Design of Some Impedance Matching Networks

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VLSI DESIGN AND MICROELECTRONICS TECHNOLOGY

Thesis Submitted By

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CERTIFICATE OF RECOMMENDATION

This is to certify that the thesis entitled **"Design of Some Impedance Matching Networks"** submitted by **Sahin Islam** under my guidance and supervision and be accepted in partial fulfilment of the requirement for awarding the degree of **Master of Technology** in **VLSI design and Microelectronics Technology** of Jadavpur University. The research results presented in this thesis have not been included in other paper submitted for the award of any degree to any other Institute or University.

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CERTIFICATE OF APPROVAL*

The foregoing thesis is hereby approved as a credible study of an Engineering Subject carried out and presented in a manner of satisfactory to warrant its acceptance as a pre-requisite to the degree for which it has been submitted. It is to be understood that by this approval, the undersigned do not necessarily endorse or approve any statement made, opinion expressed or conclusion drawn therein but approve the thesis only for the purpose for which it has been submitted.

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DECLARATION OF ORIGINALITY AND COMPLIANCE OF ACADEMIC ETHICS

I hereby declare that this thesis contains literature survey and original research work done by the undersigned candidate, as a part of his degree of **Master of Technology** in **VLSI design and Microelectronics Technology** under department **of Electronics and Telecommunication Engineering** of Jadavpur University. All the information in this document has been obtained and presented in accordance with academic rules and ethical conduct. I also declare that as required by these rules and conduct, I have fully cited and referenced all materials and results that are not original to this work.

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Date: 24|08|2022

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Chapter 1

Introduction

Impedance is an important part of a larger design process for a microwave component or system. The basic idea of impedance matching is illustrated in figure, which shows an impedance matching network placed between a load impedance and a transmission line. The matching network is ideally lossless, to avoid unnecessary loss of power, and is usually designed so that the impedance seen looking into the matching network is Z_0. Then 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.

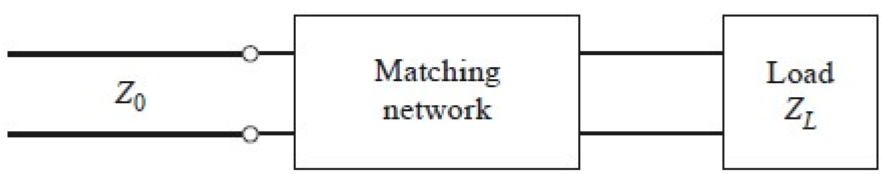

Fig. 1.1: Impedance matching network to match load with transmission line

Impedance matching 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 sensitive receiver components (antenna, low-noise amplifier, etc.) may improve the signal-to-noise ratio of the system. Impedance matching in a power distribution network (such as an antenna array feed network) may reduce amplitude and phase errors. As long as the load impedance, Z_L, has a positive real part, a matching network can always be found. Many choices are available, however, and we will discuss the design and performance of several types of practical matching networks.

Factors that may be important in the selection of a particular matching network include the following:

Bandwidth:

Any type of matching network can ideally give a perfect match (zero reflection) at a single frequency. In many applications, however, it is desirable to match a load over a band of frequencies. There are several ways of doing this, with, of course, ac corresponding increase in complexity.

Complexity:

As with most engineering solutions, the simplest design that satisfies the required specifications is generally preferable. A simpler matching network is usually cheaper, smaller, more reliable, and less lossy than a more complex design.

Implementation:

Depending on the type of transmission line or waveguide being used, one type of matching network may be preferable to another. For example, tuning stubs are much easier to implement in waveguide than are multi section quarter-wave transformers.

Adjustability:

In some applications the matching network may require adjustment to match variable load impedance. Some types of matching networks are more amenable than others in this regard.

Application:

At radio frequencies of upper VHF or higher up to microwave frequencies one quarter wavelength is conveniently short enough to incorporate the component within many products, but not so small that it cannot be manufactured using normal engineering tolerances, and it is at these frequencies where the device is most often encountered. It is especially useful for making an inductor out of a capacitor, since designers have a preference for the latter. Another application is when DC power needs to be fed into a transmission line, which may be necessary to power an active device connected to the line, such as a switching transistor or a varactor diode for instance. An ideal DC voltage source has zero impedance, that is, it presents a short circuit and it is not useful to connect a short circuit directly across the line.

Feeding in the DC via a λ/4 transformer will transform the short circuit into an open circuit which has no effect on the signals on the line. Likewise, an open circuit can be transformed into a short circuit. The device can be used as a component in a filter, and in this application, it is sometimes known as an inverter because it produces the mathematical inverse of an impedance. Impedance inverters are not to be confused with the more common meaning of power inverter for a device that has the inverse function of a rectifier. Inverter is a general term for the class of circuits that have the function of inverting an impedance. There are many such circuits and the term does not necessarily imply a λ/4 transformer.

The most common use for inverters is to convert a 2-element-kind LC filter design such as a ladder network into a one-element-kind filter. Equally, for band pass filters, a two-resonator-kind (resonators and anti-resonators) filter can be converted to a one-resonator-kind. Filters incorporating λ/4 inverters are only suitable for narrow band applications. This is because the impedance transformer line only has the correct electrical length of $\lambda/4$ at one specific frequency. The further the signal is from this frequency the less accurately the impedance transformer will be reproducing the impedance inverter function and the less accurately it will be representing the element values of the original lumped-element filter design.

Chapter 2

Concept of Quarter wave Transformer

2.1: Basic concept of quarter wave transformer:

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

One drawback of the quarter-wave transformer is that it can only match real load impedance. 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. The operation of a quarter wave transformer from both an impedance viewpoint and a multiple reflection viewpoint.

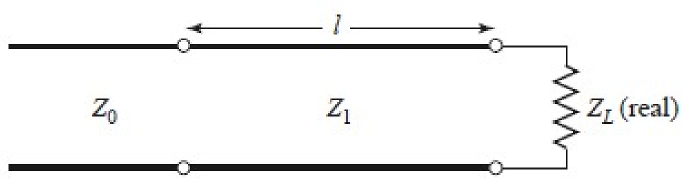

Fig. 2.1: Addition of a transmission line for matching between a load and other line

In this figure the length of matching line is l which is dependent on the wavelength (λ) of the signal. l = λ/4.

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

$$Z_1 = \sqrt{Z_o Z_L}$$

At the design frequency, f, the electrical length of the matching section is λ/4, but at other frequencies the length is different, so a perfect match is no longer obtained. The reflection coefficient can be written as,

$$\Gamma = \frac{Z_{in} - Z_o}{Z_{in} + Z_o}$$

$$= \frac{Z_1(Z_L - Z_o) + j\tan\beta\,({Z_1}^2 - Z_o Z_L)}{Z_1(Z_L + Z_o) + j\tan\beta\,({Z_1}^2 + Z_o Z_L)}$$

Putting the value of Z_1, we get

$$\Gamma = \frac{Z_L - Z_o}{(Z_L + Z_o) + j\,2\tan\beta\,(\sqrt{Z_o Z_L})}$$

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, *multi section 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.

