KADİR HAS UNIVERSITY
GRADUATE SCHOOL OF SCIENCE AND ENGINEERING
PROGRAM OF ELECTRONICS ENGINEERING

REAL FREQUENCY DESIGN OF NARROWBAND IMPEDANCE EQUALIZER WITH COMPLEX TERMINATIONS

GÖKMEN YEŞİLYURT

MASTER’S THESIS

ISTANBUL, JUNE, 2018

GÖKMEN YEŞİLYURT M.S. Thesis 2018

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Submitted to the Graduate School of Science and Engineering of Kadir Has University in partial fulfillment of the requirements for the degree of Master's in the Program of Electronics Engineering

ISTANBUL, JUNE, 2018

DECLARATION OF RESEARCH ETHICS /
METHODS OF DISSEMINATION

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ACCEPTANCE AND APPROVAL

This work entitled **REAL FREQUENCY DESIGN OF NARROWBAND IMPEDANCE EQUALIZER WITH COMPLEX TERMINATIONS** prepared by **GÖKMEN YEŞİLYURT** has been judged to be successful at the defense exam held on **08 June 2018** and accepted by our jury as **MASTER'S THESIS**.

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DATE OF APPROVAL:
8.06.2018

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REAL FREQUENCY DESIGN OF NARROWBAND IMPEDANCE EQUALIZER WITH COMPLEX TERMINATIONS

ABSTRACT

In this thesis, the real frequency design of L, Pi and T type networks with complex terminations has been analyzed. The main purpose of the proposed approach is to compare the existing methods against the proposed method on impedance matching design in narrowband with complex termination.

There are two types of models, the serial model for T type network design and the parallel model for Pi and L type networks design. The generator and load termination impedances are given as measured values. Thus, they can be considered as a resistance and a reactive element connected in series. To be able to design a Pi type narrowband matching network, these series impedance models must be transformed into parallel models. But for T type network designs, the assumed serial models can be used, there is no need for a transformation process. Then, the corresponding Pi or T type networks can be designed by using Q based method that is well-defined in the literature.

The proposed approach is a new method for designing narrowband impedance matching networks. The purpose of this new method is to configure impedance matching networks for all types L, Pi, and T without the need for termination models with component values. For comparison with existing methods, the complex terminations part and the Q based method equations have been derived, and a new Q based method (modified) is considered for the new method.

The proposed approach is illustrated by different examples of network types L, Pi and T.

Keywords: Narrowband networks, lossless networks, L type networks, Pi type networks, T type networks

KARMAŞIK SONLANDIRMALI DARBANT EMPEDANS UYUMLAŞTIRMA DEVRELERİNİN GERÇEK FREKANSLI TASARIMI

ÖZET

Bu tezde, karmaşık sonlandırma L, Pi ve T tipi uyumlaştırma devrelerinin gerçek frekanslı tasarımı analiz edilmiştir. Önerilen yaklaşımın temel amacı, mevcut olan yöntemlerin, karmaşık sonlandırma dar bantlı empedans uyumlaştırma devrelerinin tasarımı üzerine önerilen metoda göre karşılaştırılmasıdır.

İki tip model vardır: T tipi devre tasarımı için seri model ve Pi ve L tipi devre tasarımı için paralel model. Kaynak ve yük sonlandırma empedansları ölçüm değerleri olarak verilir, böylece, seri olarak bağlanmış bir direnç ve reaktif eleman olarak kabul edilebilirler. Bir Pi tipi devre tasarlayabilmek için bu seri empedans modelleri paralel modellere dönüştürülmelidir. Fakat T tipi devre tasarımları için, varsayılan seri modelleri kullanılabilir, bir dönüşüm sürecine gerek yoktur. Daha sonra, ilgili devre Pi veya T tipi, literatürde iyi tanımlanmış kalite faktörüne (Q) dayalı bir yöntemle tasarlanabilir.

Önerilen yaklaşım, dar bantlı empedans uyumlaştırma devrelerini tasarlamak için yeni bir yöntemdir. Bu yeni yöntemin amacı, devre elemanların değerleri olan sonlandırma modellerine gerek duymadan, tüm L, Pi ve T tipleri için dar bantlı empedans uyumlaştırma devrelerini yapılandırmaktır. Mevcut yöntemlerle karşılaştırmak için, karmaşık sonlandırma kısmı ve Q tabanlı yöntem denklemleri türetilmiştir ve yeni yöntem için yeni bir Q-tabanlı yöntem (değiştirilmiş) geliştirilmiştir.

Bu tez çalışmasında, önerilen yaklaşımlar L tipi, Pi tipi ve T tipi olarak farklı örneklerle açıklanmıştır.

Anahtar Sözcükler: Dar bant devreler, Kayıpsız devreler, L tipi devreler, Pi tipi devreler, T tipi devreler

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For a person who had many duties in their professional life, he needs a proactive supervisor, for directing him every time to the right way, and find quick solutions for to face troubles. He needs a patience, and he needs help for success in his academic life.

And for all these reasons, I'm lucky to have had the opportunity to work with my supervisor Assoc. Prof. Metin Şengül.

And following his advice, I'm motivated to continue my academic career in a foreign country to reach my doctorate.

Thanks Professor.

DEDICATION

To my wonderful family

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1. INTRODUCTION

The chronology of the history of this problem in the literature, the first theory submitted, the improvements, the important people who have worked on this area is analyzed, and a small summary about it was shared as below.

The theory of narrowband and broadband matching can be considered to have begun after the development of a bandwidth theory for a narrowband load impedance consisting of the parallel combination of a capacitor and a resistance by BODE (Bode, 1957), FANO (Fano, 1950) and YOULA (Youla, 1964) generalized the analytical theory of the gain bandwidth for arbitrary loads.

In the near past, CARLIN opened a new page in the literature with a new approach, the real frequency technique, eliminates most of the problems associated with conventional analytical methods and therefore renewed and active research efforts in the field of broadband and narrowband (Carlin, 1977). The most difficult aspect of this approach was the direct use of experimental real frequency load data in the design process. This new technique has made it very attractive because, in almost all microwave matching problems, the device to be adapted usually describes numerical values obtained from experimental measurements. On the other hand, the method has been qualitatively demonstrated to provide superior design performance with a simpler construction compared to other existing techniques. As a result, this approach has set a new trend in matching problem studies and has encouraged many researchers to develop various variants of the real-world frequency method.

In the present technics, the entire available analytical and real frequency design procedure concerns either only lumped circuit elements or distributed transmission lines alone in the matching network design. Virtually, the physical realization of ideal lumped and distributed network elements, especially in discrete, hybrid or monolithic microwave integrated circuit designs, presents serious application problems.

After analyzing the history of the impedance matching problem, another in-depth approach to the goal is shared as below.

Impedance matching problems are one of the fundamental problems in the design and development of communication systems, and in radio frequency circuits. For these circuits the main objective is to transfer the maximum possible power between the generator and its loads. One of the important examples on this subject is about all sensitive receivers, the maximum power transfer is decisive, no doubt, this type of circuit that carries really low signal levels, is not able to tolerate this problem. For this reason, in most cases, for the initial design extreme care must be taken to ensure that each element of the chain is suitable for its load. For the solution of this problem, naturally it is necessary to involve a design of an equalization network, called matching network, to transfer a given impedance into another specified impedance, in the set called impedance matching or equalization.

In this context, although there have been valuable contributions to the characterization of some restricted classes of mixed-element structures, a complete theory for the problems of approximation and synthesis of mixed-distributed networks is always not available. Thus, the problem is quite difficult from a theoretical as well as a practical point of view.

The stages of the thesis are submitted with their little resumes as below,

The second part, the problems of broadband matching are studied briefly and illustrated with the simple example to make the difference with the narrowband.

The third part of this thesis is devoted to the general approach in the theory of narrowband matching. The problem of narrowband matching is explained with existing approaches in the literature.

The fourth part, some knowledge about Smith Chart for the narrowband solution is briefly reviewed, to understand the more effective solution technique against the equations and the basic Q of the literature.

The fifth part, the proposed approach has been submitted with fully developed illustrated examples and also trying to demonstrate the purpose of this approach, that it is not necessary to have termination models with component values, more to the proposed approach it suffices simply to have termination impedance measurement values.

The last part, the conclusion of the proposed approach was submitted.

In the appendix, the publication, the Matlab code to design L, Pi, and T types matching networks, as well as the schematic view of the examples of circuits types L, Pi and T created on the software have been shared.

2. BROADBAND MATCHING

2.1 Broadband Matching Problem

In general, the main purpose of the matching in this subject is to design a lossless two-port circuit between the generator and the complex load impedance, such that the power transfer of the generator to the load is maximized in a band prescribed frequency. To define the ratio between the power given to the load and the usable power of the generator, the best reference is to take into account the capacity of power transfer without loss of an equalizer circuit which is measured much more effective with the gain of the power of the transducer.

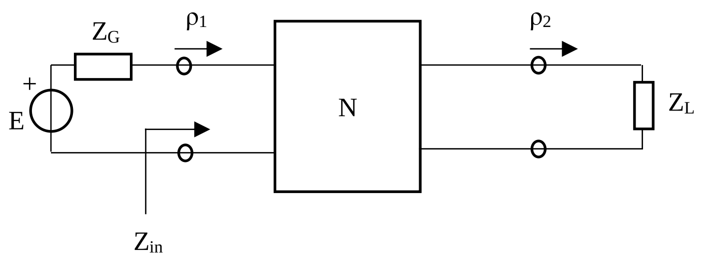


Figure. 2.1. Broadband matching problem

For reference, the basic circuit organizing is as in Figure 2.1, the ideal matching of the generator to the load implies that the power transmitted by the lossless equalizer network N to the load ($\rho_2 = \rho_1$) is equal to the usable power of the generator. It needs that the impedance of the generator conjugate the input impedance of the equalizer, which is possible only on single isolated frequencies, but not on the frequency band. If the correspondence is required over a wide frequency band, the problem has to contain the correspondence that must be obtained internally by the maximum match tolerance and the minimum bandwidth. The goal of broadband matching is to find an equalization network so that the minimum value of the transducer power gain is maximized in a corresponding band and the trade-offs between gains, bandwidth being mainly dictated by the given terminal impedances. Thus, the problem of broadband matching of termination impedances given by a lossless equalizer requires two important reason;

- One of the main considerations is the gain-bandwidth limits of the given termination impedances. With this, it is possible to achieve the best possible performance on the widest operating band.
- The second is equalization structure, it is possible to achieve the best possible performance in this way.

Depending on the termination impedances required, the matching problem can be categorized as single matching, double matching, and active two-port. The classic problem of single matching is the coupling of a purely resistive generator to a complex load which referred to like the double matching problem. Another category of problem, in which the input and output of an active device are simultaneously adapted to the load and generator data impedances, is generally referred to as an active two-port problem. A microwave amplifier is a typical example of this type of design.

In the literature, the problem of matching has been approached from different points of view and formulated in terms of various descriptive network functions (diffusion, impedance functions, etc.). In the following, to illustrate the basic ideas underlying analytic theory, some of the basic approaches will be examined in this part.

2.1.1 The classical single matching problem

In this part, following the literature for the classical single matching problem, the equations used will be examining. To begin, the review of the configuration of a circuit for a simple matching problem is illustrated in Figure 2.2, after this review, the equation will be shared step by step with a comment about using for the solution of this problem.

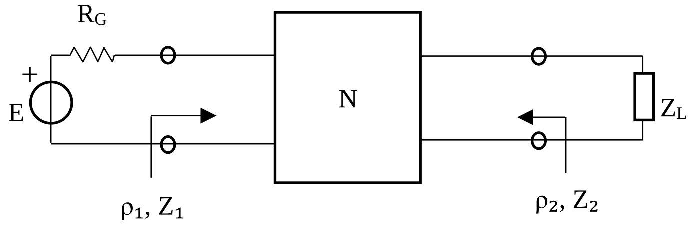


Figure. 2.2. Circuit configuration for single matching problem

In order to be formulated in terms of the normalized reflection function at ports 1 and 2, the conditions for matching the complex load Z_L to the resistive generator must be considered. To define the input reflection ρ_1 is as below,

$$\rho_1 = \frac{Z_1 - R_G}{Z_1 + R_G} \tag{2.1}$$

port 2 is terminated in the load Z_L, when Z_1 is the input impedance (as to the right of port 1). On the other hand, set the reflection function on port 2, is as below,

$$\rho_2 = \frac{Z_2 - Z_L^*}{Z_2 + Z_L} \tag{2.2}$$

port 1 is terminated with R_G, when Z_2 is the impedance (as the left of port 2), and the upper asterisk indicates complex conjugation. In the formula, on port 2, ρ_2 is the complex normalized reflection function, where the normalization is taken with respect to the complex load impedance Z_L. The following formula describes, that on the imaginary axis of the complex frequency plane the two-port are lossless.

$$|\rho_1|^2 = |\rho_2|^2 \tag{2.3}$$

After that, the function of the power gain of the transducer at real frequencies is as below,

$$T(\omega) = 1 - |\rho_1|^2 = 1 - |\rho_2|^2 \tag{2.4}$$

The purpose of broadband matching is to design a lossless network N, for this it is necessary that the Transducer power gain described by (2.4) is maximized in a required frequency band. Clearly, to minimize the modulus of the reflection coefficients $|\rho_1|$ and $|\rho_2|$, it is necessary to maximize $T(\omega)$. The matching task in this formalism then requires the determination of a feasible impedance function Z_2, so that a prescribed frequency characteristic for the ρ_1 or ρ_2 reflection function module is approximated to the desired matching band.

