DESIGNING DUAL BAND IMPEDANCE MATCHING TRANSFORMER

By

VIKAS VIKRAM SINGH

BTP report submitted in partial fulfilment of the requirements
for the Degree of B.Tech. in
Electronics & Communication Engineering on April 18, 2018

BTP Track: Research

BTP Advisor
Dr. M. S. HASHMI

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Student's Declaration

I hereby declare that the work presented in the report entitled "**DESIGNING DUAL BAND IMPEDANCE MATCHING TRANSFORMER**" submitted by me for the partial fulfilment of the requirements for the degree of Bachelor of Technology in Electronics & Communication Engineering at **Indraprastha Institute of Information Technology, Delhi**, is an authentic record of my work carried out under guidance of **Dr. M. S. Hashmi.**

acknowledgements have been given in the report to all material used. This work has not been submitted anywhere else for the reward of any other degree.

.... Place & Date: ....
(Vikas Vikram Singh)

Certificate

This is to certify that the above statement made by the candidate is correct to the best of my knowledge.

.... Place & Date: ....
(Dr. M. S. Hashmi)

ACKNOWLEDGEMENTS

First of all, I would like to thank Professor **Dr. M. S. HASHMI** for providing me an opportunity to work on this project under his supervision and guiding me throughout the project. I would also like to thank Mr. Rahul Gupta for the help and advice throughout the project. I also wish to thank my family for their continuous support and encouragement throughout my studies. I especially wish to thank my mother Mithla Singh, my father Saket Kumar Singh for continuous encouragement and support.

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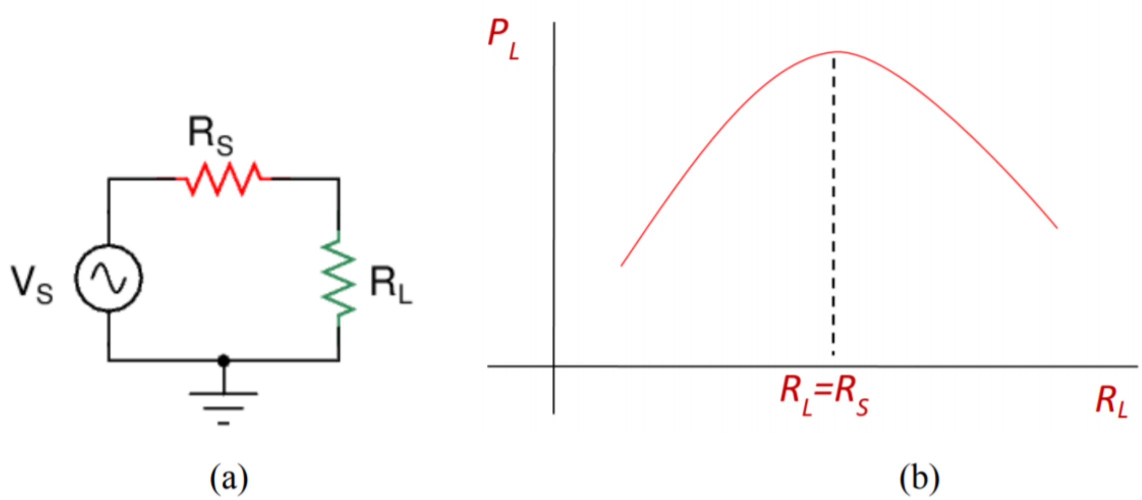
1. INTRODUCTION

Impedance matching or impedance transforming circuits (networks) are one of the most ubiquitous blocks in a host of RF/microwave components and systems. One of the motivations to use impedance matching networks comes from the ***maximum power transfer theorem***. A voltage source V_s with its source resistance connected to the load R_L is shown in Fig. 1.1(a). Variation of the power delivered to the load, P_L is shown in Fig. 1.1(b). It is evident from Fig. 1.1(b) that P_L is maximum when R_L=R_S. This result is known as the condition for maximum power transfer. If the source and the load were a complex quantity, R_L=R^*_S holds true, where asterisk denotes a complex conjugate quantity. Since, in general, in RF/microwave components such as in amplifier, the impedance looking into the gain device is different than the source impedance; a matching network is required to transfer maximum power from the source side to the load side. Two points must be noted in this context. First, need of matching is not always motivated from the maximum power transfer theorem. Sometimes, such as in low-noise amplifiers or in wideband amplifiers, matching (other than conjugate) is used to trade- off among noise, bandwidth and gain requirements [2-3]. Second, amplifiers are not the only devices where a matching circuit is required. In fact, matching network is such a ubiquitous block in RF/Microwave devices and systems that "***they are everywhere***". Some applications of matching networks in RF/microwave devices are highlighted in the next few paragraphs.



(a) (b)

Figure 1.1 (a) Voltage source connected to a load (b) Power delivered to the load versus the load value.

1.1 Application of Matching Networks

Low Noise Amplifier: Since the first aim of LNA design is to obtain minimum noise figure, input matching should be done accordingly. For an amplifier, there is an optimum source impedance, Γopt, which leads to minimum noise figure. In order to obtain minimum noise figure, input impedance of an amplifier should be matched to this optimum source impedance value. It is important to note that, however, optimum source impedance that gives minimum noise figure is not necessarily the optimum impedance for the maximum power transfer. Therefore, one must sacrifice the power gain, and therefore return loss, to obtain minimum noise figure, and vice versa in order to optimize output signal-to-noise ratio unless those impedances are same by coincidence. In addition to obtain minimum noise figure, one of the other challenges in LNA design is to obtain the maximum available gain. Therefore, conjugate matching should be done at the output of the amplifier in order to get the best S/N at the output of the system. If the load impedance is Z_L, then for maximum power transfer source impedance Z_S is equal to a conjugate of load impedance.

$$Z_S = Z_L^*$$

The reflection coefficient Γ is a normalized measure of the relationship between source impedance and load impedance. Input and output impedance matching is given by the input and output return loss. Return loss (RL) is the relationship between the reflected power wave at a port to incident power wave at the same port. A perfect match will have no reflection and an SWR of 1.

Power Dividers: A Wilkinson power divider is shown in Fig. 1.2(a). This configuration is used for splitting power equally or for power combining. It may not be apparent if the concept of matching has to do anything here. For this purpose, even- mode equivalent circuit of the Wilkinson divider is shown in Fig. 1.2(b). It is evident now that the idea in such a situation is to *match* a real load impedance of Z_0 to $2Z_0$ using a quarter-wave line [4]. It is a fact that the characteristic impedance of the quarter-wave line must be $\sqrt{(Z_0 * 2Z_0)} = \sqrt{2}\, Z_0$ and that is exactly shown in the Wilkinson power divider of Fig. 1.2(a) [2].

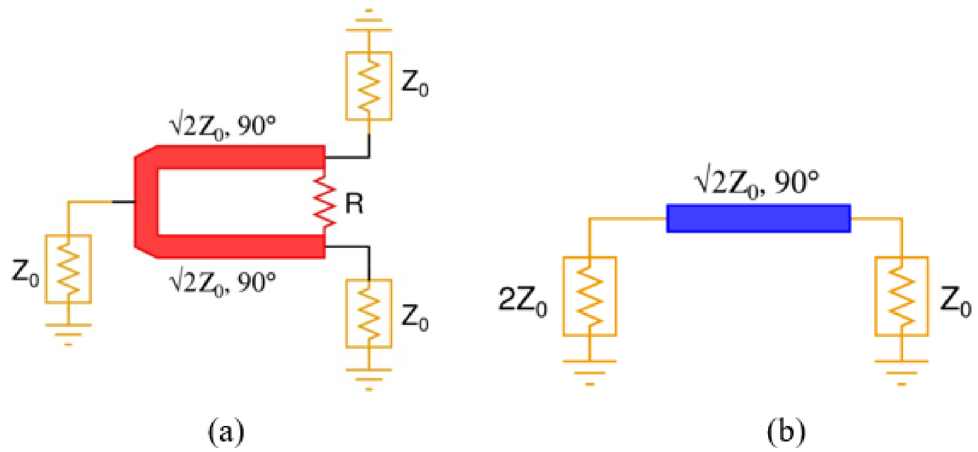


Figure 1.2 (a) A Wilkinson power divider and its (b) even-mode half circuit.

Antenna Feed Line: A patch antenna is shown in Fig. 1.4. The feed lines can be attached as shown in figure [7]. As one moves away from the center of the antenna, the impedance looking into the antenna also varies. Thus, to couple maximum power from the power amplifier (PA, serving as a source to the antenna with Z_0 as the source impedance) to the antenna, the impedance of PA and the antenna must be matched. Moreover, if the impedance level at the patch edge is matched to Z_0, it results into very thin feed lines. Therefore, an inset feed is often used.

