\frac{\cos\theta}{\cos\theta_m}\right)\right] \quad \text{.......(3.42)}$$

Then |sec θm cos θ| ≤ 1 for θm <θ < π – θm . so $|T_n(\sec\theta_m\cos\theta)| \leq 1$ over this same range.

Because cosnθ can be expanded into a sum of terms of the form cos(n − 2m)θ, the Chebyshev polynomials can be rewritten in the following useful form:

$$T_1(\sec\theta_m\cos\theta) = \sec\theta_m\cos\theta,$$

$$T_2(\sec\theta_m\cos\theta) = \sec^2\theta_m(1 + \cos 2\theta) - 1$$

$$T_3(\sec\theta_m\cos\theta) = \sec^3\theta_m(\cos 3\theta + 3\cos\theta) - 3\sec\theta_m\cos\theta$$

$$T_4(\sec\theta m\cos\theta) = \sec^4\theta_m(\cos 4\theta + 4\cos 2\theta + 3) - 4\sec^2\theta_m(\cos 2\theta + 1) + 1.$$

3.4.2 Design of Chebyshev Transformers

We can now synthesize a Chebyshev equal-ripple passband by making Γ(θ) proportional to $T_N(\sec\theta_m\cos\theta)$, where N is the number of sections in the transformer. Thus, using (3.28), we have

$$\Gamma(\theta) = 2e^{-jN\theta}[\Gamma_0\cos N\theta + \Gamma_1\cos(N-2)\theta + \cdots + \Gamma_n\cos(N-2n)\theta + \cdots]$$

$$= Ae^{-jN\theta}T_N(\sec\theta_m\cos\theta) \quad \text{........ (3.43)}$$

where the last term in the series of (5.61) is (1/2) $\Gamma_{N/2}$ for N even and $\Gamma_{(N-1)/2}$ cos θ for N odd. As in the binomial transformer case, we can find the constant A by letting θ = 0, corresponding to zero frequency. Thus,

$$\Gamma(0) = \frac{Z_L - Z_0}{Z_L + Z_0} = A T_N(\sec\theta_m)$$

………. (3.44)

so we have

$$A = \frac{Z_L - Z_0}{Z_L + Z_0} \frac{1}{T_N(\sec\theta_m)}$$

…….. (3.45)

If the maximum allowable reflection coefficient magnitude in the passband is m, then from (3.45) Γ_m= |A| since the maximum value of T_n(sec θm cos θ) in the passband is unity. Then (3.45) gives

$$T_N(\sec\theta_m) = \frac{1}{\Gamma_m}\left|\frac{Z_L - Z_0}{Z_L + Z_0}\right|$$

……… (3.46)

$$\begin{aligned}\sec\theta_m &= \cosh\left[\frac{1}{N}\cosh^{-1}\left(\frac{1}{\Gamma_m}\left|\frac{Z_L - Z_0}{Z_L + Z_0}\right|\right)\right]\\ &\simeq \cosh\left[\frac{1}{N}\cosh^{-1}\left(\left|\frac{\ln Z_L/Z_0}{2\Gamma_m}\right|\right)\right].\end{aligned}$$

………. (3.47)

Once θ_m is known, the fractional bandwidth can be calculated as

$$\frac{\Delta f}{f_0} = 2 - \frac{4\theta_m}{\pi}.$$

………… (3.48)

Chapter 4

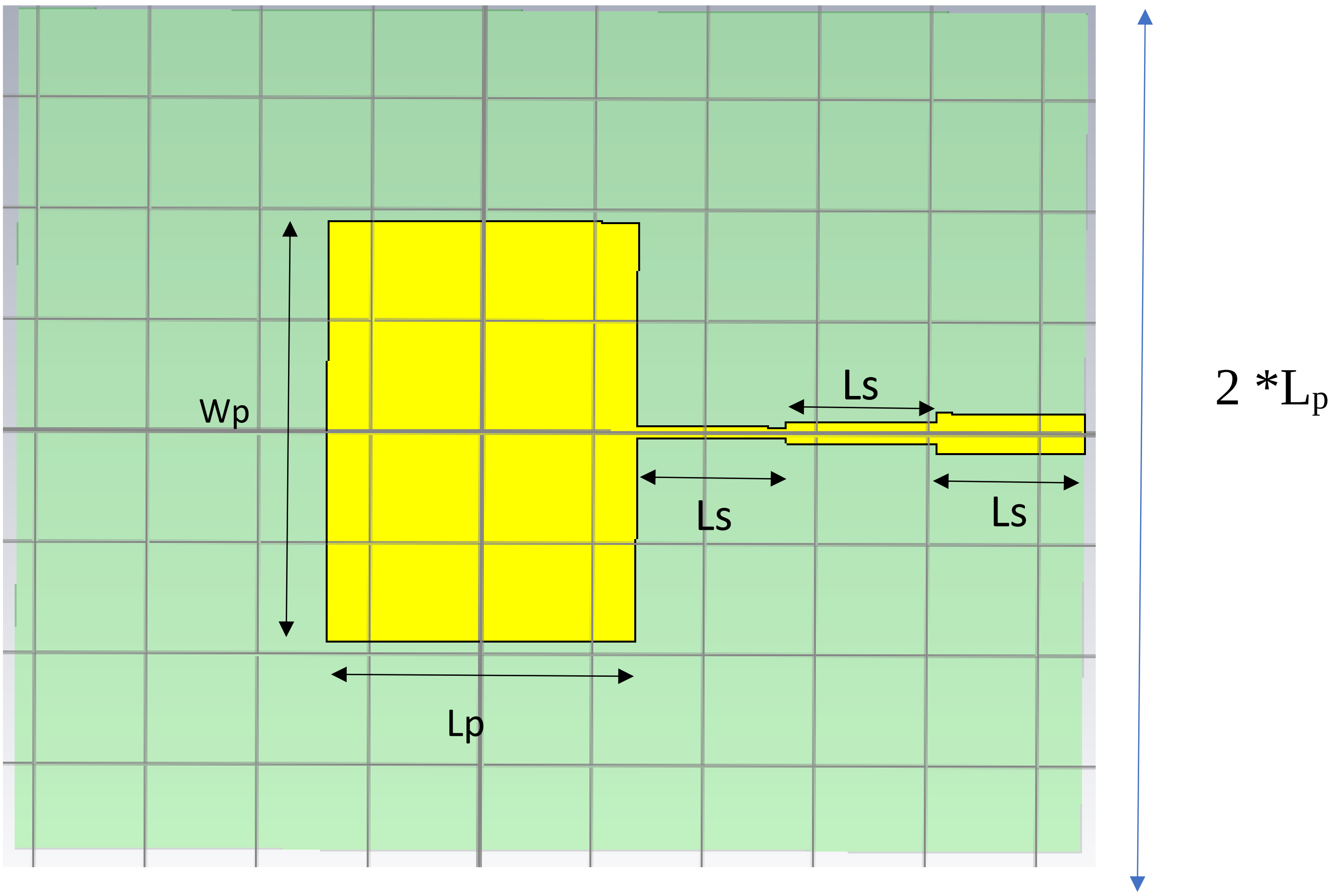
Preface

In this chapter we will design a quarter wave binomial section for the purpose of impedance matching of a microstrip patch antenna operating at 2.45 GHz fed into a 50 ohm source impedance line .