Due to the difficulties with complex standard scattering functions, another description of the matching problem is also favorably used. In this oncoming, the load impedance is mentioned as a two-port Darlington, as shown in Figure 2.3. In this representation, two-port [N L] cascades can be considered as a reactance filter with resistive terminations. Thus, one can attempt to solve the matching problem by designing a reactance filter with the constraint that a portion of the resulting filter coincides with the Darlington equivalent of the prescribed load.

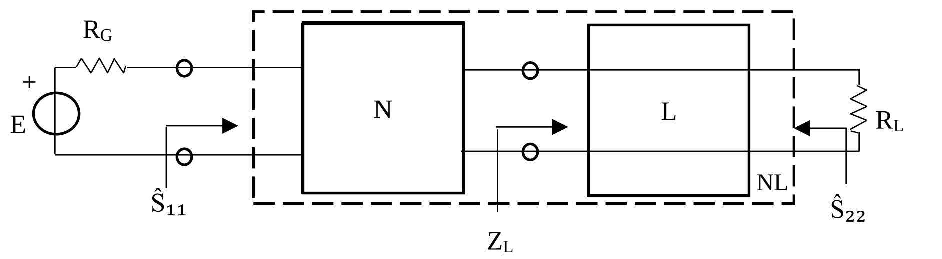


Figure. 2.3. Darlington representation of the load for a single matching problem.

For this reason, BODE has established the fundamental limits of the gain-bandwidth for the problem of matching a resistive generator to a limited load impedance comprising of the parallel configuration of the resistor and a capacitor (Bode, 1945). FANO has expended the work of BODE and fixed the problem of impedance matching between a passive load and a resistive generator (Fano, 1950). His analysis of the matching is based on the Darlington representation of load impedance and can be accepted as representative to understand the link between classical filter theory and the matching problem.

The parameters $\hat{S}_{12}$ and $\hat{S}_{22}$ of the composite network [N L], can be written (as below) by taking into account the single matching representative represented in Figure 2.3

$$\hat{S}_{12} = \frac{S_{12}\, L_{12}}{1 - S_{22}\, L_{11}} \qquad \hat{S}_{22} = L_{22} + \frac{S_{12}\, L^2_{12}}{1 - S_{22}\, L_{11}} \tag{2.5a,b}$$

where Sij and Lij represent the actual normalized scattering parameters of the equalization circuit N and the Darlington equivalent of the load L respectively.

From the first equation of (2.5), it is obvious that the zeros of L_{12} are also zeros of $\hat{S}_{12}$, if they are not canceled by those of the denominator. The zeros of the denominator term 1-$S_{22}L_{11}$ can exist only to the left of the complex plane ρ (LHP). If the denominator term has zeros on the axis jω, they are simple and will be canceled by those of S_{12} in the numerator. Indeed, in the case of a zero-axis j, say $\rho_0 = j\omega_0$, have $S_{22}\,(j\omega_0)\, L_{11}\,(j\omega_0) = 1$. Since S_{22} and L_{11} are bounded by unit, this is only possible if $|S_{22}\,(j\omega_0)| = |L_{11}\,(j\omega_0)| = 1$. In this case, S_{12} must have a simple zero for $\rho = j\omega_0$ because of the unitary property. From these considerations, it can be concluded that all the zeros of L_{12} in Re $\rho \geq 0$ must absolutely be included in those of $\hat{S}_{12}$.

Now, using the condition on the L_{12} zeros in the second equation of (2.5) to a zero of L_{12} say ρi, is as below,

$$\hat{S}_{22}(\rho i) = L_{22}(\rho i). \tag{2.6}$$

moreover, if the L_{12} zero is of multiplicity k, then the derivatives of the reflection functions $\hat{S}_{22}$ and L_{22} at this zero must be equal, is as below,

$$\hat{S}_{22}^{(k)}(\rho) = L_{22}^{(k)}(\rho) \text{ at } \rho = \rho i, \tag{2.7}$$

where the exponent (k) designates here the derivative k'th of a function. In other words, for $\hat{S}_{22}$ and L_{22}, the first k coefficients are equal. This property constitutes the basic conditions on the actual feasible bounded reflection function $\hat{S}_{22}$ satisfies the condition of (2.7) at all the transmission zeros of a given load, then the reflection function $\hat{S}_{22}$ can be realized by extracting first the Darlington equivalent of the load L then the lossless equalizer N.

In view of the approximation, to reflect the effect of the load impedance on the gain-frequency characteristic, FANO has derived some formula and the modulus of the function $\hat{S}_{22}$ reflection function (Fano, 1950). With these relationships, the $\hat{S}_{22}$ reflection function could be constructed to ensure the extraction of the prescribed load network. YOULA has solved the matching problem described above in a more general form directly by studying the standard complex diffusion parameters (Youla, 1964). He deduced the feasibility conditions of the complex normalized reflection function ρ_2 given in (2.2) using a suitable all-pass product derived from the transmission-carrying zeros. His analysis resides in the standardized, normalized complex reflection function defined by

$$\rho_2 = \Upsilon i \frac{Z_2 - Z_L^*}{Z_2 + Z_L} \tag{2.8}$$

where Υi is an all-pass function constructed from the zeros of the given load impedance. In its approach, from a feasible transfer function that satisfies the desired characteristic, a reflection function with suitable all-pass products Υi, of the form (2.8) is first generated. Then, using the formula concerned for series evaluation around the load transmission zeros of the terms in (2.8), obtained a set of equations on the indeterminate parameters of the reflection function. This set of equations constitutes the feasibility conditions on the reflection function coefficients and the load impedance given in the sense of Youla. The indeterminate parameters are then calculated by optimization, under the coefficients constraints mentioned above. Thus, a permissible solution for the reflection function is obtained. Although the method seems applicable to all general cases, the steps of the procedure, such as the explicit construction of the approximation function, the evaluation of series expansion, and the generation of the system of equations, are generally rather extensive cases almost insoluble. On the other hand, because of the necessary all-pass factors, the resulting structures cannot generally be realized without coupled coils. This is a great disadvantage from the point of view of technical realization.

In the case of a double coupling (Figure 2.4), where the impedance of the generator is also complex, it is necessary to consider the effect of the generator network G on the gain-bandwidth restrictions with that of the load L the solution to find the restrictions due to G and L requires in general that the gain-bandwidth restrictions valid for a single match at the input and at the output ports are satisfied simultaneously. This requires that the transfer function $\hat{S}_{21}$ (p) of the trio [GNL] includes all the transmission zeroes jω and RHP of both [G] and [L] and makes them suitable multiplicities. In this case, the problem is obviously much more complex than the single match.

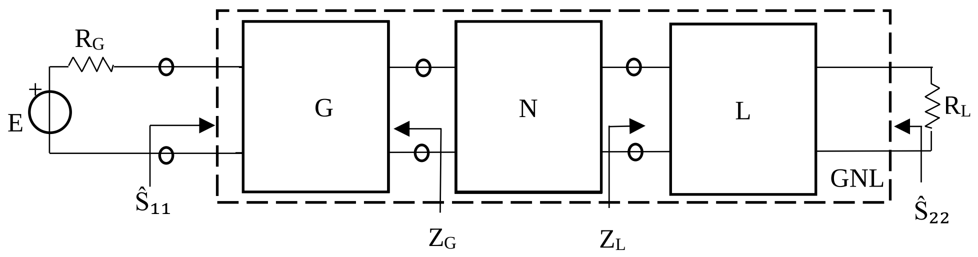


Figure. 2.4. Darlington representation for double matching problem

WOHLERS has included double matching issues in the literature, the concept of compatible impedance (Wohlers, 1965). Then, CHIEN (Chien, 1974), CHEN (Chen, 1977), SATYANARAYANA (Satyanarayana and Chen, 1980) and YOULA (Youla, 1964) used the complex standardized scattering parameters to obtain a general solution to the double-matching problem. A simplified solution to the double matching problem is later improved by CARLIN and YARMAN using true normalized scattering parameters (Carlin and Yarman, 1983), (Yarman, 1986), (Youla et al., 1984).

2.2 Real Frequency Broadband Matching

For circuit realization for load and generator impedances, explicit expression or, equivalently, the theory of analytic matching is required. In practice, however, real-frequency experimental data are generally encountered for termination impedances to be matched. In such cases, the data must be approximated by an appropriate equivalent circuit or a feasible analytic function, in order to be able to use analytic theory, which is not easy and satisfactory. Another problem also comes from the need to start with an analytical form of a gain function that is the best choice for the given end-point impedances. That is, an analytical form of the transfer that incorporates both the matching two-port terminations and the complex terminations is required in advance. Termination impedances are then processed in a complicated way to establish the theoretical bandwidth and gain restrictions that must be satisfied by some form of assumed transfer function. These limitations take the form of a set of equations that include one or two reactive elements, these restrictions being resolvable, however, for more complicated terminations, the difficulties subjected to the procedure are well-nigh not manageable. In fact, the most complicated case handled by the analytic theory in the literature is that of the simple LCR load (2 reactances and a resistor) that is matched on the low-pass band. Even in this case, the solution is very hard. For these cases, even when the procedure is applicable in principle, If the number of reactive elements in the charging network is greater than two, this can lead to unnecessarily complicated equalizers. (Carlin and Amstutz, 1981), (Yarman, 1982), (Carlin and Civalleri, 1985).

On the other hand, one can try to design matching networks using microwave circuit design packages assisted by software available in professional life. (Perlman and Schepps, 1982). This software utilizes fully digital methods and are designed for the analysis and optimization of the circuit element base. In other words, the network configuration and an accurate estimate of the element values must be provided to the software. Here, the major difficulty with matching network design is the determination of an optimal configuration that is generally doubtful. In addition, the performance function is very often nonlinear under conditions of values of elements indeterminate to optimize. To ensure optimal convergence, it is necessary to take values of elements sufficiently close to the final solution. For narrowband design problems with a small number of elements, the choice of circuit configuration and element initialization may not be very crucial. With trial and error, reasonable solutions can be found for these cases. However, for complicated problems that require broadband designs, if the optimal configuration is not known, the use of these software is difficult and would give unsatisfactory results. Thus, such design assemblies created by the software are intended the last adjustment of element values, when the circuit is synthesized in advance.

As can be concluded from the discussion above, purely numerical methods as well as analytic theory are subject to several problems and are generally unable to handle most of the corresponding issues. To circumvent the problems associated with analytic theory and purely numerical techniques, preceding studies in the literature have addressed the task of determining more practical ways to design matching networks, which can work directly on measured digital data.

To eliminate the major flaws that the analytic theory, CARLIN has proposed a new method called real frequency technique. (Carlin and Amstutz, 1981). It is not necessary to assume the analytical form of the transfer function or the equalizer configuration, since the CARLIN method directly uses the actual experimental frequency data for the generator and the load. Following various simple matching problems, the real frequency technique leads to better performance with simpler structures than compared to a model obtained with analytic theory (Yarman, 1982), (Carlin and Amstutz, 1981), As a result of these advantages, the real frequency technique has proven to be the most feasible approach to solving broadband matching problems encountered in practical applications.

Later, CARLIN, YARMAN, FETTWEIS and PANDEL developed some alternative real frequency algorithms (Carlin and Yarman, 1983), (Yarman, 1982), (Yarman, 1985), (Fettweis, 1979), (Pardel and Fettweis, 1985), (Pardel and Fettweis, 1987), (Yarman and Fettweis, 1990). And several authors have used variants of these techniques to solve matching and amplification design problems (Beccari, 1984), (Jarry and Perennec, 1987), (Hatley, 1967). In the remainder of this part, basic real frequency broadband matching techniques will be described in this part.

3. NARROWBAND MATCHING

3.1 Fundamental Properties of Narrowband Matching

In this part, the general approach in the theory of narrowband matching is reviewed. For DC circuits, according to the important theorem, to transfer the maximum power from a generator to its load, the resistance of the load must be equal to the resistance of the generator. To prove the important theorem, with calculations and drawings a basic proof is presented in Figure 3.1. For ease of calculation, the generator is fixed with a resistance and a voltage of one ohm and one volt, respectively.