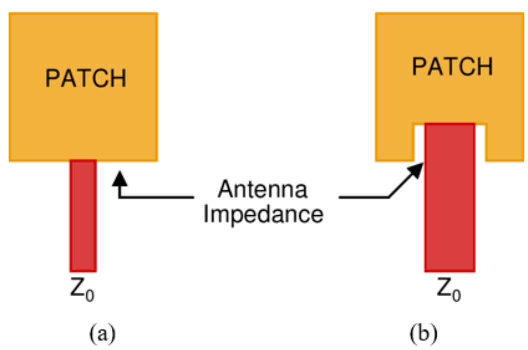


Figure 1.4 Patch antenna with feed lines. (a) edge-feed (b) inset feed

Energy Harvesting: There has been immense interest in RF energy harvesting systems. This essentially entails extraction of energy from cell phone towers or other ambient RF radiations. The motivation is to supply power to the low-energy wireless sensor nodes and wearable electronic devices whose manual maintenance would be very challenging [8-11]. Fig. 1.6 shows a typical block diagram of an RF energy harvesting system. It is obvious that an antenna would capture the ambient RF energy and pass it to a non-linear device, normally a Shottky diode, followed by a filter to suppress the ripples. This part of the system is usually termed as *rectenna*. The load includes a power management module (PMM) that would be required to store the energy. However, a very crucial block is the impedance matching network. Its purpose is to match the complex input impedance of the rectifier to the 50Ω impedance of antenna for maximum power transfer. Design of an energy harvesting system essentially involves design of the rectifier and associated matching networks.

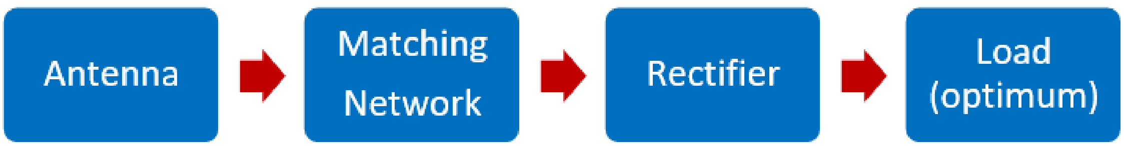


Figure 1.6 Block diagram for RF energy scavenging system.

1.2 Major Contributions and Scope of this Thesis

After an extensive research survey on impedance transformation circuits, presented in the coming chapters, it was identified that the existing state-of-art has severe limitations. It is, therefore, extremely important to investigate and propose novel solutions to address those limitations in the existing design approaches. This thesis has been pursued to contribute into this direction and to address some of the concerns. The scope, objectives, and the contributions of this thesis are briefly outlined below.

1.3 Scope of this work

With regard to the scope and limitations of this thesis, few important points must be borne in mind. Firstly, the impedance matching networks presented in this thesis are intended for establishing matching at fundamental frequencies. Passive components, such as the Wilkinson power divider, shown earlier in Fig. 1.2, or the branch-line coupler, shown earlier in Fig. 1.3, require matching at fundamental frequencies. The power amplifiers, especially the high efficiency PAs, however, require proper matching at harmonic frequencies as well. Fortunately, as reported in [14], the fundamental matching network could always be designed independent of the required harmonic terminations and, therefore, the techniques discussed in this thesis would also be of significant interests to the PA designers.

Secondly, the bandwidths of matching networks are not emphasized as is customary in the literature of multi-frequency matching network. The reason is that the notion of bandwidth is a vague idea when the loads become frequency dependent. And, more importantly, this figure of merit is not that crucial in multi-frequency design as it is in wide-band design. If the same bandwidth performance is sought from a multi-frequency circuit as it is mandated in wideband design then there is no need to go for multi- frequency design in the first place. It is perhaps this reason that the reported multi- frequency designs have as low as 30MHz bandwidth [15-17].
Finally, it must be noted as well that the impedance matching network prototypes shown in this thesis are essentially one-port devices terminated into the considered loads. Therefore, only the measured return-losses, that is, reflection in terms of S_{11} or Gamma have been depicted for them. It should also be noted that insertion loss is direct function of the return loss assuming that no radiation/conductive/dielectric losses exist in structure. These losses are often negligible at low RF frequencies and for transmission line-based structures built properly in low loss dielectric substrate.

1.4 Outline of the thesis

Chapter 1 is the background of the thesis. It also contains the application of matching networks in various areas. In this section scope, objective and outline of this thesis is also discussed. **Chapter 2** contains analysis of different topology of impedance matching transformers for single band. In this chapter all the topologies compared to each other on the basis of some parameters. It also contains the ADS simulations of those technologies. **Chapter 3** discuss about theoretical and ADS simulation analysis of dual-band quarter wavelength transmission lines. A brief of overview is provided in this section along with simulation results. It delves upon the use of dual-frequency quarter wavelength transmission line blocks in dual-frequency impedance matching networks. It is demonstrated that making use of multitudes of the existing dual-frequency quarter wavelength blocks in the literature comes very handy in the matching network design.

In **Chapter 4**, use of coupled lines to create DC isolated matching networks are discussed. A clear analytical approach to design dual-frequency matching network that caters to a wider range of load impedances are presented for real as well as frequency dependent complex impedances.
Also, Tri-band matching network are discussed in small detail as part of future work. Afterwards, Summary and conclusion is presented in the end that highlights the achievements of this thesis and some future possibilities of research in this area.

2. INVESTIGATION AND ANALYSIS OF DIFFERENT IMPEDANCE MATCHING TECHINQUES

2.1 Lumped Components Impedance Matching Network

The main challenge of LNA design is to extend the perfect impedance match over required bandwidth to get a low noise figure and moderate gain as well as the other characteristics. we begin with the simplest solution: An L-network, consisting of a single capacitor and a single inductor. On smith chart, the series capacitance will move the load anti clockwise along the constant resistance circle while the series inductance will move the load clockwise along the constant resistance circle. The shunt capacitance shifts the load anti-clockwise along the constant conductance circle while the shunt inductance shifts the load clockwise along the constant conductance circle. The constant conductance and the constant reactance are orthogonal. So a very precise combination of capacitance and inductance can move any load into the centre of the smith chart. In total eight combinations of L matching networks exist; one of them can be specifically matched to 50 Ω impedance. Some considerations must be taken into account while selecting any of L matching networks to match input impedance. It depends upon the values of the components and the ease of applying DC bias. For instance, the network 4 topology can be selected for a transistor matching network because DC bias can be applied at the grounded end of the inductor and the DC block is doubled through the series capacitor.

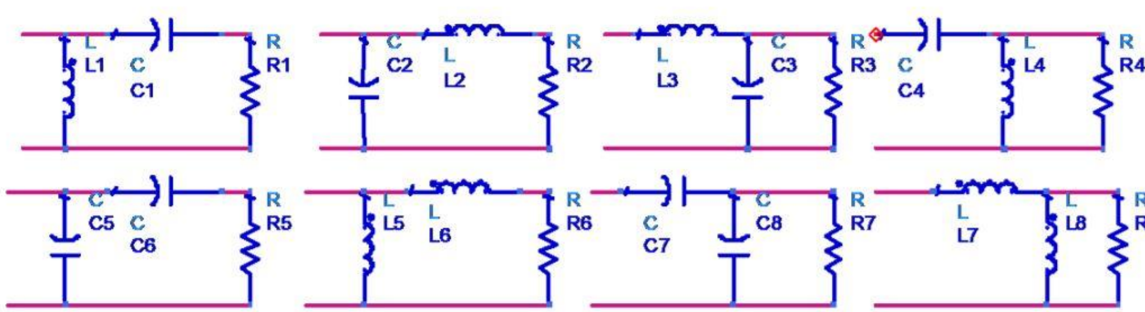


Figure 2.1: L-Type Possible configurations

L-networks have two major drawbacks:

1. They are narrow-band.

2. Capacitors and inductors are difficult to make at microwave frequencies.

2.1.1 ADS simulation

Load Impedance = 50-j*40-ohm, Source Impedance = 25-j*30-ohm, Frequency = 2 GHz.