2.2: Single section transformer:

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

$$\Gamma_1 = \frac{Z_2 - Z_1}{Z_2 + Z_1},$$
$$\Gamma_2 = -\Gamma_1,$$
$$\Gamma_3 = \frac{Z_L - Z_2}{Z_L + Z_2},$$
$$T_{21} = 1 + \Gamma_1 = \frac{2Z_2}{Z_1 + Z_2},$$
$$T_{12} = 1 + \Gamma_2 = \frac{2Z_1}{Z_1 + Z_2}.$$

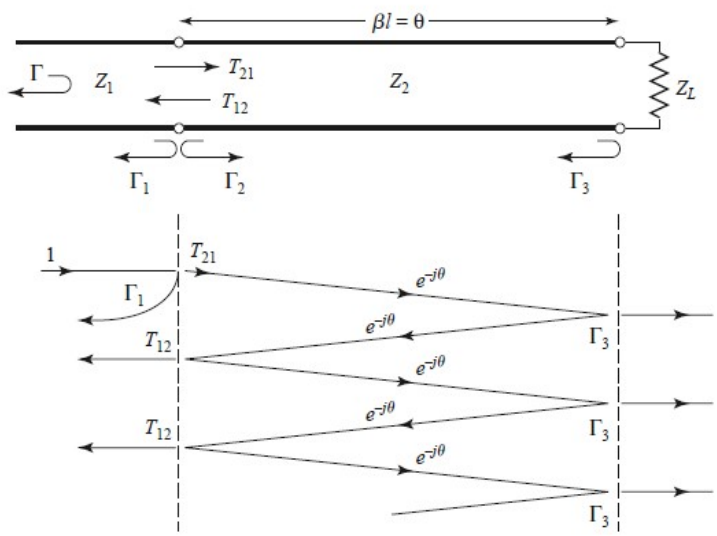

Fig. 2.2: Partial reflection and transmission in single- section matching network

The total reflection as an infinite sum of partial reflections and transmissions as follows:

$$\Gamma = \Gamma_1 + T_{12}T_{21}\Gamma_3 e^{-2j\theta} + T_{12}T_{21}\Gamma_3^2\Gamma_2 e^{-4j\theta} + \cdots$$
$$= \Gamma_1 + T_{12}T_{21}\Gamma_3 e^{-2j\theta}\sum_{n=0}^{\infty}\Gamma_2^n\Gamma_3^n e^{-2jn\theta}.$$

Using the formula of summation of infinite series, we get

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

$$\Gamma = \frac{\Gamma_1 + \Gamma_3 e^{-2j\theta}}{1 + \Gamma_1\Gamma_3 e^{-2j\theta}}.$$

As we know that,

$$\Gamma_2 = -\Gamma_1,\ T_{21} = 1 + \Gamma_1,\ \text{and}\ T_{12} = 1 - \Gamma_1$$

2.3: Multi section transformer:

Now consider the multi section transformer shown in Figure 2.3, which consists of *N* equal-length (*commensurate*) sections of transmission lines. We will derive an approximate expression for the total reflection coefficient. Partial reflection coefficients can be defined at each junction, as follows:

$$\Gamma_0 = \frac{Z_1 - Z_0}{Z_1 + Z_0},$$
$$\Gamma_n = \frac{Z_{n+1} - Z_n}{Z_{n+1} + Z_n},$$
$$\Gamma_N = \frac{Z_L - Z_N}{Z_L + Z_N}.$$

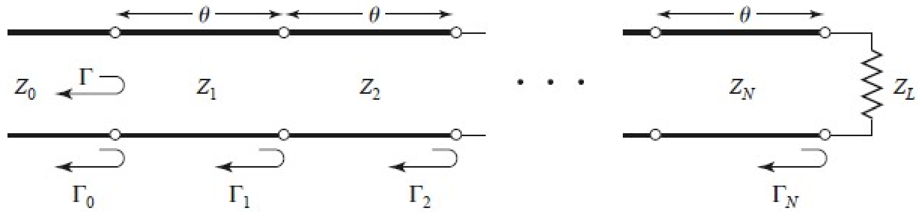

Fig. 2.3: Partial reflections in multi section matching network

We also assume that Z_N 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. The overall reflection coefficient will be,

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

Further assume that the transformer can be made symmetrical, therefore we can write as,

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

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

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

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 Γ_N and using enough sections (N). This should be clear from the realization that a Fourier series can approximate an arbitrary smooth function if enough terms are used. In the next two sections we will show how to use this theory to design multi section transformers for two of the most commonly used pass band responses: the *binomial* (maximally flat) response, and the *Chebyshev* (equal-ripple) response.

Chapter 3

Edge Impedance of a Patch

3.1 Introduction:

In this chapter, a rectangular patch antenna is designed at 2.45 GHz ISM band. Edge impedance of this patch is measured and plotted with respect to frequency. Microstrip patch antenna (Fig. 3.1) consist of two conducting plates (upper conductor part is called patch and lower conductor part is called ground surface) and a very thin substrate with height h (0.003 $\lambda_0 \leq$ h $\leq$ 0.05 λ_0, where λ_0 is free space wavelength). The relative permittivity (ϵ_r) of substrate should be in the range of [2.2, 12]. The patch geometry can be rectangular, circular, square, triangular, elliptical etc. [1]. A simple patch structure is illustrated in Fig. 3.1.

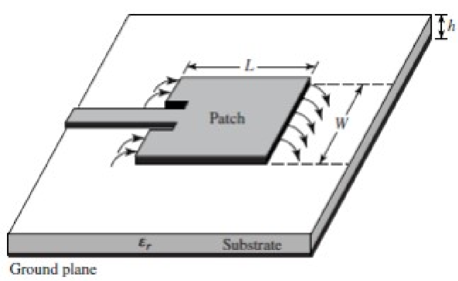

Fig. 3.1: Microstrip Patch antenna with microstrip line feed [1]

3.2 Theoretical calculation of microstrip antenna:

There are several methods to analyze microstrip antennas such as transmission-line model, cavity model and full wave techniques. Transmission line is the easiest one. In transmission-line model, microstrip antenna is considered as two radiating apertures or slots of width W separated by the length of patch L. Due to finite length and width of patch, fringing fields present along both length and width. This fringing is a function of ratio between length or width and height of substrate and dielectric constant of substrate. As (L/h) is much greater than 1, fringing is reduced. However, it should be considered because it impacts the resonant frequency of antenna [1].

The fringing fields from patch experience a non-homogeneous two-dielectric environment (typically air and substrate). For that reason, the effective dielectric constant is introduced to account for wave propagation and fringing. The value of effective permittivity ϵ_{eff} is $1 < \epsilon_{eff} < \epsilon_r$ where relative permittivity of substrate should be much greater than 1.

As W/h is much greater than unity, the effective permittivity can be written as [1],

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r + 1}{2}\left[1 + 12\,\frac{h}{W}\right]^{-\frac{1}{2}}$$

Due to presence of fringing fields, electrical dimensions of microstrip patch antenna greater than physical size. The extended length ΔL can be expressed as [1],

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

For dominant mode (TM_{10}), physical length L is taken as $\lambda/2$. The effective length of patch will be, $L_{eff} = L + 2\Delta L$ and resonant frequency at dominant mode is given by [1],

$$(f_r)_{10} = \frac{c}{2L_{eff}\sqrt{\epsilon_{eff}}}$$

$$L_{eff} = \frac{c}{2(f_r)_{10}\sqrt{\epsilon_{eff}}}$$

Therefore, the physical length (L) of the patch is given by,

$$L = \frac{c}{2L_{eff}\sqrt{\epsilon_{eff}}} - 2\Delta L$$

The width of an efficient radiator is given by [1],

$$W = \frac{c}{2f_r}\sqrt{\left(\frac{2}{\epsilon_r + 1}\right)}$$

3.3 Design of Microstrip patch antenna at 2.45 GHz

By using the above equations of microstrip antenna, we get the length and width of patch is 28.75 mm and 45 mm respectively. A 28.75 mm X 45 mm patch element is designed by using copper with thickness of 0.035 mm. 30 mil (0.762 mm) Arlon AD 430 ($\epsilon_r = 4.3$ and $tan\delta = 0.003$) is used as substrate material. A ground plane is designed with copper material. The dimensions of ground and substrate layers are 80 mm X 100 mm. Fig. 3.2 shows the top view of this patch.

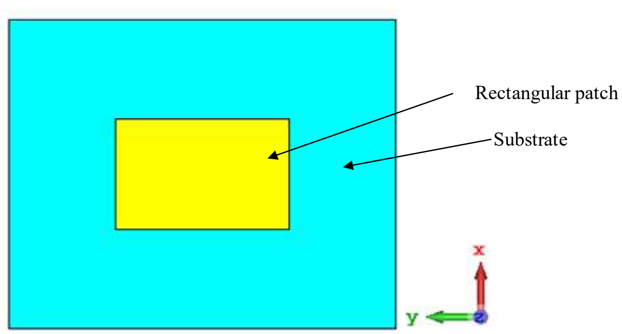

Fig. 3.2: Top view of patch designed in CST Microwave Studio

3.4 Edge impedance of patch using discrete port

A discrete port is patch at placed at an edge of the designed patch (refer to Fig. 3.2) as shown in Fig. 3.3. It is added to measure impedance at this edge of patch.

