IŞIK UNIVERSITY
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CONSTRUCTION OF MULTIVARIABLE BROADBAND MATCHING NETWORKS: AN INTEGRATED DESIGN TOOL

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Abstract

One of the major problems in high frequency applications is designing matching circuits. Real frequency design of broadband matching networks is a commonly preferred approach since experimental termination data is directly utilized.

In this work, design of matching circuits with real frequency techniques was examined. In order to design mixed circuits with lumped and distributed components, an integrated design tool was implemented by combining parametric and scattering approaches.

This new approach was implemented as a MATLAB based design tool. This tool was integrated with .NET and a standalone, redistributable, user friendly general purpose matching circuit design tool was created.

ÇOK DEĞİŞKENLİ GENİŞBANT
UYUMLAMA DEVRELERİNİN TASARIMI: ENTEGRE BİR TASARIM ARACI

Özet

Yüksek frekans uygulamalarında, uyumlama devresi tasarımı temel problemlerden bir tanesidir. Uyumlama devresi tasarımında kullanılan gerçek frekans teknikleri deneysel yük verisini doğrudan kullanabilmekte olup yaygın olarak tercih edilen bir yaklaşımdır.

Bu çalışmada gerçek frekans yöntemleri ile uyumlama devresi tasarım yaklaşımları incelendi. Toplu ve dağılmış elemanlardan oluşan karma devrelerin tasarımı için empedans ve saçılım parametreli yaklaşımları birleştirilerek entegre bir tasarım aracı geliştirildi.

Geliştirilmiş olan yaklaşım MATLAB tabanlı bir tasarım yazılımı halinde gerçeklendi. Yazılım .NET ile entegre edilerek, tek başına dağıtılabilecek ve çalıştırılabilecek genel amaçlı bir uyumlama devresi tasarım programı elde edildi.

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Chapter 1
Introduction

One of the major problems in designing a communication system is to match a given device to the system so that the optimum power gain performance can be achieved over the broadest bandwidth possible. Engineers benefit from coupling circuits to design equalizers or matching networks that transfer a given impedance to a new one, which is called impedance matching or equalization.

The theory of broad-band matching began after the development of a gain-bandwidth theory for a restricted load impedance consisting of the parallel combination of a capacitor and resistor by Bode [1]. Fano and Youla generalized this to arbitrary loads [2,3]. After these developments, there have been lots of researchers who developed their techniques to solve broadband matching problems.

In order to solve matching problems, several computer programs have already been developed, however; since they are based upon brute force methods, they have some deficiencies to overcome the complicated broadband matching problems.

In 1977, the Real Frequency Technique, which eliminates most of the problems due to the analytic methods and purely numerical CAD techniques, was developed [4]. The most important property of this method is the direct utilization of the experimental real frequency load data in the design process. This made the new technique very attractive, because in almost all microwave matching problems, the device to be matched is usually described by numerical data obtained as a result of experimental measurements. Besides this technique, yields better design performance results with simpler structures over other techniques. Because of these advantages, most of the researchers have studied this technique and developed several methods to match a given device to load and generator impedances. Some of the well known techniques are the following:

- Line segment technique for modeling a complex load to a resistive generator
- Direct computational technique for solving double matching problem
- Broadband matching based on parametric representations of Brune functions
- Real frequency matching via scattering approach

In the past, all the available analytic and real frequency design techniques dealt with only lumped circuit components or only distributed networks with transmission lines alone in the matching network design. However, in practice, especially in the microwave discrete, monolithic or hybrid integrated circuit (MIC) designs, the physical realization of ideal lumped and distributed network elements presents serious implementation problems. For a complete characterization of MIC layouts, we need to model al the physical parameters and the possible parasitic effects inherent to the implementation process, and to take them into account in the design procedure. To do that, we must work on lumped, distributed and mixed elements in the network design.

It is shown that filter and matching designs with mixed lumped and distributed elements have advantageous over only lumped or only distributed designs. Unfortunately, there is no analytic solution for solving mixed matching problems, yet. Analytic solution for mixed design requires characterization of the mixed element structures using transcendental or multivariable functions. Although there are some analytic characterizations for restricted class mixed networks, the complete theory for the approximation and synthesis problems of mixed networks is still not available. So, one needs to solve mixed design problems not analytically but using computer aided design procedures.

The main objective of this study is to examine some of the real frequency techniques and understand the idea behind them and by combining the broadband matching based on parametric representations of Brune functions and real frequency matching via scattering approach to implement a standalone, user friendly tool that can design lumped, distributed and mixed matching and amplifier networks. The abilities of the tool that will be implemented are expected to be as follows;

- Lumped matching design.
- Distributed matching design.

- Mixed matching design with low pass ladders with unit elements (LPLUs).
- Multistage amplifier design with lumped, distributed or mixed lumped-distributed matching equalizers.

This tool will be able to synthesis the optimized lumped networks using Darlington synthesis [5] and distributed networks with transmission lines by using Richards Extractions algorithm [6] to get the component values.

In the second chapter, the matching problem is examined. Some fundamental information which will be used to derive some equations related to the matching problem is also given in second chapter. Scattering description, canonic representation of scattering matrix, parametric representation of driving point impedance, characterization of distributed networks and calculating transducer power gains are also the subjects that will be discussed in the second chapter.

The third chapter is devoted to multivariable description of mixed lumped and distributed two-port networks. In this chapter, mixed networks are examined and are characterized for designing matching and amplifier circuits. We propose a method that starts with parametric approach and turns again to scattering approach for designing networks consisting of low pass ladders with unit elements.

In the fourth chapter, broadband amplifiers are discussed. At the beginning of the chapter, historical and introductory information about amplifiers were mentioned. Later, we studied how to solve matching active devices with the load and generator by using parametric approach.

The fifth section is devoted to a user friendly, standalone application that combines all the theory which is mentioned in this study. We present flow diagrams of graphical user interface and the engine that runs under that interface.

In the sixth section, integration with MATLAB and .NET is examined. A basic example is given to illustrate how to call dynamic-link libraries (dlls) that were compiled by MATLAB in .NET projects.

Chapter 7 is devoted to examples that were examined by the integrated matching tool.

Chapter 2
Broadband Matching of Complex Terminations

2.1 Matching Problem

What we call matching is, designing a lossless two-port network, consisting of passive elements, to be inserted between the complex generator and the load impedance, such that the transfer of power from source to the load is maximized over a prescribed frequency. The power transfer capability of the matching network is best measured with the transducer power gain which is defined as the ratio of power delivered to load to the available power from the generator.

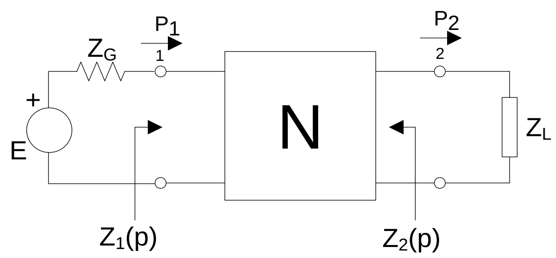


Figure 2.1 Broadband matching problem

Referring to Figure 2.1, ideal matching of the generator to the load impedance through a lossless network implies that $P_1 = P_2$. This can be done only when $Z_G = Z_1^*$, that is complex conjugate matching. We can conjugately match this network only at a single isolated frequency, but not over a frequency band. If we want to match this network over a finite frequency band, we will need to incorporate a tolerance on the match over a prescribed bandwidth so that the matching problem can be solved. The aim is to maximize the minimum value of transducer power gain over a required bandwidth, which implies a trade-off between transducer power gain and bandwidth.

The gain bandwidth limitations are mainly dictated by the given terminating impedances. So, the matching problem involves two main considerations;

Determining the gain-bandwidth limitations of given terminating impedances and finding the most practical equalizer structure to reach the best possible performance.

Depending upon the terminating impedances, the matching problem can be classified as single matching problem, double matching problem or active two-port problem:

- *Single matching problem* is the matching of purely resistive generator to the arbitrary complex load impedance.

- *Double matching problem* is the matching of an arbitrary complex load to an arbitrary complex generator.

- *Active two-port problem* is the matching of input and output of an active device to the complex load and generator impedances simultaneously. Typical application for this class of problem is the design of microwave amplifiers.

In the literature, these classes of matching problems have been formulated in terms of various network functions (scattering, impedance, transfer-scattering etc.) and tried to be solved analytically [1,2,3,7]. Analytic approach requires an explicit expression or a circuit realization for the terminating impedances to be matched. Also an analytic form of the gain function incorporating the complex terminations is required to be processed along with the analytic termination expressions so that theoretic gain-bandwidth restrictions are satisfied. These restrictions take the form of a set of integral expressions which should be solved simultaneously. Due to the difficulties in analytic approach the most complicated case treated in the literature is limited to that of a simple single matching problem with RLC load which is matched over a low pass band. For more complicated loads, the difficulties subject to the procedure are almost unmanageable.

In practical matching problems, one usually encounters with experimentally measured numerical real frequency data for the terminating impedances. A major problem at this point is the approximation of numerical data in analytic form, which is by no means easy and satisfactory.

Because of these reasons analytic approach is subject to severe restrictions and usually is not capable of handling complicated matching problems. Therefore, all the studies in the literature have addressed the task of determining more practical means for designing matching networks, which can directly process numerical load data. In this line the so called “Real Frequency Matching” techniques have been developed [8, 9, 10]. In the literature several variants of these techniques have been studied to solve matching and amplifier design problems [11, 12]. The main idea in these techniques is to characterize the unknown matching network in terms of a set of real parameters and determine these parameters, hence the matching network, by satisfying gain-bandwidth restrictions via a nonlinear optimization routine. Major advantages of these techniques can be stated as:

- Direct utilization of the experimental real frequency load and generator data without any need to approximate an analytic form.
- No need to assume a fixed equalizer network topology.

2.2 Scattering Description of Lossless Two-Ports

We can characterize the terminal behavior of lossless two-port networks by any one of the matrices such as impedance, admittance, chain and scattering matrices. Although impedance and admittance matrices are very useful, it is not guaranteed that the impedance or admittance of a given network exists, since they are defined with respect to the zero or infinite loadings at the ports. However, the scattering matrixes are defined with respect to the finite loadings at the ports, so one can be ensured of the existence of scattering matrixes for all networks.

Below you can find the basic definition and properties of scattering matrixes associated with the lossless two-port network.

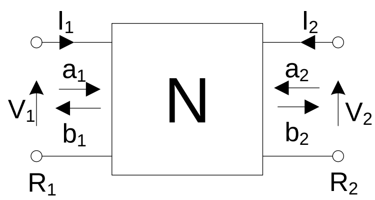


Figure 2.2 Definition of port variables in a two-port

Here, let N be a two-port network and R_i denotes a positive real normalizing resistance related to port i. The port variables, in the Figure, a_i and b_i are defined as

$$a_i = \frac{V_i + R_i I_i}{2\sqrt{R_i}}, \qquad b_i = \frac{V_i - R_i I_i}{2\sqrt{R_i}} \tag{2.1}$$

The variables, a_i and b_i, are the linear combinations of the voltage V_i and current I_i and they are called normalized incident and reflected waves, respectively. The inverse relationship follows:

$$V_i = (a_i + b_i)\sqrt{R_i} \qquad , \quad I_i = \frac{a - b}{\sqrt{R_i}} \tag{2.2}$$

The scattering matrix S of the two-port N is defined by

$$b = Sa \ , \quad \text{where} \quad b = \begin{pmatrix} b_1 \\ b_2 \end{pmatrix} \text{and } a = \begin{pmatrix} a_1 \\ a_2 \end{pmatrix} \ , \quad S = \begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix} \tag{2.3}$$

By using the equations above, we can acquire the scattering parameters as below.

$$S_{11} = \left.\frac{b_1}{a_1}\right|_{a_2=0}, \quad S_{12} = \left.\frac{b_1}{a_2}\right|_{a_1=0}, \tag{2.4}$$

$$S_{21} = \left.\frac{b_2}{a_1}\right|_{a_2=0}, \quad S_{22} = \left.\frac{b_2}{a_2}\right|_{a_1=0}, \tag{2.5}$$

$a_i = 0$ specifies that the port i is perfectly matched, i.e. port i is matched with the load R_i. S_{11} and S_{22} specify the input and output *reflectance*, respectively. S_{12} and S_{21} on the other hand specify the transmittance between the ports in the forward and reverse direction respectively.

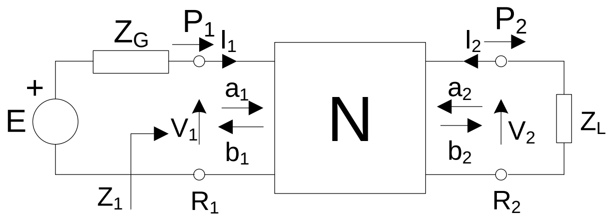


Figure 2.3 Doubly terminated two port

Referring to Figure 2.3, let Z_1 be the impedance seen at port 1 when port 2 is terminated with normalized resistor Z_L. If the voltage current relation at port 1, $V_1 = Z_1 I_1$, and the load condition at port 2 ($a_2 = 0$) are taken into account, we get

$$S_{11} = \frac{Z_1 - Z_G}{Z_1 + Z_G}, \tag{2.6}$$

which is the fundamental relation between the input impedance and the input reflectance of the two port.

Similarly, if we solve S_{12} by using (2.4), we get

$$S_{12} = \frac{V_2 - Z_L I_2}{V_1 + Z_G I_1}\sqrt{\frac{Z_L}{Z_G}} \tag{2.7}$$

Since $a_2 = 0,\ V_2 + Z_G I_2$ should be 0 to satisfy (2.1). This yields us the following equation

$$S_{21} = 2\sqrt{\frac{Z_G}{Z_L}}\frac{V_2}{E} \tag{2.8}$$

where $E = V_1 + Z_G I_1$. S_{12} is known as usual transmittance of the network. If we take the modulus square of S_{12}, we get power ration which is called transducer power gain of the two port.

$$|S_{21}|^2 = \frac{1}{4}\frac{\frac{|V_2|^2}{Z_L}}{\frac{|E|^2}{Z_G}} = \frac{P_2}{P_{A_1}} \tag{2.9}$$

where P_2 is the power delivered to port 2 and P_{A_1} is the available power from the excitation at port 1. Similarly, we can find S_{22} and S_{12} as follows,

$$S_{22} = \frac{Z_2 - Z_L}{Z_2 + Z_L} \tag{2.10}$$

And

$$S_{12} = 2\sqrt{\frac{Z_L}{Z_G}}\frac{V_1}{E'} \tag{2.11}$$

where $E' = V_2 + Z_G I_2$.

Referring to the Figure 2.3, it can easily be seen that $|a_1|^2$ is the available power delivered by the generator E and calculated as

$$|a_1|^2 = \frac{|E|^2}{4R_1} = P_{A_1} \tag{2.12}$$

Similarly, $|b_1|^2$ corresponds to reflected power at the port 1.

As we all know, $P_1 = ReV_1 I_1^*$, where * denotes the complex conjugation. If we multiply V_1, I_1 in the equation 2.2; what we get,

$$P_1 = |a_1|^2 - |b_1|^2 \tag{2.13}$$

This equation is same for port 2 and calculated as

$$P_2 = |a_2|^2 - |b_2|^2 \tag{2.14}$$

Thus, the total power dissipated, P_d in the two-port can be calculated as the difference of the total incident powers and reflected powers at both ports. That is

$$P_d = \sum_{i=1}^{2} |a_i|^2 - \sum_{i=1}^{2} |b_i|^2 \tag{2.15}$$

By using (2.3),

$$P_d = a^{*T}[I - S^{*T}S]a \tag{2.16}$$

For a passive network, the dissipated power, P_d is always non-negative for $Re\, p \geq 0$, where $p = \sigma + j\omega$. This implies that the matrix $I - S^{*T}S$ is also non-negative definite in $Re\, p \geq 0$.

If the port is lossless, $S^{*T}S$ should be equal to I for zero dissipated power that is.

$$S^{*T}S = I \text{ for } Re\, p \geq 0 \tag{2.17}$$

So, S of lossless two-port is unitary on the imaginary axis.

Due to the analytic continuation, it is also known that [13]

$$S_*^T S = I, \tag{2.18}$$

where lower asterisk denotes the paraconjugation of S.

For a lossless two-port network the scattering parameters satisfy the following conditions:

- The entries of $\boldsymbol{S}$ matrix are rational and real for real p
- S is analytic in $Re\, p \geq 0$.
- S is paraunitary satisfying $S_*^T S = I \;\; \forall$p
- If S is symmetric (S_{12}=S_{21}), then the lossless two-port is reciprocal.

Canonic Representation of Scattering Matrix

You can see the canonic representation of scattering matrix in terms of 3 canonic polynomials as below.

$$S = \frac{1}{g}\begin{pmatrix} h & \sigma f_* \\ f & -\sigma h_* \end{pmatrix} \tag{2.19}$$

where $f_* = f(-p)$ indicates the paraconjugate of a real function. The polynomials f, g and h have following properties.

- $f = f(p)$, $g = g(p)$ and $h = h(p)$ are real polynomials in the complex frequency p.
- g is strictly Hurwitz polynomial.
- f is monic, i.e. its leading coefficient is equal to unity.
- f, g and h are related by the following condition.

$$gg_* = hh_* + ff_*. \tag{2.20}$$

- σ is unimodular constant ($\sigma = \mp 1$).
- Degree of $h \leq$ degree of g.
- Degree of $f \leq$ degree of g.

The difference between the degree of g and the degree of f defines the number of transmission zeros at infinity and the degree of polynomial g referred to as degree of the two-port lossless network.

Computation of Transducer Power Gain

Referring to the doubly terminated two-port in Figure 2.3, the Transducer Power Gain can be expressed in terms of the unit normalized scattering parameters of the two-port and the source and load terminations as follows:

$$T(\omega) = \frac{(1 - |S_G|^2)(1 - |S_1|^2)}{|1 - S_G S_1|^2} \tag{2.21}$$

where S_1 is the unit normalized input reflection coefficient of the two-port when its output is loaded by Z_L.

$$S_1 = S_{11} + \frac{S_{12}S_{21}S_L}{1 - S_{11}S_L} \tag{2.22}$$

Combining these expressions we obtain

$$T(w) = \frac{(1 - |S_G|^2)(1 - |S_L|^2)|S_{21}|^2}{|1 - S_{11}S_G^2||1 - S_2 S_L|^2}, \tag{2.23}$$

where

$$S_G = \frac{Z_G - 1}{Z_G + 1}, S_L = \frac{Z_L - 1}{Z_L + 1}, S_2 = S_{22} + \frac{S_{12}S_{21}S_G}{1 - S_{11}S_G} \tag{2.24}$$

Substituting the polynomial forms of scattering parameters in Belevitch representation, the gain can directly be expressed in terms of the canonic polynomials f, g, and h as

$$T(w) = \frac{(1 - |S_G|^2)(1 - |S_L|^2)|f|^2}{|g - hS_G + \sigma S_L(h^* - S_G g^*)|^2} \tag{2.25}$$

where h^* implies complex conjugation.

2.3 Parametric Representation of Driving Point Impedance Functions

Referring to Figure. 2.4 the driving point impedance $Z_2(p)$ of a lossless two-port when its output is terminated in a resistor, can be used to characterize the network N. It is known that for a realizable network, the impedance $Z_2(p)$ is a positive real rational function (Brune function) in complex frequency variable p [14].

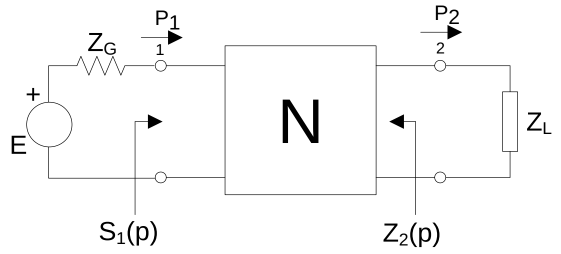


Figure 2.4 Broadband matching problem

In this technique we want to find $Z_2(p)$ to acquire all information about that network [4]. It has been shown that n^{th} order positive real impedance function $Z_2(p)$, can be written as a partial fraction expansion as

$$Z_2(p) = B_0 + \sum_{i=1}^{n} \frac{B_i}{p - p_i}, \tag{2.26}$$

where p_i's denotes the distinct poles of $Z_2(p)$ with $Re(p_i) < 0$ and B_i's are the corresponding complex residues. B_0 is a real constant.

The even part of $Z(p)$ is of the following form.

$$EvZ_2(p) = \frac{Z_2(p) + Z_2(-p)}{2} = \frac{f(p)f(-p)}{d(p)d(-p)}, \tag{2.27}$$

where $d(p)$ is the Hurwitz polynomial of $Z_2(p)$ and $f(p)$ is the real polynomial with the degree of not exceeding that of $d(p)$.

We can express $d(p)$ as

$$d(p) = D_n \prod_{i=1}^{n} p - p_i, \tag{2.28}$$

where

$$D_n = \prod_{i=1}^{n} \frac{-1}{p_i}. \tag{2.29}$$

In order to get a transformer-free network, we should design our network such that it matches at DC. That is the value of $Z_2(0) = 1$. If we choose D_n as above, we can provide $Z_2(0) = 1$.

We can calculate residues B_0 and B_i's as below.

$$B_i = -\frac{f(p_i)f(-p_i)}{p_i D_n^2 \prod_{\substack{k=1 \\ k \neq n}}^{n} (p_k^2 - p_i^2)} \tag{2.30}$$

and

$$B_0 = \begin{cases} 0, & if \deg f < n \\ \frac{1}{D_n^2}, & if \operatorname{def} f = n \end{cases} \tag{2.31}$$

We can calculate $f(p)$ as follows.

$$f(p) = p^{m_1} \prod_{i=0}^{m_2} (p^2 + a_i^2) \tag{2.32}$$

where m_1 and m_2 are nonnegative integers and a_i's are real coefficients. f polynomial provides us transmission zeros. So if we want to design a network with low pass elements only, due to the implementation problems, we should choose m_1=0 and m_2=0 to bring all transmission zeros to infinity. Below there are three possible network design and corresponding m_1's and m_2's

- m_1=0, m_2=0, $f(p) = 1$, all transmission zeros are at infinity , low pass design

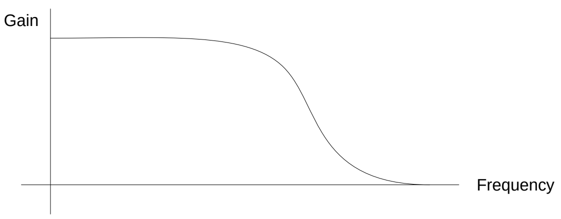


Figure 2.5 Low pass design

The difference between the $deg(g)$ and $deg(f)$ gives us the transmission zeros at infinity. In low pass design there are n transmission zeros at infinity. The more transmission zeros we have at infinity, the sharper corner we get at the transducer power gain curve.