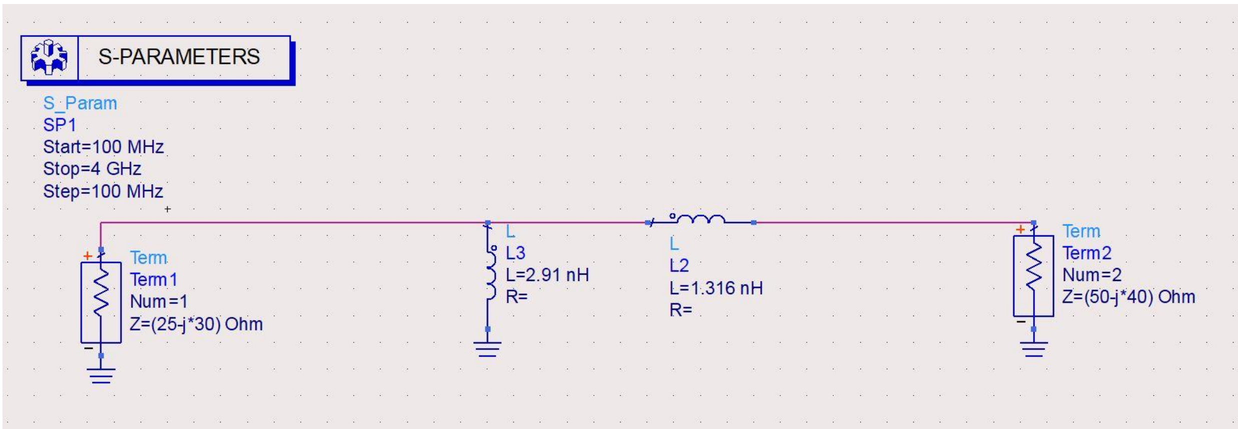


Figure 2.2: L-Type Matching Circuit

2.1.2 Simulation Results

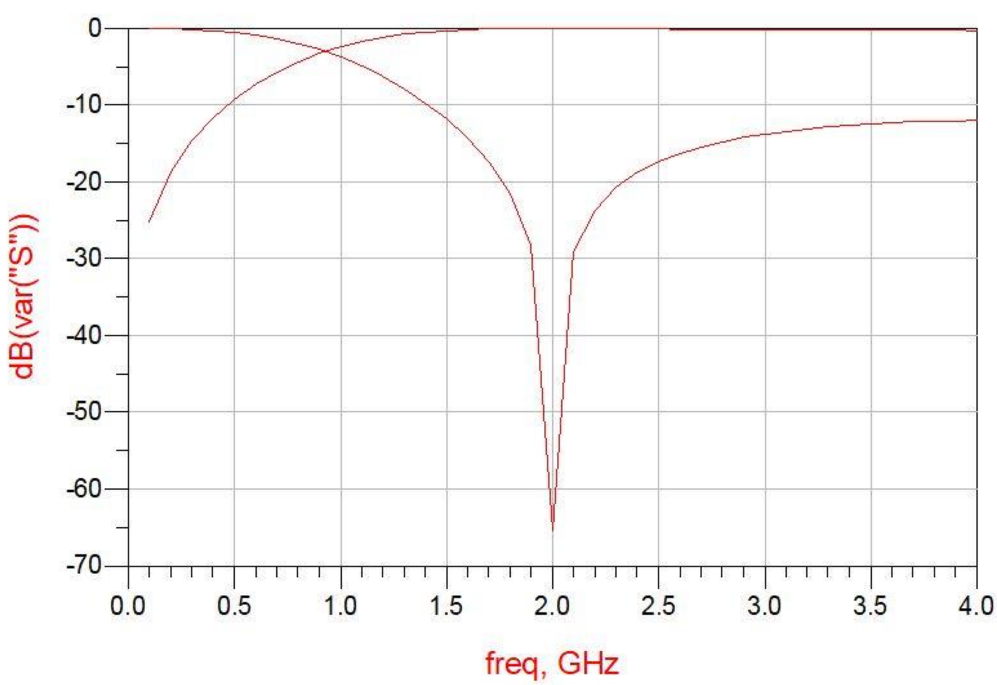


Figure 2.3: Simulation Results

Impedance matched at the required frequency.

2.2 Quarter wave Impedance Transformer

A quarter-wave transformer is a simple impedance transformer which is commonly used in impedance matching in order to minimize the energy which is reflected when a transmission line is connected to a load. The quarter-wave transformer uses a transmission line with different characteristic impedance and with a length of one-quarter of the guided-wavelength to match a line to a load. The characteristic impedance of the quarter wave line is the geometric average of Z0 and RL. Therefore, quarter wave line with characteristic impedance Z1 = $(Z0*RL)^{1/2}$ will match a transmission line with characteristic impedance Z0 to a resistive load RL. An additional feature of the quarter-wave transformer is that it can be extended to multi section designs in a methodical manner, for broader bandwidth. The Drawbacks of quarter-wave Transformer is that matching bandwidth is very narrow. In other words, we obtain a perfect match at precisely the frequency where the length of the matching transmission line is a quarter-wavelength. Also, It can be used only for real impedances loads. In quarter-0wave transformer, we would need to cut the TL to insert it.

2.2.1 ADS Simulation

We will attempt to match a real load of $R = 20\Omega$ to a transmission line with a 50Ω characteristic impedance at a frequency of 6.0 GHz.

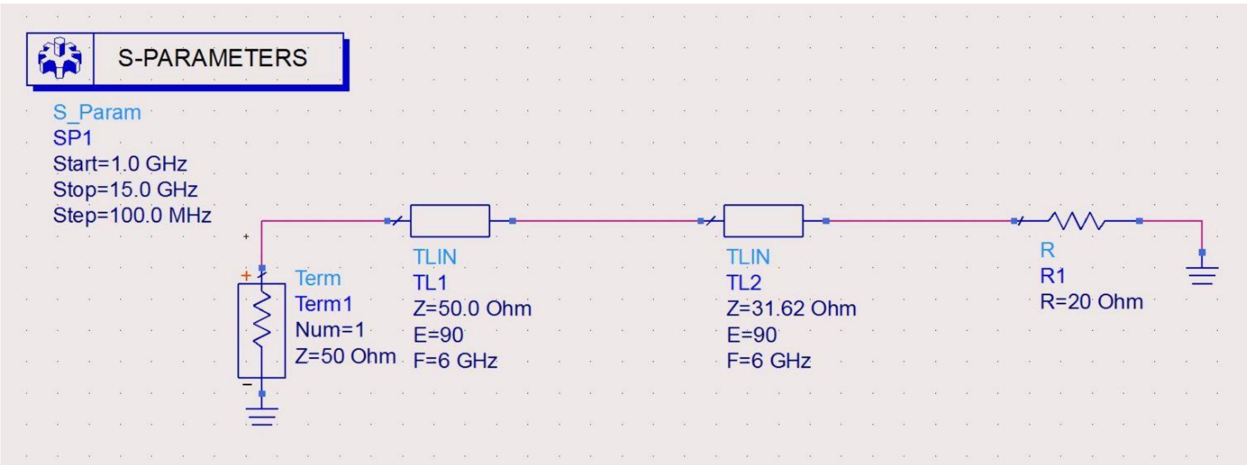


Figure 2.4: Quarter wave matching

Ideal transmission lines converted into microstrip lines using Line Calc in ADS.

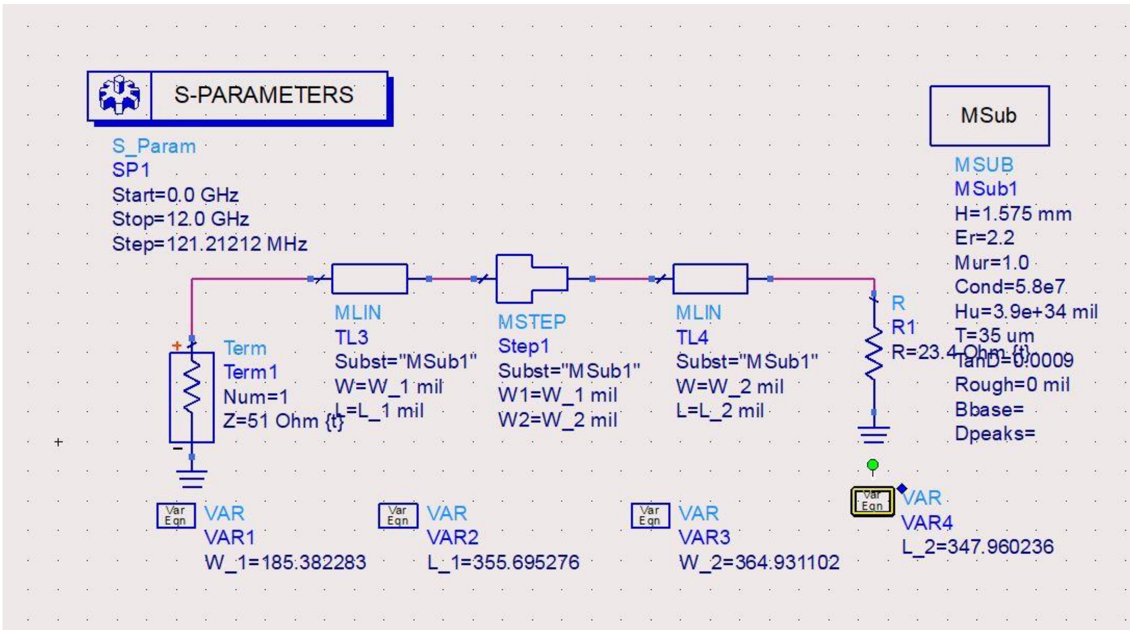


Figure 2.5: Quarter wave Matching using Microstrip Lines

2.2.2 Simulation Results

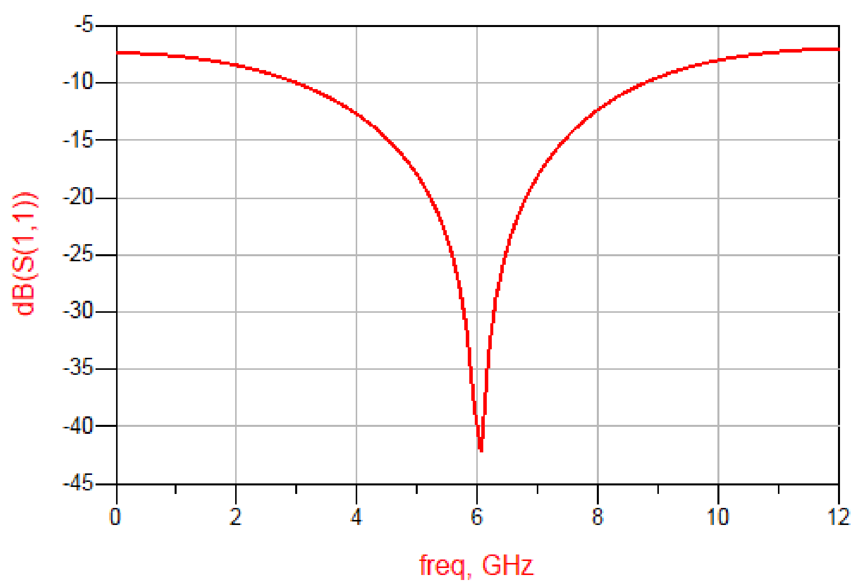


Figure 2.6: Simulation Results

Impedance matched at the required frequency.