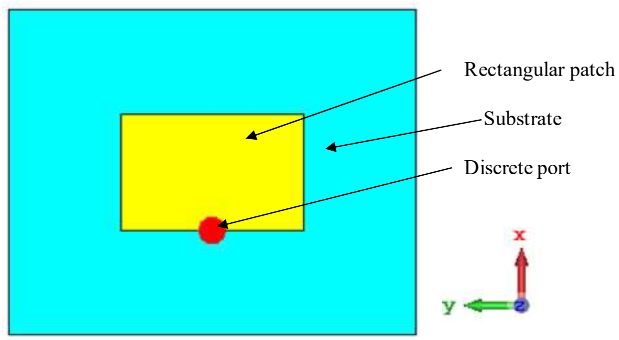

Fig. 3.3: Discrete port placed at an edge of patch

After simulating the patch structure after placing the discrete port the return loss and impedance responses are observed. Fig. 3.4 shows the S_{11} vs frequency plot of this design. Impedance (real and imaginary) vs frequency plot is depicted in Fig. 3.4. The return loss value of this patch is 5.04 dB at resonant frequency of 2.43 GHz. The impedance at that frequency is 180.29 ohm.

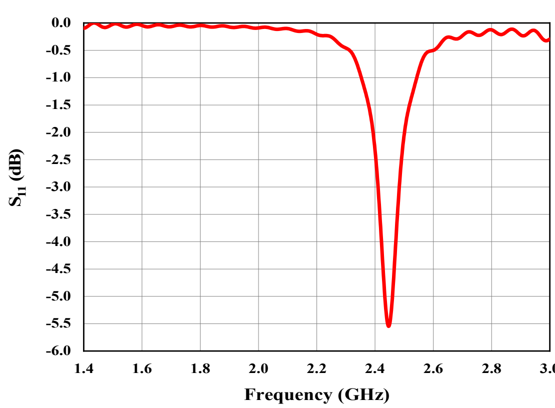

Fig. 3.4: S_{11} vs frequency of patch

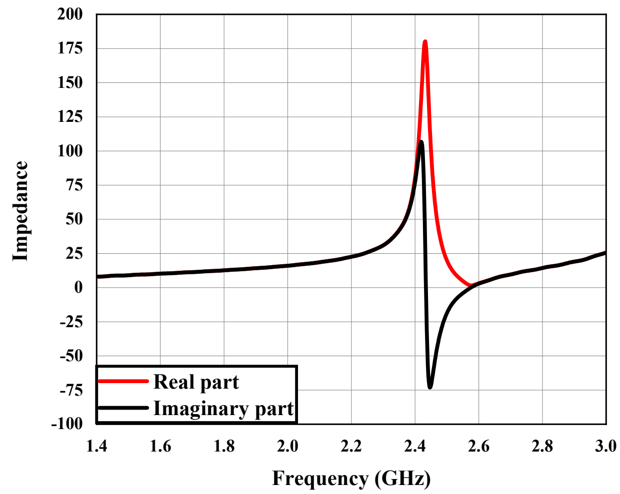

Fig. 3.5: Impedance vs frequency of patch

3.5 Conclusion

From this chapter the edge impedance of the patch is 180.29 ohm at its resonant frequency of 2.43 GHz. To match this antenna impedance with 50 ohm SMA connector, impedance matching network is required. For that reason, in the next part, matching network will be designed.

REFERENCES

[1] C. A. Balanis, “*Antenna theory: Analysis and Design*,” Wiley, 4th Edition.

Chapter 4

Single Section Matching

4.1 Introduction:

In the previous chapter a patch is designed at 2.43 GHz. Microstrip feeding technique is used in this work for exciting the patch. A single section matching network designed in this chapter between patch and 50 ohm port.

4.2 Calculation of single section matching network:

According to the transmission line theory, the input impedance of an arbitrary length transmission line (refer to Fig. 4.1) can be written as,

$$Z_{in} = Z_o \frac{Z_L + jZ_o \tan \beta l}{Z_o + jZ_L \tan \beta l}$$

where Z_o= characteristic impedance of transmission line, Z_L = load impedance, l = distance from load at which input impedance is calculating and β is the phase constant.

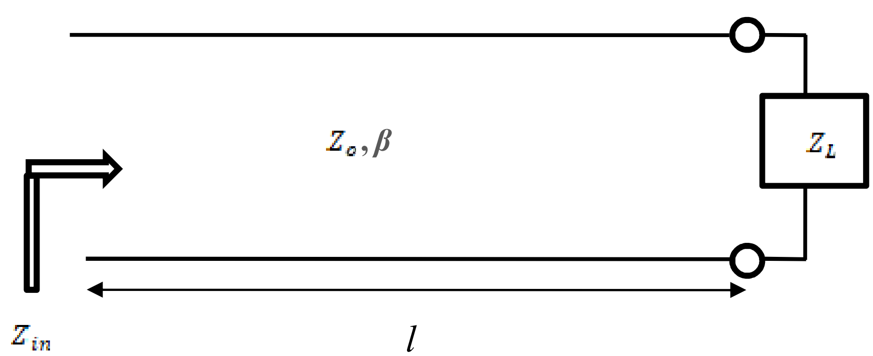

Fig. 4.1: A transmission line terminated by a load

Now for quarter wave transmission line, $l = \frac{\lambda}{4}$.

So,

$$\beta l = \frac{2\pi}{\lambda} \times \frac{\lambda}{4} = \frac{\pi}{2}$$

Therefore, input impedance for a quarter wave transmission line is,

$$Z_{in} = \frac{Z_o^2}{Z_L}$$

In this work, a quarter wave transformer is connected between patch and the 50 ohm connector. Here, Z_{in} = edge impedance of patch = 180.29 ohm and Z_L = 50 ohm. Then, the characteristic impedance of the quarter wave transformer will be,

$$Z_0 = \sqrt{Z_{in} \times Z_L} = 94.94 \text{ ohm.}$$

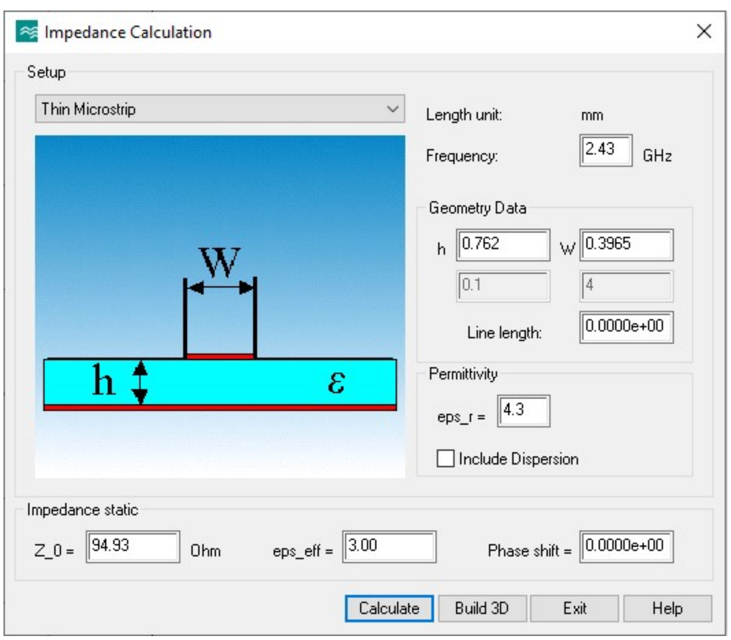

Fig. 4.2: Impedance calculation for 94.94 ohm strip line

By using microstrip line impedance calculator of CST, the width of the strip line is calculated to obtain 94.94 ohm impedance as shown in Fig. 4.2. For the Fig. 4.2, the width of the line is obtained as 0.3965 mm. As the operating frequency of this patch is 2.43 GHz, the length of the quarter wave transformer will be 14.75 mm. In similar manner, width of the line to provide 50 ohm impedance is calculated as 1.494 mm.

4.3 Design of single section matching network:

Based on the above calculations, a single section quarter wave transformer is designed in CST Microwave Studio 2016 as shown in Fig. 4.3 to match impedance between an edge the patch and 50 ohm line.