- $m_1=n$, $f(p) = p^n$, all transmission zeros are at DC , high pass design

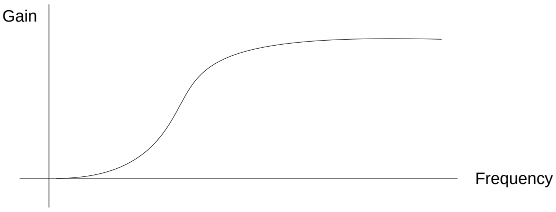


Figure 2.6 High pass design

The more transmission zeros at DC, the sharper corner we get at the transducer power gain curve.

- $m_1 \neq 0$, $m_1 < n$ some transmission zeros at DC, some at infinity, band pass design

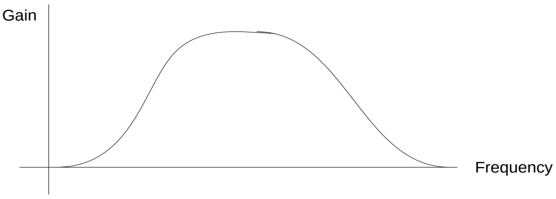


Figure 2.7 Band pass design

If we want to design a lumped network, difference of $deg(g)$ and $deg(f)$ gives the number of transmission zeros at infinity. This situation is same for the distributed design.

The roots of the polynomial $d(p)$ gives us the poles of $Z_2(p)$. All the roots are distinct and for all complex pole, there is a complex conjugate of that pole.

$$\begin{Bmatrix} p_i = -\alpha_i + j\beta_i \\ p_i^* = -\alpha_i - j\beta_i \end{Bmatrix}, for\ i = 1 \ldots . n_2 \text{ and } \qquad p_n = -\alpha_0 \text{ for n odd,} \tag{2.33}$$

where $\alpha_i > 0$, and n_2 is defined by

$$n_2 = \begin{cases} \dfrac{n}{2} & , if\ n\ is\ even \\ \dfrac{(n-1)}{2} & , if\ n\ is\ odd \end{cases} \tag{2.34}$$

By using these equations we can generate $Z_2(\omega)$ as follows.

$$R_2(w) = -\sum_{i=1}^{n} \frac{p_i B_i}{\omega^2 + \omega_i^2} + B_0 \tag{2.35}$$

$$X_2(\omega) = -\omega \sum_{i=n}^{n} \frac{B_i}{\omega^2 + p_i^2} \tag{2.36}$$

Note that $Z_2(\omega)$ is generated and it is not in polynomial form. It is actual impedance value in complex form. If we want to generate $Z_2(p)$ in polynomial form, we must write it as

$$Z_2(p) = \frac{n(p)}{d(p)} \tag{2.37}$$

where n is the numerator polynomial and d is the denominator polynomial. We find $n(p)$ and $d(p)$ by writing partial fraction expansion as one fraction. After finding n and d, we can find h and g polynomials as

$h(p) = n(p) - d(\mathrm{p})$ and $g(p) = n(p) + d(p)$. In order to weight h and g we multiply them by $D_n/2$.

It is also possible to add a foster a section before the network. In general, foster section can be written as

$$Z_F(jw) = jX_F(w) \tag{2.38}$$

where $X_F(w)$ is defined in partial fraction expansion form as

$$X_F(w) = b_\infty \omega + \frac{b_0}{\omega} + \sum_{i=1}^{N_F} \frac{2b_i\omega}{\omega_i^2 - \omega^2} \tag{2.39}$$

and residues (b_∞, b_0, b_i) are all nonnegative real constants. For practical low-pass and band-pass design, it is sufficient to consider only a series of branches whose singularities are located at 0 and ∞. That is

$$X_F(w) = b_\infty \omega + \frac{b_0}{\omega} \tag{2.40}$$

Since foster section is a purely imaginary value, it changes the imaginary part of the $Z_2(\omega)$ as

$$X_2(\omega) = X_F - \omega \sum_{i=n}^{n} \frac{B_i}{\omega^2 + p_i^2} \tag{2.41}$$

When foster section is inserted, n and d polynomials also change and this must be taken into account.

After setting all parameters and calculating residues, n and g polynomials, we can write gain expression as

$$T(\omega) = \frac{(1 - |S_G|^2)(1 - |S_1|^2)}{|1 - S_G S_1|^2} \tag{2.42}$$

where

$$S_1 = \frac{H}{H_*} \frac{Z_L - Z_2^*}{Z_L + Z_2} \tag{2.43}$$

S_G is the reflection at the generator

$$S_G = \frac{Z_G - 1}{Z_G + 1} \tag{2.44}$$

and

$$H = \frac{f_*}{d} \tag{2.45}$$

Characterization of Distributed Networks

At microwave frequencies it is not easy to realize lumped elements. So, distributed networks composed of transmission lines are required. The idea, in order to design distributed network, is similar to the one in the lumped case. However, there must be some modifications in the f polynomial and the complex frequency p.

- p $(p = \sigma + j\omega)$ should be replaced by Richard variable $(\lambda = \Sigma + j\omega)$.

where $\lambda = tanhp\tau$, and τ is the commensurate one-way delay of the transmission lines. For a distributed network

- $f(p)$ should be replaced by $f(\lambda)(1 - \lambda^2)^{\frac{u}{2}}$, Here u is the number of commensurate lines. With the use of Richards variable in network functions all the realizability conditions of a network consisting of transmission lines remains unchanged and the impedance or scattering based representation of a distributed two-port can be done in the new Richards domain following the same lines as discussed above.

2.4 Real Frequency Design of Matching Network

In the real frequency design of a lossless matching network to match complex terminations which are defined in real frequency data, we may utilize the scattering or impedance based two-port network functions to optimize the gain of the system. In the optimization step a linear least square sense error minimization for the gain of the system is utilized. In the scattering based description, although the numeric of the problem is well behaved due to the unity bounded scattering functions, the procedure inherently involves polynomial factorization steps. In the impedance based parametric approach on the other hand, the two-port driving point impedance and hence the gain function can be explicitly expressed in terms of the singularities of the network to be designed. This advantage makes the impedance based parametric approach the most efficient tool for real frequency design of matching networks. Therefore, the design algorithms studied in this work are mainly based on the parametric representation approach.

2.4.1 Parametric Real Frequency Design Algorithm

Given the real and imaginary part measurement data for the termination impedances over the frequency band of interest, the design algorithm involves the following computational steps:

Step 1: Decide on the complexity of the matching network; degree and type of the network (lumped or distributed),

Step 2: Topology of the network type and location of transmission zeros (high pass, low pass or band pass) are defined by setting m_1 and m_2 values used in 2.32.

Step 3: According to degree of the network, poles of the positive real impedance function are computed as follows.

$\begin{Bmatrix} p_i = -\alpha_i + j\beta_i \\ p_i^* = -\alpha_i - j\beta_i \end{Bmatrix}, for\ i = 1 \ldots . n_2$ and $p_n = -\alpha_0$ for n odd,

where

$$n_2 = \begin{cases} \dfrac{n}{2} & , if\ n\ is\ even \\ \dfrac{(n-1)}{2} & , if\ n\ is\ odd \end{cases}$$

User should enter the input variables α_i, β_i, so that poles can be generated.

Step 4: If desired, foster sections are given to algorithm and we get an input vector for generating impedance function like

$$\Theta = [\propto_0, \{\alpha_i, \beta_i, i = 1, \ldots n_2\}, \{b_\infty, b_0\}]$$

Step 5: A least square optimization routine is called and tried to converge a prescribed desired gain level, defined by equation 2.42, by changing the parameter values of the input vector which are used to generate poles of the impedance function. At this step the objective function to be minimized may be chosen as

$$\delta = \sum_{i=1}^{N_W} [T(w_i, \Theta) - T_0]^2 \tag{2.46}$$

where N_W denotes the number of frequencies over the pass band and T_0 is the desired flat gain level.

Step 6: Once the nonlinear search routine terminated and the optimum set of pole values are found, generate the driving point impedance function and obtain its realization employing Darlington synthesis.

The parametric real frequency design approach outlined above has following substantial advantages for the construction of matching networks.

- Since the analytic form of impedance function is directly processed the realizability of the resulting impedance function is automatically ensured.
- Direct control of the transmission zeros and hence the topology of the matching network is possible.

- Since the optimization is carried out directly on the poles of the impedance function and the residues are explicitly computed the method offers high numerical stability. There exist no need to polynomial factorization and numerical inaccuracies.

- The gain function can be expressed explicitly in terms of the unknown poles of the system. Thus explicit determination of the derivatives of the objective function with respect to optimization parameters is possible if a gradient based search routine is used, which improves the computational efficiency and convergence.

2.4.2 Matching Examples

In the following the application of the real frequency parametric approach is illustrated by two matching network design examples.

Example 1: Design of a lumped matching network between the generator and the load impedances defined as shown in Figure 2.8.

The purpose is to match Z_G to Z_L over a prescribed bandwidth to maximize the power gain.

In this example, the generator impedance Z_G consists of a resistance 1Ω in series with an inductance which is 1 Henry. The load impedance Z_L is given by a parallel R_L and C_L where R_L=1Ω and C_L=1F in series with an inductance L_L which has the value of 2H (Figure 2.8).

The frequency band of the double matching problem is taken over B: $0 \leq w \leq 1$.

Step 1: *Topology of the network and initial guess for input parameters of the algorithm.*

Complexity of the network is taken as three and the poles of the $Z_2(p)$ are given initially as

$p_0 = -0.2 + j0.4$

$p_2 = -0.2 - j0.4$

$p_2 = -0.6 + j0.8$

$p_3 = -0.6 - j0.8$.

Step 2: *Optimization of the transducer power gain.*

In the optimization, the flat-gain level T0=0.98 is picked over the frequency band $0 \le w \le 1$, and Z(p) is computed by using (2.37) as follows.

$$Z_2(p) = \frac{0.7004p^3 + 1.6774p^2 + 1.5166p + 0.5450}{p^4 + 2.3949p^3 + 2.4614p^2 + 1.4873p + 0.5450}$$

By using $Z_2(p)$, we can find the optimized poles as below.

$p_0 = -0.2306 + j\ 0.6821$

$p_1 = -0.2306 - j0.6821$

$p_2 = -0.9669 + j0.3411$

$p_3 = -0.9669 - j0.3411$

Step 3: Finally $Z_2(p)$ is synthesized by using zero shifting algorithm described in [5] and the values of the capacitors and inductors are found as follows.

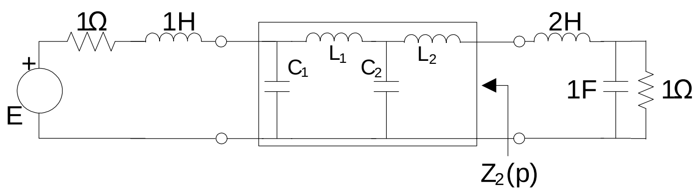


Figure 2.8 Lumped matching example

C_1=1.4277F, C_2=1.33033F, L_1=2.3652F, L_2=0.4203F

The transducer power gain of the matching design is depicted in the Figure 2-9.

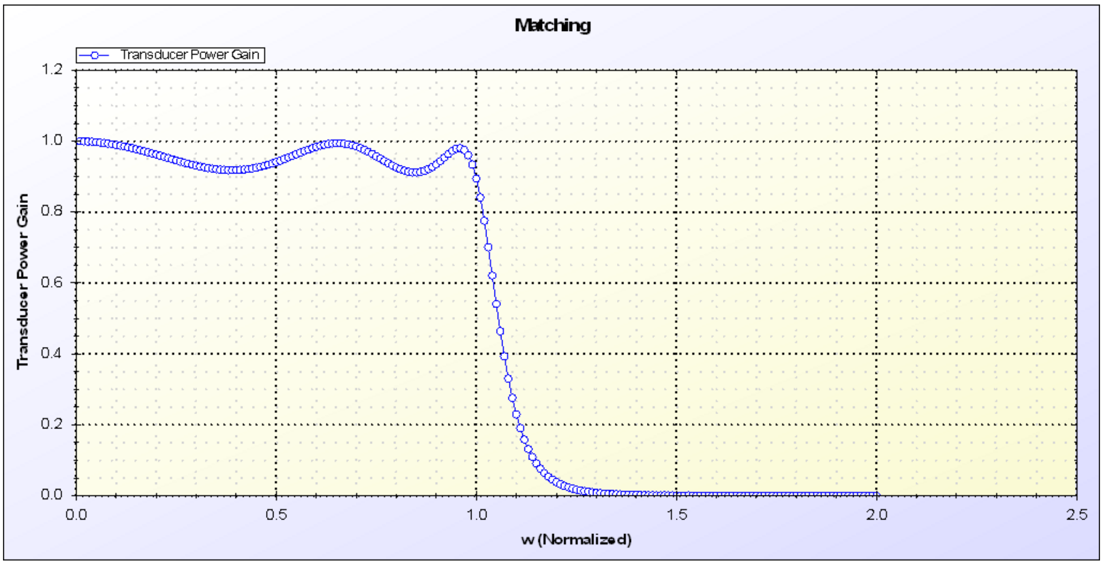


Figure 2.9 Lumped matching example transducer power gain

In this design, we get an average gain and minimum gain as 0.9551 and 0.8947, respectively

Example 2: Design of a distributed matching network between the generator and the load impedances defined as shown in Fig 2-10

The purpose is to match Z_G to Z_L over a prescribed bandwidth to maximize the power gain.

In this example, the generator impedance, $Z_{G,}$ and the load impedance, $Z_{L,}$ are same with the previous example except that we replaced p with λ .

The frequency band of the double matching problem is taken over B: $0 \leq w \leq 1$.

Step 1: *Topology of the network and initial guess for input parameters of the algorithm.*

Complexity of the network is taken as three. Because of the low pass design, we put three transmission zeros to infinity. In other words, the number of transmission lines is taken as three. Commensurate delay length, is taken as 0.707.

The poles of the $Z_2(\lambda)$ are given initially as

$\lambda_0 = -0.43$

$\lambda_1 = -0.269 + j0.88$

$\lambda_2 = -0.269 - j0.88$

Step 2: *Optimization of the transducer power gain.*

Before begin the optimization process, we changed p ($p = \sigma + j\omega$) with ($\lambda = \Sigma + j\omega$) and replaced $f(p)$ with $f(\lambda)(1 - \lambda^2)^{\frac{u}{2}}$

In the optimization, the flat-gain level T0=0.97 is picked, as we did in lumped case, over the frequency band $0 \le w \le 1$, and $Z(\lambda)$ is computed by using (2.37) as follows.

$$Z_2(\lambda) = \frac{0.0330\lambda^3 + 0.7315\lambda^2 + 0.8609\lambda + 0.1818}{\lambda^3 + 1.3373\lambda^2 + 0.5520\lambda + 0.1818}$$

The optimized poles are found as.

$\lambda_0 = -0.9595$

$\lambda_1 = -0.1889 + j0.3921$

$\lambda_2 = -0.1889 - j0.3921$

Step 3: Finally $Z_2(\lambda)$ is synthesized by using zero shifting algorithm described in [5] and the values of the transmission lines are found as indicated in Figure 2.10.

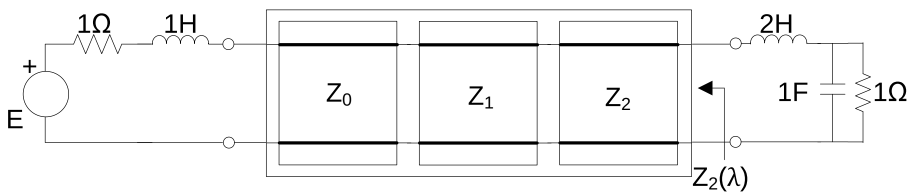


Figure 2.10 Single stage distributed matching circuit

Z_0=0.5885Ω, Z_1=0.1093Ω, Z_2=0.0338Ω

Below, transducer power gain is depicted.

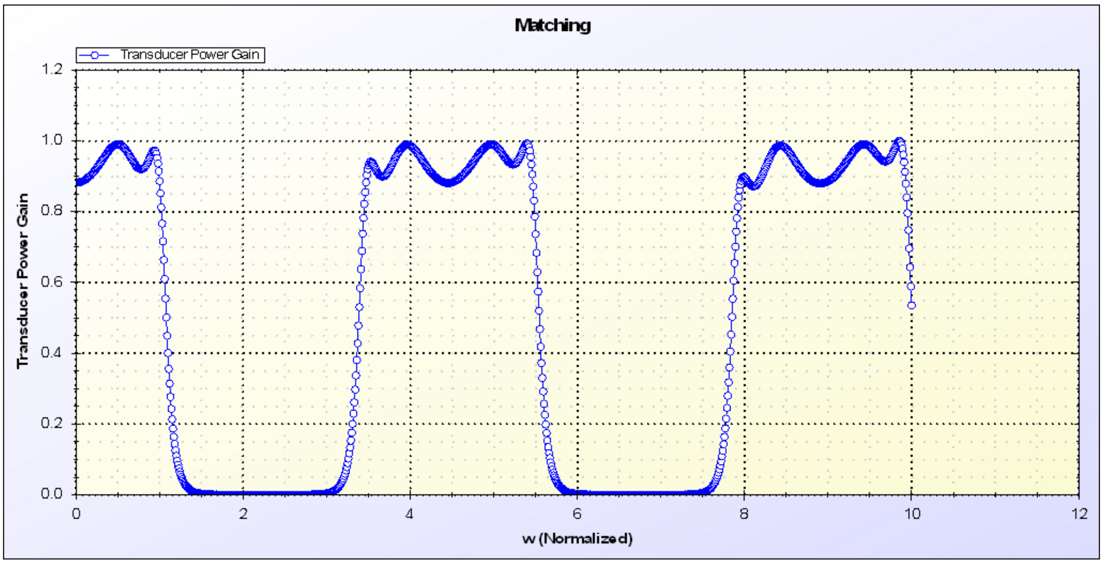


Figure 2.11 Single stage distributed matching transducer power gain

In this design, we get an average gain and minimum gain as 0.9434 and 0.8781, respectively. The gain response is periodic as expected from a distributed type matching network.

Chapter 3
Multivariable Description of Mixed Lumped and Distributed Two-Port Networks

3.1 Introduction

Up to know, we consider only lumped or only distributed networks. Now we consider the networks that have lumped and distributed components at the same time. At microwave and millimeter wave frequencies, use of lumped elements alone in the circuit realization presents serious implementation problems, because of the difficulties regarding the physical interconnection of components and the associated parasitic effects. Therefore, it is inevitable to use distributed structures composed of transmission lines between the lumped elements. Since these transmission lines are also considered in the design process, the performance of the network will be better. So it can be concluded that the cascade of reciprocal two-port networks connected by means of equidelay ideal transmission lines constitutes a useful model.

Designing mixed lumped and distributed element networks has gained a great deal of importance for a long time in the literature. But, a complete design theory for the mixed element networks still does not exist. Although some classical network concepts have been extended to cover some classes of mixed element two-ports, the problem of approximation and synthesis of arbitrary mixed element networks could not be resolved completely.

In literature, a special interest has been devoted to the mixed element networks composed of lumped reactances and ideal uniform lossless transmission. That is, the structure of interest consists of cascaded lossless lumped two-ports connected with ideal transmission lines (UEs).

Microwave filters and matching networks composed of this kind of cascaded structures have obviously the properties of both lumped and distributed networks and offer advantages over those designed with lines or lumped elements alone. One of the most important advantages is the harmonic filtering property of the mixed structure. Furthermore, the required physical circuit interconnections are provided by nonredundant transmission line elements which also contribute to the filtering performance of the structure. In this case, description of the distributed sections can be made in terms of Richards variable λ, ($\lambda = \tanh(p\tau)$), whereas the lumped sections are described in terms of the complex frequency variable p. In mathematical terms, description of lossless two-ports constructed with lumped and equal length transmission lines can be made by using complex two-variable functions. In fact, since the complex variables p and λ are not independent, the problem is actually a single variable one. However, if one assumes that p and λ are independent variables, then the problem can be treated using multivariable functions.

In this section two-variable characterization of mixed lumped and distributed element cascaded networks will be discussed. For the real frequency design of such matching networks a semi analytic construction approach will be investigated.

3.2 Construction of Lumped-Distributed Two-Ports

A typical distributed two-port with lumped discontinuities can be modeled as a lossless two-port formed with cascade connections of lumped elements and commensurate transmission lines (Unit Elements) as shown in Figure 3.1.

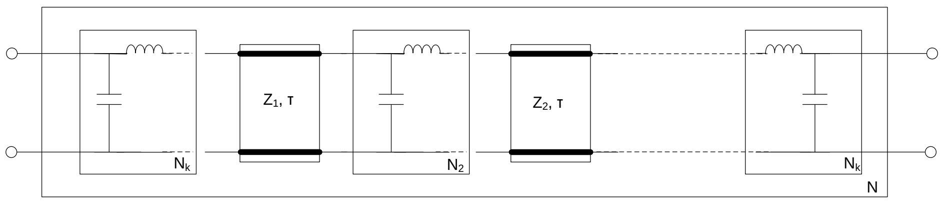


Figure 3.1 Generic form of lumped-distributed two-ports

Lossless two-ports, constructed with lumped and distributed elements at the same time, can be described in terms two variable scattering matrix $= S(p, \lambda)$. Here, $p = \sigma + j\omega$ is the complex frequency variable associated with lumped elements and $\lambda = \Sigma + j\omega$ is the Richard variable, associated with the equal length transmission lines, also called Unit Elements (UE). In Belevitch representation, the scattering parameters are given by,

$$S_{11} = \frac{h(p,\lambda)}{g(p,\lambda)}, \qquad S_{12} = \sigma \frac{f(-p,-\lambda)}{g(p,\lambda)}$$
$$S_{21} = \frac{f(p,\lambda)}{g(p,\lambda)}, \qquad S_{22} = -\sigma \frac{h(-p,-\lambda)}{g(p,\lambda)} \tag{3.1}$$

or

$$\mathrm{S} = \frac{1}{g}\begin{pmatrix} h & \sigma f_* \\ f & -\sigma h_* \end{pmatrix} \tag{3.2}$$

where $f_* = f(-p,-\lambda)$ indicates the paraconjugate of a real function.