With regard to alternating current or waveforms varying in time, as prescribed in this same theorem when the load impedance (Z_L) is equal to the conjugate complex, transferred maximum power from a generator to its load is realizable, the real part stays idem, the change is simply in the conjugated part, it is necessary to take the opposite reactance. For example, for the impedance generator, if the simple formula was $Z_G = R + jX$, according to the theorem the conjugated complex part should be $Z_G = R - jX$.

Following Figure 3.1, the theorem for maximum power is much more understandable, it is obvious that for the maximum transfer to occur, the load impedance must be clearly the complex conjugate of the generator. In Figure 3.2, it is represented schematically. If the Z_G (the generator) is to be taken in hand, with an inductive series reactive component (+ jX), which drives the impedance of the complex conjugated charge inductor consisting of a capacitor reactance (- jX), in series with the R_L, the complex components are in series and cancel, after this cancellation, the R_G and the R_L are equal. The maximum transfer will be acquired; Thus, simply refer to a condition by speaking of a generator driving its complex conjugate in which any generator reactance resonates with an equal and opposite

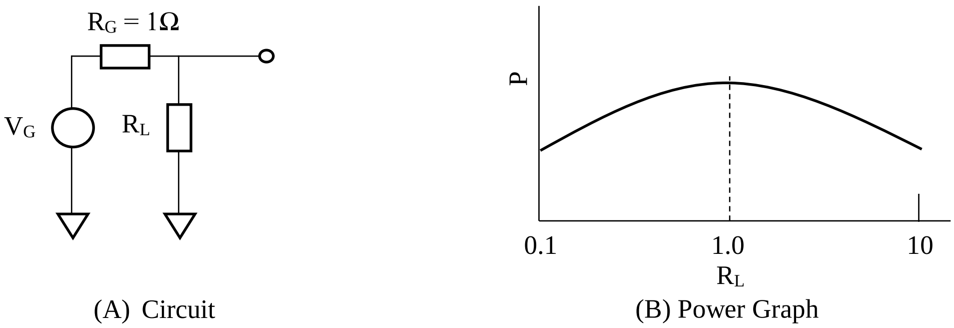


Figure. 3.1 Theorem (Power)

charge reactance, thus leaving only equal resistance values for the charge of the generator and endings.

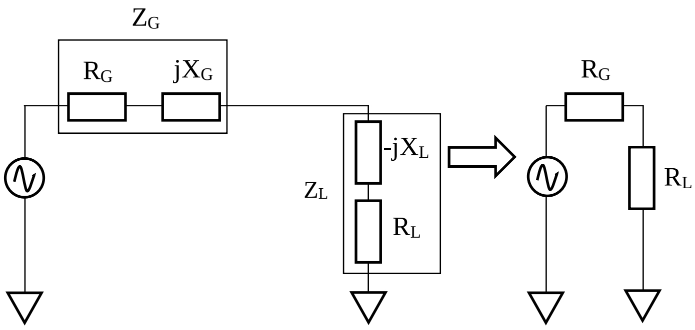


Figure. 3.2. The result of the equivalent circuit of the generator impedance driving its complex conjugate

3.2 The L Network Matching

In the matching circuits, the circuit L is very simple, and a circuit frequently used, as shown in Figure 3.3. This circuit is named after connection way of its components, which resembles the shape of the letter L, as the drawings show on this subject, there is a kind of possibility of creating these circuits, the components are L and C, there is the possibility of creating it in low-pass (Figures 3.3A and 3.3B) and there is also the choice to choose high-pass (Figures 3.3C and 3.3D).

To understand more deeply before moving on to the phase of the equations, which will be used to design the networks in Figure 3.3, let's try to refine on an existing matching network, after this analysis will understand exactly how the matching of impedance is realized in this part.

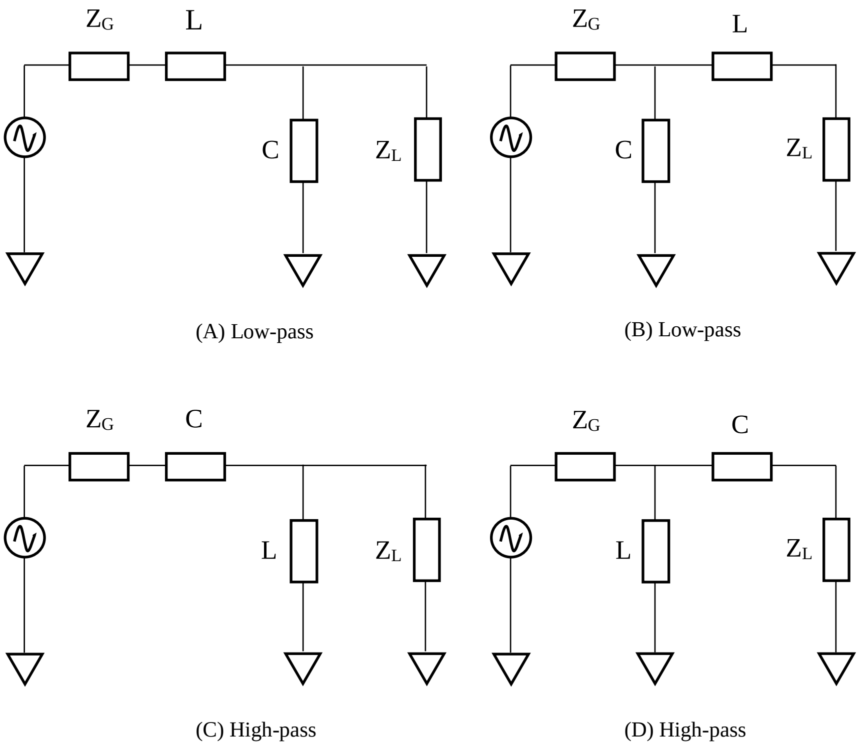


Figure. 3.3. The L type networks

The following equations will help us very easily to design the impedance matching circuit of Figure 3.3,

$$Q_S = Q_P = \sqrt{\frac{R_P}{R_S} - 1} \tag{3.1}$$

$$Q_S = \frac{X_S}{R_S} \tag{3.2}$$

$$Q_P = \frac{X_P}{R_P} \tag{3.3}$$

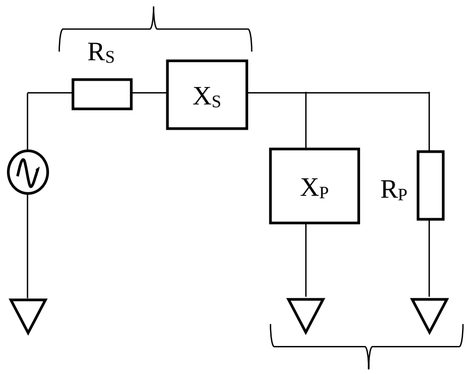
where, as shown in Figure 3.4:

Q_S = For the part (Q) values of series,

Q_P = For the part (Q) values of shunt,

R_P = For the shunt part, resistance.

$Q_S = X_S / R_S$



$Q_S = Q_P = \sqrt{\frac{R_P}{R_S} - 1}$ $Q_P = R_P / X_P$

Figure. 3.4. The L type network design (Summary)

X_P = the shunt reactance,

R_S = the series resistance,

X_S = the series reactance.

For the quantities X_P and X_S, as a result of theorem, X_P or X_S can be either capacitive reactance, or inductive reactance, it must not be of the same type, they must be chosen opposite, for example, if one chooses that X_P is capacitor, the X_S must be considered as an inductor, and the opposite can be chosen for both.

3.2.1 The L type method with a basic example

For a basic L type circuit, to matching the circuit, taking 120-ohm generator and 1200-ohm load at 150 MHz. To transfer a DC voltage from the generator to the load, and the path need an inductor in the series leg, for reference, the circuit in Figure 3.3(A) is chosen,

With the equation (3.1), (3.2) and (3.3) respectively,

$$Q_S = Q_p = \sqrt{\frac{1200}{120} - 1} = 3$$

$$X_S = Q_S * R_S = 3*120 = 360\text{-ohm (Ind.)}$$

$$X_P = \frac{Rp}{Qp} = \frac{1200}{3} = 400\text{-ohm (Cap.)}$$

Last step, defining the value of the component at 150 MHz:

$$L = \frac{X_S}{\omega} = \frac{400}{2\Pi(150*10^6)} = 424 \text{ nH}$$

$$C = \frac{1}{\omega X_P} = \frac{1}{2\Pi(150*10^6)*(400)} = 2.65 \text{ pF}$$

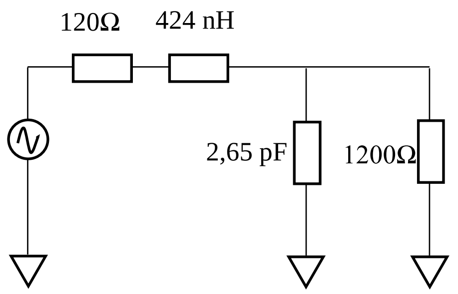


Figure. 3.5. The basic example of L type network circuit

3.3 Properties of Complex Load

In the circuits, they take into account that the inputs and outputs are regularly complex, with this information, it can be understood that it means that they contain resistive and reactive components ($R \pm jX$). In many circuits it is the same case, for example it can be seen on transmission lines, antennas and mixers and most other generators and loads.

3.3.1 First approaches to manage complex impedances: Absorption

The impedance matching network absorb all stay reactances. By careful placement of the elements concerned, so that the main capacitors are in parallel with parasitic capacitances, and the main inductances are in series with all the plateau inductances. Values of the spurious components are then subtracted from the values of the calculated elements, which creates new element values (C ', L'), these elements are smaller than the values of the calculated elements.

3.3.2 Second approaches to manage complex impedances: Resonance

For resonance, the frequency of interest must be resonant of a parasitic reactance with an equal and opposite reaction. Illustrated this approach with the following example, the details of this example are important to understand how they can design simple impedance matching networks between complex loads. For it to be absorbed into the network, they must be resonant with an equal and opposite reactance,

3.3.2 Illustrate the resonant approach with an example.

For a complex load circuit, to trying the resonant approach, taking 100-ohm generator and 1000-ohm load at 100 MHz, and 80 pF for the capacitor in shunt legs. To transfer a DC voltage from the generator to the load, the circuit in Figure 3.3(C) is chosen.

Firstly, it is necessary to eliminate the parasitic capacitor of 80 pf ($C_{misplace}$) by making it resonate with a parallel inductance with 100 MHz.

$$L = \frac{1}{\omega^2 C_{misplace}} = \frac{1}{(2\Pi(100*10^6)^2 *(80*10^{-12})} = 31.66 \text{ nH}$$

After eliminated the misplace capacitance, second step is proceeding with matching the network between 100-ohm generator and seeming 1000-ohm load.

$$Q_S = Q_p = \sqrt{\frac{1000}{100} - 1} = 3$$

$$X_S = Q_S * R_S = 3*100 = 300\text{-ohm}$$

$$X_P = \frac{Rp}{Qp} = \frac{1000}{3} = 333\text{-ohm}$$

Third step, defining the value of the component

$$C = \frac{1}{\omega X_S} = \frac{1}{2\Pi(100*10^6)*(300)} = 5.30 \text{ pF}$$

$$L = \frac{X_P}{\omega} = \frac{333}{2\Pi(100*10^6)} = 530 \text{ nH}$$

The last step, this circuit can be simplified, two parallel inductors to be replacing by a single inductor, according,

$$L_{new} = \frac{L_1 * L_2}{L_1 + L_2} = \frac{(530*31.66)}{(530 + 31.66)} = 30 \text{ nH}$$

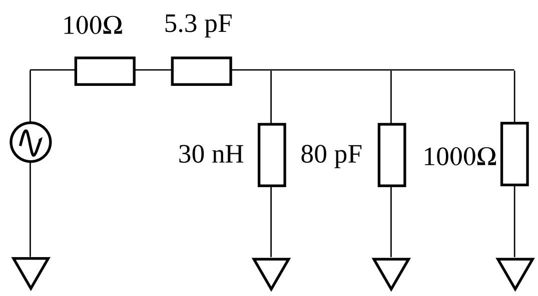


Figure. 3.6. The example of the resonant approach

3.4 The Pi Network Matching

To describe the Pi network, think of two back-to-back L networks configured to match the load and the generator to a virtual resistance at the junction between the two networks. The configuration of the Pi network is shown in Figure 3.7. In the Figure concerned the meaning of X_{S1} and X_{S2} have negative signs, it is merely symbolic, the purpose of these negative signs and to indicate that the X_S values are opposite of reactance of X_{P1} and X_{P2}, respectively. To explain this configuration more clearly, if X_{P1} is a capacitor X_{S1} must be an inductor, it may be the opposite too. On the other hand, if X_{P2} is an inductor, X_{S2} must be a capacitor, it can be the opposite in this case too. These signed do not indicate negative reactances (capacitors).