2.3 Multi section Binomial Impedance Matching Transformer

The passband response 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. Thus, such a transformer is also known as maximally flat. This type of response is designed, 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 if we let

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

Then the magnitude $|\Gamma(\theta)|$ is

$$|\Gamma(\theta)| = |A||e^{-j\theta}|^N\,|e^{j\theta} + e^{-j\theta}|^N = 2^N|A||\cos\theta|^N$$

2.3.1 ADS Simulation

We will attempt to match a real load of $R = 20\Omega$ to a transmission line with a 50Ω characteristic impedance at a frequency of 6.0 GHz. With given that the bandwidth of the 4-section transformers is defined by $\Gamma m = 0.1$. Assuming TEM wave propagation in the transmission lines, and the transmission line dielectric constant is εr = 9.0.

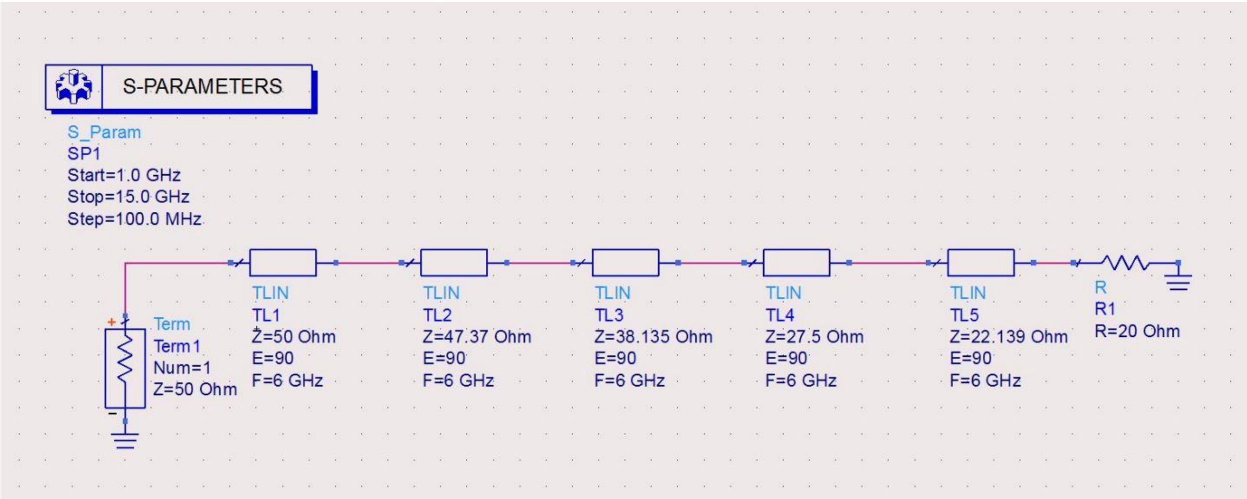


Figure 2.7: Binomial Matching

Transmission lines converted into microstrip lines using Line Calc in ADS.

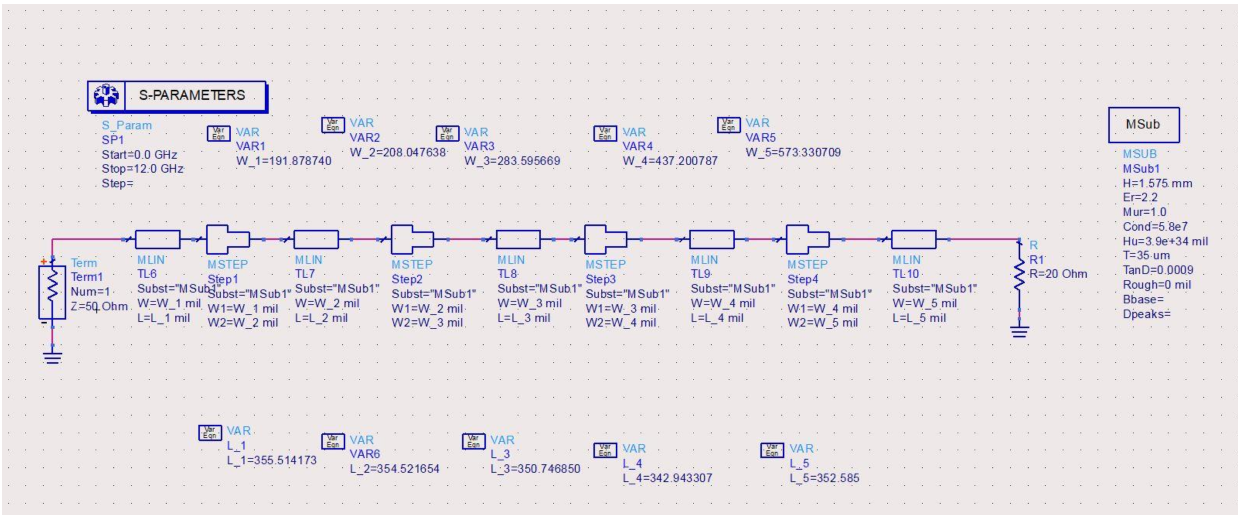


Figure 2.8: Binomial Matching using Microstrip Lines

2.3.2 Simulation Results

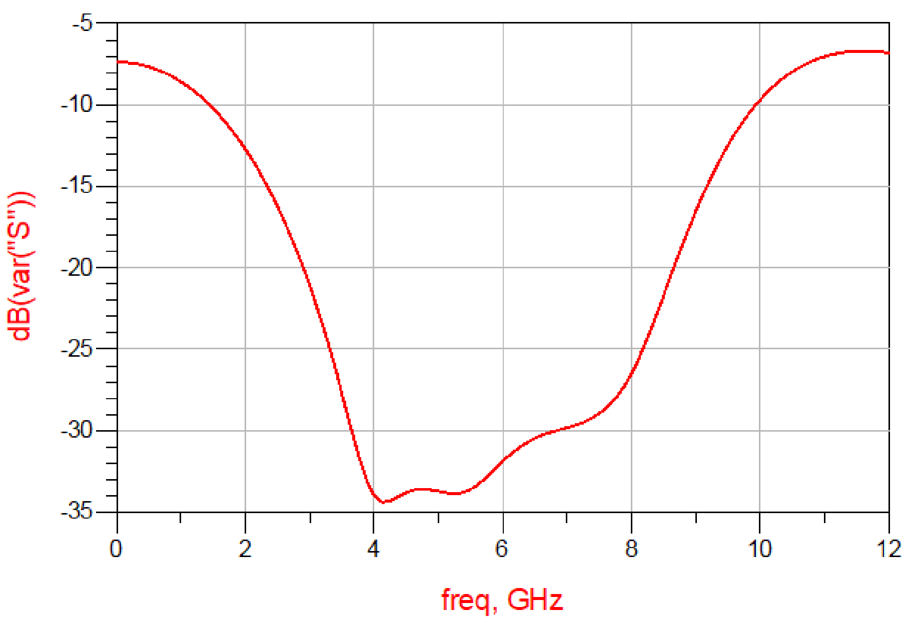


Figure 2.9: Simulation Results

Impedance matched at the required frequency.

2.4 Multi section Chebyshev's Impedance Matching Transformer

In contrast with the binomial matching transformer, the Chebyshev transformer optimizes bandwidth at the expense of passband ripple. If such a passband characteristic can be tolerated, the bandwidth of the Chebyshev transformer will be substantially better than that of the binomial transformer, for a given number of sections. The Chebyshev transformer is designed by equating $\Gamma(\theta)$ to a Chebyshev polynomial, which has the optimum characteristics needed for this type of transformer.

2.4.1 ADS Simulation

We will attempt to match a real load of $R = 20\Omega$ to a transmission line with a 50Ω characteristic impedance at a frequency of 6.0 GHz. With given that the bandwidth of the 4-section transformers is defined by $\Gamma m = 0.1$. Assuming TEM wave propagation in the transmission lines, and the transmission line dielectric constant is $\varepsilon r = 9.0$.