$g(p,\lambda)$ and $h(p,\lambda)$ can be defined as;

$$g(p,\lambda) = \sum_{i=0}^{n_\lambda} g_i(p)\lambda^i \text{ and } h(p,\lambda) = \sum_{i=0}^{n_\lambda} h_i(p)\,\lambda^i \tag{3.3}$$

$$g_i(p) = \sum_{j=0}^{n_p} g_{ji} p^j \text{ and } h_i(p) = \sum_{j=0}^{n_p} h_{ji}\, p^i \tag{3.4}$$

where n_p and n_λ are the total number lumped and distributed elements respectively, or equivalently the two variable polynomials can be expressed as

$$g(p,\lambda) = p^T \wedge_g \lambda = \lambda^T \wedge_g^T p \tag{3.5}$$

where

$$\wedge_g = \begin{pmatrix} g_{00} & g_{01} & \cdots & g_{0n_\lambda} \\ g_{10} & g_{11} & \cdots & \cdot \\ \vdots & \vdots & \ddots & \vdots \\ g_{n_p 0} & \cdot & \cdots & g_{n_p n_\lambda} \end{pmatrix}, \quad \begin{matrix} p^T = \left(1\; p\; p^2 \ldots p^{n_p}\right) \\ \lambda^T = \left(1\; \lambda\; \lambda^2 \ldots \lambda^{n_\lambda}\right) \end{matrix} \tag{3.6}$$

and,

$$h(p,\lambda) = p^T \wedge_h \lambda = \lambda^T \wedge_h^T p \tag{3.7}$$

where

$$\wedge_h = \begin{pmatrix} h_{00} & h_{01} & \dots & h_{0n_\lambda} \\ h_{10} & h_{11} & \dots & . \\ \vdots & \vdots & \ddots & \vdots \\ h_{n_p0} & . & \dots & h_{n_pn_\lambda} \end{pmatrix}. \tag{3.8}$$

In this representation, sub matrix, obtained by deleting the first row and first column of the $\wedge_h$, is the connectivity matrix of the numerator of the input reflectance. Whereas, the sub matrix, obtained by deleting the first row and first column of the $\wedge_g$ is called the connectivity matrix of the denominator of the input reflectance.

If the two-port includes UE's in cascade mode, then f is defined as

$$f = f_0(p, \lambda)(1 - \lambda^2)^{\frac{u}{2}} \tag{3.9}$$

where u denotes the number of unit elements (UE).

In the above representation, n_p is the number of lumped elements and n_λ is the number of unit elements. Since the two-port is lossless,

$$S(p, \lambda)S^T(-p, -\lambda) = I \tag{3.10}$$

where I is the unity matrix. Employing the Belevitch form and the losslessness condition we have,

$$gg_* = hh_* + ff_*. \tag{3.11}$$

Again, lower asterisk denotes paraconjugation.

The polynomials f, g and h have following properties.

- $f = f(p, \lambda)$, $g = g(p, \lambda)$ and $h = h(p, \lambda)$ are real polynomials in the complex frequency p and λ.

- g is strictly Hurwitz polynomial.
- f is monic, i.e. its leading coefficient is equal to unity.
- σ is unimodular constant ($\sigma = \mp 1$).

The input impedance $Z_1(p,\lambda)$ of the lossless two-port when its output is terminated in a resistor is written as

$$Z_1(p,\lambda) = \frac{g(p,\lambda) + h(p,\lambda)}{g(p,\lambda) - h(p,\lambda)} = \frac{n(p,\lambda)}{d(p,\lambda)} \tag{3.12}$$

where the polynomials $n(p,\lambda)$ and $d(p,\lambda)$ can as well be expressed in similar forms of (3.3 and 3.4).

When some dealing with mixed network design, the following boundary cases may be remarked:

a) For the mixed element lossless two-ports, the polynomial $f(p,\lambda)$ defines the transmission zeros of the cascade and it is given by

$$f(p,\lambda) = f_0(p) f_1(\lambda) \tag{3.13}$$

where, $f_0(p)$ and $f_1(\lambda)$ contains the transmission zeros due to the lumped sections and unit elements in the cascade, respectively. If, n_λ unit elements are considered in the cascade topology, then $f_1(\lambda) = (1-\lambda^2)^{\frac{n_\lambda}{2}}$. In most of the practical cases, $f_0(p)$ is chosen as an even/odd real polynomial, which corresponds to reciprocal lumped structures. A particularly useful case is obtained for $f_0(p) = 1$, which corresponds to a low-pass structure having transmission zeros only at infinity. Another practical case is to choose $f_0(p) = p^{n_p}$, which corresponds to a high-pass structure.

b) When the distributed elements, transmission lines, are removed from the mixed structure, one would end up with a lumped network whose scattering matrix can fully be described independently in terms of the canonical real polynomials $h(p,0)$ and $g(p,0)$. So, for the topologies where all the transmission zeros are at the infinity, we can generate a lumped-only network with parametric approach and use the $h_0(p)$ and $g_0(p)$ as the first column of

$\wedge_h$ and $\wedge_g$, respectively. In this case $g_0(p)$ is strictly Hurwitz and the losslessness condition holds.

$$g_0(p)g_0(-p) = h_0(p)h_0(-p) + f_0(p)f_0(-p) \tag{3.14}$$

The resulting lumped element two-port can also be described in terms of the driving point function $Z_1(p,0)$, which can be assumed to be minimum reactive and has only simple poles, (A Foster function can always be added to end up with a more general positive real impedance function). In this case, $Z_1(p,0)$, can be expressed as follows: [7]

$$Z_1(p,0) = \frac{n(p,0)}{d(p,0)} = B_0 + \sum_{i=1}^{n} \frac{B_i}{p - p_i} \tag{3.15}$$

where, p_i's denote the complex poles with $Re(p_i) < 0$,

$$B_i = -\frac{f(p_i)f(-p_i)}{p_i D_n^2 \prod_{\substack{k=1 \\ k \neq n}}^{n} (p_k^2 - p_i^2)}, \quad B_0 = \begin{cases} 0, & if \deg f < n \\ \frac{1}{D_n^2}, & if \operatorname{def} f = n \end{cases} \tag{3.16}$$

and D_n is a nonzero real constant. Here note that the poles in the partial fraction expression define the strictly Hurwitz denominator polynomial $d(p{,}0)$ of $Z_1(p,0)$ as

$$d(p,0) = D_n \prod_{i=1}^{n} (p - p_i) \text{ where } p_i = -\alpha_i^2 + j\beta_i \tag{3.17}$$

For a reciprocal two-ports, $f(p,0)$ takes the form

$$f(p,0) = p^{m1} \sum_{i=0}^{m_2} a_i p^{2i} \tag{3.18}$$

and for practical unit terminated low-pass LC ladder realizations choosing

$D_n = \prod_{i=1}^{n} \frac{-1}{p - p_i}$, we can write the polynomials $f(p,0)$, $g(p,0)$ and $h(p,0)$ as

$$f_0(p) = 1 \tag{3.19}$$

$$g(p,0) = d(p,0) + h(p,0), h(p,0) = \frac{d(p,0)}{2}[Z_1(p,0) - 1]$$

c) When the lumped elements are removed from the mixed network, one would obtain cascade connection of transmission lines whose scattering matrix can fully be described independently in terms of the canonical real polynomials $\mathrm{h}(0,\lambda)$ and $g(0,\lambda)$. So, for the topologies where all the transmission zeros are at the infinity, we can generate a distributed-only network with parametric approach and use the $\mathrm{h}(\lambda)$ and $\mathrm{g}(\lambda)$ as the first column of $\Lambda_h$and Λ_g, respectively.

The boundary polynomials $h(0,\lambda)$ and $f(0,\lambda)$ define the cascade of Unit Elements in the composite structure, where $g(0,\lambda)$ is strictly Hurwitz and

$$g(0,\lambda)g(0,-\lambda) = h(0,\lambda)h(0,-\lambda) + \left(1-\lambda^2\right)^{n_\lambda} \tag{3.20}$$

The impedance function $Z_1(0,\lambda)$ describing the distributed element case is given as,

$$Z_1(0,\lambda) = \frac{n(0,\lambda)}{d(0,\lambda)} = B_0 + \sum_{i=1}^{n} \frac{B_i}{\lambda - \lambda_i} \tag{3.21}$$

where, λ_i's denote the complex poles with $Re(\lambda_i) < 0$, and the residues B_i and $d(0,\lambda)$ can also be expressed in similar forms of (3.16) and (3.17), where all functions are written in λ-variable instead of *p*-variable. For this case the polynomials $f(0,\lambda)$, $g(0,\lambda)$ and $h(0,\lambda)$ are given as

$$f(0,\lambda) = (1-\lambda^2)^{\frac{n_\lambda}{2}} \tag{3.22}$$

$$g(0,\lambda) = d(0,\lambda) + h(0,\lambda), \qquad h(0,\lambda) = \frac{d(0,\lambda)}{2}[Z_1(0,\lambda) - 1]$$

3.3 Construction of Two-Variable Scattering Functions

By making use of the aforementioned properties, once the complexity of two-port and the form of $f(p,\lambda)$ is selected, the single variable boundary polynomials $\{h(p,\lambda), g(p,\lambda)\}$ and $\{h(0,\lambda), g(0,\lambda)\}$ are readily obtained from the lumped and distributed parts of the two-port by making use of (3.14) and (3.20) respectively. Having chosen these independent polynomial coefficients, the remaining coefficients of $\{h(p,\lambda), g(p,\lambda)\}$ can be generated in terms of these independent entries using the cascade connectivity information and the paraunitary losslessness condition

$$g(p,\lambda)g(p,-\lambda) = h(p,\lambda)h(p,-\lambda) + f(p,\lambda)f(-p,-\lambda) \tag{3.23}$$

Utilizing the boundary conditions given by (a), (b) and (c) above in the paraunitary relation, and by equating the coefficients of the same powers of the complex frequency variables, we end up with an equation set, which we call *Fundamental Equation Set* (FES) [6].

Based on the above discussion, we propose the following procedure for the generation of two-variable polynomials, which in turn define the scattering and impedance functions for mixed lumped-distributed structures.

Proposed Semi-Analytic Construction Procedure:

- Assuming that a regular cascaded structure as in Figure 3.1, pick the total number of lumped and distributed elements (n_p, n_λ) in the circuit topology and fix the form of $f(p,\lambda)$.
- Select the poles of impedance functions $Z_1(p,0)$ and $Z_1(0,\lambda)$ as the unknown free parameters of the problem.
- Generate the polynomials $\{h(p,0), g(p,0)\}$ and $\{h(\lambda,0), g(\lambda,0)\}$ in terms of the poles of $Z_1(p,0)$ and $Z_1(0,\lambda)$ by utilizing (3.15, 3.17, 3.19) and (3.21, 3.22).
- Establish the coefficient constraints reflecting the connectivity information for the cascade topology and then solve FES (3.23) for the remaining unknowns of the two-variable polynomials $\{h(p,\lambda), g(p,\lambda)\}$.

From the physical implementation point of view, practical circuit configurations consist of commensurate lines incorporating lumped discontinuities in low-pass or high-pass type lumped elements. For those regular structures the proposed semi-analytic approach has been successfully applied and explicit two-variable characterization for a variety of practical mixed element circuits have been obtained [3-13, 15]

3.4 Solutions for Low Order Low-Pass Ladders with Unit Elements (LPLU)

In the following, solutions for the elementary low order low-pass ladders with unit elements are summarized [15].

Second order LPLU

In this design we have only one lumped and one distributed component like below

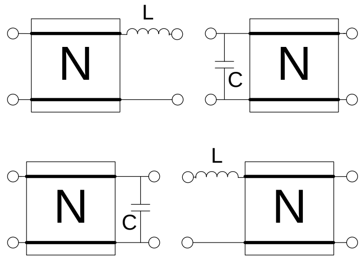


Figure 3.2 Second order LPLU networks

$$\wedge_h = \begin{pmatrix} 0 & h_{01} \\ h_{10} & h_{11} \end{pmatrix} \quad \text{and } \wedge_g = \begin{pmatrix} 1 & g_{01} \\ g_{10} & g_{11} \end{pmatrix}$$

where

$$g_{11} = g_{10}g_{01} - h_{10}h_{01}, \qquad h_{11} = g_{11}h_{10}/g_{10}$$

Third order LPLU

In this design we have two lumped and one distributed components like below

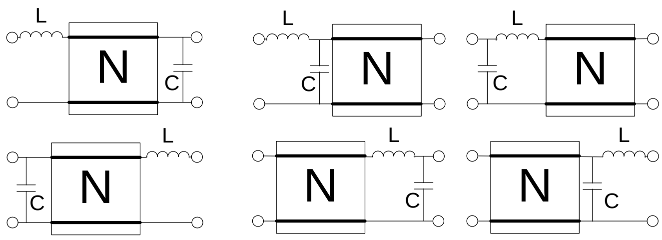


Figure 3.3 Third order LPLU networks

$$\wedge_h = \begin{pmatrix} 0 & h_{01} \\ h_{10} & h_{11} \\ h_{20} & 0 \end{pmatrix} \quad \text{and } \wedge_g = \begin{pmatrix} 1 & g_{01} \\ g_{10} & g_{11} \\ g_{20} & 0 \end{pmatrix}$$

where h_{01}, h_{10}, h_{20} are independent coefficients which defines also g_{01}, g_{10}, g_{20} and

$$g_{11} = g_{10}g_{01} - h_{10}h_{01},\ h_{11} = g_{11}g_{20}/h_{20}$$

Fourth order LPLU

In this design we have two lumped and one distributed components like below

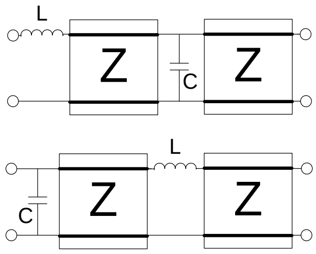


Figure 3.4 Fourth order LPLU networks

$$\Lambda_h = \begin{pmatrix} 0 & h_{01} & h_{02} \\ h_{10} & h_{11} & h_{12} \\ h_{20} & h_{21} & 0 \end{pmatrix} \quad \text{and} \quad \Lambda_g = \begin{pmatrix} 1 & g_{01} & g_{02} \\ g_{10} & g_{11} & g_{12} \\ g_{20} & g_{21} & 0 \end{pmatrix}$$

where

$$g_{11} = g_{10}g_{01} - h_{10}h_{01},\ h_{11} = h_{02}\beta/\alpha + h_{20}\alpha/\beta,\ \alpha = g_{01} - \mu h_{01},$$
$$\beta = g_{10} - \mu h_{10},\ \mu = h_{20}/g_{20},\ g_{12} = (g_{11}g_{02} - h_{11}h_{02})/\alpha,$$
$$g_{21} = (g_{11}g_{20} - h_{11}h_{20})/\beta,\ h_{12} = \mu g_{12},\ h_{21} = \mu g_{21}$$

Fifth order LPLU

In this design we have two lumped and one distributed components like below

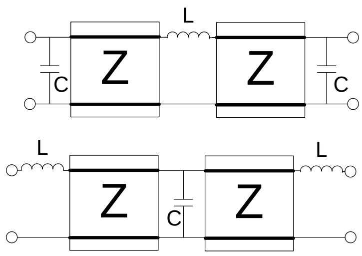


Figure 3.5 Fifth order LPLU networks

$$\Lambda_h = \begin{pmatrix} 0 & h_{01} & h_{02} \\ h_{10} & h_{11} & h_{12} \\ h_{20} & h_{21} & 0 \\ h_{30} & 0 & 0 \end{pmatrix} \quad \text{and} \quad \Lambda_g = \begin{pmatrix} 1 & g_{01} & g_{02} \\ g_{10} & g_{11} & g_{12} \\ g_{20} & g_{21} & 0 \\ g_{30} & 0 & 0 \end{pmatrix}$$

where

$$g_{11} = g_{10}g_{01} - h_{10}h_{01},\ h_{11} = h_{02}\beta/\alpha + h_{20}\alpha/\beta,\ \alpha = g_{01} - \mu h_{01},$$
$$\beta = g_{10} - \mu h_{10},\ \mu = h_{30}/g_{30},\ g_{12} = (g_{11}g_{02} - h_{11}h_{02})/\alpha,$$
$$g_{21} = (g_{11}g_{20} - h_{11}h_{20} - g_{01}g_{30} + h_{01}h_{30})/\beta,\ h_{12} = \mu g_{12}, h_{21} = \mu g_{21}$$

3.4 Generalized Real Frequency Design Algorithm for Mixed Lumped and Distributed Element Matching Networks

The extension of the real frequency technique for designing equalizers with mixed elements require a unique characterization of the matching network in terms of a number of independent free parameters. In the parametric approach discussed above, the real normalized scattering parameters are generated from *the poles* of the *partially defined impedance functions*. In other words, the mixed element structure is assumed to be separable into its lumped and distributed parts, which can completely be defined in terms of the poles of the corresponding impedance functions, provided that the transmission zeros are defined. Thus, the presented parametric description leads us to an efficient generalization of the Real Frequency technique applicable to mixed lumped and distributed element matching networks. The algorithm can be outlined as follows:

Step 1: For prescribed transmission zeros of mixed element structure, i.e. $f(p,\lambda)$, starting from the left half plane poles of $Z_2(p,0)$ and $Z_2(0,\lambda)$, we generate the canonic polynomials $h(p,\lambda)$ and $g(p,\lambda)$ by utilizing the connectivity information supplied for the cascade structure. For this purpose, compute residues of impedance functions as in (3.16) and generate the impedance functions $Z_2(p,0)$ and $Z_2(0,\lambda)$ using (3.15) and (3.21). The polynomials $h(p,\lambda)$ and $g(p,\lambda)$ are then constructed via (3.22, 3.19) and the proposed semi-analytic construction procedure.

In this procedure, $Z_2(p,0)$ is the positive real impedance function for only lumped section, where as $Z_2(0,\lambda)$ is for only distributed section.

$$Z_2(p,0) = \frac{n(p,0)}{d(p,0)} \text{ and } Z_2(0,\lambda) = \frac{n(0,\lambda)}{d(0,\lambda)}$$

h and g polynomials are computed using following relations.

$$h(p,0) = n(p,0) - d(\mathrm{p},0) \tag{3.25}$$

$$g(p,0) = n(p,0) + d(p,0) \tag{3.26}$$

and

$$h(0,\lambda) = n(0,\lambda) + d(0,\lambda) \tag{3.27}$$

$$g(0,\lambda) = n(0,\lambda) + d(0,\lambda) \tag{3.28}$$

In order to weight h and g we multiply them by $D_n/2$. Then using the construction procedure above we generate $h(p,\lambda)$ and $g(p,\lambda)$.

Step 2: Then, we form the unknown parameter vectors; $\theta_p = \{\alpha_i, \beta_i\}$ and $\theta_\lambda = \{\alpha_i', \beta_i'\}$ which define the left half plane poles of $Z_2(p,0)$ and $Z_2(0,\lambda)$ dictated by the relations (3.15) and (3.21). These parameters are chosen as the independent unknowns of the problem and determined to optimize the gain of the system by a nonlinear search routine. At this step unknowns of the problem are expressed as

For the lumped elements unknown free parameter set $\theta_p = \{\alpha_i, \beta_i\}$, where

$\begin{Bmatrix} p_i = -\alpha_i + j\beta_i \\ p_i^* = -\alpha_i - j\beta_i \end{Bmatrix}, for\ i = 1 \ldots . n_2$ and $p_0 = -\alpha_0$ for n_p odd with

$$n_2 = \begin{cases} \dfrac{n_p}{2} & , if\ n_p\ is\ even \\ \dfrac{(n_p - 1)}{2} & , if\ n_p\ is\ odd \end{cases}$$

For the distributed elements unknown free parameter set $\theta_\lambda = \{\alpha_i', \beta_i'\}$ where;

$\begin{Bmatrix} \lambda_i = -\alpha_i' + j\beta_i' \\ \lambda_i^* = -\alpha_i' - j\beta_i' \end{Bmatrix}, for\ i = 1 \ldots . n_2$ and $\lambda_0 = -\alpha_0'$ for n_λ odd with

$$n_2 = \begin{cases} \dfrac{n_\lambda}{2} & , if\ n_\lambda\ is\ even \\ \dfrac{(n_\lambda - 1)}{2} & , if\ n_\lambda\ is\ odd \end{cases}$$

Step 3: For a typical double matching configuration, the transducer power gain can be expressed in terms of the canonic polynomials of the equalizer to be designed and the unit normalized reflectances of the load and generator as follows:

$$T(w) = \frac{(1 - |S_G|^2)(1 - |S_L|^2)|f|^2}{|g - hS_G + \sigma S_L(h^* - S_G g^*)|^2} \tag{3.29}$$

where the superscript * denotes complex conjugation in two variable polynomials, i.e. $h^* = h(-jw, -j\Omega)$, where w is discrete frequency data inside the prescribed bandwidth

Step 4: Optimization of the gain in least square sense can be carried out by using the objective function

$$\delta = \sum_{i=1}^{N_F} \left[T\left(w_i, \theta_p, \theta_\lambda\right) - T_0\right]^2 \tag{3.30}$$

where N_F is the number of sampling frequencies and T_0 is the desired flat gain level over the pass band. Since there is no restriction on the real parameters of θ_p and, θ_λ any unconstraint nonlinear optimization routine can be used.

Step 5: Once the desired gain level is achieved, generate $Z_2(p, 0)$ and $Z_2(0, \lambda)$ in terms of the resultant θ_p and θ_λ parameters respectively. Then realize these impedance functions using pole-zero and Richards extractions to end up with the lumped and distributed element values which appear in the prescribed order of cascade structure.