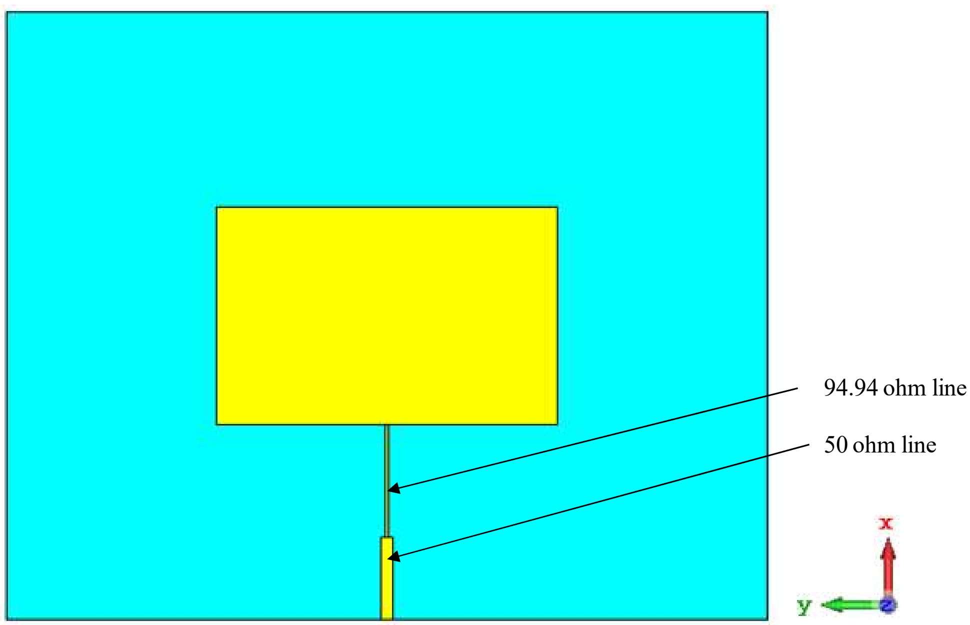

Fig. 4.3: Single section matching network design in CST

The return loss vs frequency of the design shown in Fig. 4.3 is plotted in Fig. 4.4. The return loss value is 21.10 dB at the resonant frequency 2.44 GHz. This antenna is now radiated over 2.43 GHz to 2.45 GHz. The fractional bandwidth is 0.82% around its resonant frequency.

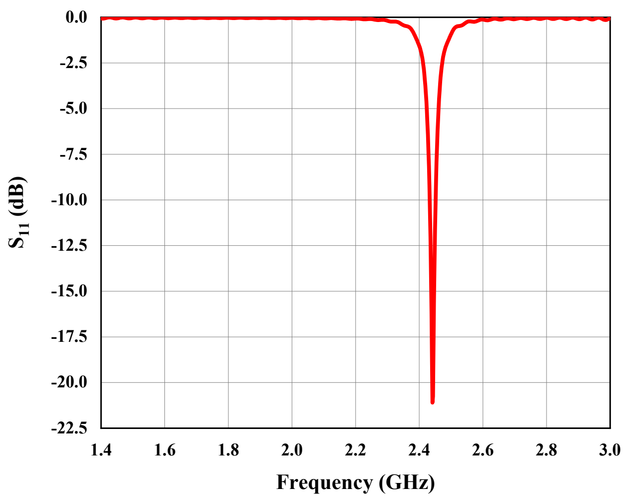

Fig. 4.4: S_{11} vs frequency plot for patch with single section matching network

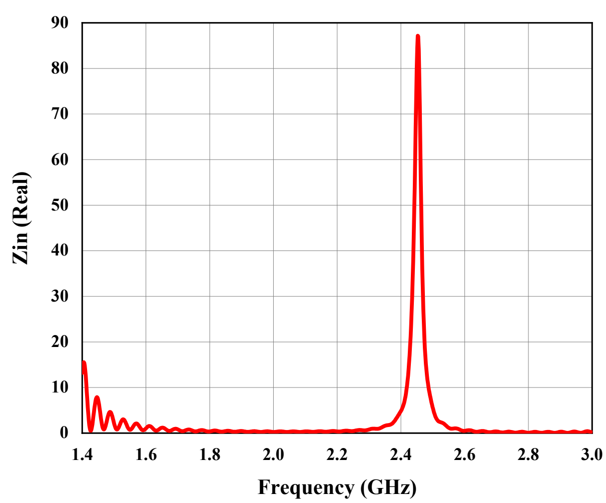

Fig. 4.5: Real part of impedance vs frequency plot

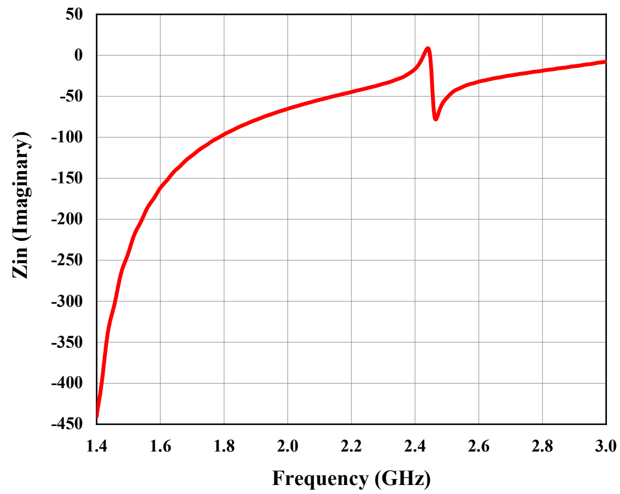

Fig. 4.6: Imaginary part of impedance vs frequency plot

The real and imaginary parts of impedance vs frequency are plotted in Fig. 4.5 and 4.6 respectively. At resonant frequency, real part of the impedance is 16.79 ohm. So, from these plots it is observed that the impedance of the patch is not properly matched with 50 ohm.

4.4 Conclusion

After implementing single section matching network, it is observed that the return loss value is improved at resonant frequency. However, the fractional bandwidth is very low. Therefore in the next chapter, multi section matching network will be designed.

Chapter 5

Multi Section Matching

5.1 Introduction:

In the previous chapter a single section matching network was designed by using microstrip line. In spite of having less complexity in single section matching technique, the fractional bandwidth is very less. It may arise problem in communication purpose. In this chapter multi section matching technique is used for achieving better bandwidth. In the first step, a two-section matching network is designed. In the next step, three-section network has been studied.

5.2 Two section matching network design:

5.2.1: *Calculation of two section matching network*:

Here, two quarter waver transformers are used to match the edge impedance of patch with 50 ohm line as shown in Fig. 5.1.

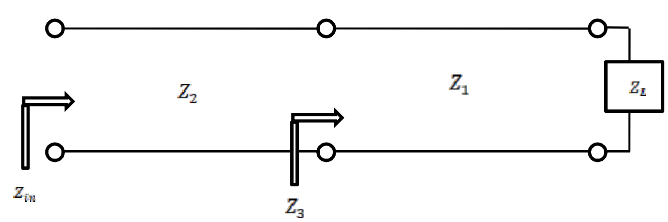

Fig. 5.1: Two section impedance matching network

As calculated earlier, the lengths of all transformers are 14.75 mm. The calculations of all variables mentioned in Fig. 5.1 are shown below.

Step 1: In the first step, the characteristic impedance Z_1 is calculated while taking the value of Z_3 is taken as 70 ohm. That means, by using quarter wave transformer with characteristic impedance Z_1, the load impedance Z_L= 50 ohm is trying to match with Z_3= 70 ohm. Therefore,

$$Z_1 = \sqrt{Z_L Z_3} = \sqrt{50 \times 70} = 59.16 \text{ ohm.}$$

Step 2: In the next step, the characteristic impedance Z_2 is calculated. That means, by using quarter wave transformer with characteristic impedance Z_2, impedance Z_3= 70 ohm is trying to match with edge impedance Z_{in}= 180.29 ohm. Therefore,

$$Z_2 = \sqrt{Z_{in} Z_3} = \sqrt{180.29 \times 70} = 112.34 \text{ ohm.}$$

Step 3: Width of the lines providing 59.16 ohm and 112.34 ohm are calculated by using Microstrip line impedance calculator in CST. The line width of 59.16 ohm is obtained as 1.114 mm whereas for 112.34 ohm is 0.2457 mm.