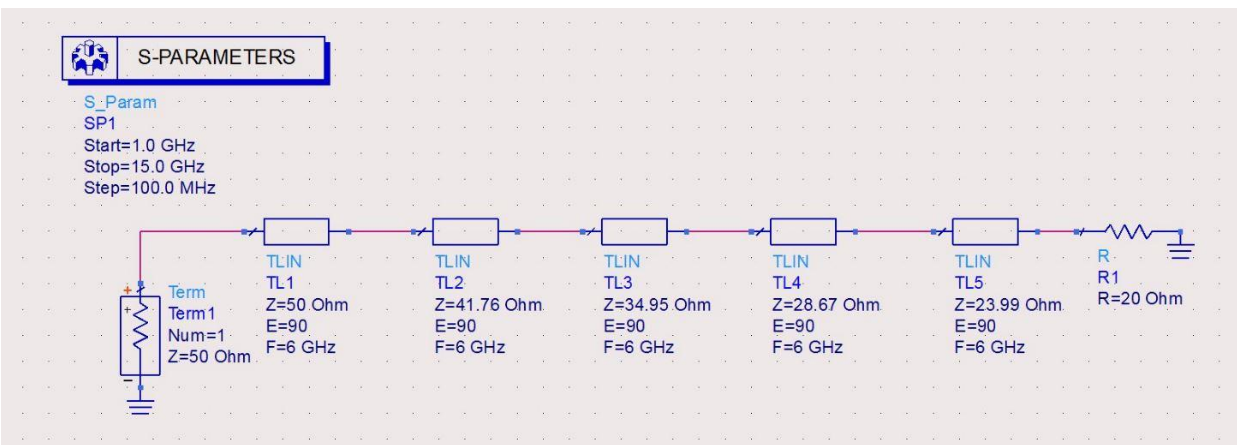


Figure 2.10: Chebyshev Matching

Transmission lines converted into microstrip lines using Line Calc in ADS.

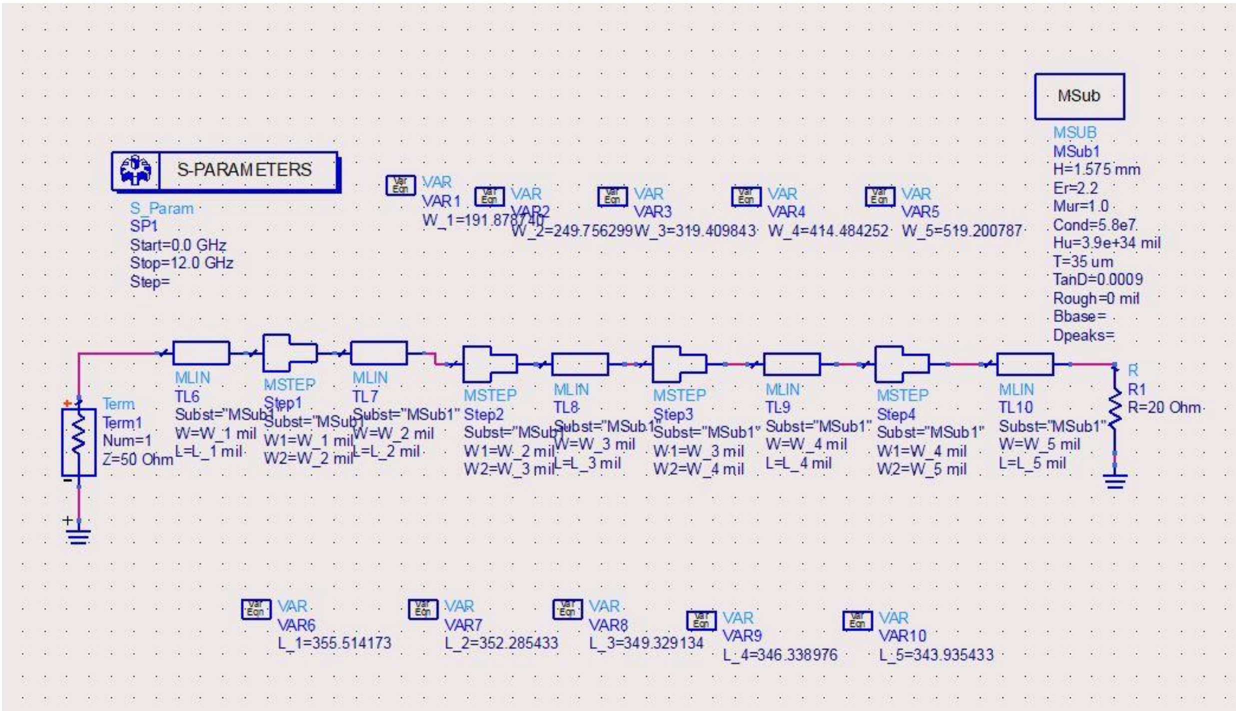


Figure 2.11: Chebyshev Matching using Microstrip Lines

2.4.2 Simulation Results

Figure 2.12: Simulation Results

Impedance matched at the required frequency.

3. Dual Band Quarter Wavelength Transmission Lines

Dual-band microstrip transmission lines have many aspects in modern communication systems. It is the key element to build resonators, band-pass or band-stop filters in terms of compact and durable size. There are several papers which discuss dual-band microstrip transmission lines. Many authors proposed conventional branch line coupler with short circuited stubs and open stubs pi shaped transmission lines or by employing T-shaped quarter wavelength microstrip transmission lines. Both use the concept of ABCD matrix to evaluate the impedances and electrical lengths of the transmission line and stubs. In this chapter, the design of the proposed dual-band transmission lines will consist of open and short-circuited stubs at both ends of the transmission line. Specifically, these stubs are asymmetrical with the transmission line and located at both ends of the transmission line. In the design, transmission line is connected with two short-circuited stubs and two open stubs. In each side both of the short and open stubs have similar electrical length and impedance. These stubs can be also treated as the equivalence of discrete capacitors and inductors for the resonant circuits at microwave frequencies. The advantage of the design is the flexibility of choosing the transmission lines and the stubs in asymmetric shapes.

3.1 Theoretical Analysis

The design consists of two open subs, two short-circuited stubs along with one transmission line. The design is aimed for realizing a quarter wavelength transmission line with an impedance of 50Ω. One of the open stubs impedance and electrical length are equal to one of the short-circuited stubs impedance and electrical length and both of the stubs are allocated at the same side of the transmission line. Similarly, the rest of the open and shorted stubs also have the same impedance and electrical length different from the previous stubs and are connected to the opposite side of the transmission line. The impedance and electrical length of the transmission line itself are different from all of the stubs. So, the resulting design becomes an asymmetric transmission line. The described design has been shown in Fig.3.1.

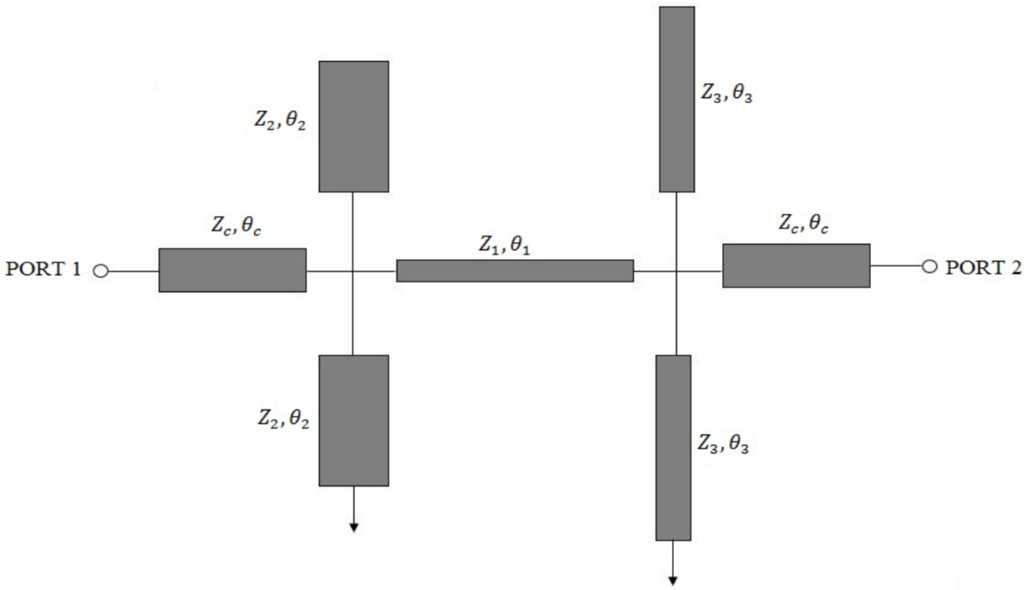


Figure 3.1: Structure of the proposed asymmetric open and short-circuited stubs loaded microstrip transmission line

In Fig.3.1, the left side of the transmission line has two stubs with impedance Z2 and ϴ2. The transmission line has an impedance of Z1 and an electrical length value of ϴ1, respectively. Finally, the right side of the transmission line has two stubs with impedance value Z3 and electrical length ϴ3. Now to obtain the design equations the ABCD matrix method has been derived in the following manner considering for each of the stubs and the transmission line.