To define a value to the virtual resistor (R), take into account that the value must be lower than R_G or R_L, it must be remembered that this value is connected to the arm of the series of each section L, if it was not the case it could be of any value wish. From time to time, the virtual resistance value (R) is defined by the Q (quality coefficient) loaded in the circuit that you specify at the beginning of the design process. To define the needs, the Q responsible for this network will be defined as:

$$Q = \sqrt{\frac{R_H}{R} - 1} \tag{3.4}$$

the definition of the initials in the equations

$R_H = R_G$ or R_L are the largest terminating impedance
R = Resistance (virtual).

Even though it is not entirely accurate, it is a mostly accepted Q-determinant formula for this circuit and is certainly close enough for most practical work.

For some Pi networks, impedance matching between the 120-ohm generator and the 1200-ohm load will be performed. Choosing for each application will depend on any factor as described below,

- For the first step it is to eliminate misplace reactances.
- Define the harmonic filtering needs.
- Define the need to turn on or off the DC voltage.

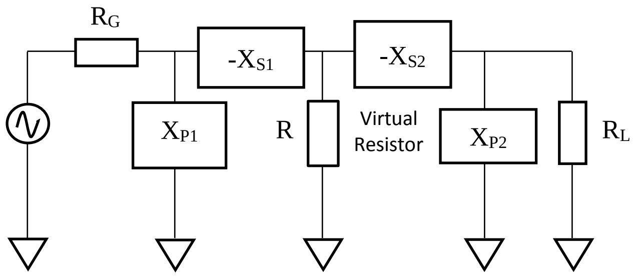


Figure. 3.7. Two back-to-back L type networks (Pi network).

3.5 The T Network Matching

The design of the network T is the same as for the network Pi, but with the T, you match the load and the generator, through two L-type networks, to a virtual resistance. As shown in Figure 3.8, both types of L will have their shunt legs connected.

For the T network, when a high Q arrangement is needed, this is to say that the network wants to be used to match two low value impedances. Section L determines the loaded Q of the network, which will occur at the end with the smallest termination resistance. It should not be forgotten that each terminating resistor is in the serial branch of each network. From this the formula for determining the Q charged of the network T is:

$$Q = \sqrt{\frac{R}{R_{small}} - 1} \tag{3.5}$$

the definition of the initials in the equation,

R = Resistance (virtual)
R_{small} = Resistance (Smallest terminating).

That the formula Q submitted for the network T is the same formula Q for the network Pi, after which, in order to produce the T-networks, the L-sections must be reversed, for this type of network, it must be certain that it is necessary to redefine the Q-formula to take into account the new localization of the resistance with respect to these networks L. Moreover, equation 3.1 is a general equation or one can derive from other equation for special applications like the equation 3.4 and 3.5

$$Q = \sqrt{\frac{R_P}{R_S} - 1} \tag{3.6}$$

the definition of the initials in the equations

R_P = the L network, resistance in the shunt branch.

R_S = the L network, resistance in the series branch.

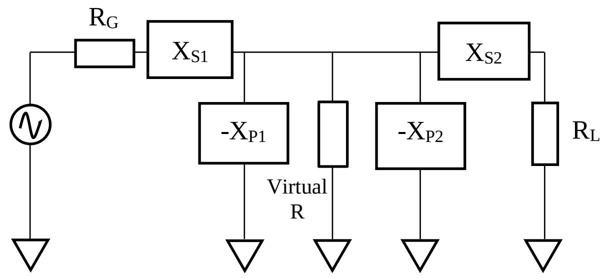


Figure. 3.8. Two back to back L type networks (T network).

3.6 The Example of Impedance Matching

An L-type impedance matching circuit is shown in Figure 3.9, the component values between the generator and the load are 120-ohm and 1200-ohm, respectively. the 120-ohm generator directly charges the load of 1200-ohm, without the matching circuit, about a third of the available signal from the generator is gone even before the start. In order not to lose this signal, adding a matching circuit will eliminate this loss and allow a maximum power transfer to the load. This is done by forcing the 120-ohm generator to see 120-ohm in the load. In the next step, the matching networks are studied with formulas and figures on how to reach a maximum power transfer to the load.

In Figure 3.9, analyzed the simplicity of how the match happens. In Figure 3.10. The first step of the analysis is to establish the load impedance when the -j400-ohm capacitor is placed through the 1200-ohm load resistor.

Following this result, the shunt configuration of the -j400-ohm capacitor and the 1200-ohm resistor resembles an impedance of 120 -j360-ohm. This is a series configuration of a 120-ohm resistor and a +j360-ohm capacitor, as indicated in Figure 3.11. In fact, if you have connected a signal generator to circuits similar to Figures 3.10 and 3.11, following the same characteristics, you cannot tell the difference between the two.

For the next step, an apparent impedance of the 120-j360-ohm series for a load, to complete the impedance matching to the 120-ohm generator, it consists of adding an equal and opposite reactance (+j360-ohm) in series with the network of Figure 3.11.

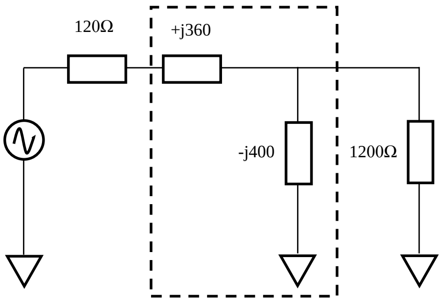


Figure. 3.9. A circuit where the generator is 120-ohm and the load is 1200-ohm.

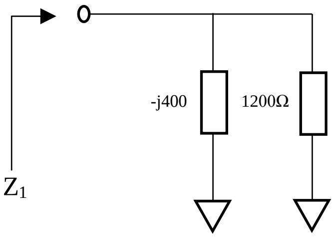


Figure. 3.10. Shunt configuration of R_L and X_G

For calculate, utilization of the formula is like below,

$$Z = \frac{X_G * R_L}{X_G + R_L}$$

$$= \frac{-j400 * (1200)}{-j400 + 1200}$$

$$= 380 \angle -71.56^0$$

$$= 120 \text{ -j360-ohm}$$

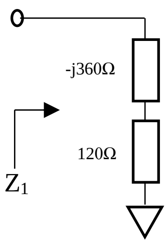


Figure. 3.11. Equivalent circuit of Figure. 3.10.

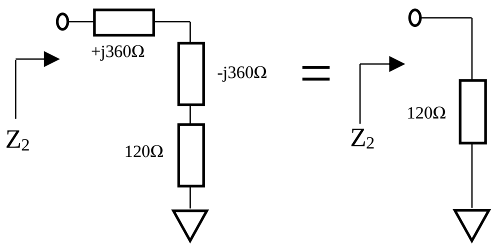


Figure. 3.12. Completing the match

The addition of +j360-ohm inductance causes the capacitor cancellation of -j360-ohm, leaving only an apparent load resistance of 120-ohm. This is illustrated in Figure 3.12. The actual network configuration of Figure 3.9 has not changed. All the steps are analyzed in small parts of the network in order to understand the function of each component.

To sum up the example, (120-ohm generator), the function of the parallel component of the impedance matching network is to transform a larger impedance into a smaller value with a real part equal to the real part of the other termination impedance. The serial impedance matching element then resonates with or cancels any reactive component present, thereby allowing the source to cause an apparently equal charge for optimal power transfer.

4. THE SMITH CHART

4.1 Fundamentals of Impedance and The Smith Chart

For RF circuit design, impedance matching is very important, because in circuits the RF part has a function of maximizing power transfer by staying in the same phase. The impedance matching has a role in controlling the initial impedance in a reference impedance (50-ohm), or in a requested reference by constructing an impedance matching circuit.

The impedance matching process involves important steps, for technicians who design this type of circuit must follow a basic engineering principle.

First, the input impedance or the output impedance of the RF parts may be necessary to match a defined impedance that is not the reference impedance, the purpose of this matching is saving and improve performance. However, for testing the part of the RF circuits, it is necessary that the input and output impedances are nevertheless at the same value of the standard reference impedance, this reference is also adopted by the manufacturers of test equipment. Thus, when the actual design of the RF circuit must match to an impedance that is not at the defined reference, a rather complicated process must be used for this.

Secondly, the design of the impedance matching network has the possibility of having active or passive parts. It is preferable to create an impedance matching network design using passive parts rather than active parts, the reasons for this choice, the passive parts are simpler to build, and it is more cost effective than the parts active. Inductors, capacitors and resistors are passive components, but resistances are generally not considered, because they attenuate the signal and introduce a considerable noise, nobody is used even if their use is authorized, it is therefore necessary to know the application of the passive components in the matching network. To create a narrowband impedance matching network, it can have one, two, or three part, following that, depending on the relative bandwidth the quantity of component can be increased.

Thirdly, it is essential to separate the narrowband and the broadband. Indeed, in theory, the limit between the narrow and wide bands involved is not exactly defined, but rather defined with technical design experience. An RF circuit part, to define the band if it is narrow or wide, the bandwidth must be considered, if the relative bandwidth is less than 15%, the band is narrow, otherwise, if the bandwidth is higher than 15%, the band is wide. In the case where the relative operating bandwidth is about 15%, it is desirable to consider this circuit as a broadband part, the objective of this choice being to have a more efficient performance both for broadband and narrowband cases. To test the return loss R_L (S_{11} or S_{22}), the trace of the impedance Z or the return loss R_L (S_{11} or S_{22}) the use of the Smith Chart is very effective, on the figure, the answer for a frequency bandwidth is displayed. To illustrate this theory by way of an example, the frequency response of impedance Z or loss of return R_L (S_{11} or S_{22}) from the lowest frequency to the highest frequency is shown in Figure 4.1.

$$\delta f_{trace} = f_{max} - f_{min} \tag{4.1}$$

$$BW = f_H - f_L \tag{4.2}$$

$$f_C = \frac{1}{2}(f_L + f_H) \tag{4.3}$$

the definition of the initials in the equations,

δf = frequency coverage,
f_{max} = highest frequency,
f_{min} = lowest frequency,
Δf = BW = frequency bandwidth,

f_H = the high-frequency end (bandwidth wanted),

f_L = the low-frequency end (bandwidth wanted),

f_C = central frequency.

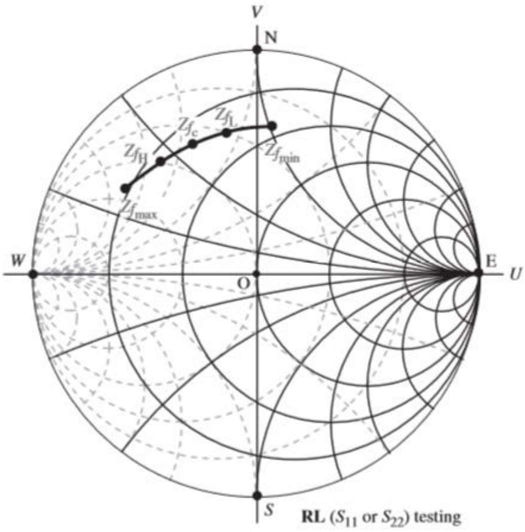


Figure. 4.1. Smith Chart from RL (S_{11} or S_{22}) testing (Li, 2013)

all points in the displayed line match to a frequency. The values of the impedance Z and the loss of return R_L (S_{11} or S_{22}) can be read simultaneously on each point:

Zf_{min} = impedance (lowest frequency),

Zf_{max} = impedance (highest frequency),

Zf_L = impedance (low-frequency end of bandwidth wanted),

Zf_H = impedance (high-frequency end of bandwidth wanted),

Zf_C = impedance (central frequency).

In the case of broadband radio frequency circuit, creates a circuit design, requires considering the response for all frequency bandwidths. In contrast, in the case of narrowband radio frequency circuits, creates a circuit design, simply requires the central frequency, because the frequency response at the center frequency is a good approximation of that for the entire bandwidth.

Fourth, in impedance matching studies, the Smith Chart is an important tool. For a simple matching study, the design of a network can be calculated manually using proven equations. However, with Smith Chart, the calculation is much simpler. In the studies of this case, the complex impedance is considered for the generator or the load, the pure resistance is no longer a limit for the impedance, to describe the complex impedance, it is necessary to know that it has two parts, one which is real, and the other which imaginary, for the imaginary part, there is a standard reference (50-ohm) which facilitates the calculations.

4.2 Implementation of an Impedance Matching Network

In circuits, the original radio frequency part generally has a different impedance than the reference impedance (50-ohm), which should be adapted. For this reason, the location of the impedance matching network must be added between the original radio frequency part and a network analyzer that provides the reference impedance (50-ohm).

To better understand the configuration of impedance matching network implementation, this configuration is shown in Figure 4.2, it is necessary to analyze the process, this configuration consists of three parts: there is an analyzer, the RF part, and the impedance matching network, the task of the analyzer is to read the return loss or impedance, for the original RF part it must be matched, and the last one must build the original matching circuit. For the last phase of the process, Smith Chart will help to execute the adjustment of the loss of efficiency and the construction of the impedance network

About the impedance matching circuit, the shape of the circuit is like arms and branches, the arms are the horizontal part, where the components are in series, and the branch is the vertical part, where the components are in parallel. Capacitors and inductors are on each arm or branch. In the case of the narrowband, each arm or branch contains only one part, either a capacitor or an inductor.