4.1 Designing a multisection Transformer

The use of simulation software is an essential part to achieve the goal. The simulation of the microstrip patch antenna is done through CST - Computer Simulation Technology. CST Studio Suite is a commercial finite element method (FEM) solver for electromagnetic structures and is one of the finest applications used for the designing of microstrip patch antenna, complex RF electronic circuits, metamaterial structures ,filters, transmission lines etc. It is an interactive simulation system, whose basic mesh element is a tetrahedron that allows solving any arbitrary 3D geometry, especially those with complex curves and shapes in a fraction of time. Different port schemes are available in this simulation software such as lumped port, wave port etc. The accurate simulation results of microstrip feed with coplanar waveguide are coming out through wave port. The optimization of the antenna parameters through CST is very useful and easy to optimize the parameters accurately . In the first step of simulation through CST , we define the geometry of the system by putting the material properties available in CST window. Also we assign suitable excitation port (wave port , lumped or floquet) is assigned to microstrip antenna. Now, the input frequency range and the number of frequency points of the output frequency response are put in the simulation software. Finally, if all the parameters are validated through validation check, then simulation results such as S parameters, VSWR, gain, group delay and far field radiation pattern can be displayed and can be used as a reference to the fabrication .

4.2 Antenna Section Geometry



(2 * Lp + 3*Ls)

Fig 4.1. Simulated Structure in CST

Symbols

Wp = Width of the patch

Lp = Length of the patch

Ls= Length of each section

W_1 = Width of the first section

W_2 = Width of the second section

W_3 = Width of the third section

4.2.1 Calculations of each section length

For the above patch, resonant frequency f = 2.45 GHz

C= $3*10^{10}$ cm/sec

Also, we know the wavelength of microstrip patch is given by the formula,

$$\lambda = \frac{c}{f\sqrt{\varepsilon_R}}$$

For FR4 substrate, ε_R = 4.2

So, we get

Wavelength $\lambda = 300 / (2.45 * \sqrt{4 \cdot 2})$

$= 59.45$ mm

So, length of each section (approx.) = (59.45 /4) = 14.86 mm

4.2.2 Dimensions

Parameters	Value(mm)
W_p	37.8
L_p	27.78
L_s	13.4
W_1	1
W_2	1.9
W_3	3.6

4.3 FR4 Substrate

"FR" stands for flame/fire retardant. FR-4 is a grade designation assigned to glass-reinforced epoxy laminate sheets, tubes, rods and printed circuit boards. It is a composite material composed of woven fibre glass cloth with an epoxy resin binder that is flame resistant.

FR-4 glass epoxy is famous and flexible high-pressure thermoset plastic laminate material with high strength to weight ratios.

If the material has to be flame or fire retardant, there are certain requirements to be fulfilled for the material to be certified, as FR. When an equipment using FR4 grade PCB and if there is some kind of overvoltage or short circuit in the equipment, then the PCB made of organic material can catch fire but it should have the ability to retard the fire by itself; that means, it should have a self-extinguishing property.

The FR4 substrate is manufactured by compressing an epoxy resin at high pressure and a glass fiber mat (or mats) is embedded within the structure. The glass fiber gives the strength to substrate and increases the dielectric constant of the composite material. The weave is usually more densely packed in the one direction, and so the material is inherently anisotropic, with a small variation in the dielectric constant in different planes. Furthermore, the manufacturing technique employed introduces inconsistency in board thickness, which can cause variation in microstrip circuit parameters. Typical FR4 board characterisation is generally carried out by the manufacturers at 1MHz. At microwave frequencies, the bulk dielectric constant value is typically similar to the value at 1 MHZ, decreasing slightly at the frequencies above a few GHz.

In most of the PCB applications, FR4 epoxy glass substrates are mostly preferred. The low cost and excellent mechanical properties makes it ideal for a wide range of electronic equipment applications. With the increasing development of microwave systems in consumer markets , there is a considerable interest in minimising the cost of these systems, which makes FR4 a good choice for microwave systems.

There are lots of commercial substrate materials are promptly available for the use at RF and microwave frequencies for the design of microstrip and printed antennas. The substrate can be preferred based on the desired material characteristics for optimal performance over the specific frequency range. Dielectric constant, thickness and loss tangent are the commonly used parameters. Normally the dielectric constant ranges from 2.2 to 12 for the operations at frequencies ranging from 1 to 100 GHz. The microstrip patch antenna design depends upon the substrate thickness. The thick substrates with low dielectric constants are the desired ones to obtain the larger bandwidth and higher efficiency due to loosely bound fringing fields. While thin substrates with large dielectric constants reduce the overall size of the antenna, however due to high loss tangents thin substrates are less efficient that results with narrow bandwidth. Therefore, substrate selection is an important matter which has to be done in the beginning to get the desirable features for a given application.

4.4 Fabrication

Standard photolithographic techniques are generally used for printed antennas fabrication on microwave substrate materials. The first essential part of fabrication is to select a proper substrate material that is most suitable for the antenna design operation. In designing the antenna we should consider the dielectric constant, loss tangent, homogeneity, and the dimensional strength of the substrate. A substrate having high loss tangent adversely affects the efficiency specially at high frequencies . We should select the dielectric material based on its dielectric constant of the substrate which depends on the application of the antenna and the radiation characteristics specifications. If we use high dielectric constant substrates, they can cause surface wave excitation and low bandwidth performance, thus degrading the performance of the antenna. In photolithographic process, a computer aided design (CAD) of the geometry is initially made and a negative mask of the geometry generated is printed on a butter paper. A single side or double-sided copper cladded substrate of suitable dimension is properly cleaned using acetone and dried in order to avoid the discontinuity caused by the impurities. Any disparity in the etched structure will shift the resonant frequency from the predicted values, especially when the operating frequency is very high. A thin layer of negative photo resist material is coated using dip coating technique on copper surfaces and it is dried by baking process. Then, the mask that we have printed earlier is placed onto the photo resist and the antenna is exposed to UV rays for about 19 minutes.

After the proper UV exposure, the layer of photoresist material that was exposed to UV rays hardens which is immersed in developer solution for few minutes in the next step. The portion that became hard will remain intact , the other parts will get washed away. To clearly view the hardened photo resist portions on the copper coating the board is dipped in the dye solution in order.