5.2.2: *Design of two section matching network*:

After estimating the dimensions of the strip lines, the antenna with two section matching network is designed in CST as shown in Fig. 5.2.

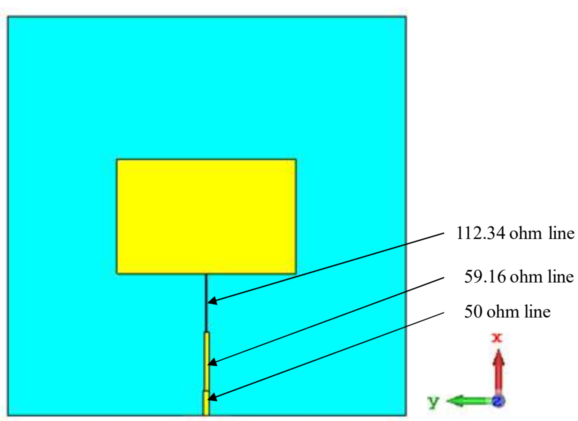

Fig. 5.2: Top view of patch with two section matching

The return loss vs frequency of the design shown in Fig. 5.2 is plotted in Fig. 5.3. The return loss value is 31.06 dB at the resonant frequency 2.44 GHz. This antenna is now radiated over 2.42 GHz to 2.45 GHz. The fractional bandwidth is 1.23% around its resonant frequency. So, the fractional bandwidth is increased by applying two section matching network instead of single section network.

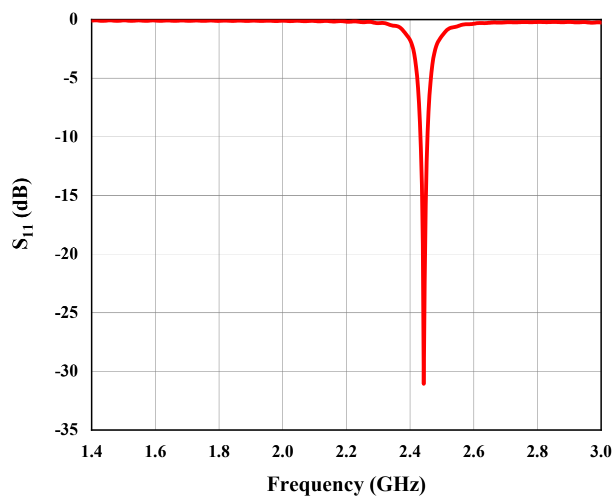

Fig. 5.3: S_{11} vs frequency for two section network

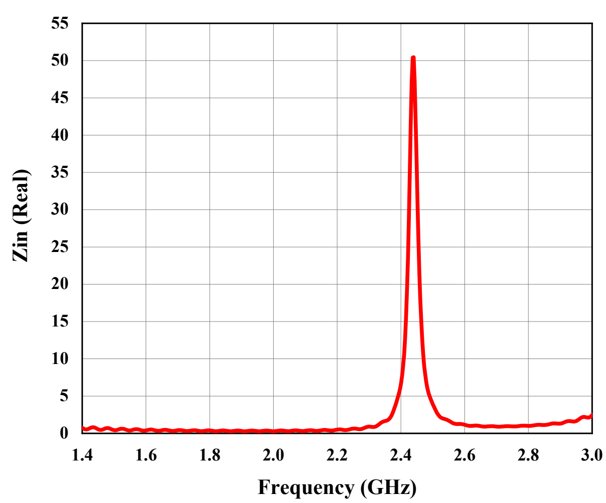

Fig. 5.4: Real part of impedance vs frequency for two section network

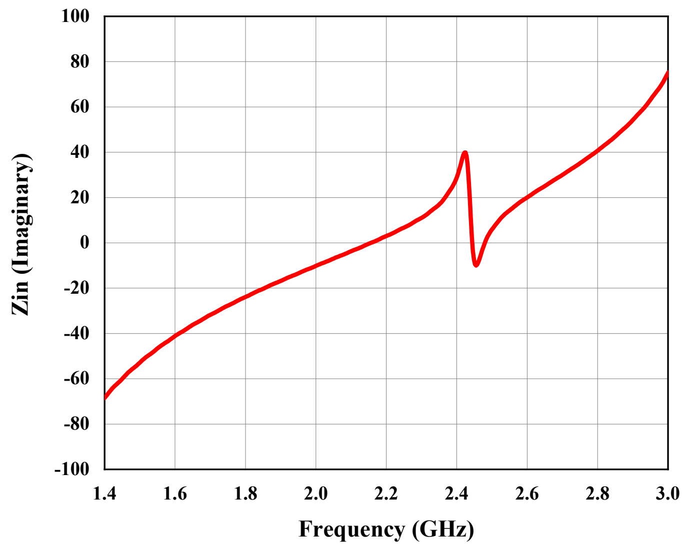

Fig. 5.5: Imaginary part of impedance vs frequency for two section network

The real and imaginary parts of impedance vs frequency are plotted in Fig. 5.4 and 5.5 respectively. At resonant frequency, real part of the impedance is 44.88 ohm. So, from these plots it is observed that impedance value is nearly matched with 50 ohm.

5.3 Three section matching network design:

5.3.1: *Calculation of three section matching network*:

Here, three section quarter waver transformers are used to match the edge impedance of patch with 50 ohm line as shown in Fig. 5.6.

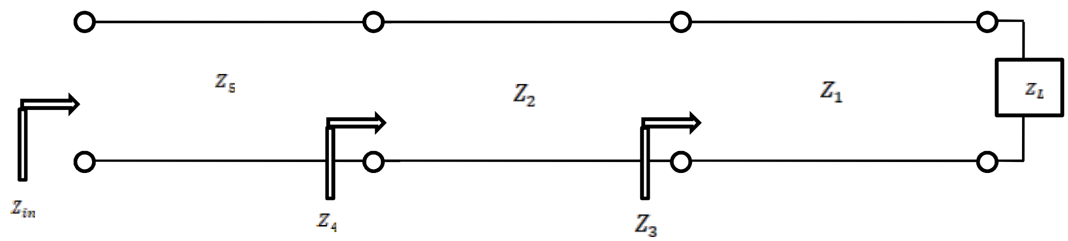

Fig. 5.6: Three section impedance matching network

As calculated earlier, the lengths of all transformers are 14.75 mm. The calculations of all variables mentioned in Fig. 5.6 are shown below.

Step 1: In the first step, the characteristic impedance Z_1 is calculated while taking the value of Z_3 is taken as 70 ohm. That means, by using quarter wave transformer with characteristic impedance Z_1, the load impedance Z_L= 50 ohm is trying to match with Z_3= 70 ohm. Therefore,

$$Z_1 = \sqrt{Z_L Z_3} = \sqrt{50 \times 70} = 59.16 \text{ ohm.}$$

Step 2: In the next step, the characteristic impedance Z_2 is calculated while taking the value of Z_4 is taken as 110 ohm. That means, by using quarter wave transformer with characteristic impedance Z_2, impedance Z_4= 110 ohm is trying to match with edge impedance Z_3= 70 ohm. Therefore,

$$Z_2 = \sqrt{Z_4 Z_3} = \sqrt{110 \times 70} = 87.75 \text{ ohm.}$$

Step 3: In the next step, the characteristic impedance Z_5 is calculated. That means, by using quarter wave transformer with characteristic impedance Z_5, impedance Z_4= 110 ohm is trying to match with edge impedance Z_{in}= 180.29 ohm. Therefore,

$$Z_5 = \sqrt{Z_{in} Z_4} = \sqrt{180.29 \times 110} = 140.82 \text{ ohm.}$$

***Step 4*:** Width of the lines providing 59.16 ohm, 87.75 ohm and 140.82 ohm are calculated by using Microstrip line impedance calculator in CST. The line widths are for 59.16 ohm, 87.75 ohm and 140. 82 ohm is obtained as 1.114 mm, 0.484 ohm and 0.114 mm respectively.

5.3.2: *Design of three section matching network*:

After estimating the dimensions of the strip lines, the antenna with three section matching network is designed in CST as shown in Fig. 5.7.