The arms or the branches, labeled 1, 2, 3, 4, etc., are added successively to the impedance matching circuit shown in Figure 4.2. Arms or the branches marked 1, 2, 3, 4 and so forth added successively into the impedance matching circuit. In the beginning, the impedance matching circuit is vacant, so that point A is directly connected to point B. Line S_{11} or Impedance is displayed on the analyzer screen. It represents the original RF part impedance.

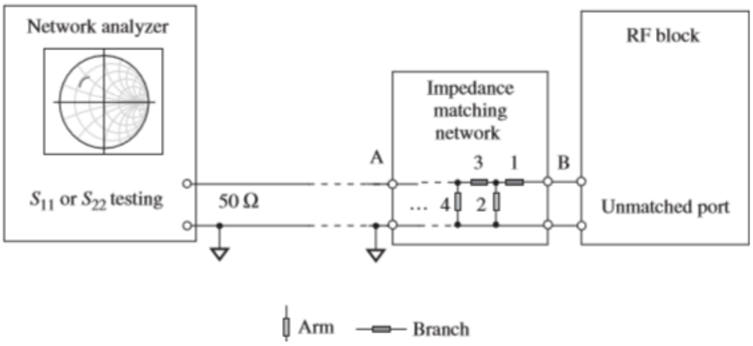


Figure. 4.2. Configuration for impedance matching from input impedance of an original RF part to50-ohm (Li, 2013)

As the first arm labeled with 1 is added between the points A and B, the line S_{11} or impedance moves to another point on the Smith Chart. When other branches or arms are added in the impedance matching circuit, the impedance moves to other points of the Smith Chart consequently. It is expected that the impedance eventually moves to the center of the Smith Chart (50-ohm). At this stage, the impedance matching process is completed, and the resulting arm and branch entity is the requested impedance matching network. Generally, for point A to reach the reference impedance (50 ohms), it is necessary to increase the quantity of arms or branches.

4.3 The Design of The Impedance Matching Network Created by One Part

4.3.1 The one part included in series in the circuit

To create a part in serial design in the impedance matching network, it is desirable to use the impedance instead of the admittance, to demonstrate the variation of its electrical characteristics, the equations are shared as below,

$$Z_{os} = R_{os} + jX_{os} \tag{4.4}$$

$$Z = Z_{os} + \Delta Z = R + jX \tag{4.5}$$

the definition of the initials in the equations,

Z_{os} = initial impedance before included part,
R_{os} = real part of Z_{os},
X_{os} = imaginary part of Z_{os},
Z = resultant impedance after the included part,
ΔZ = variation of impedance,
R = real part of Z,
X = imaginary part of Z.

An ideal inductance or an ideal capacitance can be used in the part that will be included in the impedance matching network, the real part of the impedance will remain fixed, the change will be in the imaginary part of the resulting impedance, described with the following equations,

$$\Delta R = R - R_o = 0 \tag{4.6}$$

$$\Delta X = X - X_o \tag{4.7}$$

the definition of the initials in the equations,

ΔR = changing of resistance,
ΔX = changing of reactance.

the part that will be inserted with an inductor in series in the impedance matching network,

$$\Delta X = \Delta X_L = + L\omega \tag{4.8}$$

the resulting impedance is as below, calculation from equations (4.4) to (4.8),

$$Z = R + jX = Z_{os} + Z_L = R_{os} + j(X_{os} + X_L) \tag{4.9}$$

the part that will be inserted with a capacitor in series in the impedance matching network,

$$X = X_C = -\frac{1}{C\omega} \tag{4.10}$$

the resulting impedance is as below, calculation from equations (4.4) to (4.7),

$$Z = R + jX = Z_{os} + \Delta Z_C = R_{os} + j(X_{os} + \Delta X_C) \tag{4.11}$$

the definition of the initials in the equations,

L = inductance of the part included to network,
C = capacitance of the part included to network,
ω = angular frequency,
ΔZ_L = inductive impedance included,
ΔZ_C = capacitive impedance included,
ΔX_L = the inductive reactance included,
ΔX_C = the capacitive reactance included.

On Smith Chart, analyzing the effects when inductance or capacitance in series is included in the impedance matching network, Figure 4.3 illustrates the direction in which the impedance is drawn after these effects. the impedance change in Smith Chart has basic rules, which must be respected:

The impedance matching network that will be inserted with an inductor in series,

- for the first rule, the inductive reactance included must be positive ($\Delta X = \Delta X_L$), the operating frequency must increase the amplitude of the inductive reactance if it also increases;
- the original impedance P moves clockwise along the circle r = constant impedance, if the insertion of the inductor L_S is in series. the value of the inductor defines the length of the movement arc.

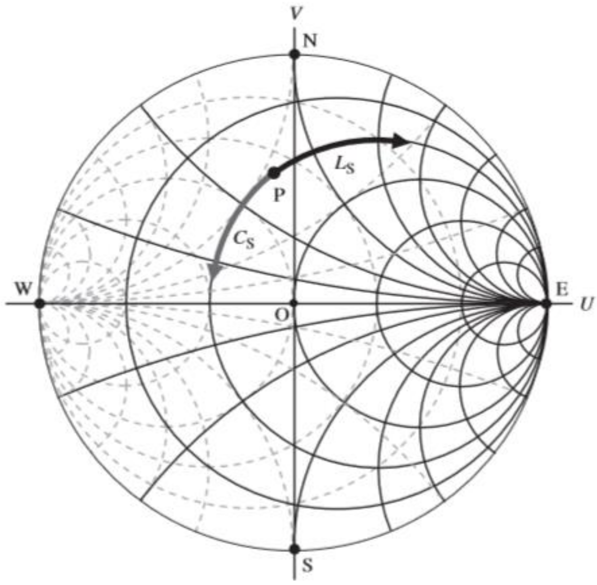


Figure. 4.3. The insertion effects of directions of an inductor or a capacitor in series on the Smith Chart (Li, 2013)

4.3.2 The one part included in parallel in the circuit

To create a part in parallel design in the impedance matching network, it is desirable to use the admittance instead of the impedance, to demonstrate the variation of its electrical characteristics, the equations are shared as below,

$$Y_{op} = G_{op} + jB_{op} \tag{4.12}$$

$$Y = Y_{os} + \Delta Y = G + jB \tag{4.13}$$

the definition of the initials in the equations,

Y_{op} = initial admittance before included part,
G_{op} = real part of Y_{op},
B_{op} = imaginary part of Y_{op},
Y = resultant admittance after the included part,
ΔY = variation of admittance,
G = real part of Y,
B = imaginary part of Y.

An ideal inductance or an ideal capacitance can be used in the part that will be included in the impedance matching network, the real part of the impedance will remain fixed, the change will be in the imaginary part of the resulting impedance, described with the following equations,

$$\Delta G = G - G_{op} = 0, \ \Delta B = B - B_{op} \tag{4.14}$$

the part that will be inserted with an inductor in parallel in the impedance matching network

$$\Delta B = \Delta B_L = -\frac{1}{L\omega} \tag{4.15}$$

the resulting admittance is as below, calculation from equations (4.12) to (4.14), and (4.17),

$$Y = Y_{op} + \Delta Y_L = G + jB = G_{op} + j(B_{os} + \Delta B_L) \tag{4.16}$$

the part that will be inserted with a capacitor in parallel in the impedance matching network

$$\Delta B = \Delta B_C = C\omega \tag{4.17}$$

the resulting admittance is as below, calculation from equations (4.12) to (4.14),

$$Y = Y_{op} + \Delta Y_C = G + jB = G_{op} + j(B_{op} + \Delta B_C) \tag{4.18}$$

the definition of the initials in the equations,

L = inductance of the part included,
C = capacitance of the part included,
ω = angular frequency,
ΔY_L = inductive admittance included,
ΔY_C = capacitive admittance included,

ΔB_L = imaginary part of inductive included,

ΔB_C = imaginary part of capacitive included.

On Smith Chart, analyzing the effects when inductance or capacitance in parallel is included in the impedance matching network, Figure 4.4 illustrates the direction in which the impedance is drawn after these effects. the impedance change in Smith Chart has basic rules, which must be respected:

The impedance matching network that will be inserted with an inductor in parallel,

- for the first rule, the admittance included must be a negative imaginary part ($\Delta Y = \Delta Y_L$), the operating frequency must decrease the amplitude of the admittance when its increases;
- the original impedance P moves counterclockwise along the circle g = constant impedance, if the insertion of the inductor L_P is in parallel, the value of the inductor defines the length of the movement arc.

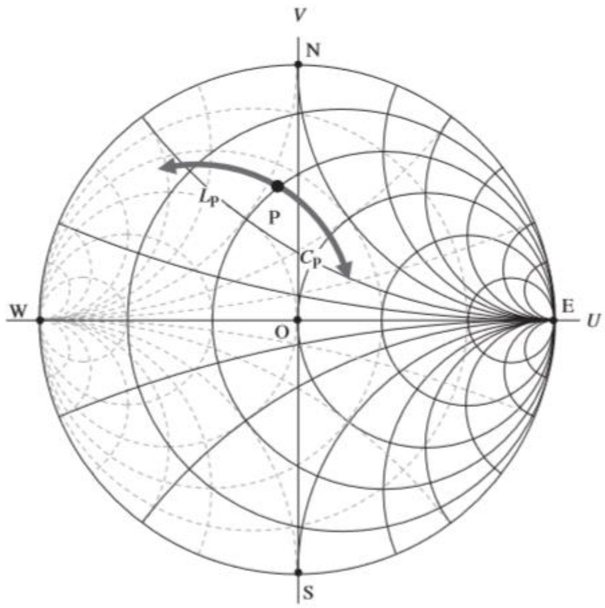


Figure. 4.4. The insertion effects of directions of an inductor or a capacitor in parallel on the Smith Chart (Li, 2013)

As a matter of fact, Figures 4.3 and 4.4 can be combined into Figure 4.5, where the insertion effects of directions of an inductor, a capacitor or a resistor in series and in parallel are also depicted, though it makes sense only theoretically and not in practical engineering design.

Theoretically,

- the original impedance P moves along the arc with x = constant to a higher resistance circle, if the insertion of the resistor R_S is in series, the value of the resistor defines the distance moved.
- the original impedance P moves along the arc with x = constant to a lower resistance circle, if the insertion of the resistor R_P is in parallel, the value of the resistor defines the distance moved.

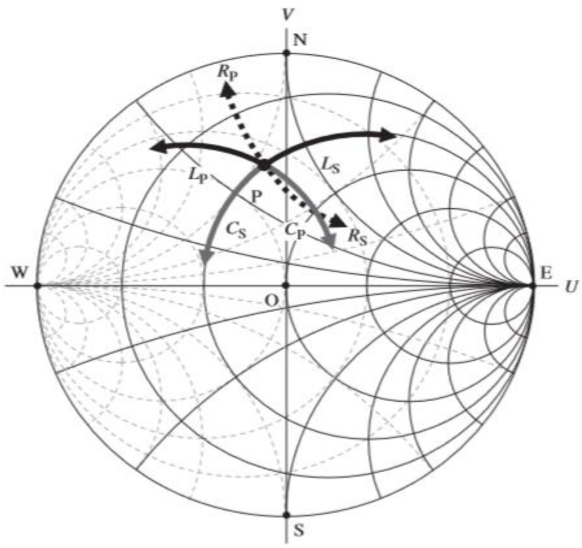


Figure. 4.5. The insertion effects of directions of an inductor, a capacitor or a resistor on the Smith Chart (Li, 2013)

In real life, to build an impedance matching network that simply contains one part is almost impossible for the inductor or the capacitor, the reason for this impossibility is the unrealizable derivation of the impedance of origin to the reference impedance (50-ohm), the only possibility with one part is the original impedance which must lie on both circles with r = 1 and g = 1, or in two narrow ring areas adjacent to both circles.

However, for the operating part of a narrowband system, there is the possibility of matching the original impedance to the desired value with 2 circuit parts. For some design of the circuit part of RF, the implementation of two-part impedance matching networks is quite common.

4.4 The Design of The Impedance Matching Network Created by Two Part

4.4.1 Sectors in a Smith Chart

For configuration of a matching network, the impedance of origin and the value of the impedance wanted is very important. Generally, if the impedance request is equal to the reference impedance (50-ohm), the location of the impedance of origin to make matching of the Smith Chart, that is the unique work which will guide the configuration of the matching of the impedance.

Following a method, to formulate impedance values, the Smith Chart can be divided into four sectors (Li, 2005). Not just the configuration, with this method suggest, the value of the parts in a matching network can be determined and calculated without difficulty. For the Smith Chart, the boundaries of the 4 sectors are illustrated in Figure 4.6, and Table 4.1 illustrates the expansion of impedance in the defined sectors.