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ \frac{-j}{Z_2}\cot\theta_2 & 1 \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ \frac{j}{Z_2}\tan\theta_2 & 1 \end{bmatrix} \times \begin{bmatrix} \cos\theta_1 & jZ_1\sin\theta_1 \\ \frac{j}{Z_1}\sin\theta_1 & \cos\theta_1 \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ \frac{-j}{Z_3}\cot\theta_3 & 1 \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ \frac{j}{Z_3}\tan\theta_3 & 1 \end{bmatrix}$$

each element of the ABCD matrix is derived to be,

$$A = \tan 2\theta_3 + \frac{2Z_1}{Z_3}\tan\theta_1$$

$$B = jZ_1\sin\theta_1$$

$$C = j\left[\frac{\cos\theta_1}{Z_2}(\tan\theta_2 - \cot\theta_2) - \frac{Z_1}{Z_2 Z_3}\sin\theta_1(\tan\theta_2 - \cot\theta_2)(\tan\theta_3 - \cot\theta_3) + \frac{\cos\theta_1}{Z_3}(\tan\theta_3 - \cot\theta_3) + \frac{\sin\theta_1}{Z_1}\right]$$

$$D = \cot\theta_1 + \frac{2Z_1}{Z_2}\cot 2\theta_2$$

By assuming the proposed quarter wavelength transmission line's characteristic impedance Zc is equal to 50Ω, the ABCD matrix should be equal to:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 0 & jZ_c \\ \frac{j}{Z_c} & 0 \end{bmatrix}$$

Through the equivalency of the matrix element, we obtain for the dual-band operation at the low frequency and high frequency the electrical length of the transmission line can be implied by,

$$Z_1 \sin\theta_{1f1} = \pm Z_c$$

$$Z_1 \sin\theta_{1f2} = \pm Z_c$$

where Θ1f1 and Θ1f2 are the electrical lengths of the lines at the two-desired operating frequencies. The relation between low and high frequency electrical lengths is as,

$$\theta_{1f2} = n\pi + \theta_{1f1} \qquad n = 2$$

From the relation of dual-band frequency ratio with respect to electrical lengths

$$\frac{\theta_{1f1}}{\theta_{1f2}} = \frac{f_1}{f_2}$$

The low frequency electrical length can be calculated in the following way,

$$\theta_{1f1} = \frac{f_1}{f_2 - f_1} n_2 \pi \qquad at\ n = 2$$

where f1 and f2 are low and high frequency respectively for the dual-band operation. The impedance values of the transmission line and the stubs can be calculated through the following equations.

$$Z_1 = \frac{Z_c}{\sin\theta_1}$$

$$Z_2 = -2Z_1 \frac{\cot 2\theta_2}{\cot\theta_1}$$

$$Z_3 = -2Z_1 \frac{\tan\theta_1}{\tan 2\theta_3}$$

Each of the impedance value has been considered within the range between 20Ω to 120Ω and the electrical lengths are between 15° to 360°. The realizable frequency ratio for the design can be increased by increasing the value of n.

3.2 Simulations

The desired values of impedances for stubs and transmission line have been calculated at different frequency ratios. With the calculated design parameters circuit simulations have been done in ADS.

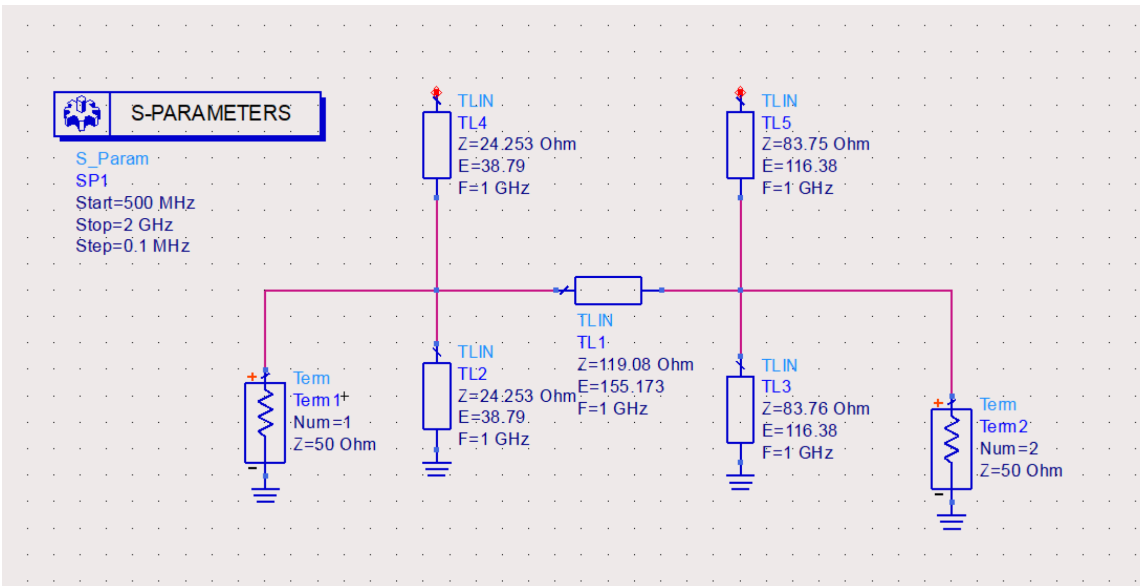


Figure 3.2: ADS Simulation Circuit

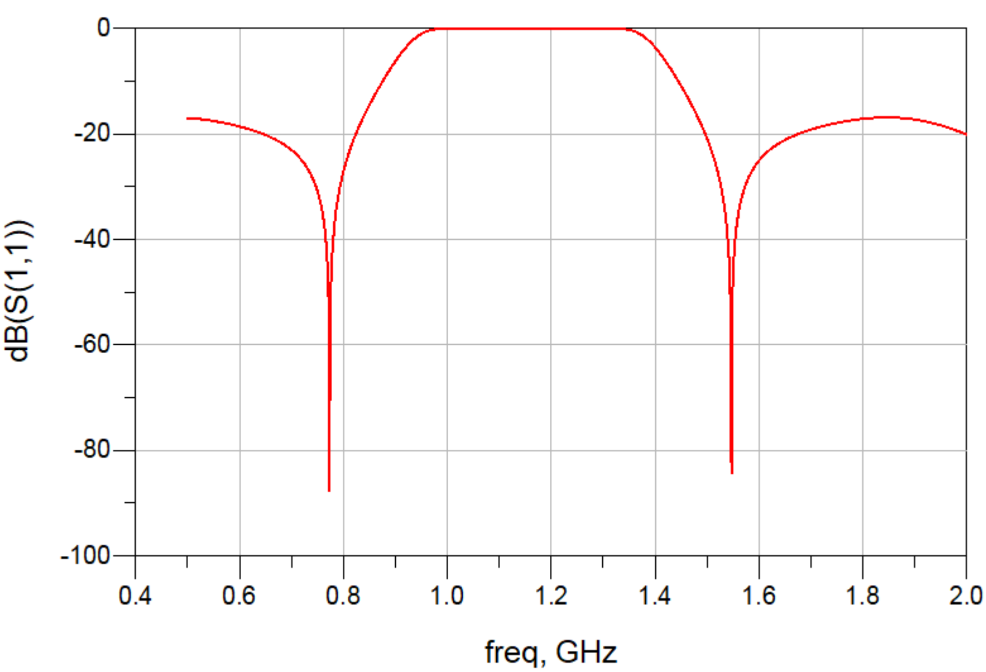


Figure 3.3: S-Parameters Simulation Results

4. Dual Band Impedance Matching Network Using Quarter Wavelength Transmission Lines Blocks

Impedance matching based on quarter wavelength transmissions has been studied extensively in the context of single frequency or wideband matching problem [2]. In addition, most of these reports have focused on matching real load impedances to real source impedances. A quarter wavelength block consists of several distributed/lumped elements and its electrical parameters, e.g. ABCD, are equivalent to that of a quarter wavelength transmission line at a frequency. As per requirement of multi-frequency applications, a large number of dual-frequency quarter wave-length blocks have been reported as well [32-34]. It is evident that the current state-of-the-art in dual-frequency component design is replete with applications of numerous types of dual-frequency quarter wavelength blocks. A pertinent question then arises that if a dual-frequency quarter wavelength transmission block could be used for impedance matching purposes, that too for frequency dependent complex loads (FDCLs). Another important aspect of this discussion is that a quarter wavelength transmission line is a symmetric network. What happens to this symmetry when dual-frequency quarter wavelength blocks are synthesized from seemingly asymmetric structures? Moreover, while requirement of bandwidth is not that critical in multi-frequency applications but it would be instructive to get some intuitive feeling about this. This chapter of the thesis aims at providing answer to these important questions.

4.1 Proposed Matching Network

A conventional quarter wavelength transmission line is used for matching real load impedances to real source impedances. The FDCL is assumed to have values as $Z_L=R_1+jX_1$ @f_1 and $Z_L=R_2+jX_2$ @f_2. Once again Y_{in1} is the admittance looking into section A, Y_{in2} is the admittance looking into section B, and Z_{in3} is the impedance looking into the combination of sections A and B.