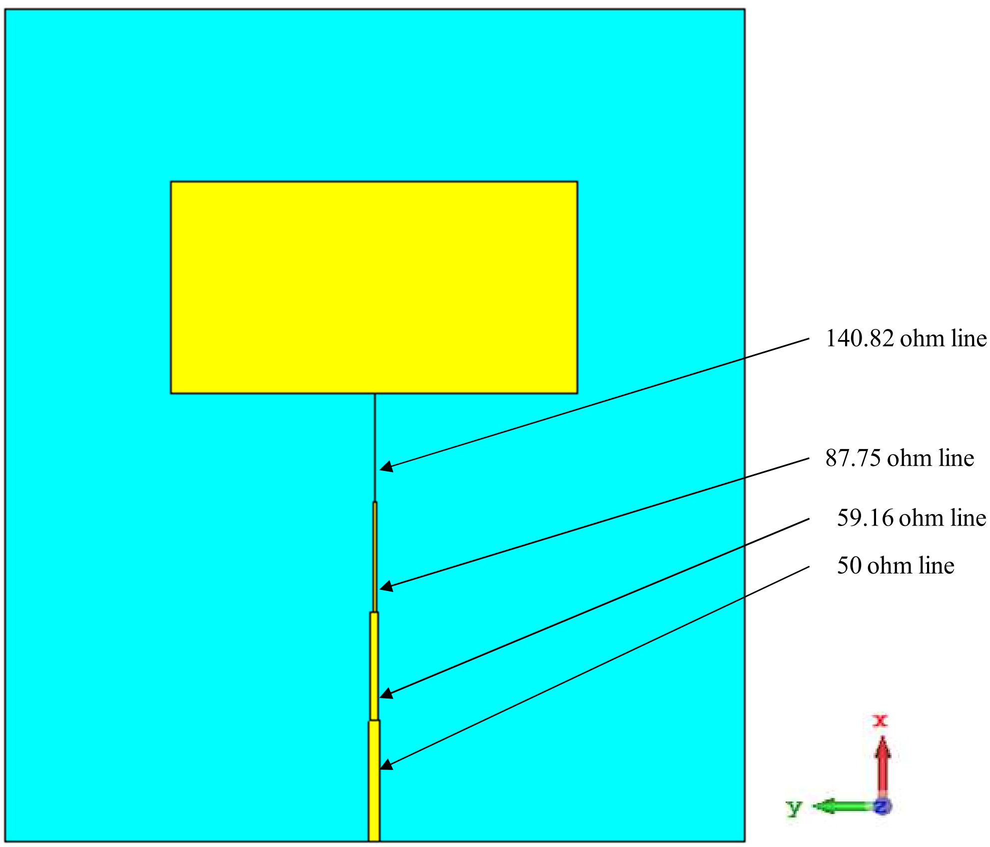

Fig. 5.7: Top view of patch with three section matching

The return loss vs frequency of the design shown in Fig. 5.7 is plotted in Fig. 5.8. The return loss value is 14.63 dB at the resonant frequency 2.43 GHz. This antenna is now radiated over 2.41 GHz to 2.44 GHz. The fractional bandwidth is 1.61% around its resonant frequency. So, the fractional bandwidth is increased by applying three section matching network instead of two section network.

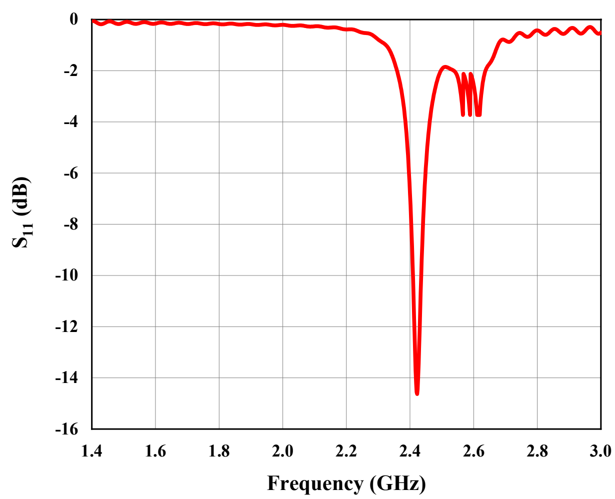

Fig. 5.3: S_{11} vs frequency for three section network

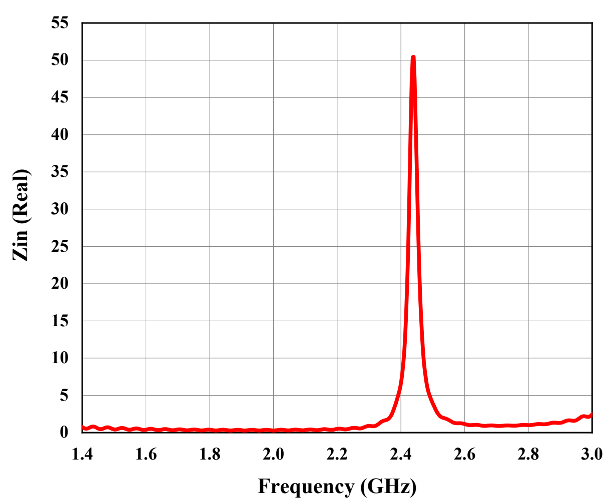

Fig. 5.4: Real part of impedance vs frequency for three section network

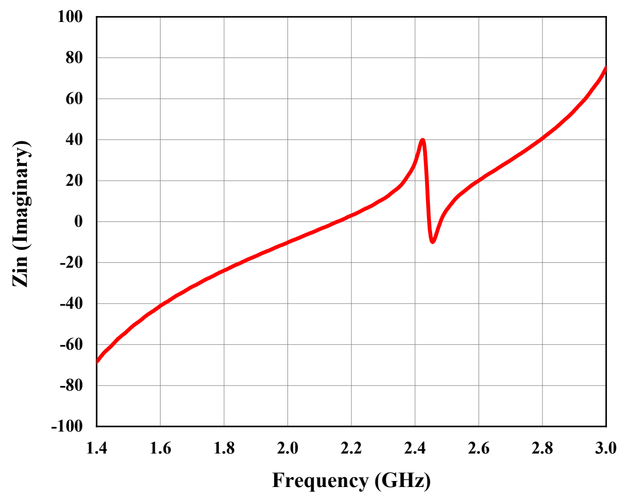

Fig. 5.5: Imaginary part of impedance vs frequency for three section network

The real and imaginary parts of impedance vs frequency are plotted in Fig. 5.4 and 5.5 respectively. At resonant frequency, real part of the impedance is 52.25 ohm. So, from these plots it is observed that impedance value is nearly matched with 50 ohm.

5.3 Conclusion:

In this chapter, two-section and three-section impedance matching networks are designed to match edge impedance of patch with 50 ohm line. It is observed that the matching is improved and bandwidth is increased with the increment of number of sections.

Chapter 6

Fabrication and Measurement

6.1 Introduction:

In previous chapters it is observed that, two-section and three-section impedance matching networks provide better matching and larger bandwidth than single section matching network. However, the widths of the lines for two and three section networks are too small to fabricate. Therefore, the patch antenna with single section matching network is fabricated.

6.2 Fabrication:

The top view of the fabricated antenna is shown in Fig. 6.1.

Fig. 6.1: Fabricated patch antenna with single section matching network

The process of fabrication is mentioned below.