After developing phase, the unwanted copper portions are etched off using Ferric Chloride ($FeCl_3$) or Hydrochloric acid solution ($HCL + H_2O_2$) solution to get the required antenna geometry on the substrate. The etched board is rinsed in running water to remove any etchant .The copper parts under the hardened photoresist remains intact, whether the others gets washed away. Next the laminate is cleaned carefully using acetone solution, which removes the hardened photoresist left .

In the final step the input SMA connector port is connected by using soldering method.

Fig. 4.2 Fabricated antenna Design

Chapter 5

In this chapter we will include the simulated and practically determined results of the proposed antenna structure. We will also include S-parameters, VSWR, Input impedance calculated from Smith Chart plot and radiation pattern for the given structure.

5.1 Simulation and Measurements

Figure 5.1 Antenna Measurement setup

5.1.1 S11 parameter

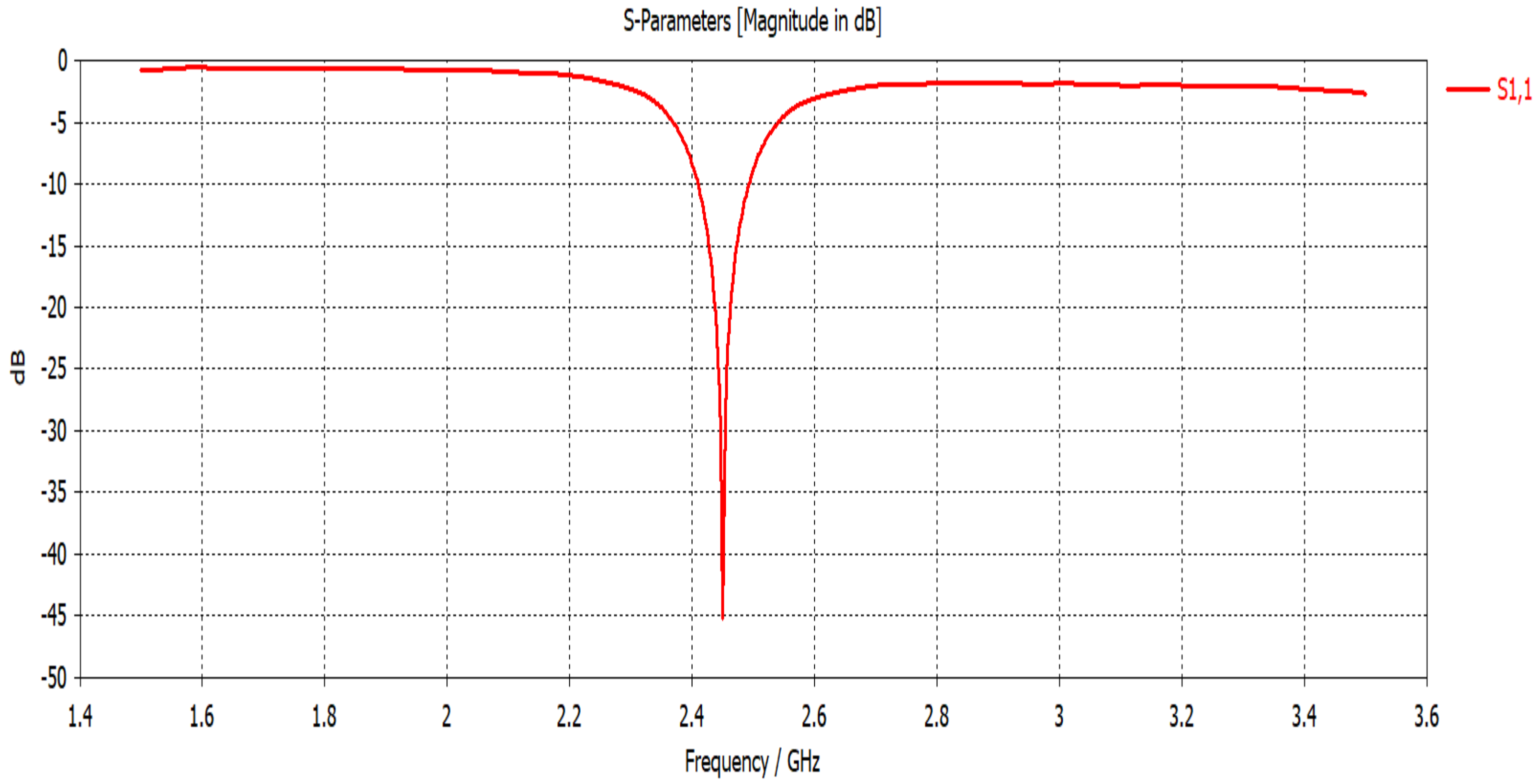


Figure 5.2a Simulated S_{11} (in dB) vs Frequency (GHz)