It is apparent from Fig. 4.1 that the transformer A converts the FDCL to Y_{in1} such that, $Y_{in1}|_{f1} = Y_{in1}|_{f2}$ that is, $Y_{in1}|_{f1}$ = G - jB if then $Y_{in1}|_{f2}$ = G + jB. The Section B, which is primarily a dual-frequency susceptance, cancels out the imaginary parts of Y_{in1} at the two specified frequencies. Thus, as a last step there is a need to match $Z_{in3} = 1/G$ at the two frequencies. It is well known that a real impedance, such as Z_{in3}, can be matched to another real impedance, such as Z_0, using a λ/4 line; the only additional requirement in the present case is to have a dual-frequency λ/4 line, which is what the section C essentially is.

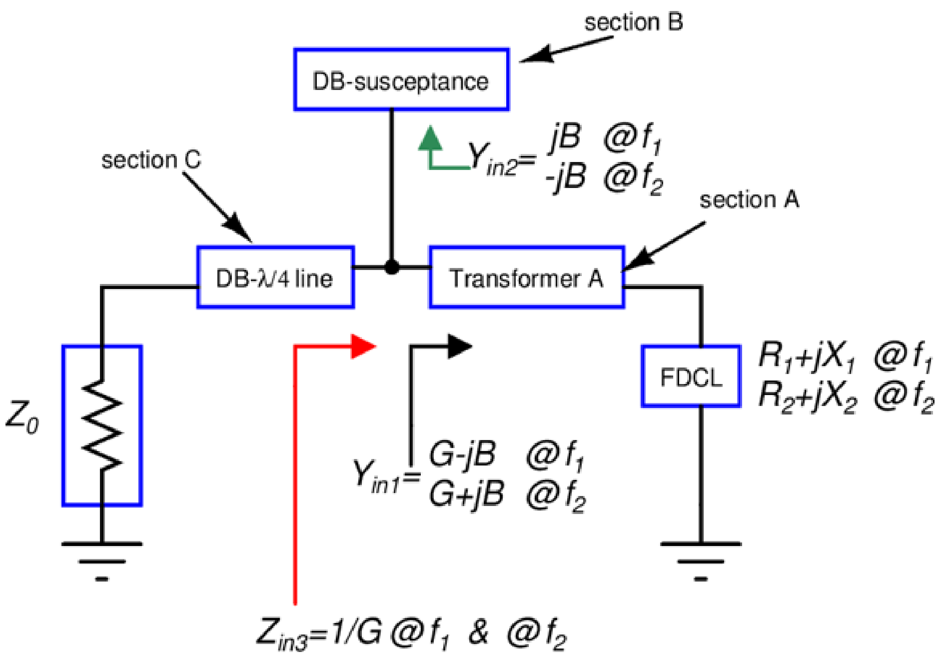


Figure 4.1: Pictorial illustration of the proposed matching scheme (DB: Dual-Band)

It must be noted at this juncture that most of the previously reported schemes have used sections A and B in the same manner as being used here. The main aspect of the proposed scheme is the incorporation of section C, the dual-frequency λ/4 line, into the conventional structure. The strength and uniqueness of the proposed impedance network, therefore, is the flexibility of choosing any dual-frequency λ/4-line from a variety of those reported in the literature to suit specific applications.

4.2 Implementation of proposed scheme

One possible implementation of the proposed matching network is shown in Fig 4.2. Section A, adjacent to the FDCL, $Z_L= R_1+jX_1$ at f_1 and R_2+jX_2 at f_2, is a transmission line having characteristic impedance Z_1 and electrical length θ_1. Section B is either a short or open stub having characteristic impedance Z_2 and electrical length θ_2 whereas section C is an L-shaped dual-frequency λ/4-line. The term Z_{Tr_in} is the input impedance that the load sees into the transformer, and we will return to its significance later.

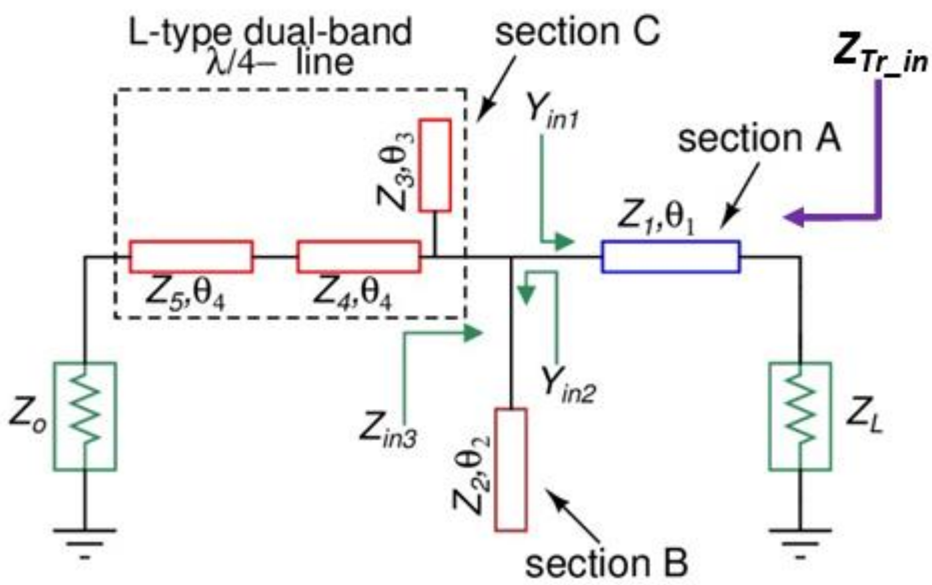


Figure 4.2: Implemented dual-frequency matching network based on the scheme shown in Fig. 4.4.

4.3 Design Process

A. Design of Section 1

The admittance $Yin1$ is given by $Yin1 = Gin1 + Bin1$. All the frequency dependent terms in design equation have squared magnitude i.e. always positive except the numerator. Subsequently, a simplified expression for $\theta 1$ given in (4) can be deduced from the numerator of (3). $\theta 1 = (n + 1)/ (r + 1)\ \pi$, Where, the term r is the frequency ratio $f2/f1$. Furthermore, for dual-band operation, the value of $\theta 1$ needs to satisfy $Yin1|f1 =$ *Conjugate of* $Yin1|f2$.

B. Design of Section 2

First, the term $ZA2$ can be defined as the input impedance of the transmission line having characteristic impedance $Z1$ and electrical length $\theta 2$, i.e. equals to $\theta 1$, $ZA2 = Z1\ (ZL + jZ1tan\theta 1)\ /(\ Z1 + jZLtan\theta 1)$. Then the admittance $Yin2$ can be given by: $Yin2 = 1/\ [Z3(ZL\ 1 + +jZ1\ \tan\theta 1\ )(Z1 + jZLtan\theta 1)]$.Now, this expression can be simplified as $Yin2 = Gin2 + jBin2$. Once again, for dual-band operation, $Yin2|f1 =$ *Conjugate of* $Yin2|f2$. Now, for the matching, $\mathrm{Re}(Yin1) = Re(Yin2)$ at the node $N1$ and this essentially means $Gin1 = Gin2$. The simplifications lead to the design equation for dual band matching network for the odd-mode impedance Zo in terms of free variable ρ, the ratio of even-mode and odd-mode impedances, of the coupled-line.

C. Design of Section 3

This section cancels the net imaginary part during the matching, i.e, $Yin1 = Yin2$, at node $N1$. At this node, $jIm(Yin1 + Yin2) = j(Bin1 + Bin2)$ @ $f1$ and $jIm(Yin1 + Yin2) = -j(Bin1 + Bin2)$ @ $f2$. This cancellation can be achieved using a dual-band short or open-stub at this node. Mathematically, the overall cancellation of imaginary part of $Im(Yin1 + Yin2)$ at node $N1$ can be expressed as: $Im(Yin3) = -j(Bin1 + Bin2) = j(1=Z2)tan\theta 2@f1 = j(Bin1 + Bin2) = j(1=Z2)tan(r\theta 2)@f2$. The expressions can be simplified to determine the values of $Z3$ and $\theta 2$ for dual-band operations. $\theta 2 = (n + 1) / (r + 1)\ \pi$ and $Z3 = -tan\theta 2=(Bin1 + Bin2)$.

D. Design Procedure

For a given load and source impedances of ZL and ZS, the design starts with section 1. The odd-mode characteristic impedance Zo of the coupled-line is determined using above sections. During this stage, ρ, $Z1$, and $Z3$ are considered free variables whereas $\theta 1$ is dependent on the frequency ratio r and is calculated. For example, the values of Zo for $\rho = 3$ and $r = 3$ are depicted in Figs. 4.6 and 4.7. The results in Fig. 4.6 provide the values of Zo for ZL varying from 10 to 40 Ω, whereas Fig. 4.7 shows the range of Zo for ZL varying from 60 to 260 Ω. The curves apparently give the realizable values of odd mode impedance (Zo) for the chosen frequency and transformation ratios. The free variables $Z1$ and $Z3$ can be varied to develop realizable matching between higher values of ZL and the source impedance while also maintaining the realizability of Zo as shown in Fig. 4.7. Now, all the required parameters to design section 1 and section 2 of the impedance matching network is available and therefore the net imaginary impedance at node $N1$ can be computed using above sections. Finally, these parameters are used to design the length and characteristics impedance of the stub (section 3) that is used to nullify this imaginary impedance.