Chapter 4
Real Frequency Design of Broadband Amplifiers

4.1 Amplifier Design Problem as Active Device Matching

Amplifier design at microwave frequencies differs significantly from the conventional low-frequency approaches and require special considerations especially in the small signal operation of the circuit. Once the proper biasing conditions are satisfied, the use of appropriate matching networks at the input and output of active components to reduce the reflections and to provide maximum possible power transfer over the broadest possible bandwidth is the main design objective. Another major issue is avoiding simultaneously undesired oscillations. Therefore, in the design of broadband microwave amplifiers, a fundamental problem is to realize front-end and back-end matching networks so that the transfer of power from source to load is maximized over a prescribed frequency range as illustrated in Figure 4.1. In such a case the problem is actually one of a single or double matching problem with an active gain block whose input and output impedance or reflection data should be matched properly. Hence the amplifier design problem can be treated as a typical active device matching where the real frequency matching techniques can directly be applied along with proper power gain requirements and stability considerations.

In this section transistor amplifier design is discussed based on the terminal characteristics of transistors as represented by numerical Scattering parameters data. The power gain expressions and the stability considerations for the active device will be mentioned and the development of matching networks to yield maximum flat gain level over the frequency band of interest will be obtained by proper extension of the real frequency matching network design approach.

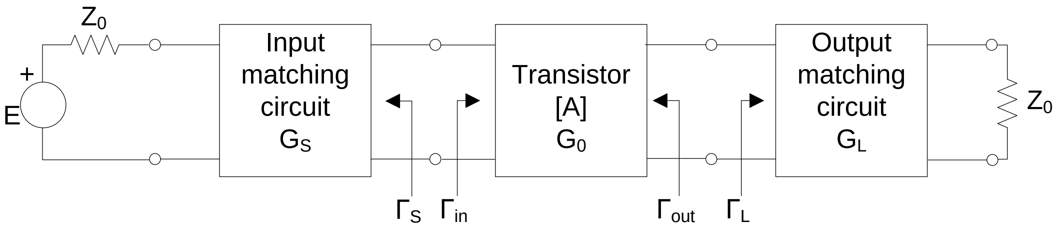


Figure 4.1 The general transistor amplifier circuit

Referring to the general amplifier block representation shown in Figure 4.1 let the scattering matrix of the transistor is denoted by A and Γ_S and Γ_L indicate the input and output matching network reflectances. The transducer power gain of the system is given by [16]

$$G_T = \frac{P_L}{P_{avs}} = \frac{|A_{21}|^2(1-|\Gamma_S|^2)(1-|\Gamma_L|^2)}{|1-\Gamma_S\Gamma_{in}|^2|1-A_{22}\Gamma_L|^2}$$

Or equivalently alternative form

$$G_T = \frac{P_L}{P_{avs}} = \frac{\left|A_{21_k}\right|^2\left|S_{21_{k+1}}\right|^2}{\left|1-\check{S}_{22_k}A_{11_k}\right|^2\left|1-\check{A}_{22_k}S_{11_{k+1}}\right|^2} \text{ [6]} \tag{4.3}$$

where

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0} \tag{4.4}$$

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0} \tag{4.5}$$

Z_0 being the real normalization characteristic impedance.

$\mathit{\Gamma}_{in}$ and $\mathit{\Gamma}_{out}$ are being the input and output reflectances of the active devices when the other terminal is loaded.

$$\Gamma_{in} = \frac{V_1^-}{V_1^+} = A_{11} + \frac{A_{12}A_{21}\Gamma_L}{1-A_{22}\Gamma_L} = \frac{Z_{in}-Z_0}{Z_{in}+Z_0} \tag{4.6}$$

$$\Gamma_{out} = \frac{V_2^-}{V_2^+} = A_{22} + \frac{A_{12}A_{21}\Gamma_S}{1-A_{11}\Gamma_S} = \frac{Z_{out}-Z_0}{Z_{out}+Z_0} \tag{4.7}$$

As you can see here, the maximum transducer power gain is $|S_{21}|^2$ which occurs when input and output of the network perfectly matched, $\Gamma_L = \Gamma_S = 0$. When we have conjugate matching output of the transistor, that is to say when Γ_{out}=Γ_L^*. The transducer power gain expression for maximum power transfer case is modified as follows:

$$T_k' = T_k(w)\frac{1}{1-\left|\check{A}_{22_k}\right|^2}$$

Stability considerations

We must take into account the stability of the amplifier network. Oscillation is possible if one of the input and output impedance has negative real part. In order to prevent this, both $|\Gamma_{\mathrm{in}}|$ and $|\Gamma_{\mathrm{out}}|$ must be greater than 1. Since $|\Gamma_{\mathrm{in}}|$ and $|\Gamma_{\mathrm{out}}|$ depends on source and load matching networks, the stability of the amplifier depends on and $|\Gamma_{\mathrm{S}}|$ and $|\Gamma_{\mathrm{L}}|$. According the values of $|\Gamma_{\mathrm{S}}|$ and $|\Gamma_{\mathrm{L}}|$, we can split stability in to two types, conditional stability and unconditional stability.

- **Unconditional stability:** The network is unconditionally stable if $|\Gamma_{\mathrm{in}}| < 1$ and $|\Gamma_{\mathrm{out}}| < 1$ for all passive source and load impedances.
- **Conditional stability:** The network is said to be conditional stable, if $|\Gamma_{\mathrm{in}}| < 1$ and $|\Gamma_{\mathrm{out}}| < 1$ only for a certain range of load and source impedances. This case is also known as potentially unstable.

In order to get unconditionally stable design, following conditions should be satisfied.

$$|\Gamma_{\mathrm{in}}| = \left|S_{11} + \frac{S_{12}S_{21}\Gamma_{\mathrm{L}}}{1 - S_{22}\Gamma_{\mathrm{L}}}\right| < 1 \tag{4.9}$$

$$|\Gamma_{\mathrm{out}}| = \left|S_{22} + \frac{S_{12}S_{21}\Gamma_{\mathrm{S}}}{1 - S_{11}\Gamma_{\mathrm{S}}}\right| < 1 \tag{4.10}$$

If the transistor is unilateral, i.e. $S_{12} = 0$, this conditions reduces to the simple results as $|S_{11}| < 1$ and $|S_{22}| < 1$ for unconditional stability condition. Otherwise, equations 4.3.1 and 4.3.2 defines a range for Γ_{S} and Γ_{L} values where the amplifier will be stable. In order to get stable region for Γ_{S} and Γ_{L} we must find input and output stability circles. The stability circles are defined as the loci in the Γ_{S} or Γ_{L} plane for which $|\Gamma_{\mathrm{in}}| = 1$ or $|\Gamma_{\mathrm{out}}| = 1$, respectively. After the derivation of stability circles, we get the center of the output stability circle, C_L , and the radius of the output stability circle, R_L, as below. The derivation of stability circles can be found at [16].

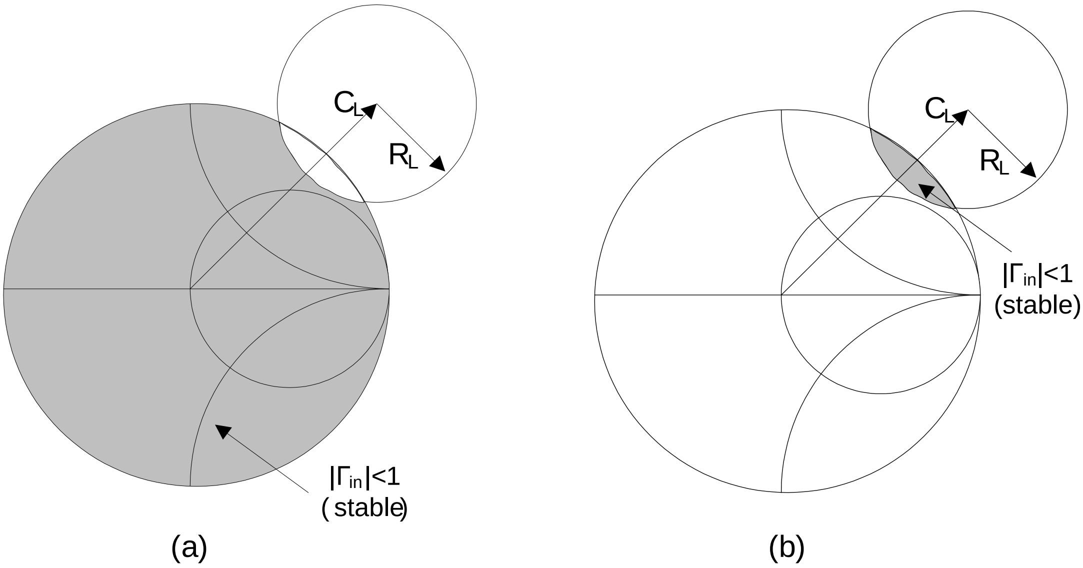


Figure 4.2 Output stability circles for a conditionally stable device (a) $|S_{11}|<1$ (b) $|S_{11}|>1$

$$C_L = \frac{\left(S_{22} - \Delta S_{11}^*\right)^*}{|S_{22}|^2 - |\Delta|^2} \text{ (center)} \tag{4.11}$$

$$R_L = \left|\frac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2}\right| \text{ (radius)} \tag{4.12}$$

where

$$\Delta = S_{11}S_{22} - S_{12}S_{21}. \tag{4.13}$$

Similarly, we can find the center and the radius of the input stability circuits, C_S and R_S, as following

$$C_S = \frac{\left(S_{11} - \Delta S_{22}^*\right)^*}{|S_{11}|^2 - |\Delta|^2} \text{ (center)} \tag{4.14}$$

$$R_S = \left|\frac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2}\right| \text{ (radius)} \tag{4.15}$$

In order for an amplifier to be unconditionally stable, following necessary and sufficient conditions should be satisfied.

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1 \tag{4.16}$$

$$|\Delta| < 1 \tag{4.17}$$

You can find the proof of these conditions in [17]. If the device is conditionally stable, operation points, Γ_S and Γ_L, must be chosen in the stable region and this must be tested over a prescribed band for broadband matching. If design goes to unstable region for one of the frequencies inside the band, we can prevent unstable situation by decreasing the power gain. The transistor can usually be made to be unconditionally stable by using resistive loading [16].

Another unconditionally stability test, which has only a single parameter, μ, has been established as follow. [16] .

$$\mu = \frac{1 - |S_{11}|^2}{|S_{22} - S_{11}^*\Delta| + |S_{21}S_{12}|} > 1 \tag{4.18}$$

4.2 Multistage Amplifier Design via Parametric Real Frequency Approach

The problem is to transfer power from source to load with minimum loss. To do that, we use front and backhand equalizers to match the network to the generator or load. You can see an amplifier network with only front hand matching network in Figure 4.3.

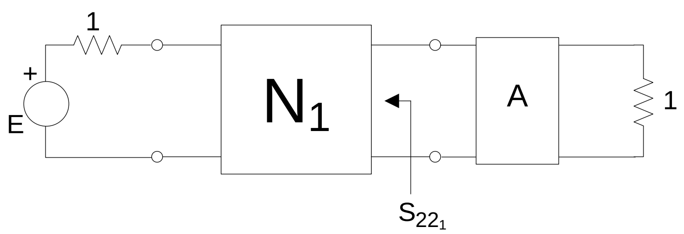


Figure 4.3 Single stage amplifier equalized at the input

We will use same steps, as we did in matching problem before, to design passive stages. However, in amplifier problem we need active devices which cannot be changed after fabrication. The characteristic of those active devices are given us as bunch of measurements related to S parameters in different frequencies. What we should do is to match generator impedance and load impedance to this active device over a prescribed frequency bandwidth.

Transducer power gain of such network, with only front hand equalizer is as the following

$$T_1(\omega) = T_G \frac{|A_{21}|^2}{\left|1 - S_{22_1} A_{11}\right|^2} \tag{4.20}$$

where

$$T_G = 1 - \left|S_{11_1}\right|^2 \tag{4.21}$$

In general, designers use front hand and back hand equalizers simultaneously. In this situation, there is no analytical solution. We need to split this problem into two basic steps and solve problem sequentially

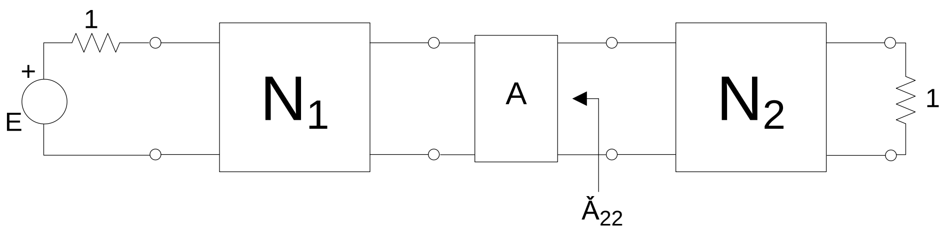


Figure 4.4 Single stage amplifier equalized at the input and the output

As you can see this network is very same with the above network except back hand equalizer, N_2. In this problem, we think $T_1(\omega)$ as T_G like below.

$$T(\omega) = \frac{T_1(\omega)\left|S_{21_2}\right|^2}{\left|1 - \check{A}_{22} S_{11_2}\right|^2} \tag{4.22}$$

where

$$\check{A}_{22} = A_{22} + \frac{A_{21} A_{12} S_{22_1}}{1 - S_{22_1} A_{11}}. \tag{4.23}$$

Transistor parameters are measured in the case where they are terminated with the 1Ω resistor. However, here an active element A is connected after passive network N_1, not 1Ω. So we must replace A_{22} with $\check{A}_{22}$.

We can sequentially process this algorithm for multistage amplifiers. $T(\omega)$'s that we found after each iteration are thought as T_G for next stage.

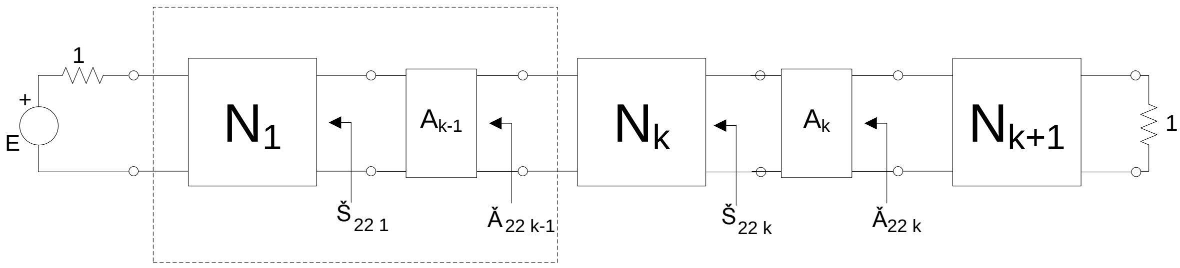


Figure 4.5 Multistage amplifier

$$T_k(\omega) = T_{k-1} \frac{\left|A_{21_k}\right|^2}{\left|1 - Š_{22_k} A_{11_k}\right|^2} \frac{\left|S_{21_{k+1}}\right|^2}{\left|1 - \breve{A}_{22_k} S_{11_{k+1}}\right|^2} \tag{4.24}$$

where

$$Š_{22_k} = S_{22_k} + \frac{\left|S_{21_k}\right|^2 \breve{A}_{22_{k-1}}}{1 - \breve{A}_{22_{k-1}} S_{11_k}} \tag{4.25}$$

As mentioned before, the output of the active devices is terminated with 1Ω.

For maximum gain design it is known that the input and output of the transistor should be conjugately matched. If we adopt this condition in the above procedure and assume conjugate matching at the output of transistor the procedure is expected to yield faster convergence with less iterations. In this case, for maximum gain condition with conjugate matching, the transducer power gain $T(\omega)$ expression for the amplifier should be modified as

$$T_k^{'}(\omega) = T_k(\omega) \frac{1}{1 - \left|\breve{A}_{22_k}\right|^2} \tag{4.26}$$

Here, $T_k(\omega)$ is the gain of the stage with real terminations defined by the generic expression equation 4.24.

In the design procedure above, the active device A_i is assumed to represent a stabilized transistor module including the transistor and the stabilizing feedback circuitry. Or the procedure is assumed to be applied for the stable frequency range of the transistor.

4.2.1 Amplifier Design Algorithm

Step 1: A transistor with numerical scattering data over a prescribed frequency band, is chosen from the database. It is recommended that, a stabilized transistor should be chosen or the frequency range for which unconditional stability is ensured should be utilized for the small signal gain optimization.

Step 2: Complexity, degree, of the back-end and front-end equalizers of the amplifier are decided.

Step 3: Topology of the network is defined by setting m_1 and m_2 values used in 2.32 for both back-end and front-end matching networks.

Step 4: According to degree of the front-end network, poles of the positive real impedance function is computed as the following.

$\begin{Bmatrix} p_i = -\alpha_i + j\beta_i \\ p_i^* = -\alpha_i - j\beta_i \end{Bmatrix}, for\ i = 1 \ldots . n_2$ and $p_n = -\alpha_0$ for n odd,

where

$$n_2 = \begin{cases} \dfrac{n}{2} & , if\ n\ is\ even \\ \dfrac{(n-1)}{2} & , if\ n\ is\ odd \end{cases}$$

User should enter the input variables α_i, β_i, so that poles can be generated.

Step 5: If desired, foster sections are given to algorithm and we get an input vector for generating impedance function like

$$[\propto_0, \{\alpha_i, \beta_i, i = 1, \ldots n_2\}, \{b_\infty, b_0\}]$$

Step 6: Optimization routine is called and tried to converge desired gain by changing the values inside the input vector which are used to generate poles of the impedance function characterizing the front-end matching network.

Step 7: The above steps from 4 to 6 are repeated for the overall gain optimization and the construction of the back-end impedance.

Step 8: Once the initial results for the matching networks of the amplifier are obtained, the overall gain of the system is re-optimized by final trimming iterations.

Thus for maximum gain convergence, the impedance functions for the front-end and back-end matching networks are obtained and they are synthesized by pole-zero shift approach to end up with the amplifier equalizer element values [5].

Chapter 5

Multivariable Broadband Matching Tool

We prepared a standalone tool to design matching and amplifier networks with a user friendly interface. During the process of implementation, we used the numerical power of MATLAB and .NET framework. We can say that, the engine of the tool was written in MATLAB and was dressed up with .NET. There is no need for MATLAB installation on target computer on which this program will be installed and executed.

Below, you can find the design problems that can be solved via this tool.

- Single stage lumped matching design.
- Single stage distributed matching design.
- Mixed element matching design with LPLU type matching network..
- Single stage amplifier design with only lumped elements at front and back matching networks
- Single stage amplifier design with only distributed elements at front and back matching networks
- Single stage amplifier design with lumped at front and distributed at back matching network or vice versa.
- Multistage amplifier design with lumped, distributed or mixed LPLU type matching network.

Before the introduction of user interface and how to perform design problem as mentioned before, it will be meaningful to explain the engine of the tool, which is the m files that are used to process complex operations.

5.1 Engine of the Tool

Engine of the tool was written in MATLAB. Interface passes parameters to the a dynamic-link library (dll) which has been compiled by MATLAB and shows the result to the user. Five main .m files have been implemented for specific problems, like lumped.m, distributed.m, amp.m, mixed_matching.m and mixed_amp.m. All these .m files, call other modules which had been written in MATLAB according to the algorithm of the specific job. Below, you can find the summarized information rough flow diagrams of the main .m files.

lumped.m: This is the file that is responsible for lumped matching problems. It receives some parameters related to the optimization requirements or initial parameters for the matching algorithm. You can find the flow diagram of the lumped.m file

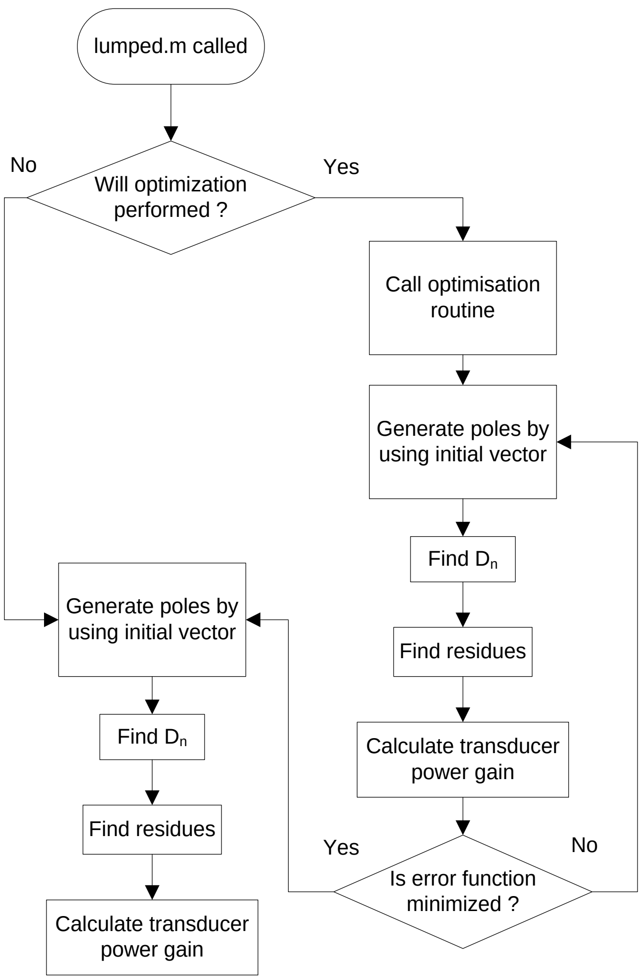


Figure 5.1 Flow diagram of lumped.m file

distributed.m: This is the file that is responsible for distributed matching problems. It does same think as lumped.m file does. However, it receives some additional parameters related with the distributed elements. You can find the flow diagram of the lumped.m file