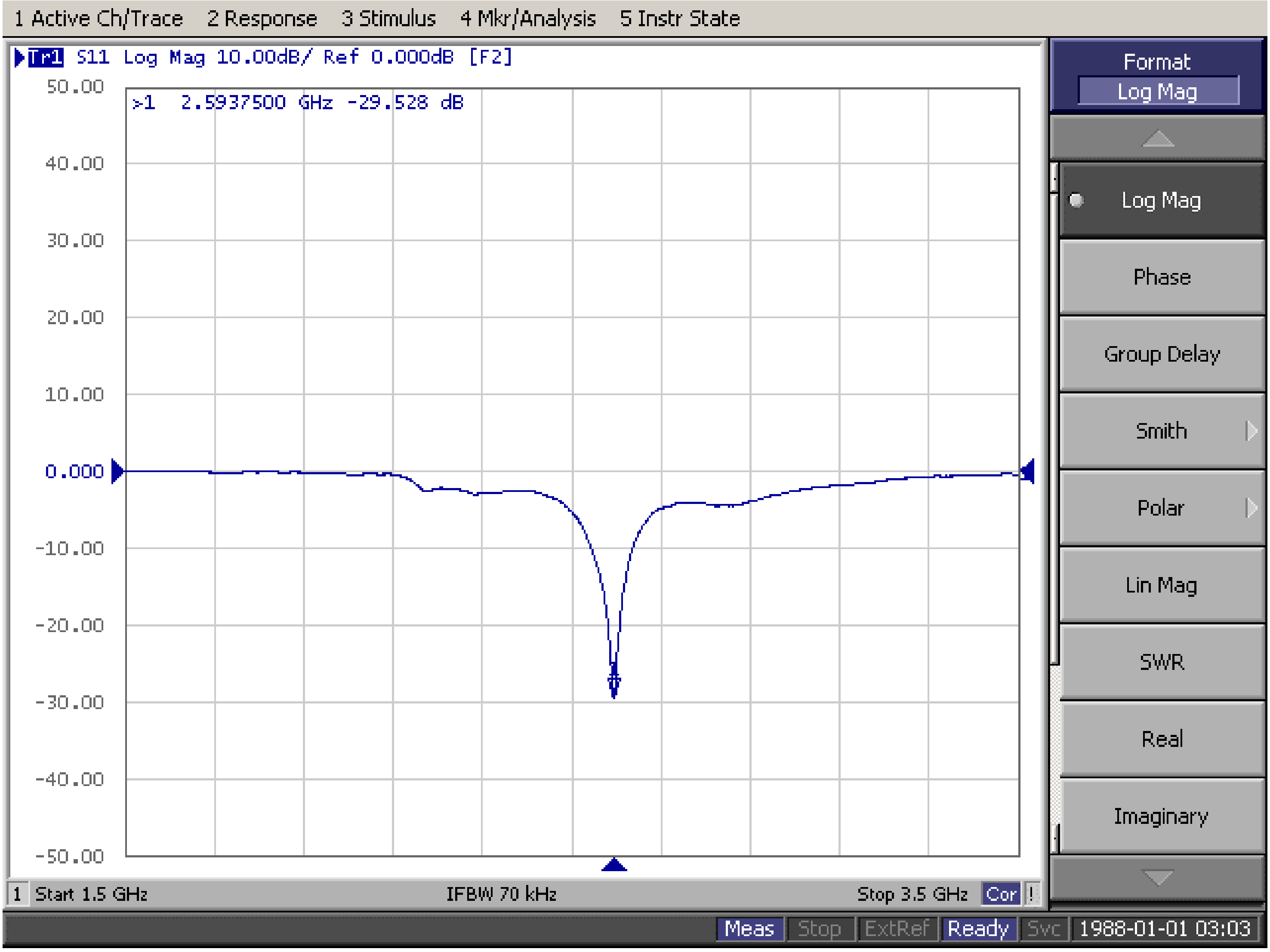
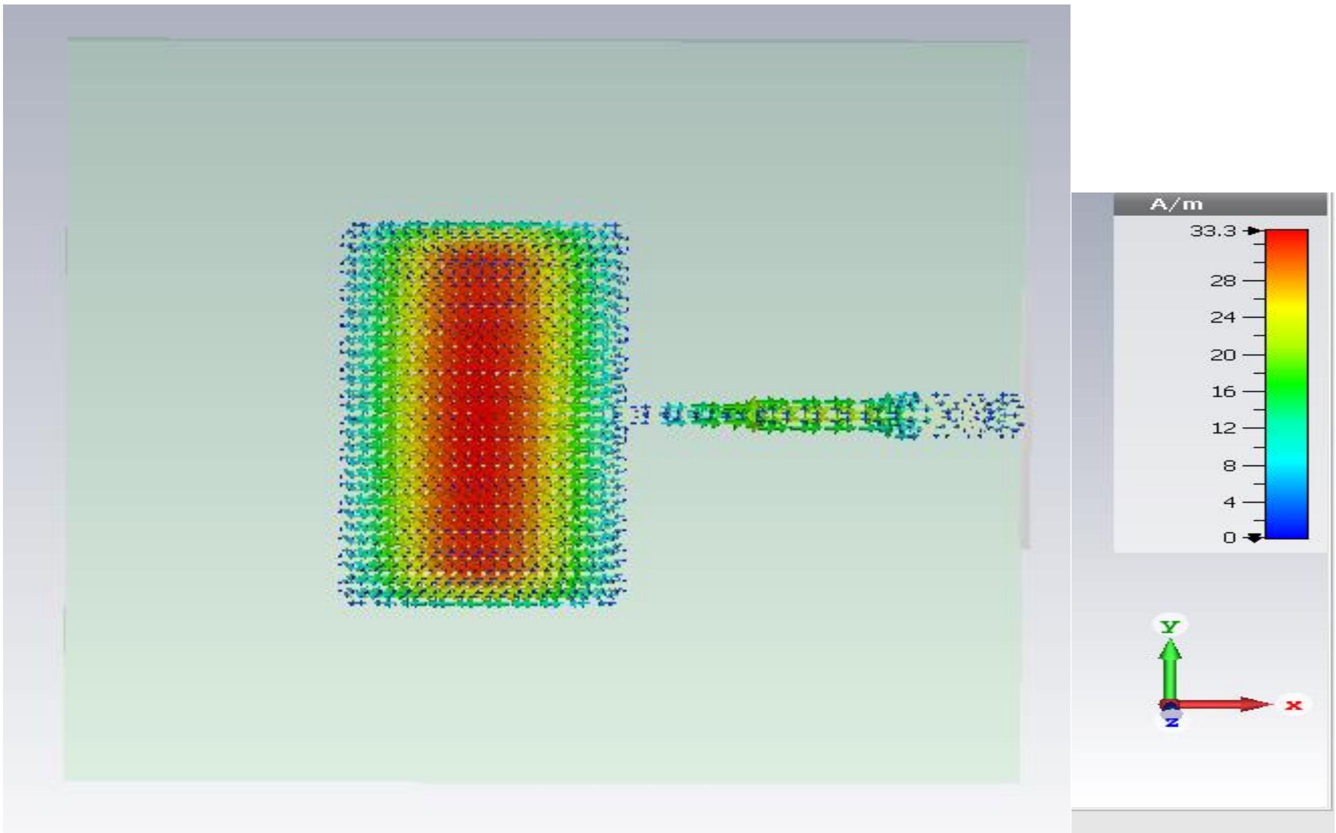


Figure 5.2b Calculated S_{11} plot from Vector Spectrum Analyzer in practical experiment.

Here, we can see The simulated S_{11} value at 2.45 GHz is - 45.1 dB. And in practical experiment the measured value of S11 is -29.5 dB.