4.4 Case studies and results

Here we have varied the odd-mode impedance of Section 1 against the load impedances to get the realizable range of load impedances.

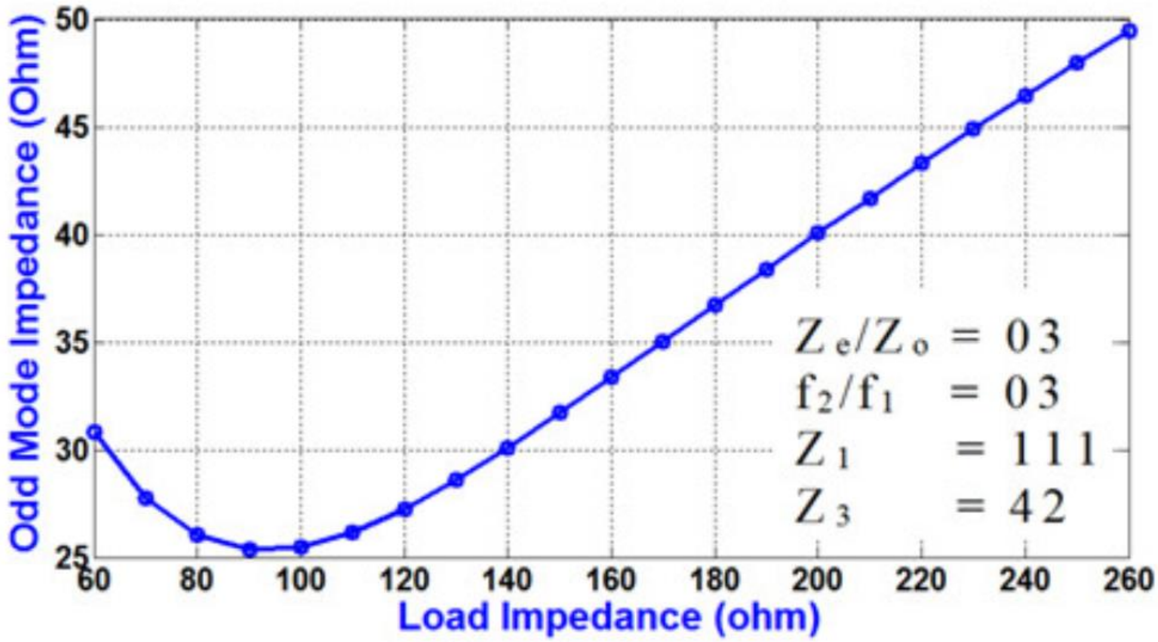


Figure 4.3: Realizable range of load impedances

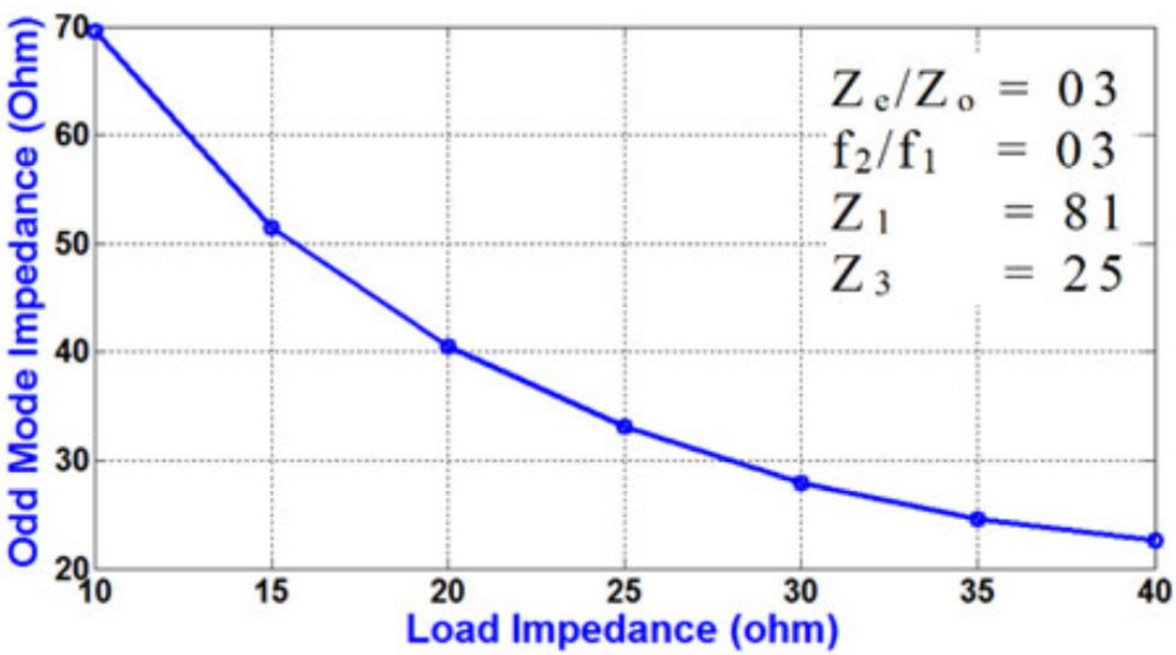


Figure 4.4: Realizable range of load impedances

Table 4.1

Case Studies and Design Example

Case	Frequencies (GHz)	Z_L (Ω)	Section 1		Section 2		Section 3		10dB BW (MHz)	
			Z_e, Z_o (Ω)	Θ1 (deg)	Z_1, Z_2 (Ω)	Θ2 (deg)	Z_3 (Ω)	Θ3 (deg)	@f1	@f2
Case 1	f1 =1 f2 =2	60	Z_e = 133.56 Z_o = 58.12	60	Z1 = 80 Z2 = 80	60	66.61	120	130	130
Case 2	f1 =1 f2 =2	100	Z_e = 155.16 Z_o = 67.52	60	Z1 = 100 Z2 = 100	60	86.6	120	100	100
Case 3	f1 =1 f2 =2	200	Z_e = 57.45 Z_o = 25	60	Z1 = 100 Z2 = 104	60	46.81	120	60	60

4.5 Conclusion

A novel, simple and successful design methodology for realizing a dual-frequency matching network has been presented. It has been demonstrated that any dual- frequency quarter wavelength block could be utilized for realizing a dual-frequency impedance transformation for frequency dependent complex loads (FDCLs). The reported design also enables the use of mirror image of dual-frequency blocks and this can potentially aid in achieving simpler board layout. Finally, an intuitive explanation of relative bandwidth of different matching networks has also been provided.

5. Future Work

5.1 Tri-Band Matching Network

A tri-frequency matching network should be able to provide concurrent impedance matching at three arbitrary frequencies. Such impedance matching network is of very high value in the multi-frequency circuits and systems [32].

However, a detailed survey of multi-frequency impedance matching network presented in Chapter 2 clearly reveals that there is lack of systematic synthesis approach for tri-frequency matching networks. The existing approaches, albeit very few, are limited because of either being a lumped element approach based, coupled lines based, or limited to only real load impedances. In fact, the tri-frequency impedance matching network used in [32] is a brute-force technique.

Therefore, it is of paramount importance that a generic distributed element based technique for systematic synthesis and design to achieve concurrent impedance matching at three arbitrary frequencies is invented.

6. SUMMARY AND CONCULSION

Due to the rapid developments in multi-frequency/multi-standard wireless communication systems during the last decade, the design of RF front-ends has become a very challenging affair while designing wideband components, especially those requiring adequate operation over multi-octave bandwidth has always been considered difficult. However, the alternative architecture employing multi-frequency components is also by no means an easier task. Whether wideband or multi-frequency; the vast majority of RF front-end components employ impedance matching networks as a key component, and therefore their analysis and design is of the paramount importance. To that end, the first important contribution of this thesis work has been use of quarter wave-block was described for the design of dual- frequency impedance transformation networks. The advantage of this technique lies in its flexibility to employ any dual-frequency quarter-wavelength block from a large number of those reported in the literature. Besides, the technique is very simple and even an asymmetric looking block, such as an Pi-network, could be employed with proper orientation to suit during layout phase. A very troublesome scenario in dual-frequency impedance matching network is the limited load of range that could be matched to a real source impedance. Since, a circuit designer has no control over the input/output impedances of a transistor or other active device at given set of frequencies, the existing dual-frequency matching networks may not yield a physically realizable design. If any of the constituting line impedance is beyond a specific range, there is no other way except to change the matching network topology or to use multiple stages of matching networks. Coupled lines are very important for matching network design as they facilitate miniaturization. Therefore, future was devoted to novel dual-frequency matching networks employing parallel coupled lines. Due to the high potential of this work, here we propose several future works to be done.

The major focus of this work has been to mitigate the limited frequency- and transformation ratios of dual-frequency networks by advanced network topologies. In addition, the tri-frequency impedance matching network puts forward a scheme to establish matching at three arbitrary frequencies concurrently. The same technique could be extended to multi-frequency scenarios.

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