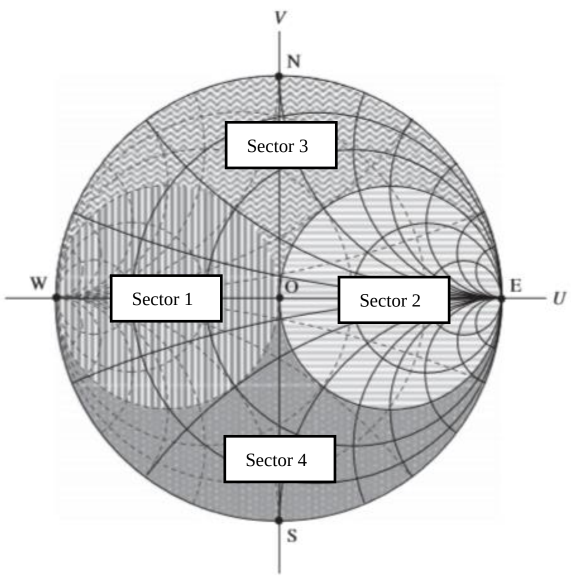


Figure. 4.6. The limits of four sectors on the Smith Chart (Li, 2013)

Table. 4.1. Range of impedance in four sectors on a Smith Chart (Li, 2013)

Sector 1	Sector 2	Sector 3	Sector 4
Low resistance or high conductance $r < 1$ $x < \|0.5\|$ $g > 1$ $-\infty < b < +\infty$	High resistance or low conductance $r > 1$ $-\infty < x < +\infty$ $g < 1$ $b < \|0.5\|$	Low resistance or low conductance $r < 1$ $x > 0$ $g < 1$ $b < 0$	Low resistance or low conductance $r < 1$ $x < 0$ $g < 1$ $b > 0$

4.5 The Design of The Impedance Matching Network Created by Three Part

For the three parts, a combination of capacitor and inductors, to configure the impedance matching network in three, this technique can reduce the scope of restrictions in a network configured in 2 parts, the only inconvenient will be the costs of another circuit part with respect to 2 parts.

4.5.1 ‘‘Pi’’ Type and ‘‘T’’ Type Configurations

All configuration of an impedance matching network created in three part, are classified into types Pi and T.

The impedance matching network of Pi-type and T-type, which consists of capacitor and inductor, has three parts, according to Table 4.2, there is a possibility of eight different configuration in Pi-type or T-type impedance matching networks

4.5.2 Recommended Configuration

Taking into account the cost.

Because of cost savings, there is the possibility of ignoring inductances of the configurations listed in Table 4.2, for an economical circuit design only configurations 1, 2, 3, and 5 will be considered, they are shown in Table 4.3.

Table. 4.2. Possible Configuration of a three-part (Li,2013)

	Pi Type	T Type
(1)	C_{P1}-C_S-C_{P2}	C_{S1}-C_P-C_{S2}
(2)	L_{P1}-C_S-C_{P2}	L_{S1}-C_P-C_{S2}
(3)	C_{P1}-L_S-C_{P2}	C_{S1}-L_P-C_{S2}
(4)	L_{P1}-C_S-L_{P2}	L_{S1}-L_P-C_{S2}
(5)	C_{P1}-C_S-L_{P2}	C_{S1}-C_P-L_{S2}
(6)	L_{P1}-C_S-L_{P2}	L_{S1}-C_P-L_{S2}
(7)	C_{P1}-L_S-L_{P2}	C_{S1}-L_P-L_{S2}
(8)	L_{P1}-L_S-L_{P2}	L_{S1}-L_P-L_{S2}

Table. 4.3. Applied and prohibited sectors of an impedance matching network of specific configurations (Li, 2013)

Configuration	Applied Sectors	Prohibited Sectors
Pi Type		
(1) C_{P1}-C_S-C_{P2}	Sector 3	Sector 1, 2, 4
(2) L_{P1}-C_S-C_{P2}	Sector 2, 3, 4	Sector 1
(3) C_{P1}-L_S-C_{P2}	Sector 1, 2, 3, 4	None
(5) C_{P1}-C_S-L_{P2}	Sector 1, 3	Sector 2, 4
T Type		
(1) C_{S1}-C_P-C_{S2}	Sector 3	Sector 1, 2, 4
(2) L_{S1}-C_P-C_{S2}	Sector 1, 3, 4	Sector 2
(3) C_{S1}-L_P-C_{S2}	Sector 1, 2, 3, 4	None
(5) C_{S1}-C_P-L_{S2}	Sector 2, 3	Sector 1, 4

Taking into account the availability of the topology.

For three-part impedance matching networks, there are restrictions as there were for two-part match matching networks.

For impedance matching networks with a special 3-parts configuration, the reference impedance (50-ohm) is only possible if the original impedance to be learned is in certain specific sectors of the Smith Chart. On the Smith Chart, for all three parts, there is a special configuration of the impedance matching network, this configuration is also shared in table 4.3, generally, Z_m the original impedance is in the 3 sectors 1, 2 or 4. Thus, following this information, the interesting configurations are 2, 3, and 5 for the Pi and T type matching networks. The configuration 1 is ignored because its forbidden sectors are 1, 2 and 4.

Take into account DC blocking, DC power supply and DC short-circuit.

For the Pi and T type impedance matching networks, the configurations to be considered are 2, 3 and 5. Table 4.4 illustrates the general performance of these plots on DC blocking, DC power supply and DC short circuit.

Following Table 4.4, it can be seen that in Pi type of impedance matching network, only configuration 2 (L_{P1}–C_S–C_{P2}) has no problem on DC blocking, DC feeding, and DC short-circuit. Configuration 3 (C_{P1}–L_S–C_{P2}) has a problem on DC blocking and in addition, it requires two RF chokes, or two $L_{infinitive}$ inductors. Configuration 5 (C_{P1}–C_S–L_{P2}) requires two RF chokes or two $L_{infinitive}$ inductors approaches to infinitive over the operating frequency range.

In T type of impedance matching network, the configuration 2 (L_{P1}–C_S–C_{P2}), 3 (C_{P1}–L_S–C_{P2}), and 5 (C_{P1}– C_S–L_{P2}) have no problem in DC blocking, DC feeding, and DC short-circuit. However, all of them need two RF chokes or two $L_{infinitive}$ inductors so that they become a four- or five-part impedance matching network indeed. Evidently, the Pi type of configuration 2 (L_{P1}–C_S–C_{P2}) is the best choice among the three-parts impedance matching networks.

Table. 4.4. Performance of the configurations 2, 3, and 5, for both Pi and T types on DC blocking, DC feeding, and DC Short-Circuit (Li, 2013)

Configuration	No Problem	Problems
Pi Type		
(2) L_{P1}-C_S-C_{P2}	No problem	
(3) C_{P1}-L_S-C_{P2}		DC blocking Two $L_{infinitive}$ are needed
(5) C_{P1}-C_S-L_{P2}		Two $L_{infinitive}$ are needed
T Type		
(2) L_{S1}-C_P-C_{S2}		Two $L_{infinitive}$ are needed
(3) C_{S1}-L_P-C_{S2}		Two $L_{infinitive}$ are needed
(5) C_{S1}-C_P-L_{S2}		Two $L_{infinitive}$ are needed

4.6 Parts in An Impedance Matching Network

One, two and three-parts configuration of passive impedance matching networks have been analyzed, it is almost impossible to configure a passive matching network with just one part. In general, narrowband designs are designed by practical engineering, for this type of design the configuration is realized in two or three parts. Due to a configuration limitation, the passive two-parts impedance matching network is likewise very popular. Of course, there is some advantage in the three-parts configuration, the configuration constraints are reduced quite seriously, and many choices of values and parts are usable.

Experimentally, for the design of a broadband matching network is more different, more than three parts of the circuit have the possibility to be used, and moreover, the passive components are not effective. After that, they almost always attenuate the signal with their resistive component. Many components in an impedance matching network have the potential to cause a significant signal attenuation.

Theoretically, the resistors can be used in an impedance matching network, but they never are, the reason is simple, they cause a attenuation and a rather important noise. Generally, capacitors and inductors are the only component of the configuration of impedance matching networks, as specified, the resistors are not used, in addition the capacitors are preferable to the inductors. Economically, for designs, using capacitors is more cost-effective than inductors. For an IC circuit design, space is also very important, for this reason, the surface of a capacitor is much smaller than that of an inductor, and in the end the Q value of an inductor is much smaller than that of an inductor. that of a capacitor in IC circuits. Following all the information revising, the capacitor is the most used passive component in the design of impedance matching network (Li, 2013).

5. THE PROPOSED APPROACH

5.1 The Rationale of The Proposed Approach

In literature, for the circuits in Figure 5.3, these can easily be designed by using the following well-known design equations (Bowick, 2008), and there are many works on this kind of matching networks. In (Sun and Fidler, 1994) (Sun and Fidler, 1996), the Q-based approach is studied systematically to design Pi type impedance matching networks; design formula based on the loaded Q are analytically obtained. In (Sun and Fidler, 1994) (Thompson and Fidler 2004) (Sun and Fidler, 1996), a method for determining the matching domain of impedance matching networks is presented. The studies are focused on the Pi networks, but the proposed method is applicable to the T networks. The analysis and design considerations for lumped element matching networks operating at high efficiency are presented in (Han and Perreault, 2006). In this work, only low-pass and high-pass single stage L matching networks are studied, since Pi and T matching networks have lower efficiency than an equivalent L matching network (Everitt and Anner, 1956).

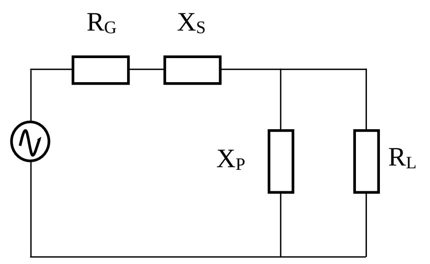


Figure. 5.1. A typical two-element L type matching network

Two-element L networks as seen in Figure 5.1 are the simplest and most widely used narrowband impedance matching circuits and can easily be designed via the following equations:

$$Q_S = Q_P = \sqrt{\frac{R_L}{R_G} - 1} \tag{5.1a}$$

$$X_S = Q_S * R_G \tag{5.1b}$$

$$X_P = \frac{R_L}{Q_P} \tag{5.1c}$$

In (Lee and Yang, 2012), a tunable impedance matching network is applied to achieve very widely tunable antenna. In this work, a T matching network is selected since Pi and T networks are confined to carry out high selectivity and a wide scope of tunable impedance dynamic ranges. An analytical tuning algorithm for a Pi network impedance tuner is proposed in (Gu et al., 2011). In Q-based design for T or Pi networks, the selection of an appropriate Q-factor is important. In (Liao et al.,2011), the relationship between reflection coefficient, load variation, and frequency drift is determined, which provides a reference for Q selection. In the Q-based approach described above to design two-element L networks and three-element Pi of T networks, the element values of the termination impedances are used. For example, if a capacitor and a resistor connected in parallel as load impedance, the capacitor and resistor element values must be known. But in practice, usually the load and generator impedances are given as measurement results, not as models with component values. Suppose that if an antenna as a load, the measurement values must be modeled at the desired frequency point. Then by using the constructed model, the Q-based approach can be implemented. But in the proposed approach, the load and generator element values are not needed; directly measurement results are used.

Generally, the generator and load impedances are defined as the measurement value at the desired frequency. Let us assume that they are measured as

$$Z_G = R_G + jX_G \tag{5.2a}$$

$$Z_L = R_L + jX_L \tag{5.2b}$$

These impedances can be assuming to be a resistor (R_G or R_L) and a reactive element connected in series, whose reactance is X_G or X_L at the corresponding frequency.

But it is necessary from equations (5.1a,b,c) and Figure 5.1 of the terminations is modeled in parallel configuration. For example, the load termination impedance (Z_L) can be modeled as a resistance ($R_{L,P}$) and a reactive element ($X_{L,P}$) connected in parallel as shown in Figure 5.2.

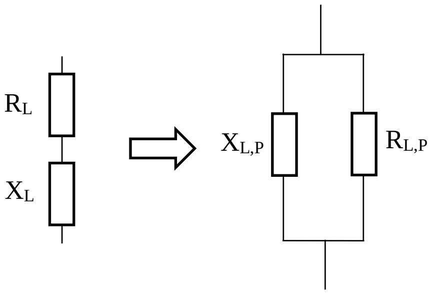


Figure. 5.2. Series to parallel transformation

The parallel resistance and reactance seen in Figure 5.2 and can be calculated as $R_{L,P} = R_L(Q^2_L + 1)$ and $X_{L,P} = X_L\left(\frac{Q^2_L + 1}{Q^2_L}\right)$ where $Q_L = \frac{X_L}{R_L}$, respectively.