5.1.2 Surface Currents



5.1.3 Z -Parameter

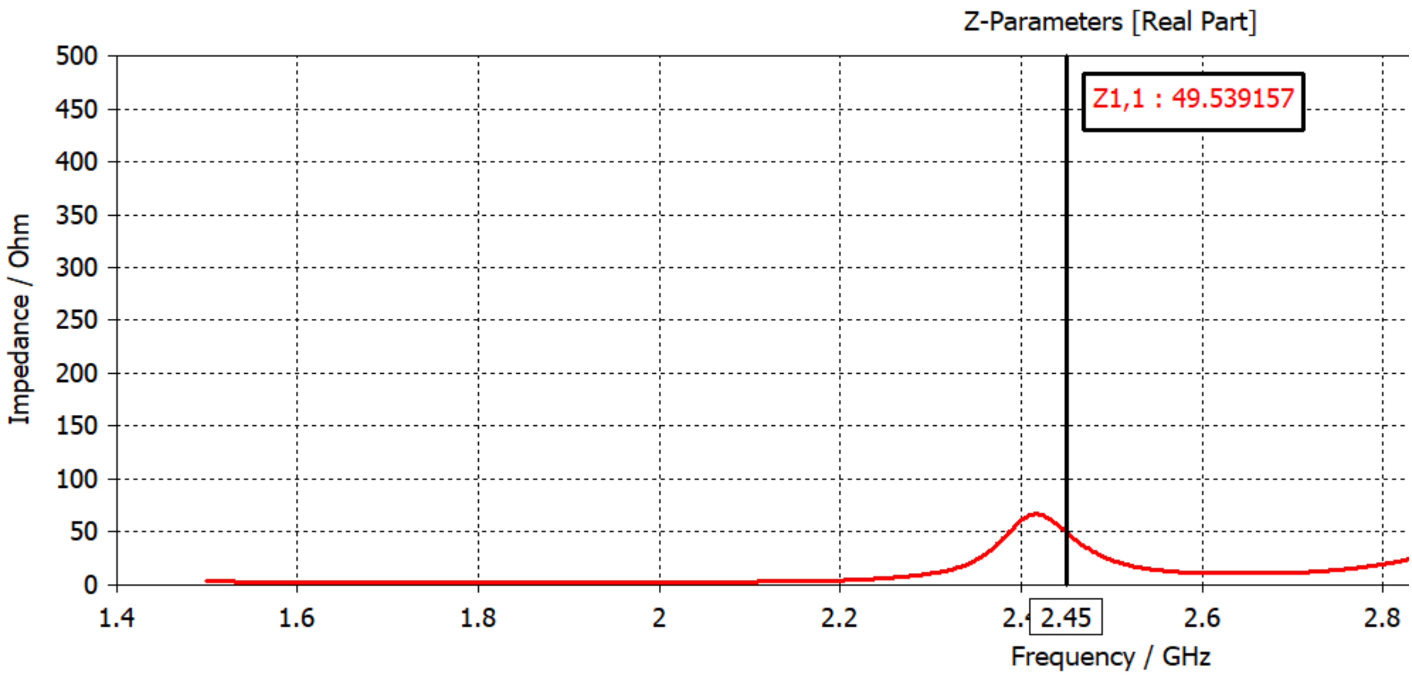


Figure 5.3 Simulated Z parameter (real part) vs frequency (GHz)

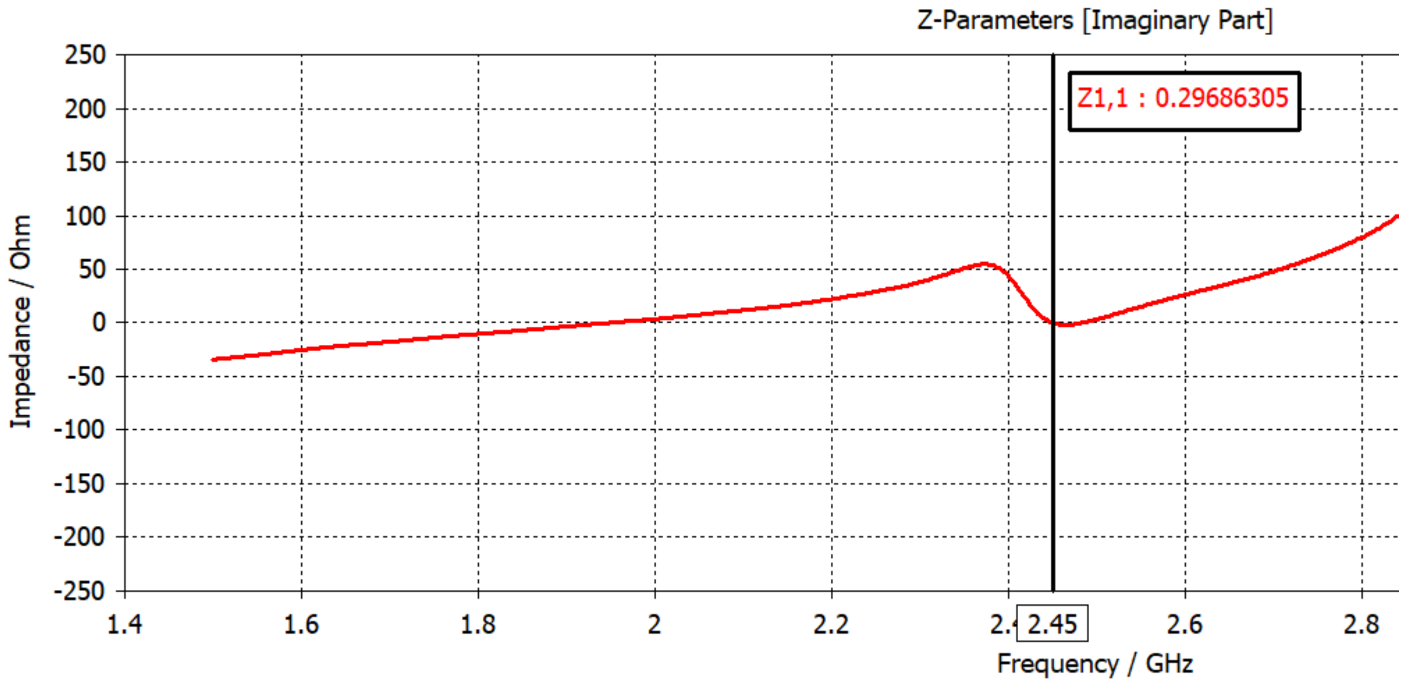
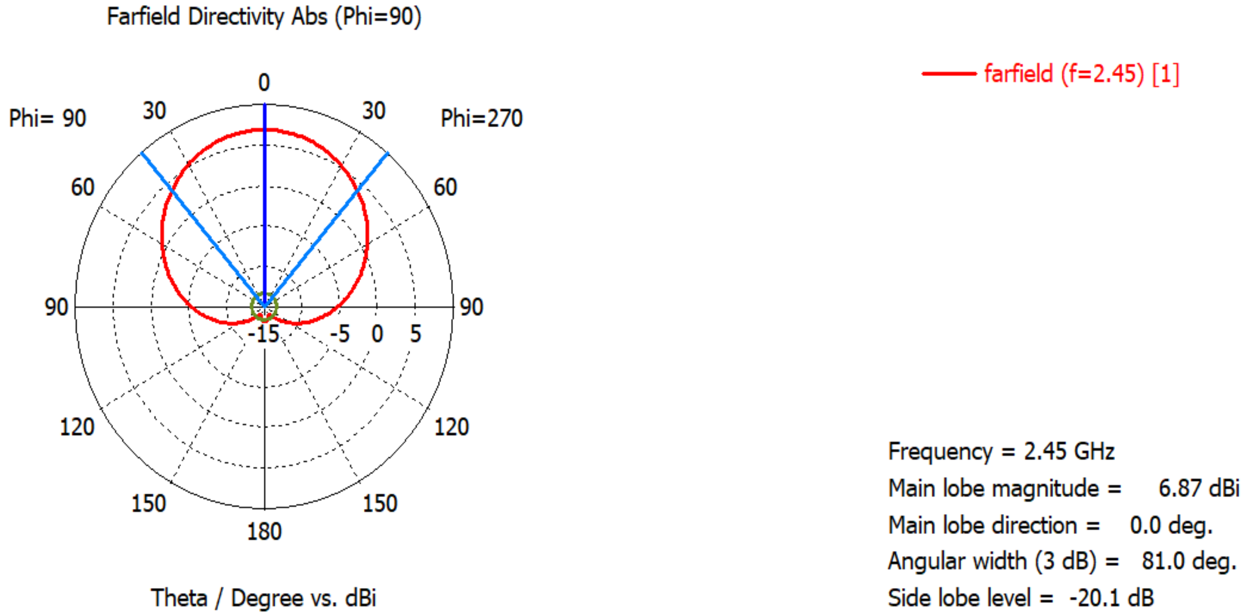