Step1: Convert the CST file into DXF file so that we can run the file on CAD tool and fill structure with black color and the designed structure will be like:

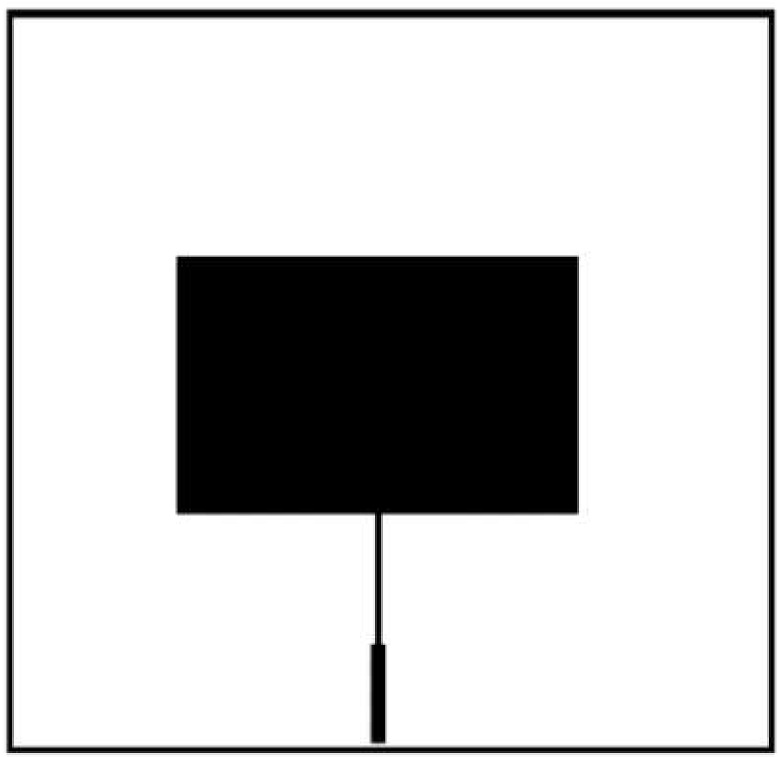

Fig. 6.2: DXF file of patch antenna with single section matching network

Step 2: Print the structure on transparent paper with the exact dimension of the structure and cut the paper along its border line.

Step 3: Now clean the sheet with acetone and let it dry. Next put photo-resist on the material by using brush and give heat on it and let it dry.

Step 4: Now put the design on the sheet and place in into the UV exposure chamber and wait for 17 minutes.

Step 5: Next by using some amount of developer on it the primary structure can be seen and then place it into ferric chloride solution for 20-25 minutes and the UV exposed copper will react with ferric chloride and will be removed and At the end a port of 50 ohm should get attached with 50 ohm line by shouldering process.

6.3 Measurement:

The return loss over frequency of simulation and measurement are plotted in Fig. 6.3. From this figure, it is observed that the return loss value for Measurement is 46.22 dB whereas return loss value for simulation is 21.10 dB. The resonant frequencies for simulation and measurement are 2.43 GHz and 2.445 GHz respectively.

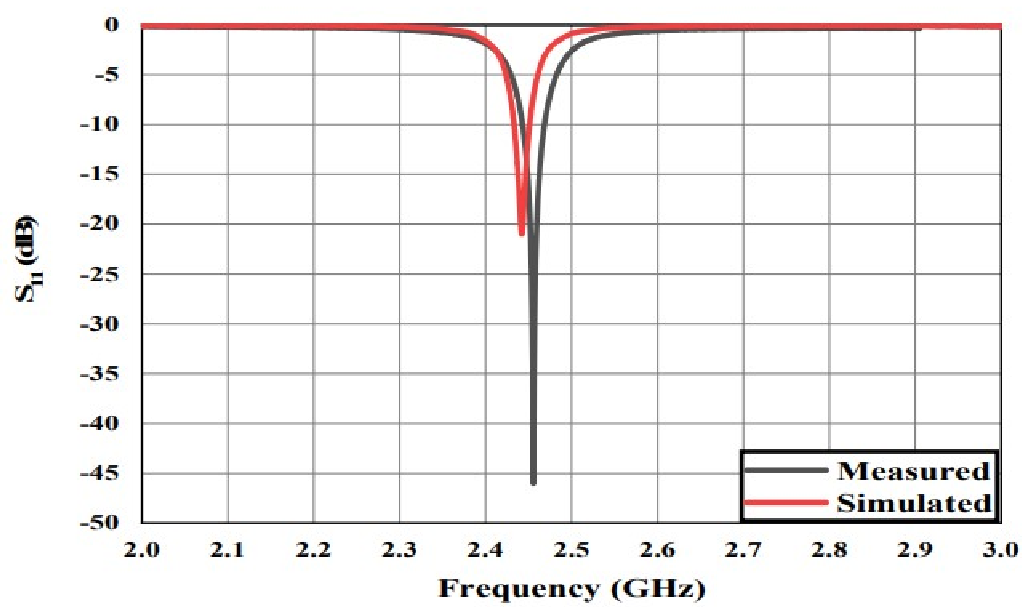

Fig. 6.3: Return loss curves for simulation and measurement

Chapter 7

Conclusion

In this thesis the importance of matching section between source to load is described. The reason why we want to match the line is that we would like maximum power traveling down the line to be absorbed by the load at the end of line.

The two-section and three-section impedance matching networks provide better matching and larger bandwidth than single section matching network. However, the widths of the lines for two and three section networks are too small to fabricate. Therefore, the patch antenna with single section matching network is fabricated.

A comparison table in terms of bandwidth of different number of impedance matching sections is shown below.

TABLE 7.1: Comparison between different number sections

No. of sections	Fractional bandwidth (%)
1	0.82
2	1.23
3	1.61

Chapter 8

Reference

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Addendum

Binomial Multi Section Matching

1.Introduction:

The pass band 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 center frequency. Such a response can be obtained with a reflection coefficient of the following form:

$$\Gamma\ (\theta) = A\ (1+e^{-2j\ \Theta})^N$$

Where A is given by

$$A= 2^{-N}\ \frac{Z_L-Z_O}{Z_L+Z_O}$$

The characteristic impedances, Z_n, can be found, but a simpler solution can be obtained using the following approximation. Because we assumed that the Γ_n are small, so, $\Gamma_n = \frac{Z_{n+1}-Z_n}{Z_{n+1}+Z_n} = \frac{1}{2} = \ln\frac{Z_{n+1}}{Z_n}$

Since $\ln(x) \cong 2(x - 1)/(x + 1)$ for x close to unity. So above equation can be simplified as:

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

Then the reflection coefficient magnitude is:

$$|\Gamma(\theta)| = 2^N |A|\,|\cos\theta|^N$$

Where,

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

3. Three section binomial matching network design:

Calculation of three section binomial matching network:

Here, three section binomial quarter waver transformers are used to match the edge impedance of patch with 50 ohm line as shown in Fig.a.

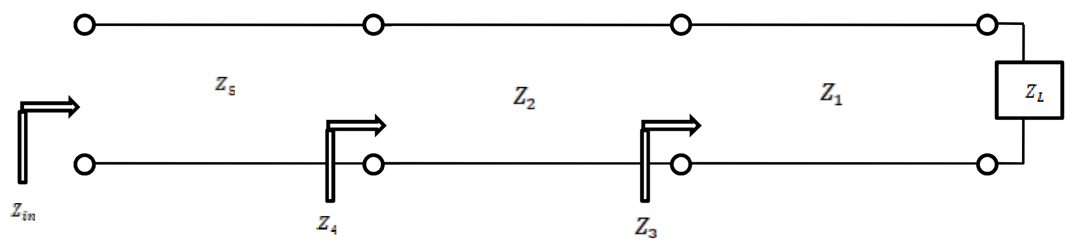

Fig. a: Three section binomial impedance matching network

By using,

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

So, the binomial coefficients are:

$$C_0^3 = \frac{3!}{0!\,3!} = 1$$

$$C_1^3 = \frac{3!}{2!\,1!} = 3$$

$$C_2^3 = \frac{3!}{1!\,2!} = 3$$

As calculated earlier, the lengths of all transformers are 14.75 mm. The calculations of all variables mentioned in Fig. a are shown below.

By using the formula to determine the characteristic impedance of each section as discussed:

$$\ln Z_{n+1} = \ln Z_n + 2^{-N} C_n^N \ln(50/180.29)$$

For n=0;

$$\ln Z_1 = \ln Z_0 + 2^{-3}\, C_0^3 \ln(50/180.29)$$

$$\ln Z_1 = \ln 180.29 + 2^{-3}\,(1) \ln(0.2773)$$

$$\ln Z_1 = 5.034$$

$$\text{So, } Z_1 = 153.58 \text{ ohm.}$$

For n=1;

$$\ln Z_2 = \ln Z_1 + 2^{-3}\, C_1^3 \ln(50/180.29)$$

$$\ln Z_2 = \ln 153.58 + 2^{-3}(3) \ln(0.2773)$$

$$\ln Z_2 = 4.553$$

$$\text{So, } Z_2 = 94.92 \text{ ohm.}$$

For n=2;

$$\ln Z_5 = \ln Z_2 + 2^{-3}\, C_2^3 \ln(50/180.29)$$

$$\ln Z_5 = \ln 94.92 + 2^{-3}(3) \ln(0.2773)$$

$$\ln Z_5 = 4.072$$

$$\text{So, } Z_5 = 58.67 \text{ ohm}$$

Now width of the lines providing 153.58 ohm, 94.92 ohm and 58.67 ohm are calculated by using Microstrip line impedance calculator in CST. The line widths are obtained as 0.08 mm, 0.4 mm and 1.14 mm respectively.