Since the resistance of the series section must be equal to the resistance of the parallel section (Figure 5.3), the following equation can be derived,

$$R_G = \frac{R_L\,(Q^2_L + 1)}{1 + Q^2} \tag{5.3}$$

where Q is the quality factor of the series and parallel sections ($Q = Q_S = Q_P$) in Figure 5.3

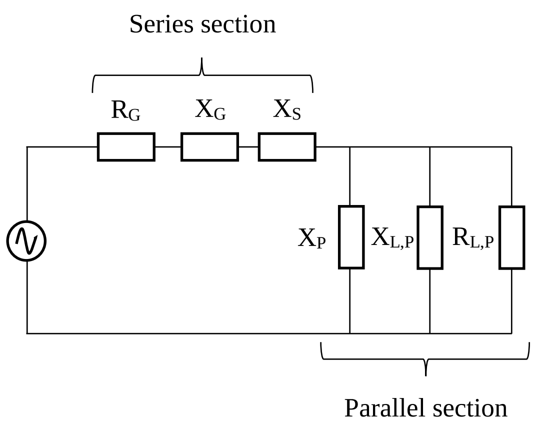


Figure. 5.3. L type network design

The qualify factor of the circuit seen in the Figure 5.3 can be found via (5.3) as

$$Q = Q_S = Q_P = \sqrt{\frac{R_L\,(Q^2_L + 1)}{R_G} - 1} \tag{5.4a}$$

Then the reactance values of the series and parallel sections can be calculated as

$$X_S = Q * R_G - \mu_1 * X_G \tag{5.4b}$$

$$X_P = \frac{R_L\,(Q^2_L + 1)}{Q - \mu_2 * Q_L} \tag{5.4c}$$

Here $\mu_1 = \pm 1$ and $\mu_2 = \pm 1$ constants must be assigned by the designer via Table 5.1

Table. 5.1. μ_1 and μ_2 constants

	μ_1	μ_2
X_P Capacitive, X_S Inductive	+1	-1
X_P Inductive, X_S Capacitive	-1	+1

Using the calculated reactances, the values of the components at the frequency concerned are obtained via the following equations,

$$C = \frac{1}{\omega X} \tag{5.5a}$$

$$L = \frac{X}{\omega} \tag{5.5b}$$

Therefore, it is now possible to design L networks via (5.4), (5.5) and Table 5.1 by directly using the measured impedance values of the generator and load impedances. There is no need to build generator and load models with component values.

5.2 The Examples of The Proposed Approach

5.2.1 The first example and its solution (L type)

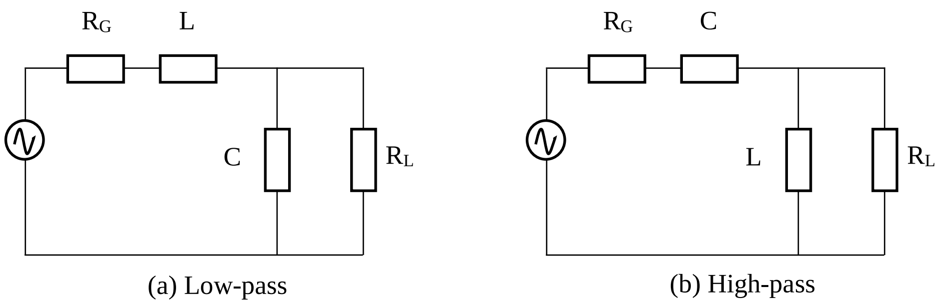


Figure. 5.4. L type network configuration

Suppose the generator and load impedances are measured at 75 MHz as $Z_G = 100 - j53.052\Omega$, $Z_L = 8.8044 + j93.4180\Omega$, respectively. If the series component is selected as an inductor ($\mu_1 = +1$) (so the parallel component will be a capacitor, $\mu_2 = -1$) as in Figure 5.4a. (R_G and R_L must be replaced with Z_G and Z_L, respectively), the calculated element values are L = 749.20 nH and C = 28.882pF.

The impedance value of the given generator can be modeled as a resistor ($R_G = 100\Omega$) and a capacitor (C_G = 40 pF) connected in series. Similarly, the given load impedance can be modeled as a resistor ($R_L = 1000\Omega$) and an inductor (L_L = 200 nH) connected in parallel. If the models obtained, the resonance method and (5.1) are used, the same values of inductance and capacitor are calculated for the network L. The power gain of the transducer of the matched system is given in Figure 5.5.

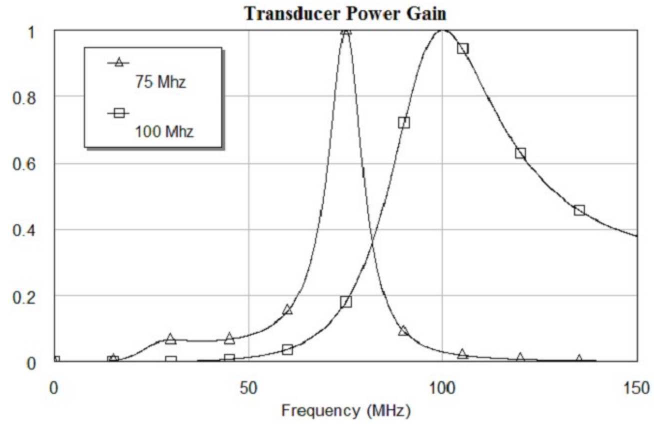


Figure. 5.5. Performance of the matched system with L type networks

Now suppose the generator and load impedances are measured at 100 MHz as Z_G = 100 + j125.66Ω, Z_L = 387.73 – j487.23Ω, respectively. If the network seen in Figure 5.4b (R_G and R_L must be replaced with Z_G and Z_L, respectively) is selected, then the series component will be a capacitor (μ_1 = -1) (so the parallel component will be an inductor, μ_2 = +1), the calculated element values are L = 373.90 nH and C = 3.7390 pF.

The impedance value of the given generator can be modeled as a resistor (R_G =100Ω) and an inductance (L_G = 200 nH) connected in series. Similarly, the given load impedance can be modeled as a resistor (R_L=1000Ω) and a capacitor (C_L=2 pF) connected in parallel. If the models obtained, the resonance method and (5.1) are used, the same values of inductance and capacitor are calculated for the network L. The power gain of the transducer of the adapted system is given in Figure 5.5.

5.2.2 The second example and its solution (Pi and T type)

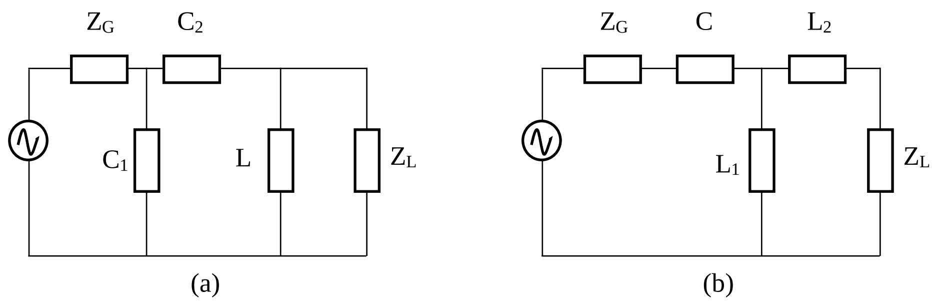


Figure. 5.6a. Designed Pi type network (C_1 = 11.341 pF, C_2 = 27.47 pF, L = 2.82 nH)
Figure. 5.6b. Designed T type network (C = 0.6210 pF, L_1 = 240.34 nH, L_2 = 54.4 nH)

Let us now design Pi and T networks. Suppose the generator and load impedance are measured at 1 GHz as $Z_G = 50 + j6.2832\Omega$, $Z_L = 71.6957 - j45.0477\Omega$, respectively. The desired Q is 5. After applying the proposed approach. The matching networks seen in Figure 5.6 are obtained. Transducer power gain curves of the matched systems are given in Figure 5.7.

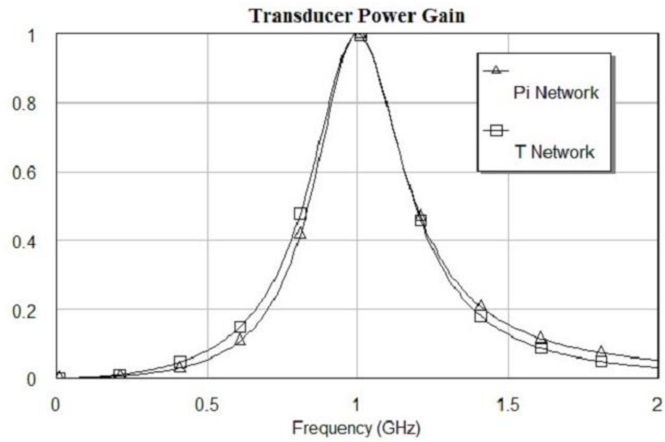


Figure. 5.7. Performance of the matched system with Pi and T types networks

Suppose at 1 GHz, the generator and load impedances are measured as $Z_G = 50 + j6.2832\Omega$, $Z_L = 71.6957 - j45.0477\Omega$ respectively. A Pi matching network with Q = 5 is desired.

The given generator impedance ($R_{G,P}$ =50.7896Ω) connected in parallel with an inductor ($L_{G,P}$ = 64.326 nH), and the given load impedance value can be modelled as a resistor ($R_{L,P}$ = 100Ω) connected in parallel with a capacitor ($C_{L,P}$ = 1 pF).

Since $R_{L,P} > R_{G,P}$, first calculate X_{S2} and X_{P2} value. The virtual resistor value can be found as

$$R = \frac{R_{L,P}}{Q^2 + 1} = \frac{100}{5^2 + 1} = 3.8462. \text{ Then}$$

$$X_{P2} = \frac{R_{L,P}}{Q} = \frac{100}{5} = 20\Omega$$

$$X_{S2} = Q * R = 5 * 3.8462 = 19.2308\Omega$$

Now calculate the Q of the L-side network of the generator as follows:

$$Q_{new} = \sqrt{\frac{R_{G,P}}{R} - 1} = \sqrt{\frac{50.7896}{3.8462} - 1} = 3.4936. \text{ Then}$$

$$X_{P1} = \frac{R_{G,P}}{Q_{new}} = \frac{50.7896}{3.4936} = 14.5379\Omega$$

$$X_{S1} = Q_{new} * R = 3.4936 * 3.8462 = 13.437\Omega$$

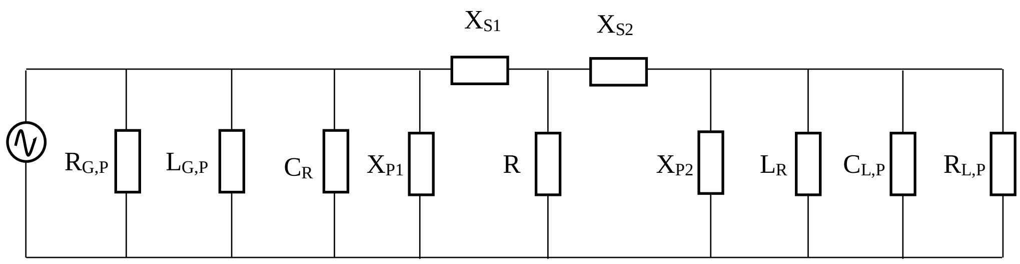


Figure. 5.8. Calculated element values for Pi type network.

As seen in Figure 5.8, connect a parallel capacitor (CR = 0.39378 pF) to resonate the inductor of the generator ($L_{G,\,P}$ = 64.326 nH). So X_{P1} must be a capacitor and X_{S1} must be an inductor. In addition, there is a need a parallel inductor (L_R = 25.33 nH) to resonate the charge capacitor ($C_{L,P}$ = 1 pF). So X_{P2} must be an inductor and X_{S2} must be a capacitor. Therefore, component values can be calculated as X_{P1} = 10.948 pF, X_{S1} = 2.1386 nH, X_{S2} = 8.2760 pF, X_{P2} = 3.1831 nH. After combining X_{P1} and C_R, X_{S1} and X_{S2}, X_{P2} and L_R, the values of the elements (X_1, X_2 and X_3, respectively) are obtained as in Figure 5.9 The power gain curve of the transducer of the adapted system is given in Figure 5.7.