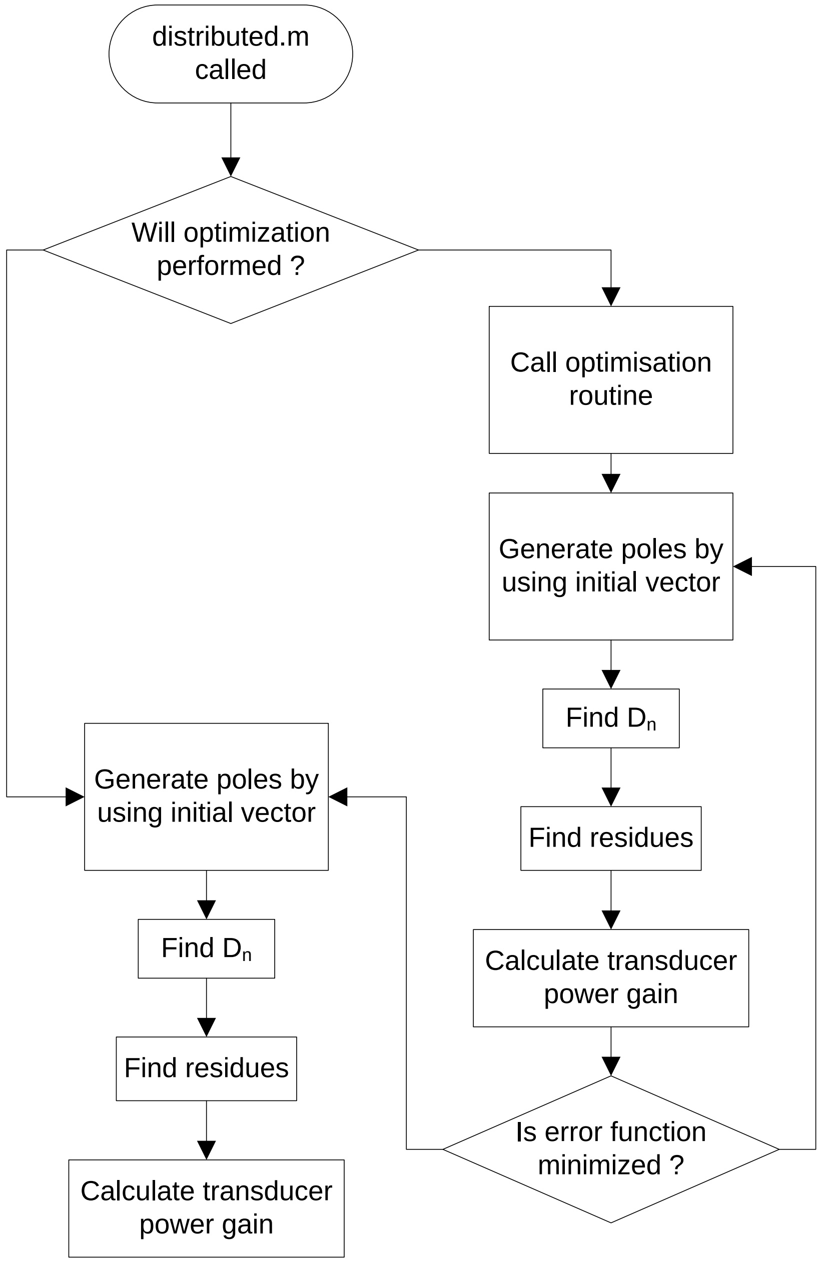


Figure 5.2 Flow diagram of distributed.m file

amp.m: This is the file that is responsible for matching the generator and load impedances to the input and the output of the transistor, respectively. You can find the flow diagram of the amp.m file

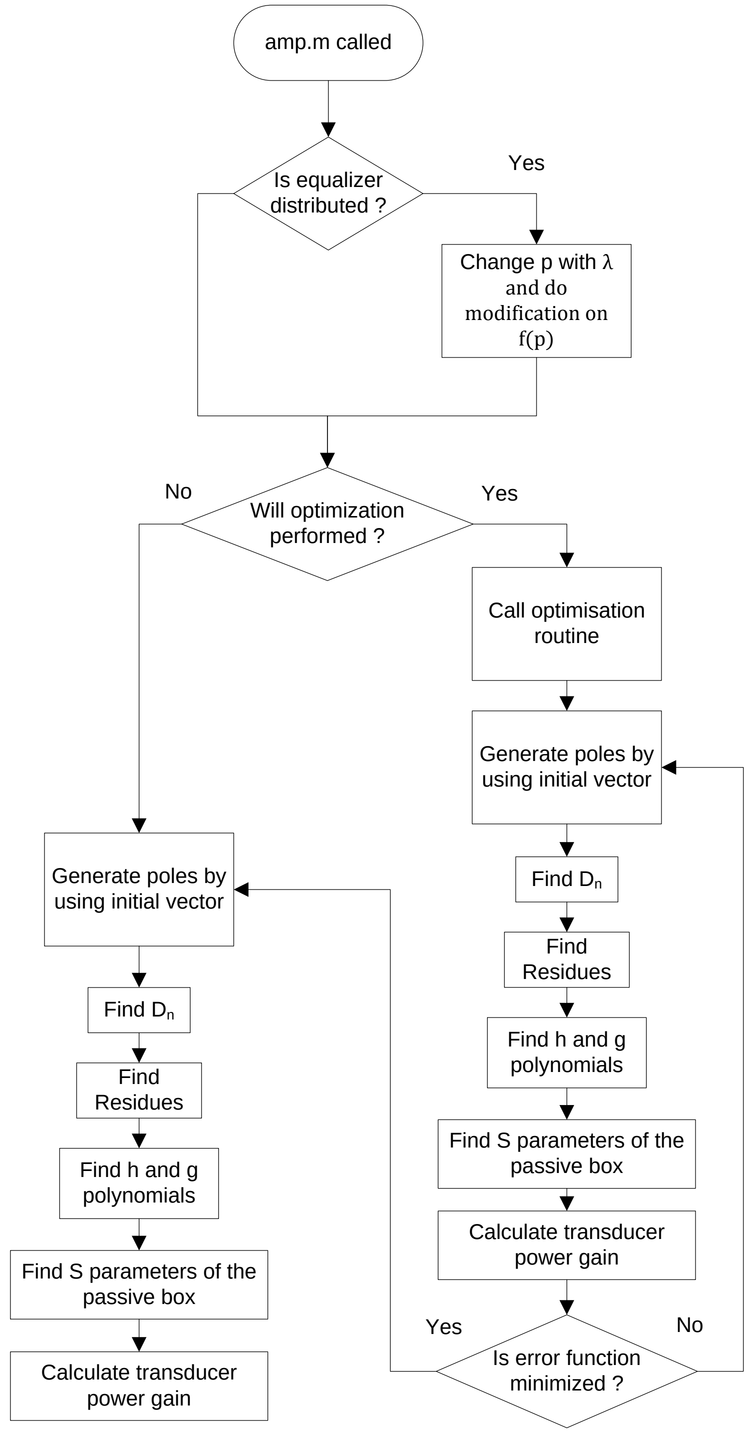


Figure 5.3 Flow diagram of amp.m file

mixedmatching.m: This is the file that is responsible for matching the generator and load impedances to the mixed passive network consisting of low pass lumped and distributed components.

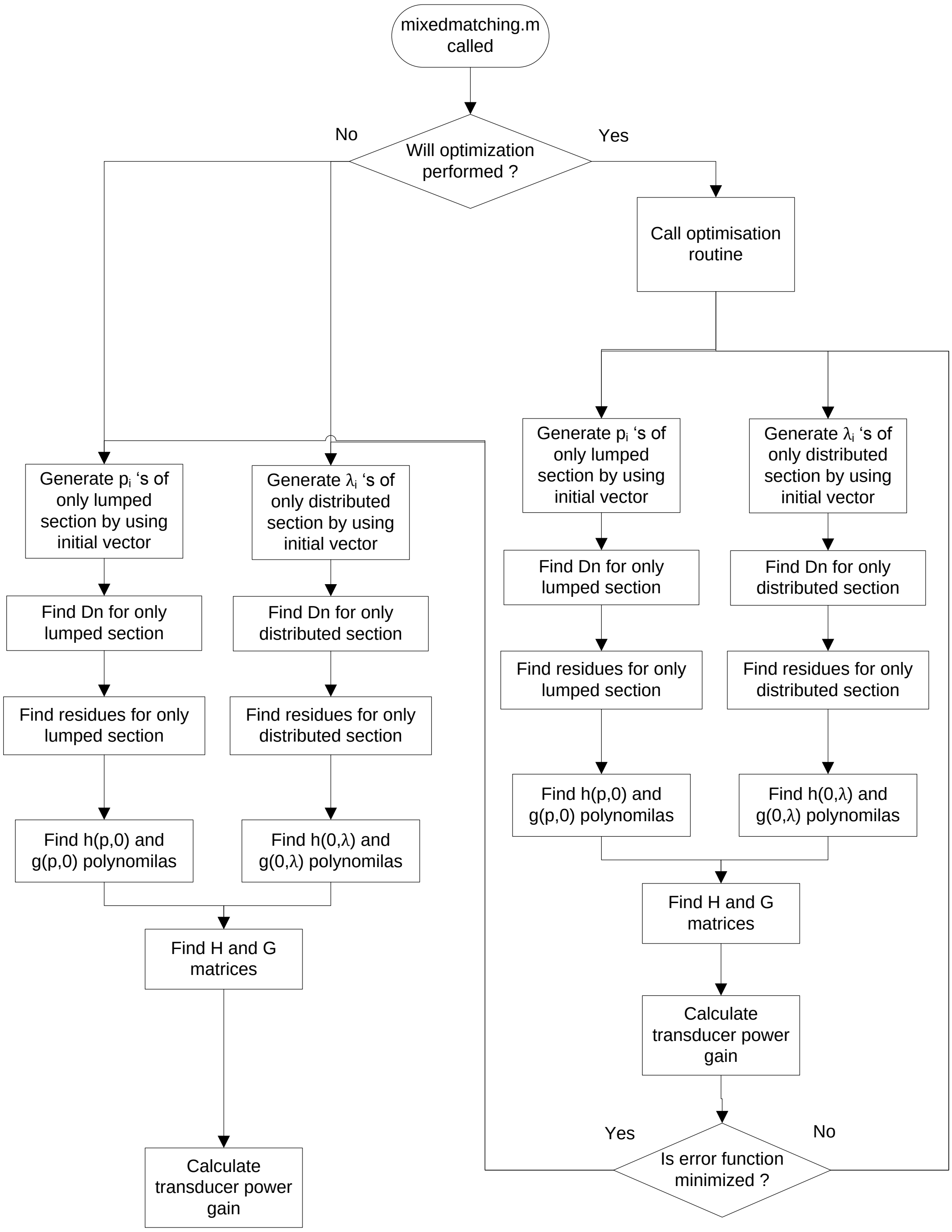


Figure 5.4 Flow diagram of mixedmatching.m file

mixedamp.m: This is the file that is responsible for matching the generator and load impedances to the mixed passive network consisting of low pass lumped and distributed components. Flow diagram of mixed matching and mixed amplifier is the same, however, the way that the transducer power gain is computed is different for both.

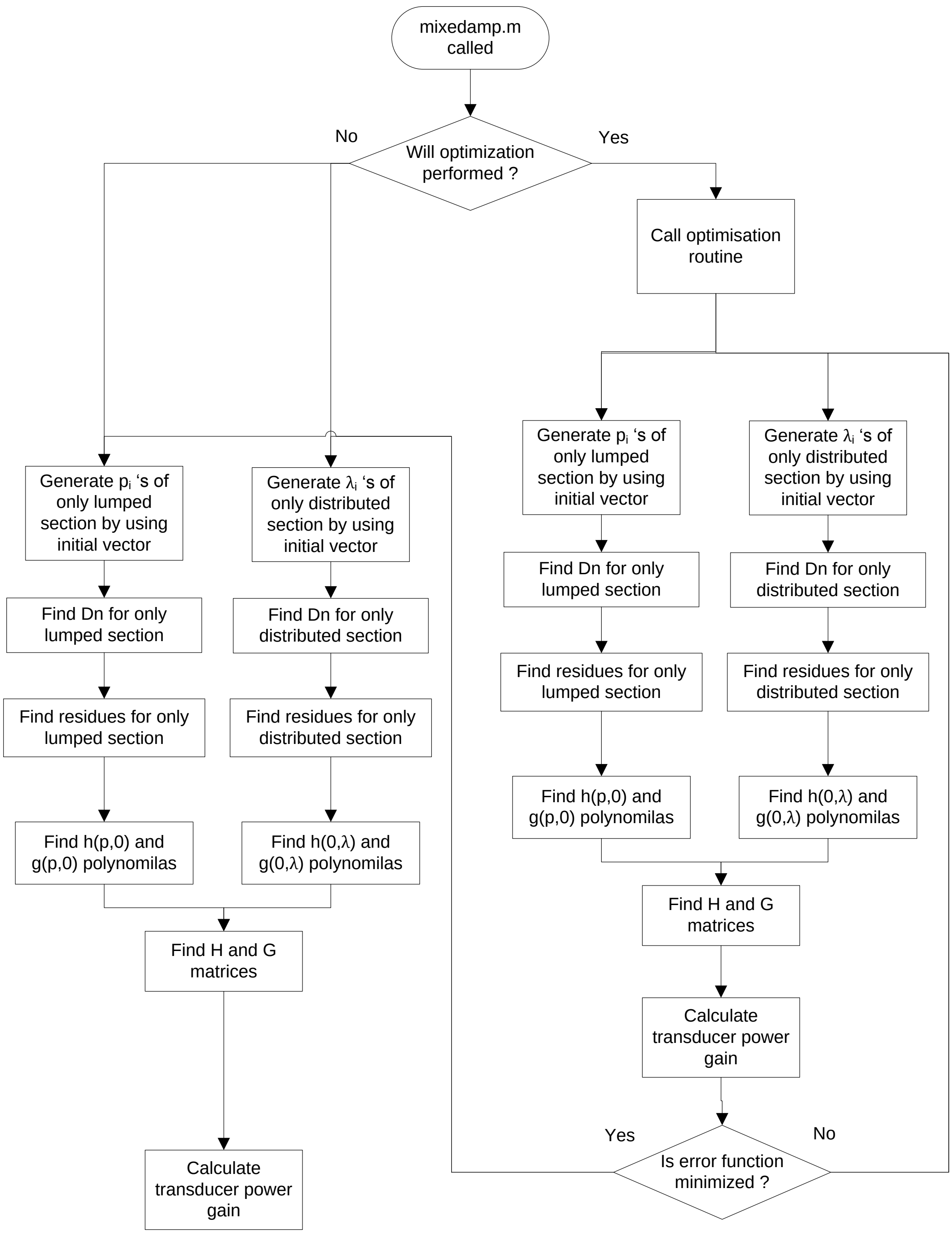


Figure 5.5 Flow diagram of mixedamp.m file

5.2 User Interfaces of the Tool

In this section, I will consider the screens of the program. You can find how to create matching networks, amplifier networks, selecting transistors, setting parameters of the passive boxes and so on. As said before, we used C#.NET and Janus components for user interface. First of all, it is meaningful to introduce overall look of the program.

5.2.1 Overview of the Main Body of the Program.

As can be seen in the Figure 5.6, program split in to four main forms, Matching, Amplifier, Mixed Matching and Mixed Amplifier. We will introduce all these windows in detail.

Figure 5.6 Main form of the tool

5.2.2 Matching Form

When user clicks the matching button, a new window is created and waits for user input to setting up the network as can be seen in the Figure 5.7. First input is the stage number of the network. When user clicks the "Create Button" link, according to the stage number, the overall view of the system is drawn with boxes and required arrays and cell arrays are created. Below, you can see a three stage matching network.

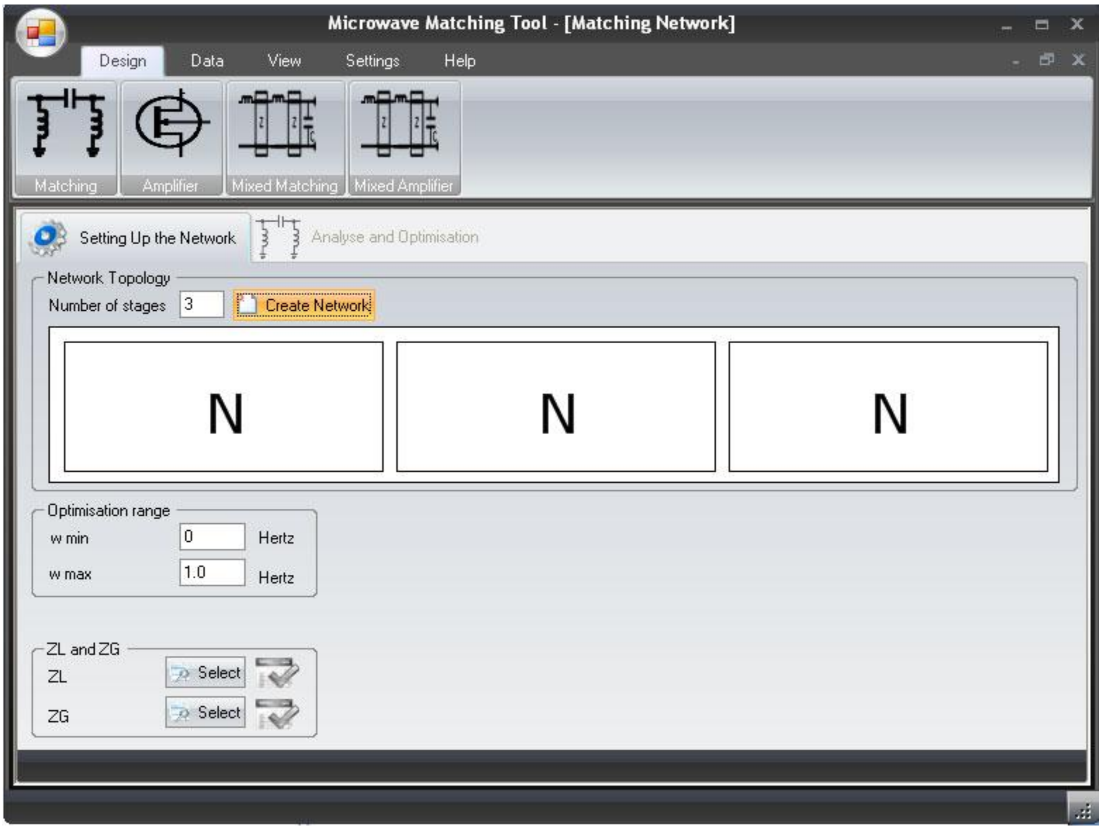


Figure 5.7 Matching window of the tool

When user clicks the one of the boxes, a new window is opened.

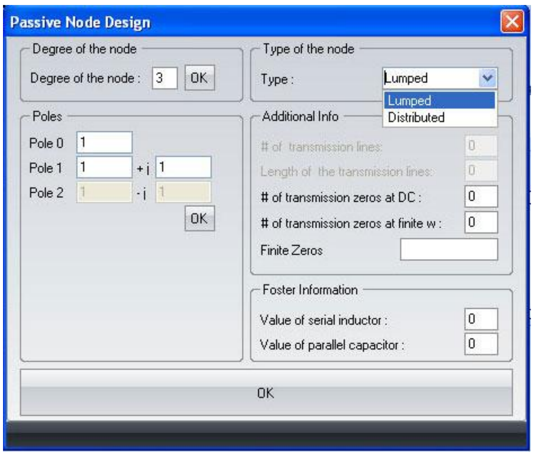


Figure 5.8 Passive box design window

In this window, user can select the complexity of the program, poles, transmission zeros and foster and type of the network. After closing the window, input parameters is saved in to the arrays or cell arrays which were created when user has entered the Create Buttons“ button.

After setting all stages, user should select a text file including the empiric data for load and generator impedances by pushing the “Select” buttons near ZL and ZG labels. Flow diagram of the events is depicted in the Figure 5.9.

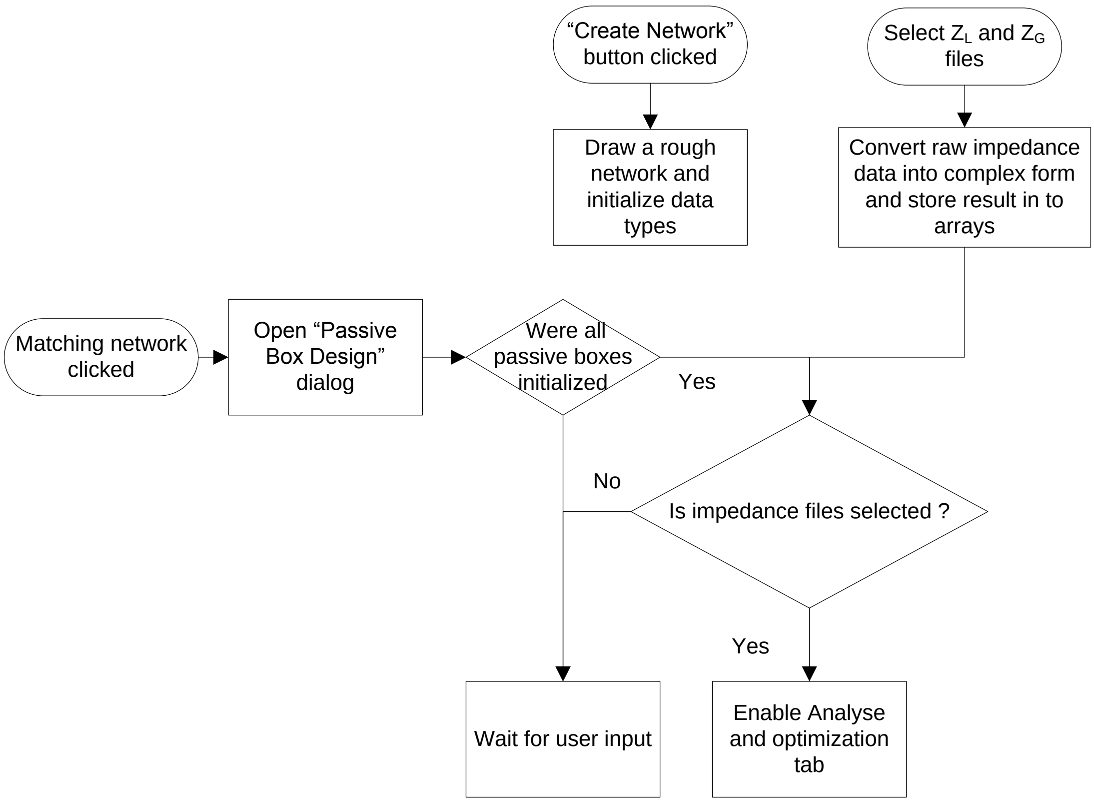


Figure 5.9 Flow diagram of matching window

Lastly, user can now analyze the circuit by pushing “Run Program” button at the “Analyze and Optimization” tab. Flow diagram of the algorithm is depicted as follows. User can re run the program to converge better transducer power gain.