4: *Design of three section matching network*:

After estimating the dimensions of the strip lines, the antenna with three section binomial matching network is designed in CST as shown in Fig. b.

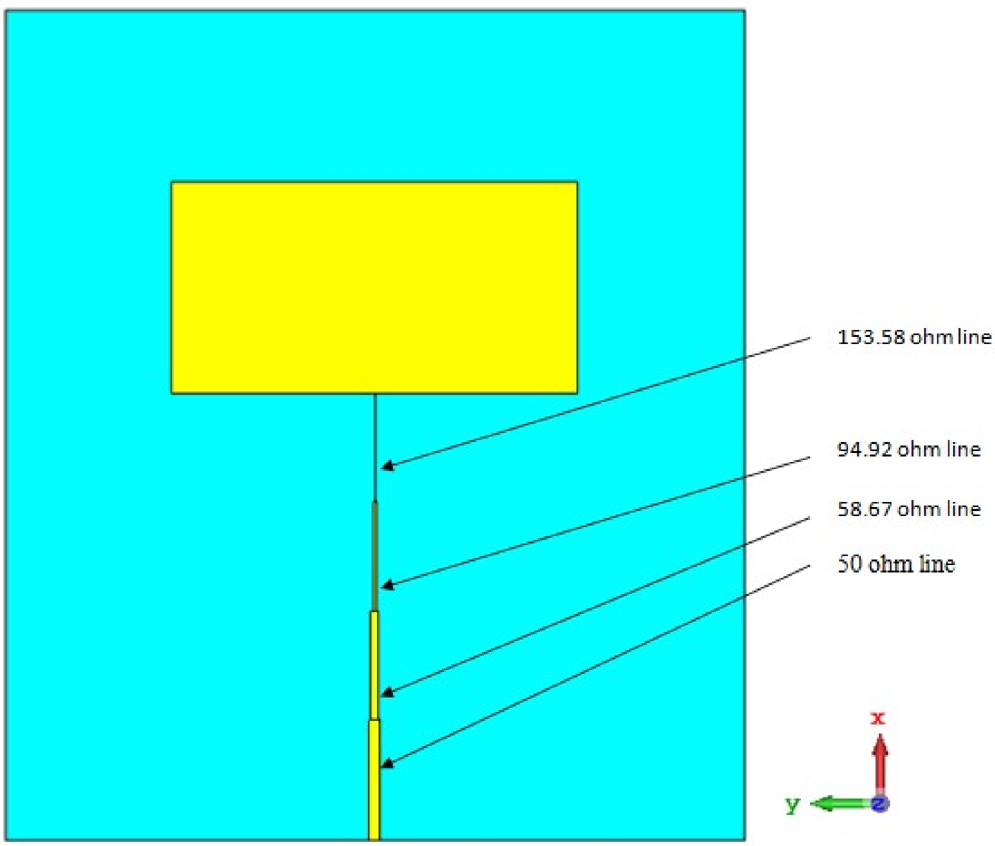

Fig. b:Top view of patch with binomialthree sectionmatching network.

The return loss value is 13.53 dB at the resonant frequency 2.43 GHz. This antenna is now radiated over 2.4 GHz to 2.45 GHz. The fractional bandwidth is 2.05% around its resonant frequency. So, the fractional bandwidth is increased by applying three section binomial matching network.

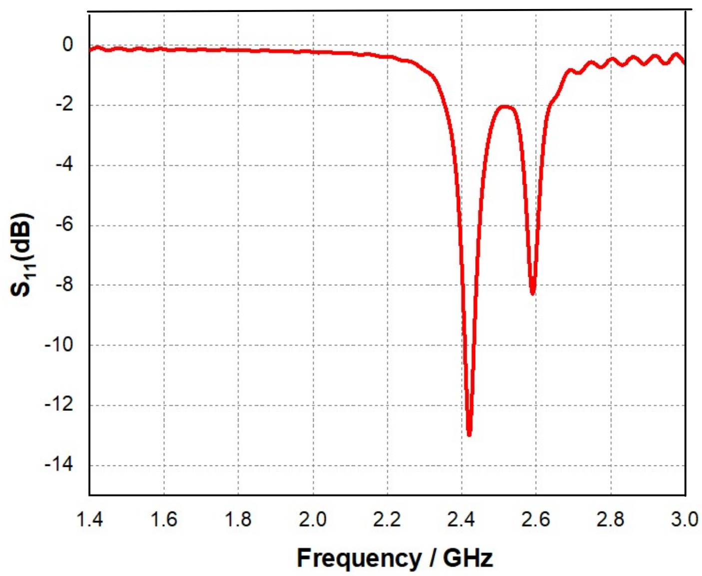

Fig: S_{11}vs frequency for binomial three section matching network.

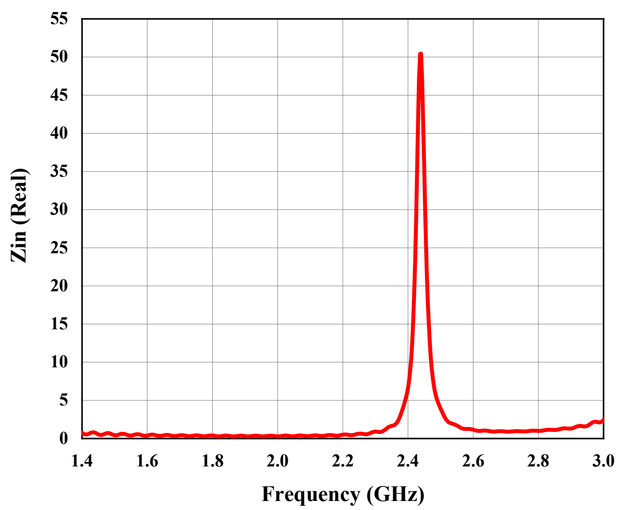

Fig: Real part of impedance vs frequency for binomial three section matching network.

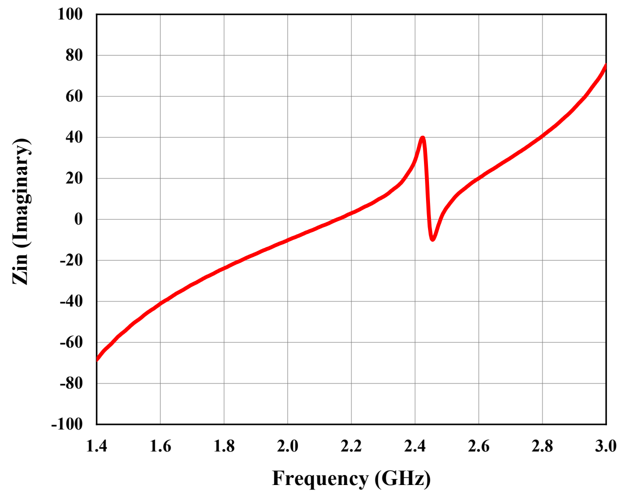

Fig.: Imaginary part of impedance vs frequency for binomial three section matching network.

The real and imaginary parts of impedance vs frequency are plotted in Fig. respectively. At resonant frequency, real part of the impedance is 52.21 ohm. So, from these plots it is observed that impedance value is nearly matched with 50 ohm.

5.Conclusion:

The advantage of binomial process is that it is not required toassume the characteristic impedance of the lines, by mathematical calculation the desirable values will come out and the analysis will beeasier and more accurate.Here, three-section binomial impedance matching network is designed to match edge impedance of patch with 50 ohm line. It is observed that the matching is improved and bandwidth is increased. The **fractional bandwidth** is **2.05%** around its resonant frequency. Therefore, the **fractional bandwidth is improved** by using binomial three section matching network.