Figure 5.4 Imaginary part of Input impedance (z11) vs Frequency(GHz)

5.1.4 Far field pattern

Simulated



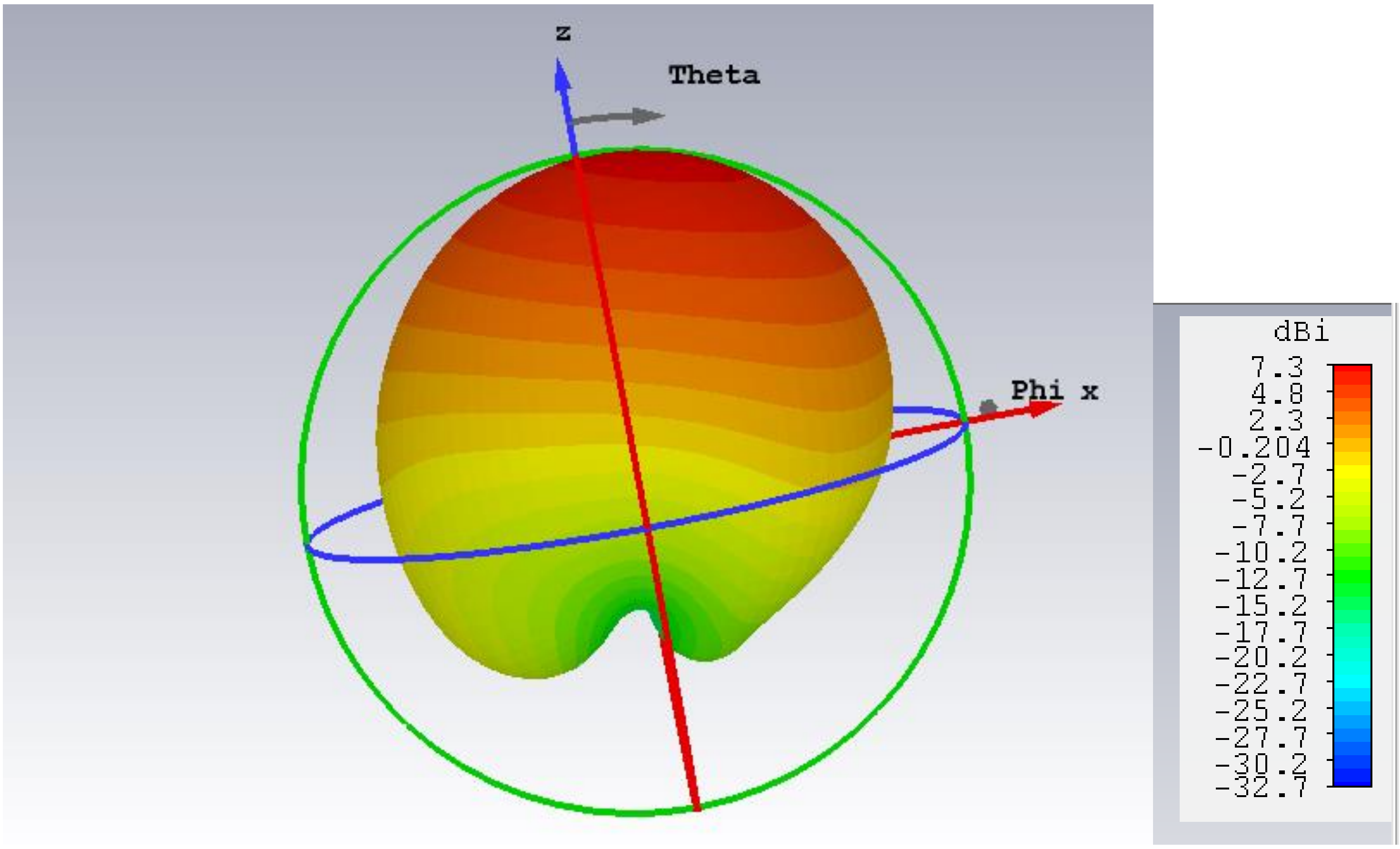


Figure 5.5 3D simulated radiation pattern of the antenna

Measured values

Deg	E(co plane) (in dB)	E(cross plane) (in dB)	H(co plane) (in dB)	H(cross plane) (in dB)
0	-31.8	-38.2	-31.8	-39.8
10	-32.1	-38.6	-32.3	-40.2
20	-32.8	-41.3	-33.2	-41.2
30	-34.2	-44.2	-34.2	-41.7
40	-36.9	-46.8	-35.7	-43.5
50	-38.1	-44.7	-37.4	-44.8

60	-37.2	-42.1	-39.2	-47.2
70	-36.8	-39.8	-40.8	-50.5
80	-38.5	-39	-41.2	-54.8
90	-40.9	-38.3	-41.6	-50.6
100	-40.5	-37.6	-41.7	-47.5
110	-38.2	-38.2	-40.8	-44.2
120	-42.1	-41.3	-43.3	-42.3
130	-40.4	-43.6	-44.2	-41.4
140	-41.7	-42.4	-45.9	-41.1
150	-43	-42.9	-47.5	-40.9
160	-49.1	-47.3	-55.2	-41.1
170	-57.1	-52.6	-52.8	-42.2
180	-53.2	-44.9	-49.2	-42.3
190	-44.7	-51.3	-50.3	-42.6
200	-43.5	-46.3	-48.7	-43.5
210	-41.2	-48.2	-50.3	-44.9
220	-41.1	-54.4	-58.7	-45.7
230	-44.3	-50.4	-61.2	-50.2
240	-42.6	-52.3	-53.2	-50.4
250	-44.8	-48.8	-53.2	-50.8
260	-42.9	-44.1	-44.7	-48.5
270	-40.8	-42.8	-43.4	-46.1
280	-39.2	-45.1	-41.7	-42.7
290	-37.2	-46.2	-39.8	-39.9
300	-35.8	-52.4	-36.6	-38.4
310	-33.9	-51.9	-34.6	-38.1
320	-33.1	-50.3	-33.2	-37.8
330	-31.7	-49.7	-31.8	-39.1
340	-30.9	-50.6	-31.5	-39.7
350	-31.2	-48.7	-32.2	-40.2
360	-32.2	-41.2	-31.6	-39.9