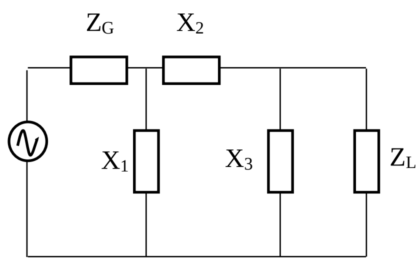


Figure. 5.9. Designed Pi type matching network (X_1 = 11.341 pF, X_2 = 27.47 pF, X_3= 2.8278 nH

Now, let us design a T network. The given generator and load impedance values can be modeled as a resistor ($R_G = 50\Omega$ and $R_L = 71,6957\Omega$) connected in series with an inductor (L_G = 1 nH) and a capacitor (C_L = 3.533 pF), respectively. Since $R_G < R_L$, first, calculate the values X_{S1} and X_{P1}. The value of the virtual resistance can be found as

$$R = R_G (Q^2 + 1) = 50 \times (5^2 + 1) = 1300\Omega. \text{ Then}$$

$$X_{P1} = \frac{R}{Q} = \frac{1300}{5} = 260\Omega$$

$$X_{S1} = Q * R_G = 5 * 50 = 250\Omega$$

Now let us calculate the Q of the load side L network as follows:

$$Q_{new} = \sqrt{\frac{R}{R_L} - 1} = \sqrt{\frac{1300}{71.6957} - 1} = 4.1391. \text{ Then}$$

$$X_{P2} = \frac{R}{Q_{new}} = \frac{1300}{4.1391} = 314.0776\Omega$$

$$X_{S2} = Q_{new} * R_L = 4.1391 * 71.6957 = 296.755\Omega$$

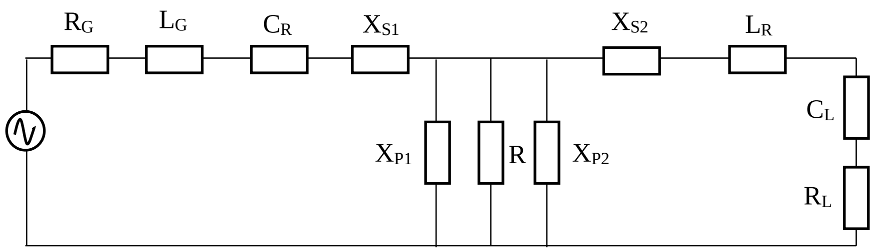


Figure. 5.10. Calculated element values for T type network.

As seen in Figure 5.10, connect a capacitor in series (C_R = 25.33 pF) to resonate the generator inductor (L_G = 1 nH). So X_{P1} must be an inductor and X_{S1} must be a capacitor. In addition, there is a need a series inductor (L_R = 7.1696 nH) to resonate the load capacitor (C_L = 3.533 pF). Thus, X_{P2} must be a capacitor and X_{S2} must be an inductor. Therefore, the component values can be calculated as X_{P1} = 41.38 nH, X_{S1} = 0.63662 pF, X_{S2} = 47.23 nH, X_{P2} = 0.50674 pF. After combining X_{S1} and C_R, X_{P1} and X_{P2}, X_{S2} and L_R, the values of the elements (X_1, X_2 and X_3, respectively) are obtained as in Figure 5.7

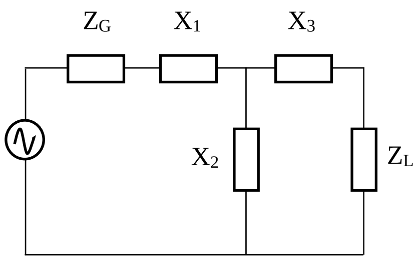


Figure. 5.11. Designed T type matching network (X_1 = 0.6210 pF, X_2 = 240.34 nH, X_3= 54.4 nH

Transducer power gain curve of the matched system is given by 5.7.

The given generator impedance value can be modeled as a resistor (R_G = 50Ω) and inductor (L_G = 1 nH) connected in series. Similarly, the given load impedance can be modeled as a resistor (R_L = 100Ω) and a capacitor (C_L = 1 pF) connected in parallel. If Pi and T networks are designed via the Q-based approach exists in literature, the same inductor and capacitor values are calculated for the networks.

6. CONCLUSIONS

Two-element L type networks are the simplest and most widely used narrowband impedance matching circuits. If the termination impedances are purely resistive, then the Q method is used to design this type of networks. Since the conjugate matching condition can be satisfied by using two-element L type networks, it is possible to transfer the maximum power at the frequency concerned. But the circuit Q is fixed by the given termination impedances. Thus, if the designer wishes to select the circuit Q, three-element Pi or T type networks must be used.

Generally, the generator and load termination impedances are described using actual frequency measurement results, and these impedances can be modeled as a resistor and as a reactive element connected in series.

Therefore, before declaring to want to design Pi type matching networks, these series impedance models must be transformed into parallel models. But in the design of T type matching networks, the given series models can be used, there is no need for a transformation step. Then, since Pi and T type matching networks can be considered as two back-to-back L type networks with virtual resistor between two L type networks, they can be designed via Q based method which is well defined in the literature for the design of L type networks.

In this thesis, a new Q-based approach has been presented. It is now possible to design L type networks by directly using the measurement values of the termination impedances. It is not necessary to obtain termination models. The proposed approach is also applied to the design of three-element Pi and T types for narrower band applications

The use of the proposed approach has been illustrated by some examples. Following these examples, the purpose of the thesis has been completed. Compared to the example of the approach proposed to the examples of the existing method, the results have been shown that the networks conceived are the same as the networks designed by the existing approach in the literature.

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APPENDIX A

A.1 Participation Certificate

ELECO 2017

10th INTERNATIONAL CONFERENCE on ELECTRICAL and ELECTRONICS ENGINEERING

Dear: Gokmen Yesilyurt

We would like to thank you for presenting your paper entitled

22- Real Frequency Design of Pi and T Matching Networks with Complex Terminations

at ELECO 2017 International Conference on Electrical and Electronics Engineering on

30 November - 2 December 2017 in Bursa, TURKEY

Best regards

Remzi CINAR
Chairman of the Chamber of
Electrical Engineers Bursa Branch

Sait Eser Karlik
Session Chairman

Prof. Dr. Hakan KUNTMAN
Conference Chairman
ELECO 2017

The Chamber of Electrical Engineers Bursa Branch

Uludağ University Faculty of Engineering Department of Electrical and Electronics Engineering

Istanbul Technical University Faculty of Electrical and Electronics Engineering

The Scientific And Technological Research Council of Turkey

IEEE Turkey Section

A.2 Matlab Code to Design L Type Matching Network

```
clc
clear

m1=input('Select series element type: -1 for C,  +1 for L : ');
if m1==1
    m2=-1;
elseif m1==-1
    m2=1;
end

if m1==1
    disp(' ')
    disp('Series element is an inductor (L)')
    disp('Parallel element is a capacitor (C)')
else
    disp(' ')
    disp('Series element is a capacitor (C)')
    disp('Parallel element is an inductor (L)')
end

% Rload=1000;
% Cload=2e-12;
% Rgen=100;
% Lgen=200e-9;
% f=100e6;
% w=2*pi*f;
% ZL=Rload/(1+i*w*Cload*Rload);
% ZG=Rgen+i*w*Lgen;
```

```
Rload=1000;
Lload=200e-9;
Rgen=100;
Cgen=40e-12;

f=75e6;
w=2*pi*f;
ZL=(Rload*i*w*Lload)/(Rload+i*w*Lload);
ZG=Rgen+(1/(i*w*Cgen));

QL=imag(ZL)/real(ZL);
Qnew=sqrt(((real(ZL)*(QL^2+1))/real(ZG))-1);

XS=Qnew*real(ZG)-m1*imag(ZG);
XP=(real(ZL)*(QL^2+1))/(Qnew-m2*QL);

if m2==1
   L=XP/w
   C=1/(w*XS)
else
   L=XS/w
   C=1/(w*XP)
end
```

A.3 Matlab Code to Design Pi Type Matching Network

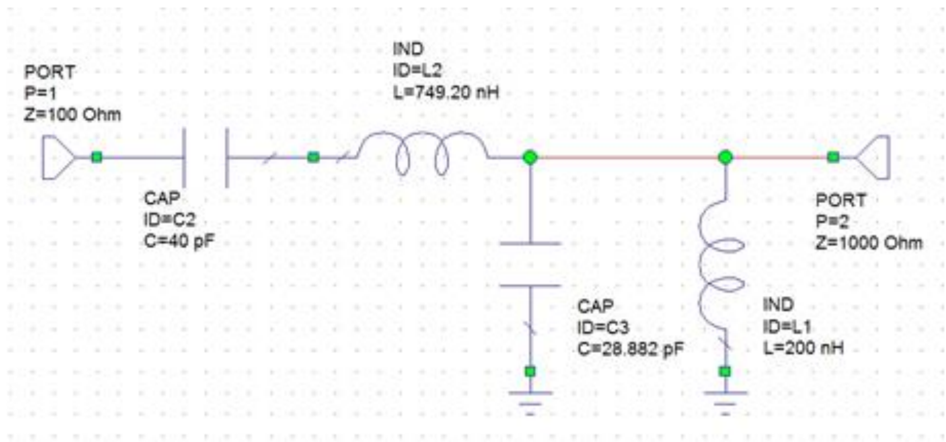
```
clc
clear

Rload=100;
Cload=1e-12;
Rgen=50;
Lgen=1e-9;

f=1e9;
w=2*pi*f;
ZL=Rload/(1+i*w*Cload*Rload);
ZG=Rgen+i*w*Lgen;
Q=5;

RG=real(ZG);
XG=imag(ZG);
RL=real(ZL);
XL=imag(ZL);
QL=XL/RL;
QG=XG/RG;

RLx=RL*(QL^2+1);
RGx=RG*(QG^2+1);

R=RLx/(1+Q^2);

XS2=Q*R
XP2=RLx/(Q-QL)

Qnew=sqrt(((RGx)/R)-1);
XS1=Qnew*R
XP1=RGx/(Qnew+QG)
```

end

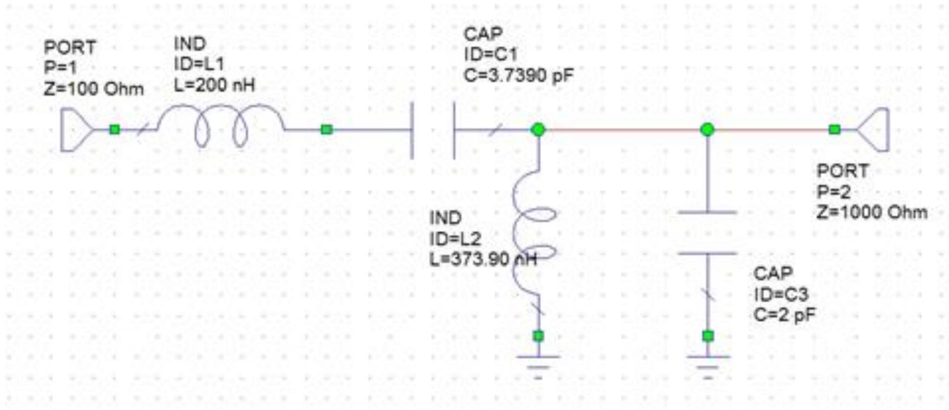
A.4 Matlab Code to Design T Type Matching Network

```
clc
clear

Rload=100;
Cload=1e-12;
Rgen=50;
Lgen=1e-9;
f=1e9;
w=2*pi*f;
ZL=Rload/(1+i*w*Cload*Rload);
ZG=Rgen+i*w*Lgen;
Q=5;

RG=real(ZG);
XG=imag(ZG);
RL=real(ZL);
XL=imag(ZL);


R=RG*(1+Q^2);

XS1=Q*RG+XG
XP1=R/Q

Qnew=sqrt((R/RL)-1);ΠΠ
XS2=Qnew*RL-XL
XP2=R/Qnew

end
```

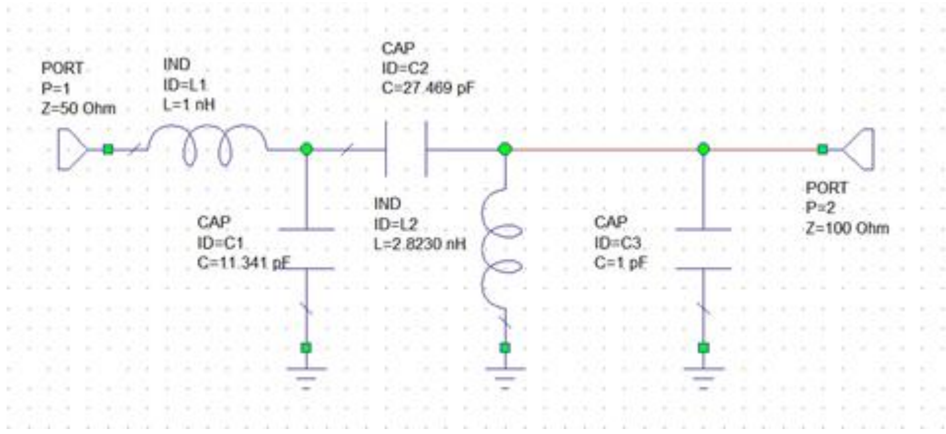
A.5 Schematic of L Type Matching Network Example (1) (AWR Microwave Office)



A.6 Schematic of L Type Matching Network Example (2) (AWR Microwave Office)



A.7 Schematic of Pi Type Matching Network Example (AWR Microwave Office)



A.8 Schematic of T Type Matching Network Example (AWR Microwave Office)