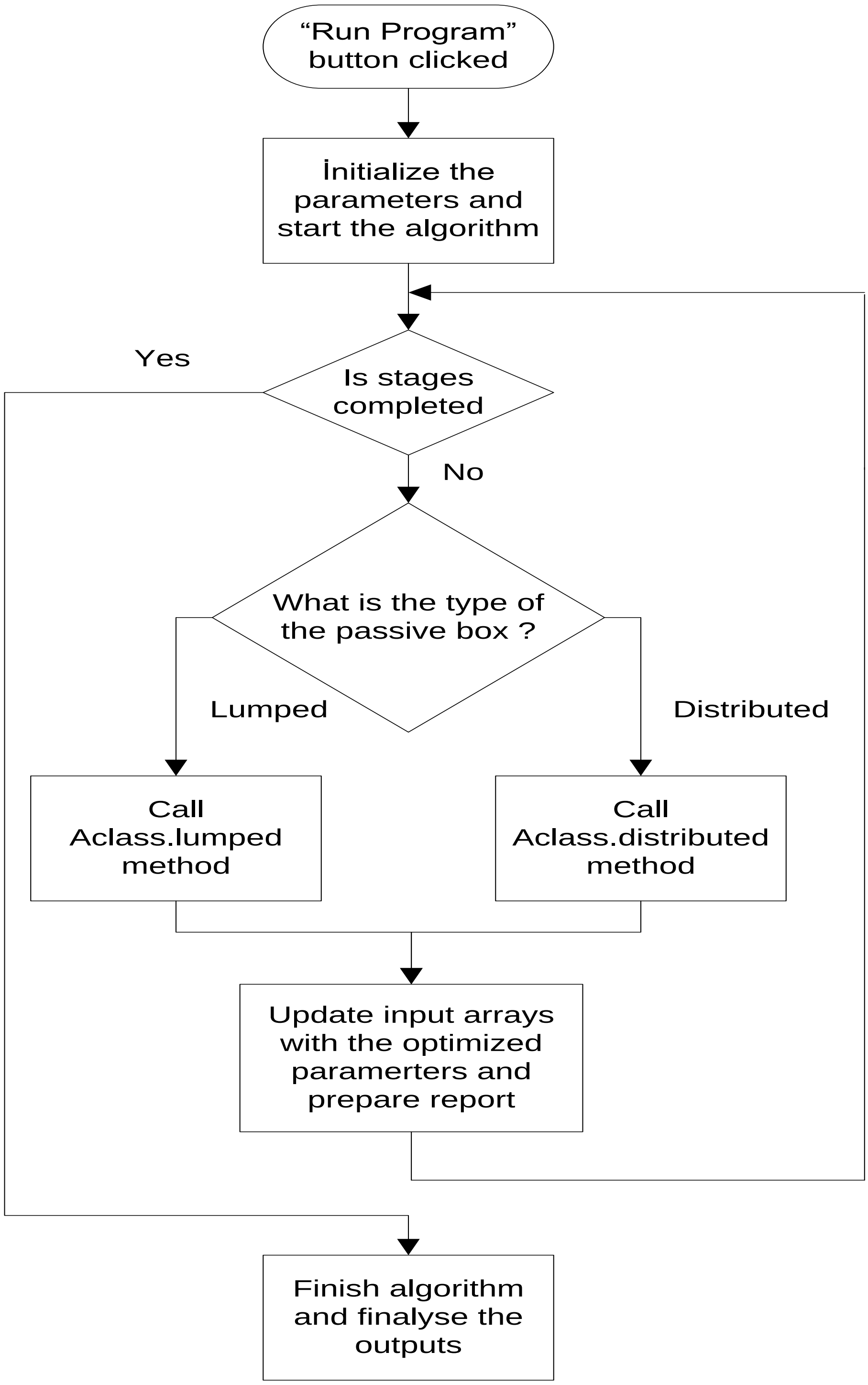


Figure 5.10 Flow diagram of matching algorithm

The appearance of the Analyze and Optimization tab depicted in the Figure 5.11.

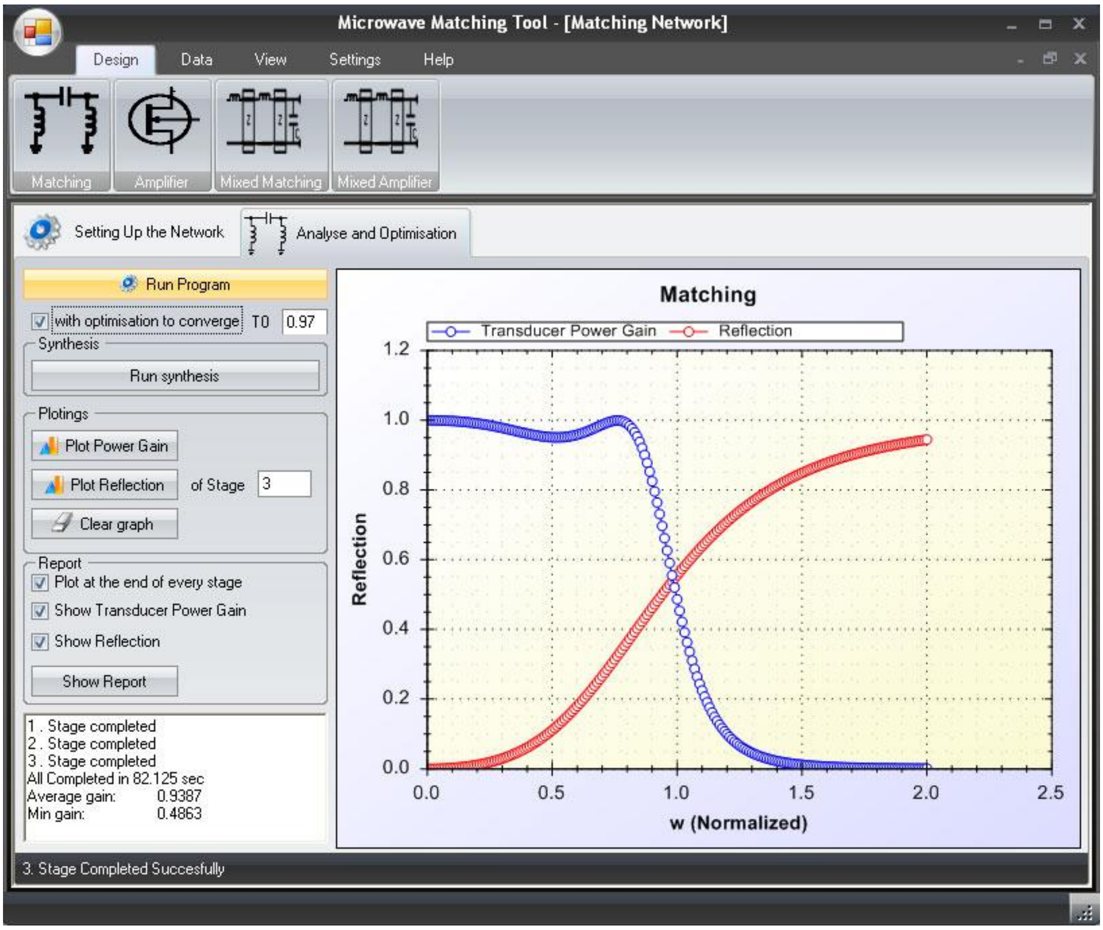


Figure 5.11 Summarized output screen of the matching window

In this window, user can see a summary result about the design process. If one wishes to see more detailed information about the problem, s/he needs to push the "Show Report" button to see a pdf report page.

5.2.3 Amplifier Form

When user clicks the amplifier button, a new window is created and waits for user input to setting up the network. First input is the stage number of the network as in matching case. When user clicks the "Create Button" link, according to the stage number, the overall view of the system is drawn with passive and active boxes and required arrays and cell arrays are created. Below, you can see a one stage amplifier with frontend and backend equalizers.

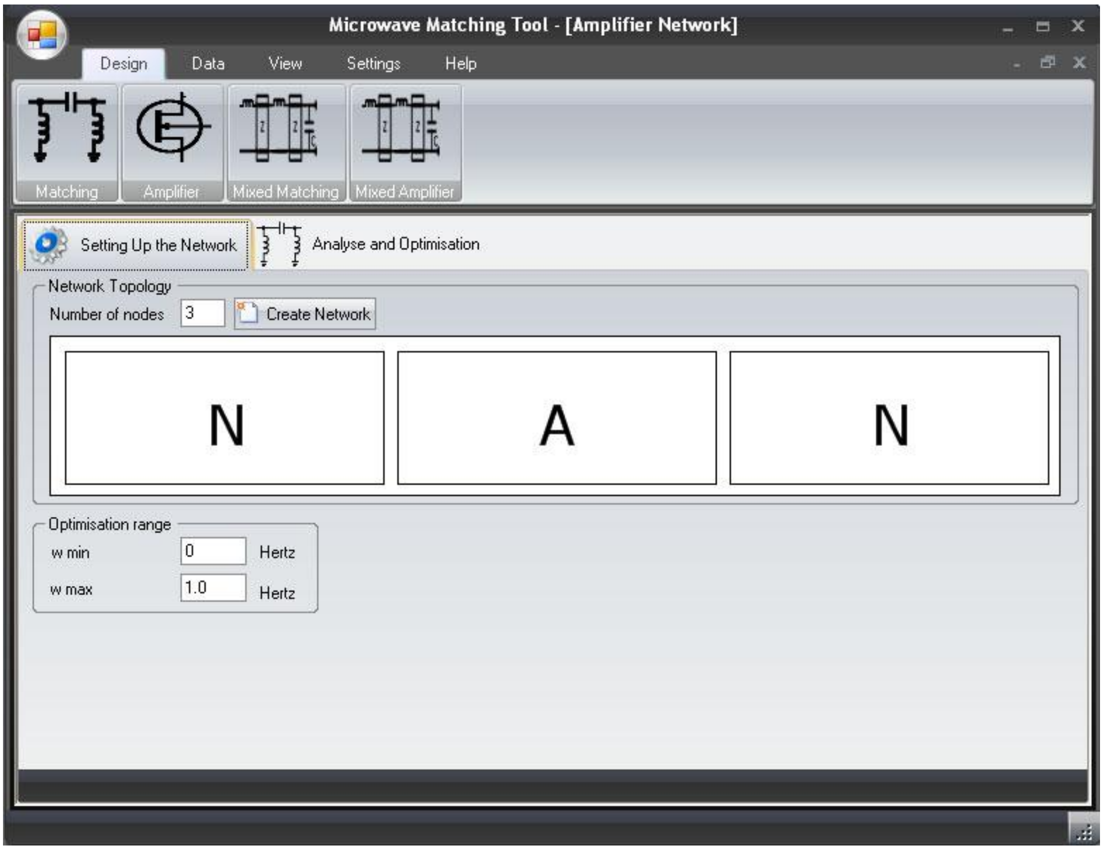


Figure 5.12 Amplifier design window of the tool

If user clicks the passive box area a passive box dialog box will be opened as in matching case. So, it will not be explained again. If user clicks the active box area, a new window, called active device selection window as in the Figure 5.13.

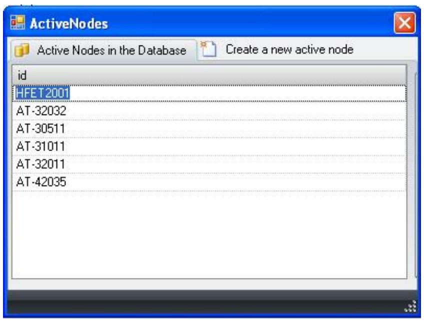


Figure 5.13 Active element selection window

When user selects an active device and close the window, all scattering parameters of the device are taken from database and saved into a cell array for later use.

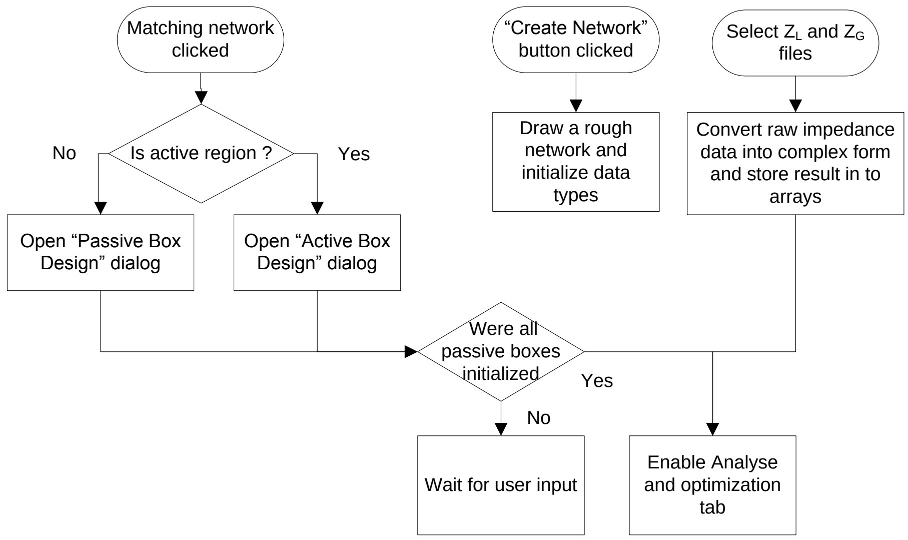


Figure 5.14 Flow diagram of the amplifier design window

After setting the stage number of the amplifier, designing passive nodes and selecting of transistors from database, user can now analyze the circuit by pushing “Run Program” button at the “Analyze and Optimization” tab.

Below you can see the flow diagram of the amplifier design.

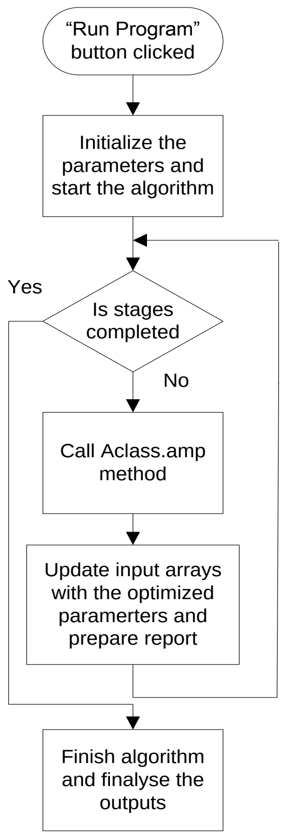


Figure 5.15 Flow diagram of the amplifier algorithm

The appearance of the Analyze and Optimization tab depicted in the Figure 5.16.

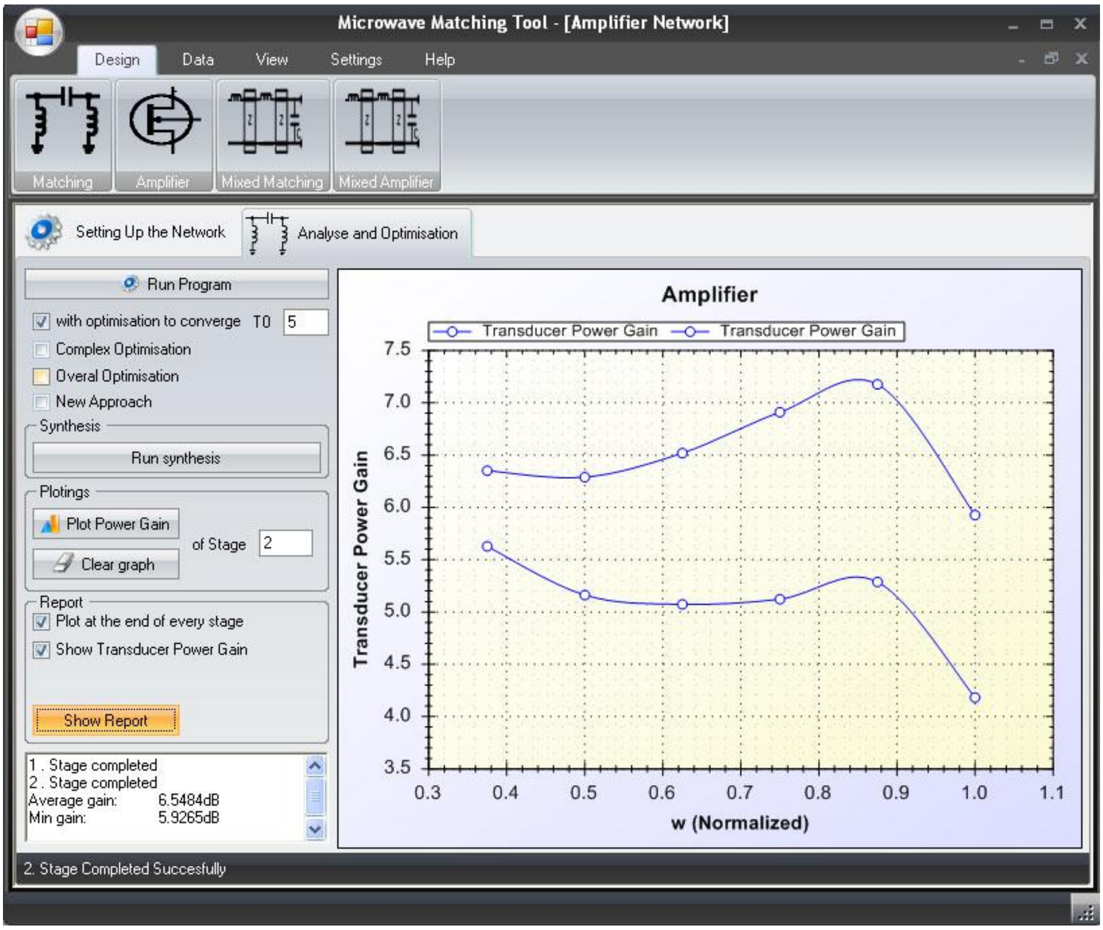


Figure 5.16 Summarized output window of the amplifier window

In this window, user can see a summary result about the design process. If some wishes to see more detailed information about the problem, s/he needs to push the "Show Report" button to see a pdf report page.

5.2.4 Mixed Matching Form

Another matching problem which this tool can perform is generating networks consisting of passive boxes which have low pass ladders with unit elements, LPLUs. When user clicks mixed matching button, a new window which have a similar interface with the one for only lumped and only distributed matching design window, is created and waits for user input to setting up the network. First input is, again, the stage number of the network as in matching and amplifier case. When user clicks the "Create Button" link, according to the stage number, the overall view of the system is drawn with passive boxes and required arrays and cell arrays are created. Below, you can see a three stage matching network.

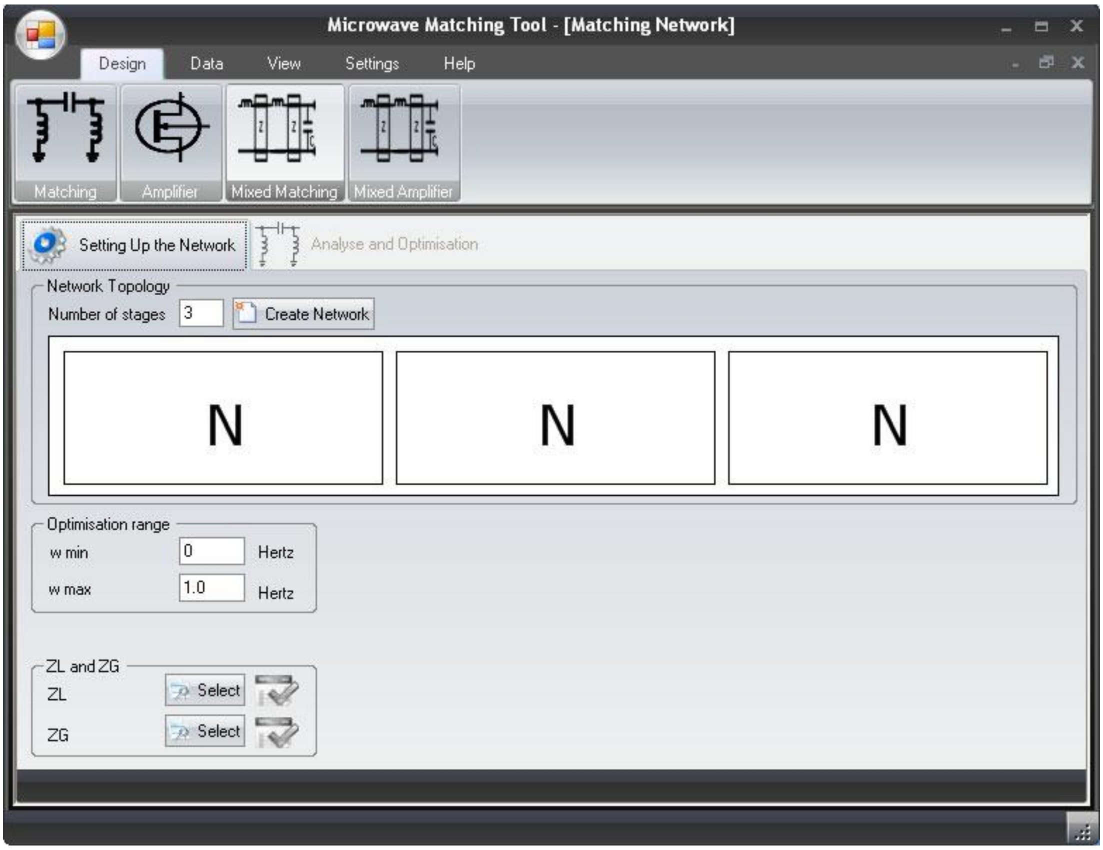


Figure 5.17 Mixed matching design window

If user clicks the passive box area a passive box dialog box will be opened as in the Figure 5.18.