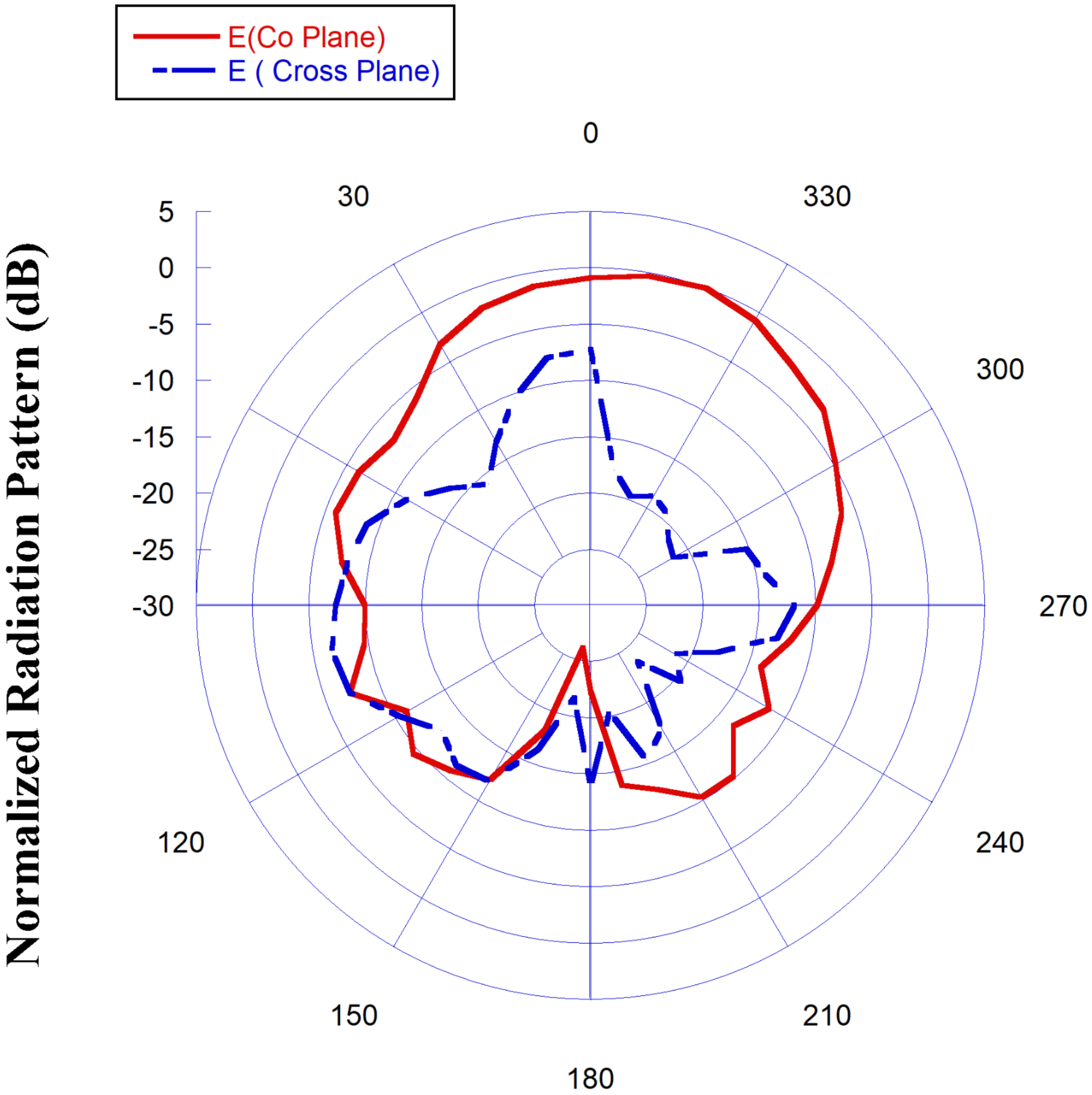


Figure 5.6 Measured normalized E-field Pattern

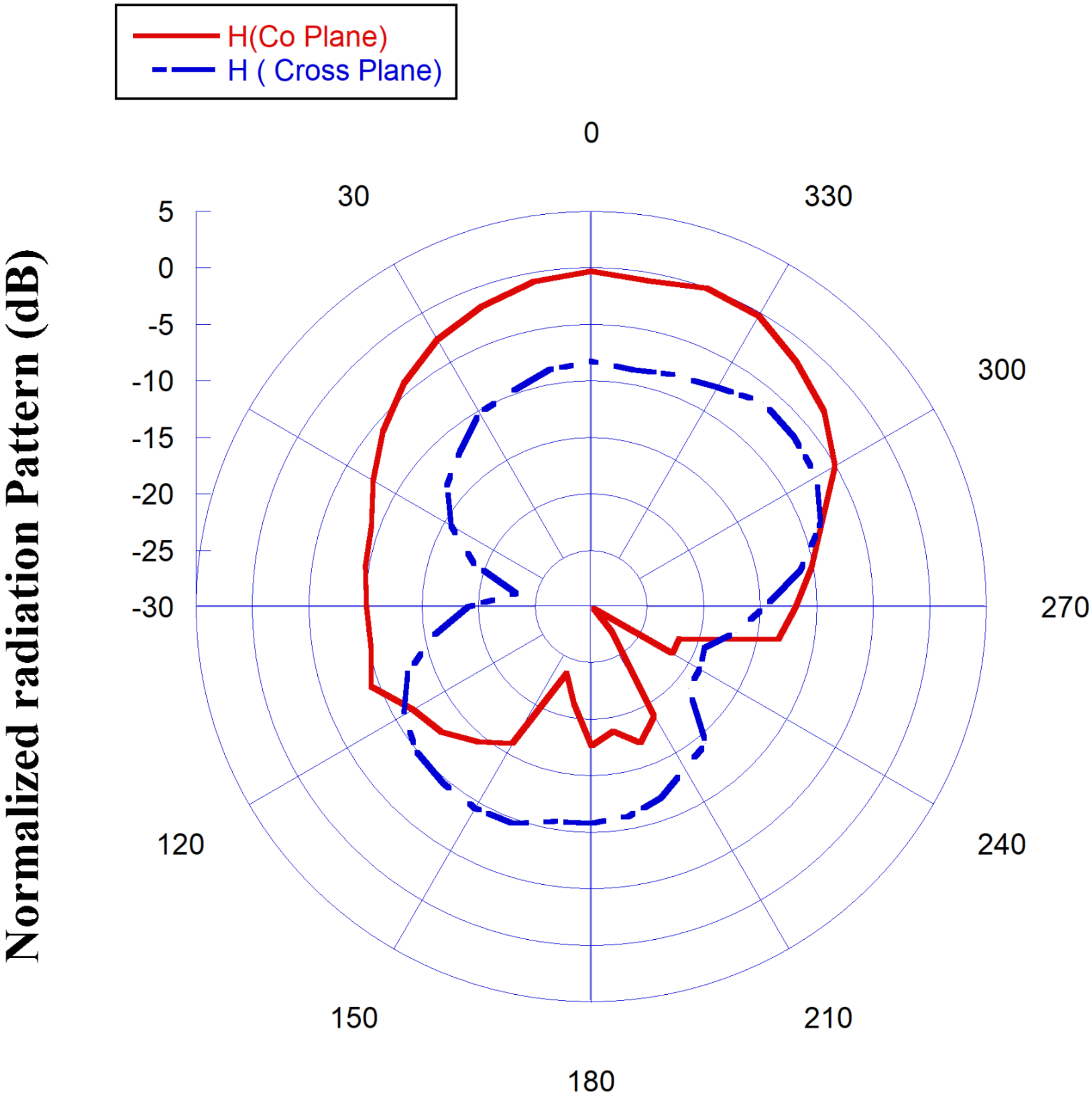


Figure 5.7 Measured normalized H-field Pattern

Here we can observe that, as we have considered the far-field values at an interval of 10 degree, we have seen the measured radiation pattern is somewhat coarse and is somewhat different from that we simulated .We can get more accurate near to simulation result by considering values at lesser intervals .

5.1.5 VSWR

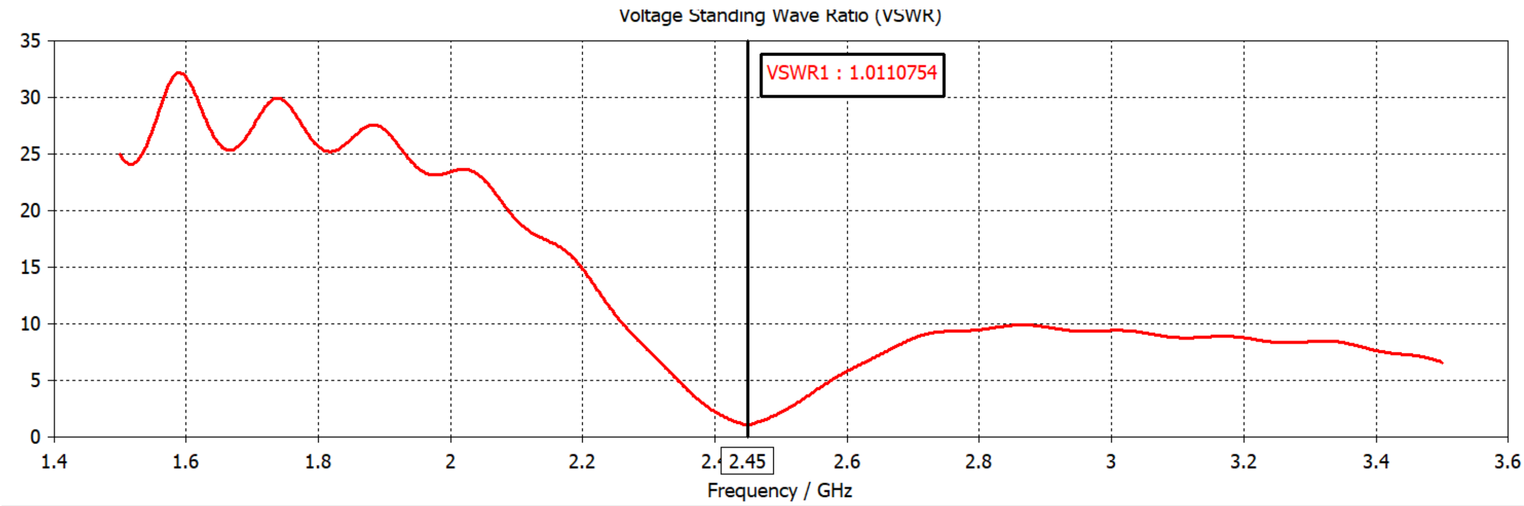
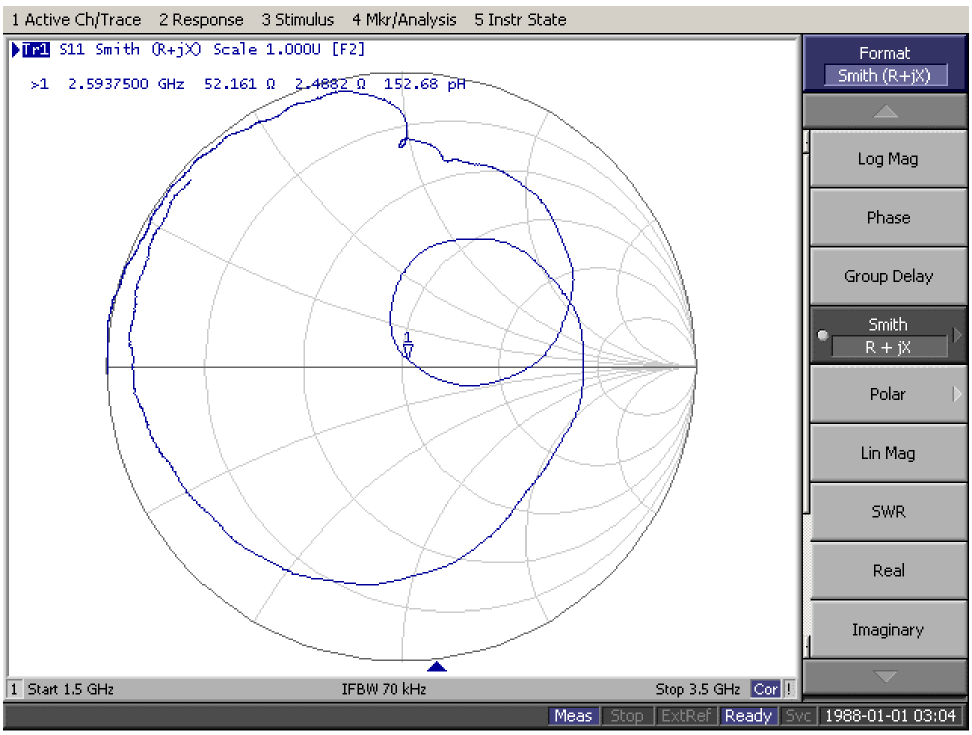


Fig 5.8 Simulated VSWR plot of the Proposed Structure

5.1.6 Smith Chart



Chapter 6

6.1 Conclusion

Here in this thesis, we have studied different types of impedance matching techniques of microstrip patch antenna. Impedance matching is a very important aspect that is to be maintained in order to get optimum radiation efficiency, thereby low reflection loss. Ideally the reactance part of the impedance should be zero, but in practical we have to make the reactance part as low as possible to get the optimum match. We have designed a quarter wave matching section having three sections for the same.

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