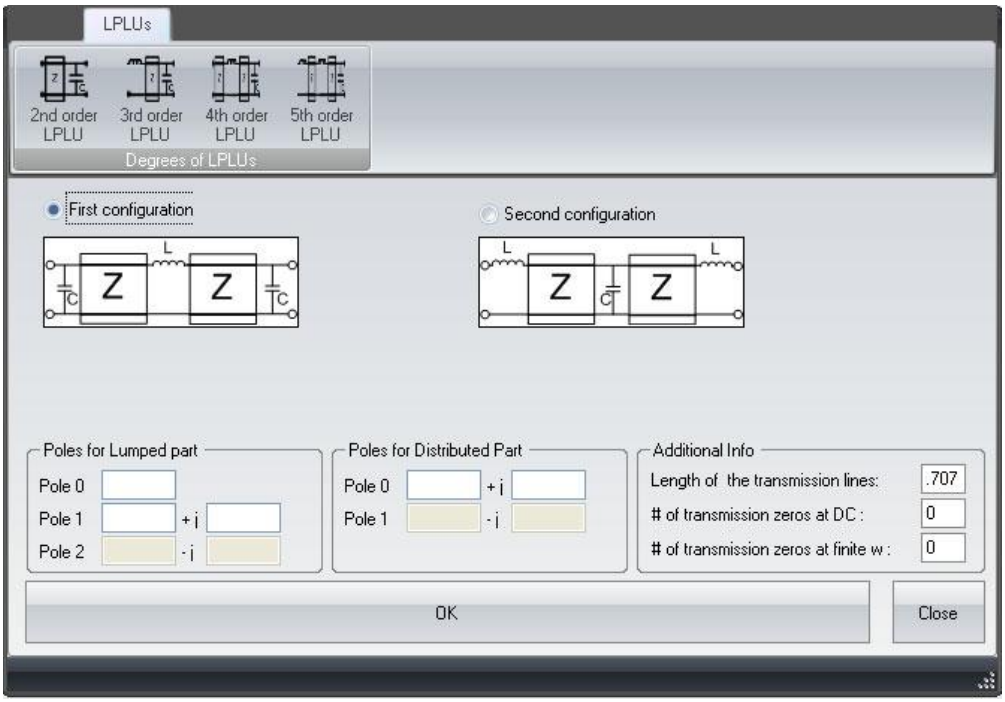


Figure 5.18 Mixed passive box design window

As we discussed in the previous sections, LPLU networks are split into only-lumped and only-distributed sections. As can be seen in the above Figure 5.20, user is asked to enter lumped and distributed poles for passive network. After completing input entry and closing the program, passive box information is saved into related arrays.

Flow diagram of the events is depicted in the Figure 5.19.

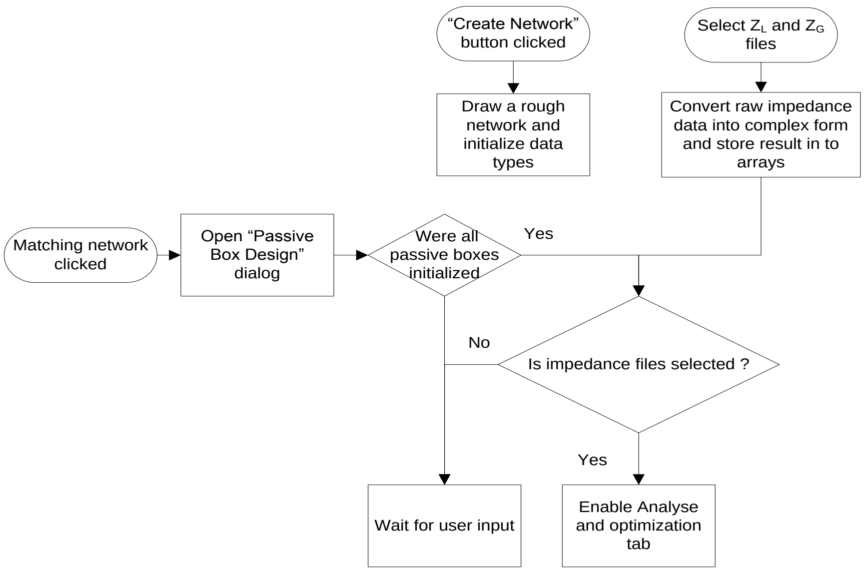


Figure 5.19 Flow diagram of the mixed matching window

Below, you can see the Analyze and Optimization tab.

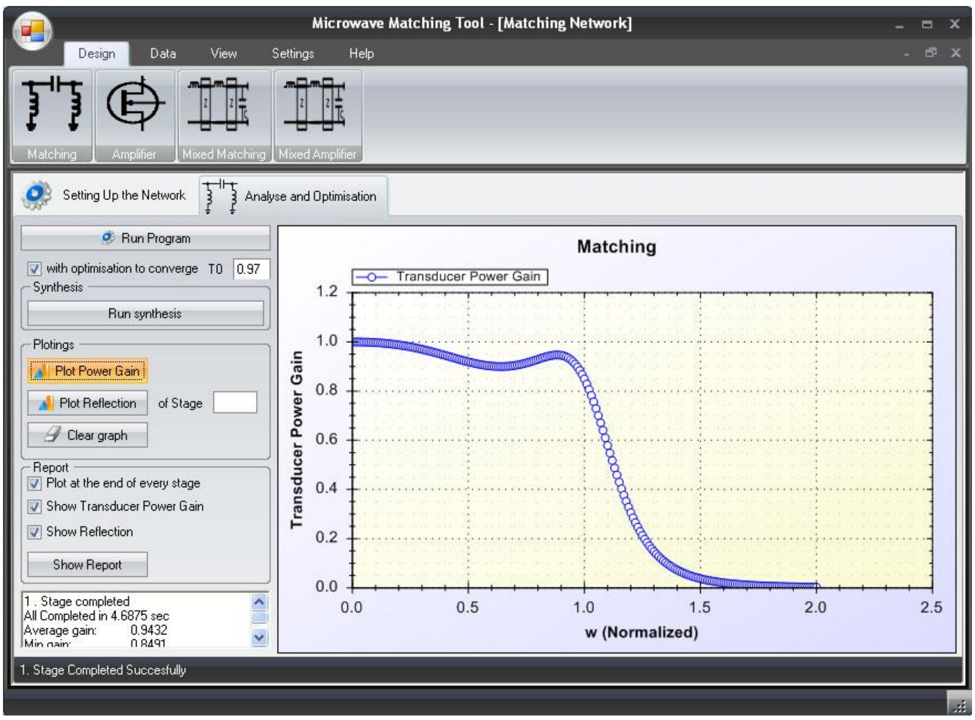


Figure 5.20 Summarized output of the mixed matching window

This is the same window with matching window which we discussed previously. When user pushes to “Run Program” button, mixed matching algorithm is run as follows.

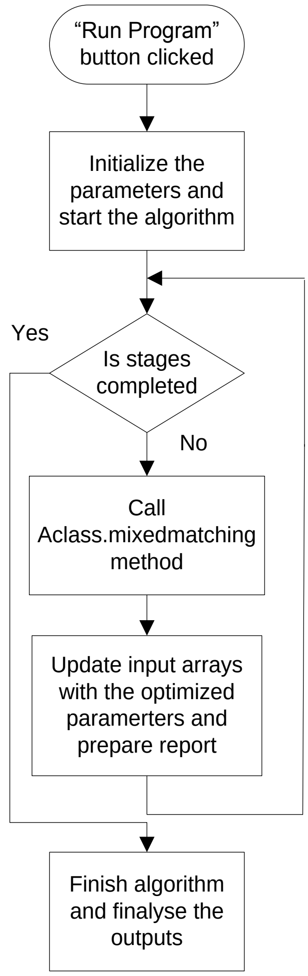


Figure 5.21 Flow diagram of the mixed matching algorithm

5.2.5 Mixed Amplifier Window

We can think that this window is a combination of the previous windows. When user clicks the passive boxes, a passive box design window is shown. We saw that window before. If user clicks the active region, an active device selection window appears. You can see the mixed amplifier window in the Figure 5.22.

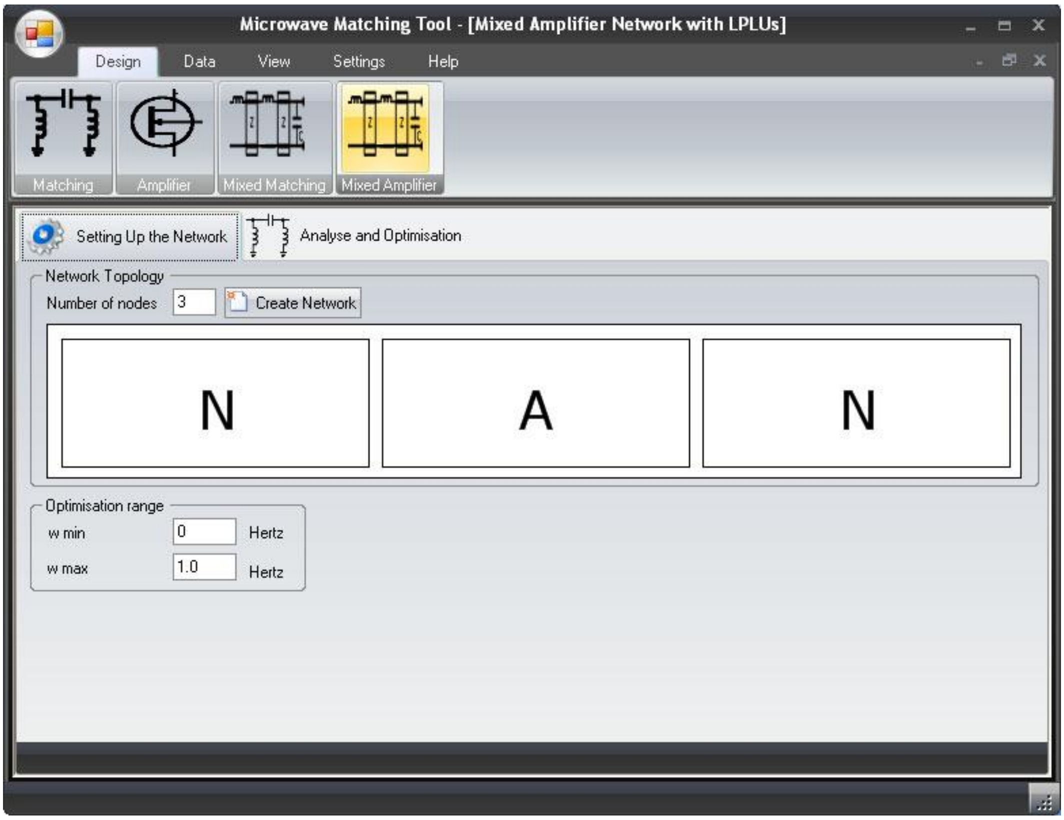


Figure 5.22 Mixed amplifier design window of the tool

When clicked the Analyze and Optimization tab, user can see an interface like below.

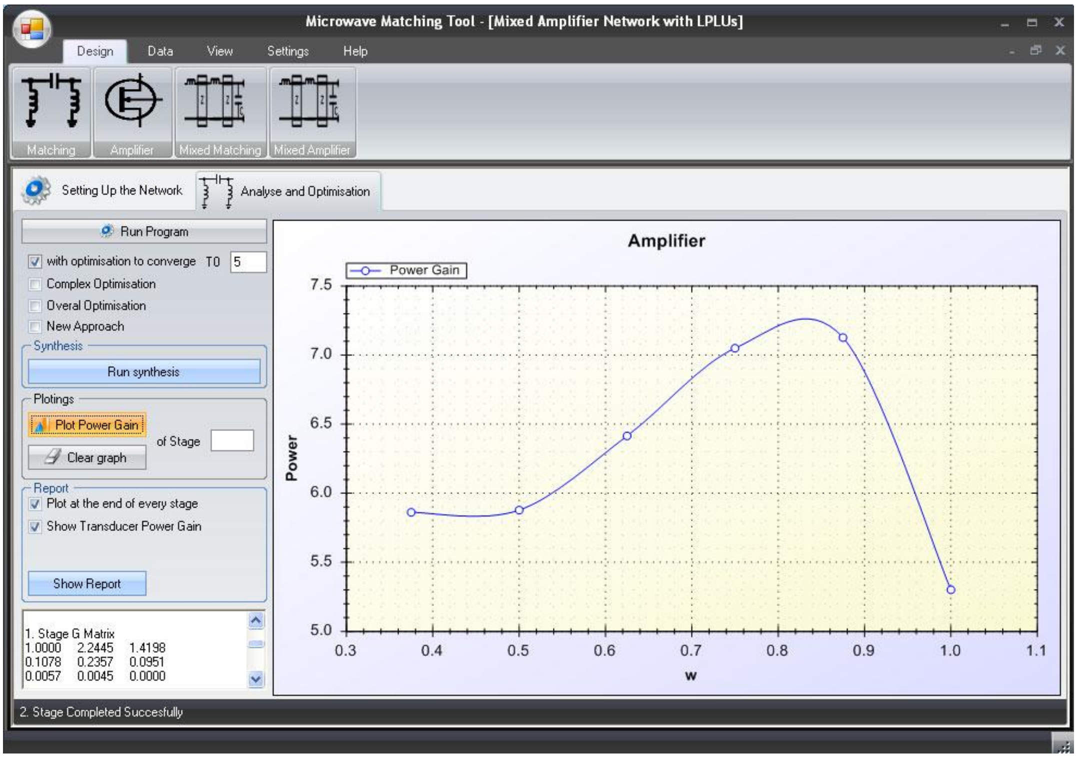


Figure 5.23 Summarized output window of the mixed amplifier window

By clicking “Run program” button, we trigger the mixed amplifier algorithm. Flow diagram of the .algorithm is depicted in the following Figure 5.24.

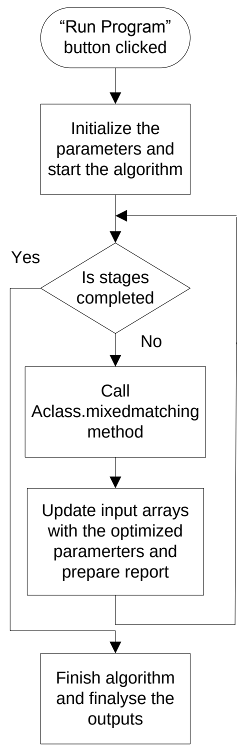


Figure 5.24 Flow diagram of the mixed matching algorithm

5.2.6 Synthesis of Lumped and Distributed Networks

After each design problem, a synthesis method which was written in C#, doSentez() is called. It is responsible for generating actual circuit and finding component values of lumped networks or distributed networks with no transmission lines. doSentez() method is a method of synth class. The synthesis method uses Darlington Zero Shifting-Long Division Algorithm [5] for lumped networks and Richards Extraction algorithm for distributed networks with only transmission lines [6]. For the distributed networks with no transmission lines; synthesis routine is run as is the lumped case. However, after finding lumped component values, they are turned out to the distributed equivalent components.

Chapter 6
Integration of MATLAB with .NET

Most of the numerical operations can be calculated using MATLAB very easily. If some wants to implement a standalone application with lots of calculations, it may not necessary to spend lots of time to create libraries for performing those calculations. You can prepare m files in MATLAB and convert those into a .dll and use it in your .NET project. Distribution of that dll is royalty free and it is not necessary to install MATLAB on the target computer in which your application will be installed.

All you need to write *deploytool* in command prompt and press ENTER key.

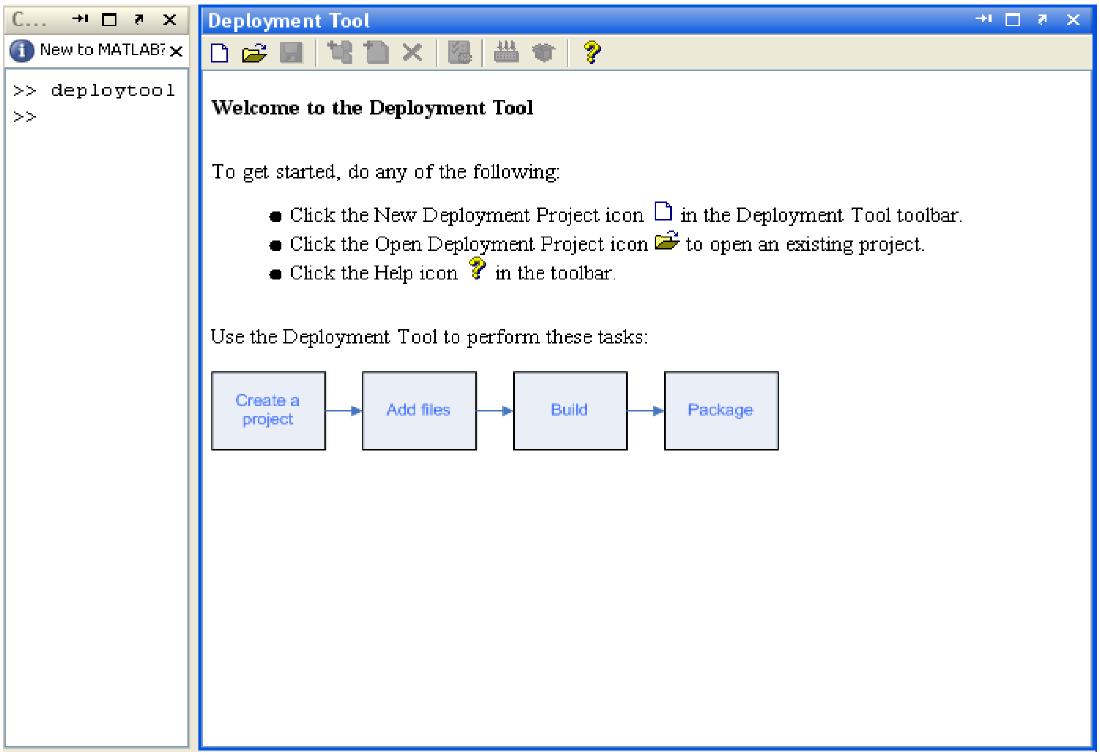


Figure 6.1 Deploy tool for integration MATLAB with .NET

As can be read in the Figure 6.1, you need to press icon to start a project.

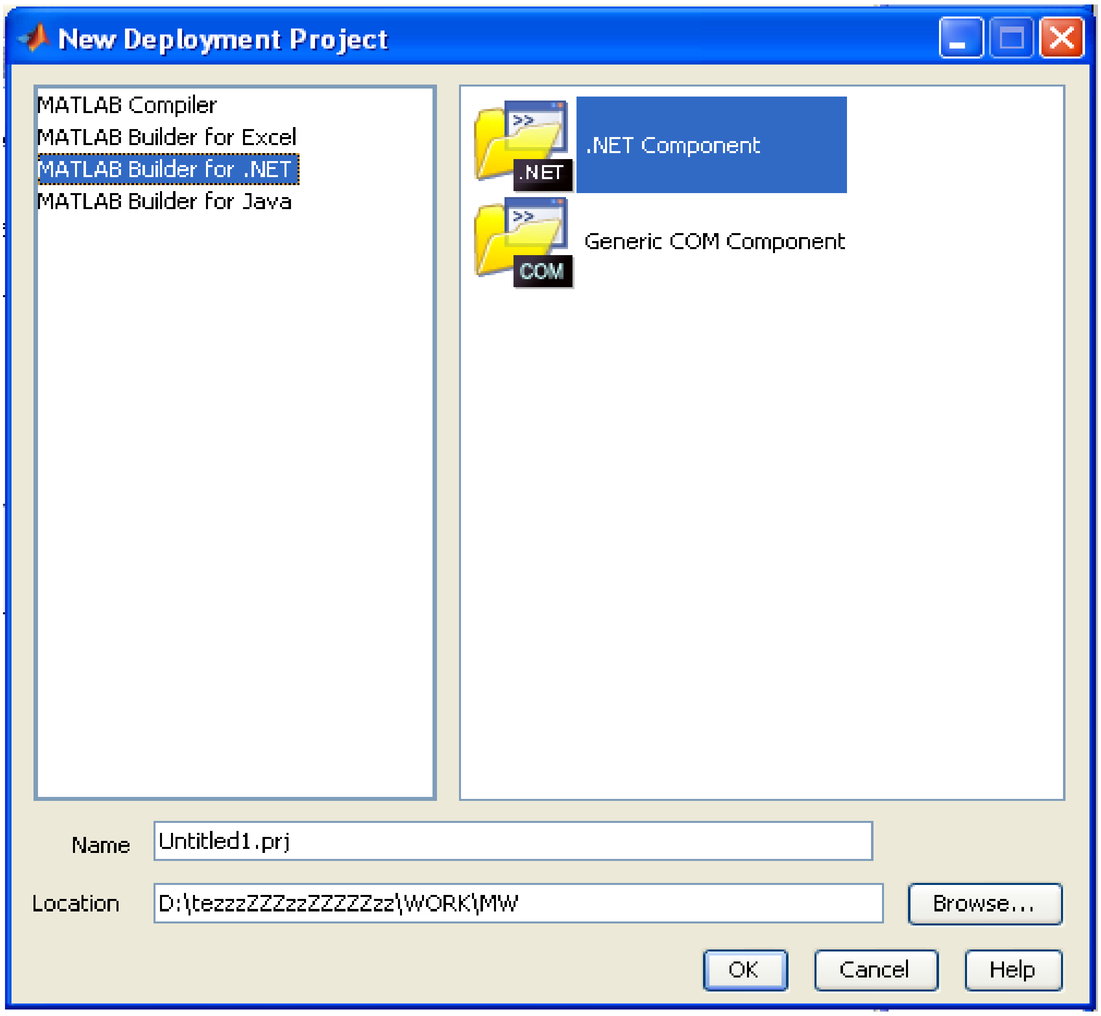


Figure 6.2 Using deployment type

You can select which type of builder and component you want from the menu as shown in Figure 6.2. Do not forget to give a meaningful name to your project. (This will be your default namespace or package name, however; you can change it later).

When you click OK, you will see a project window with only two empty directories. You need to add m files to your project.

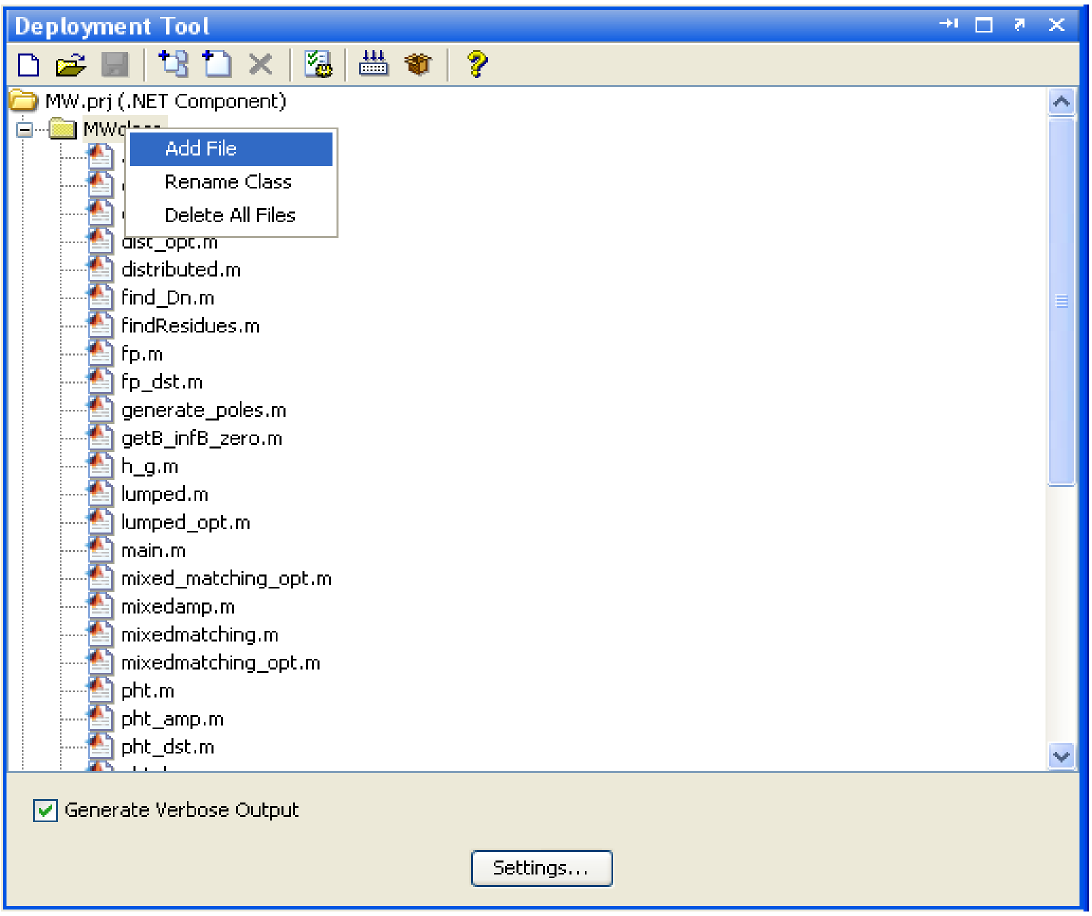


Figure 6.3 Adding a file to the deployment project

When you finish to adding all required m files (only which you wrote, not prewritten files), you can change some settings if you want by clicking icon.

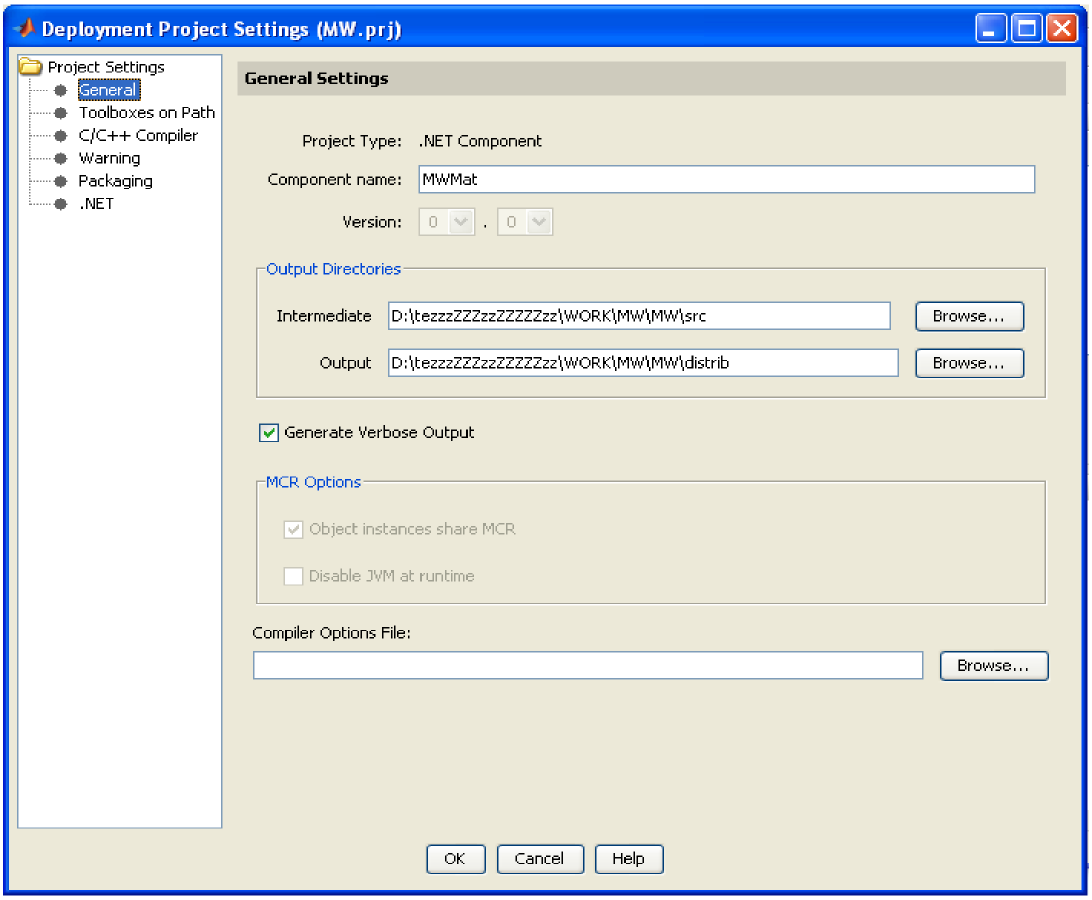


Figure 6.4 Setting class name

In this window, you can change the namespace or package name via changing Component name property. This will be the name of your dll. You can filter the toolboxes to make compilation faster and set the output locations if you want. Another important setting is the version of framework. Below you can see how to select that.

You must create a dll which is suitable with your development environment. So, you need to be sure that you selected correct one.

Figure 6.5 Selecting correct .NET framework

When you finish the configuration you can compile your project by clicking  button.

After compiling is completed you can use that dll in your .NET program. After creating a .NET project in Visual Studio, you need to add a reference to this dll.

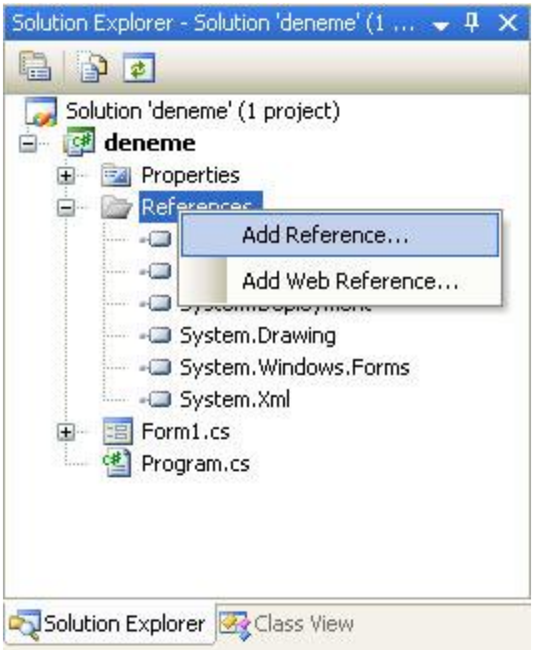


Figure 6.6 Adding reference to dll which was generated by MATLAB

When you click Add Reference, you will see a window on which you can browse your dll.

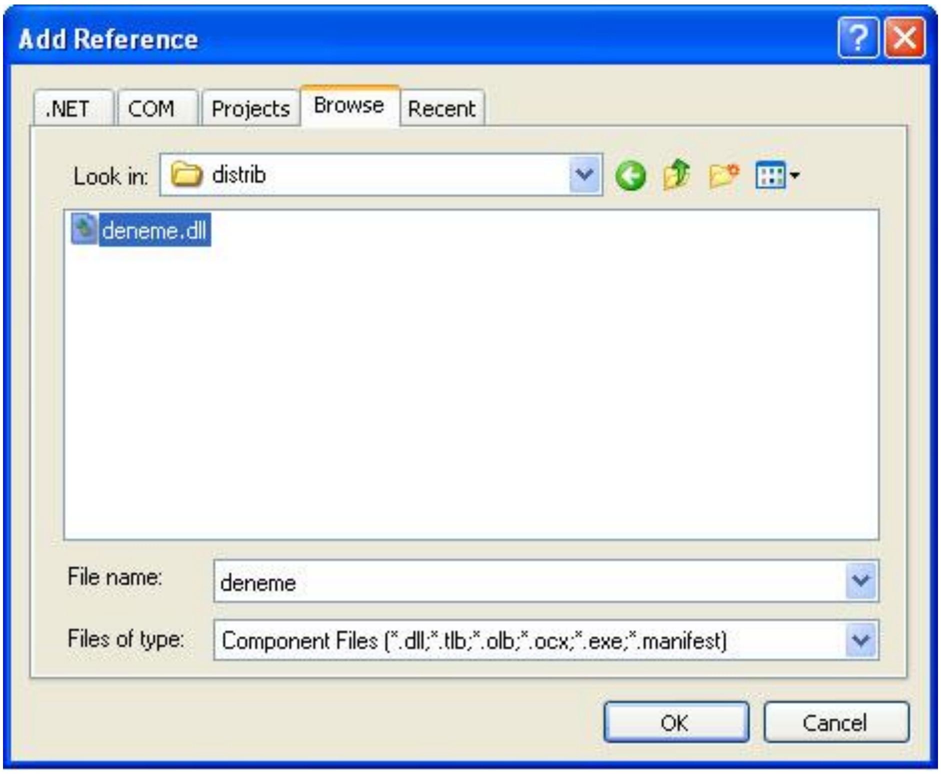


Figure 6.7 Browsing the dll

Another dll that you must use is MWArray.dll. This is dll is installed with MATLAB DotNetBuilder. In order to use MATLAB data types, cells, arrays, in your program you will use this dll. You can find it at the location *C:\Program Files\MATLAB\R2007b\toolbox\dotnetbuilder\bin\win32\v2.0*. Again be careful about the version of framework.

After referencing the required dll's we can start to implement a very simple example.

Here you can see a very simple MATLAB function

```
function [ret]=deneme(param1,param2)
ret=param1.*param2+j*param2;
```

It receives two arrays and returns a complex array.

I created a deployment project named "deneme" and added this file, then compiled it to get deneme.dll. Then, I added deneme.dll and MWArray.dll to my project and started to implement a simple example.

First, you must add these three lines at the top of the code. (Not necessarily top, but it must be in outside of the class)

```
using deneme;
using MathWorks.MATLAB.NET.Utility;
using MathWorks.MATLAB.NET.Arrays;
```

and, below you can find an example code

```
MWArray prm1 = new MWNumericArray(1, 11, new double[] {
5, 1, 2, 3, 4, 5});
MWArray prm2 = new MWNumericArray(1, 11, new double[] {
5, 1, 2, 3, 4, 5});

denemeclass d1 = new denemeclass();
MWArray[] ret = d1.deneme(1,prm1,prm2);

System.Array retreal = new
double[ret[0].NumberOfElements];
retreal =
((MWNumericArray)ret[0]).ToVector(MWArrayComponent.Real);

System.Array retimg = new
double[ret[0].NumberOfElements];
retimg =
((MWNumericArray)ret[0]).ToVector(MWArrayComponent.Imagin
ary);

for (int i = 0; i < retreal.Length; i++)
listBox1.Items.Add(retreal.GetValue(i) + "\tj" +
retimg.GetValue(i));
```

You must split the real and imaginer part of the return value in to two different Array, in my example retreal and retimg. Also you must give the number of return values at the method decleration, in my example the number of return parameters is one. Below, you can find the output

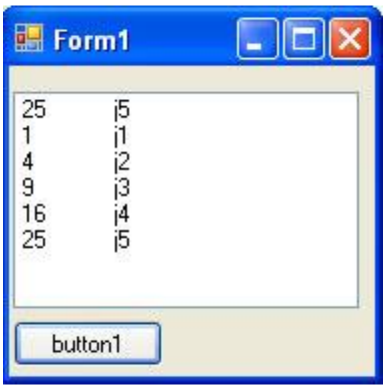


Figure 6.8 An example output

Chapter 7
Applications

In the following, we present two types of applications illustrating, matching and amplifier design examples, with only low-pass type networks. In each of the matching examples, we take generator impedance, $Z_{G,}$ and load impedance, Z_L, from a text file. In amplifier examples, Z_G and Z_L are taken as 1Ω.

7.1 Matching Examples

In this section, we will investigate several types of matching examples like

- Single stage lumped matching example.
- Single stage distributed matching example.
- Multistage lumped matching example.
- Multistage distributed matching example.
- Single stage mixed matching examples with second order LPLU.
- Single stage mixed matching examples with third order LPLU.
- Single stage mixed matching examples with fourth order LPLU.
- Single stage mixed matching examples with fifth order LPLU.
- Multistage mixed matching examples with LPLUs

7.1.1 Lumped Matching Network Design

The purpose is to match Z_G to Z_L over a prescribed bandwidth to maximize the power gain. This is the first example studied in [4].

In this example, the generator impedance Z_G consists of a resistance 1Ω in series with an inductance which is 1 Henry. The load impedance Z_L is given by a parallel R_L and C_L where R_L=1Ω and C_L=1F in series with an inductance L_L which has the value of 2H (Figure 7.1).

The frequency band of the double matching problem is taken over B: $0 \leq w \leq 1$.

Step 1: *Topology of the network and initial guess for input parameters of the algorithm.*

Complexity of the network is taken as three and the poles of the $Z_2(p)$ are given initially as

$p_0 = -0.43$

$p_1 = -0.269 + j0.88$

$p_2 = -0.269 - j0.88$.

Step 2: *Optimization of the transducer power gain.*

In the optimization, the flat-gain level T0=0.97 is picked over the frequency band $0 \leq w \leq 1$, and Z(p) is computed as follows.

$$Z_2(p) = \frac{0.7204p^2 + 0.5352p + 0.2456}{p^3 + 0.7428p^2 + 0.6716p + 0.2456}$$

By using $Z_2(p)$, we can find the optimized poles as below.

$p_0 = -0.4544$

$p_1 = -0.1442 + j0.7209$

$p_2 = -0.1442 - j0.7209$

Step 3: Finally $Z_2(p)$ is synthesized by using zero shifting algorithm described in [5] and the values of the capacitors and inductors are found as follows.

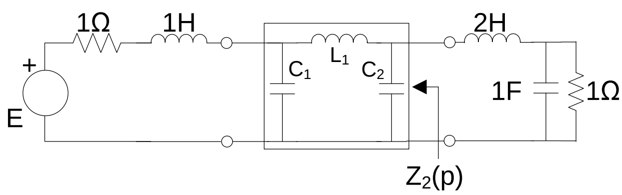


Figure 7.1 Single stage lumped matching network

C_1=1.388F, C_2=1.3463F, L_1=2.1789F

The transducer power gain of the matching design is depicted in the Figure 7.2.

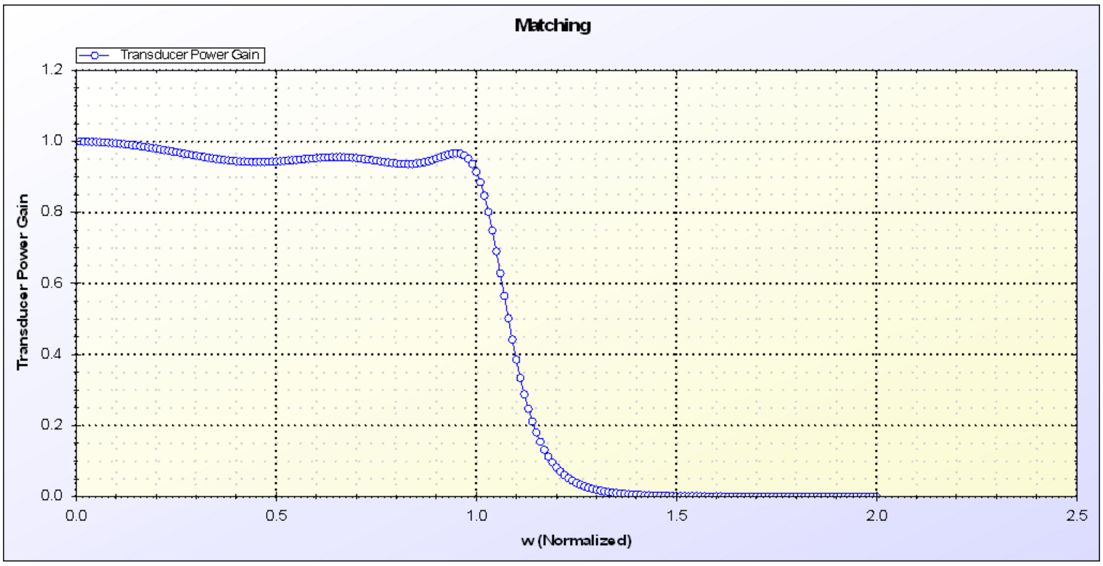


Figure 7.2 Single stage lumped matching transducer power gain

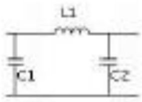
In this design, we get an average gain and minimum gain as 0.9592 and 0.9146, respectively.

You see the output of report page in the Figure 7.3

Matching Network Design

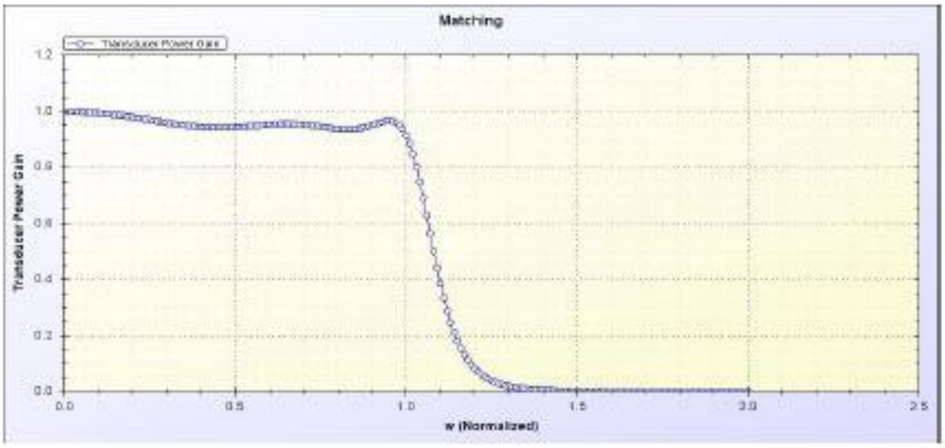
1. Stage of matching network

Type: Lumped
Degree: 3
Poles:
P1: -0.4544
P2: -0.1442 + i0.7209
P3: -0.1442 - i0.7209



z(p)=n(p)/d(p)
n(p): 0.7204 0.5351 0.2456
d(p): 1.0000 0.7428 0.6715 0.2456
h(p): -2.0358 -0.0455 -0.2777 0.0000
g(p): 2.0358 2.9789 2.4566 1.0000
Average Gain: 0.9592
Min Gain: 0.9146

C1: 1.3880F L1: 2.1789H C2: 1.3463F



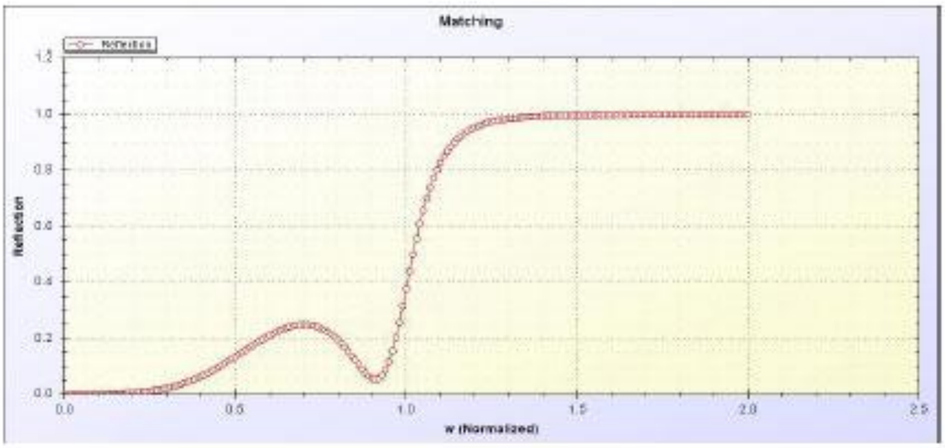


Figure 7.3 Output of the matching report page

7.1.2 Distributed Matching Example

The purpose is to match Z_G to Z_L over a prescribed bandwidth to maximize the power gain. This is the distributed version of first example studied in [4].

In this example, the generator impedance, $Z_{G,}$ and the load impedance, $Z_{L,}$ are same with the previous example.

The frequency band of the double matching problem is taken over B: $0 \leq w \leq 1$.

Step 1: *Topology of the network and initial guess for input parameters of the algorithm.*

Complexity of the network is taken as three. Because of the low pass design, we put three transmission zeros to the infinity. In other words, the number of transmission lines is taken as three. Commensurate delay is taken as 0.707.

The poles of the $Z_2(\lambda)$ are given initially as

$\lambda_0 = -0.43$

$\lambda_1 = -0.269 + j0.88$

$\lambda_2 = -0.269 - j0.88$

Step 2: *Optimization of the transducer power gain.*

Before begin the optimization process, we changed p $(p = \sigma + j\omega)$ with $(\lambda = \Sigma + j\omega)$ and used modified version of the f(p).

In the optimization, the flat-gain level T0=0.97 is picked, as we did in lumped case, over the frequency band $0 \leq w \leq 1$, and $Z(\lambda)$ is computed as follows.

$$Z_2(\lambda) = \frac{0.0084\lambda^3 + 0.6469\lambda^2 + 0.3850\lambda + 0.0918}{\lambda^3 + 0.6367\lambda^2 + 0.3722\lambda + 0.0918}$$

And here are the optimized poles.

$\lambda_0 = -0.3384$

$\lambda_1 = -0.1492 + j0.499$

$\lambda_2 = -0.1492 - j0.499$

Step 3: Finally $Z_2(\lambda)$ is synthesized and the values of the transmission lines are found as follows.

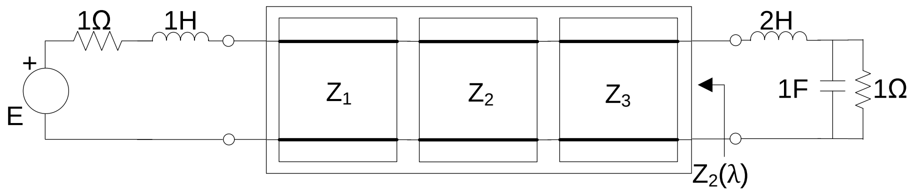


Figure 7.4 Single stage distributed matching circuit

Z_1=0.5401Ω, Z_2=0.0934Ω, Z_3=0.0159Ω

Below, transducer power gain is depicted.

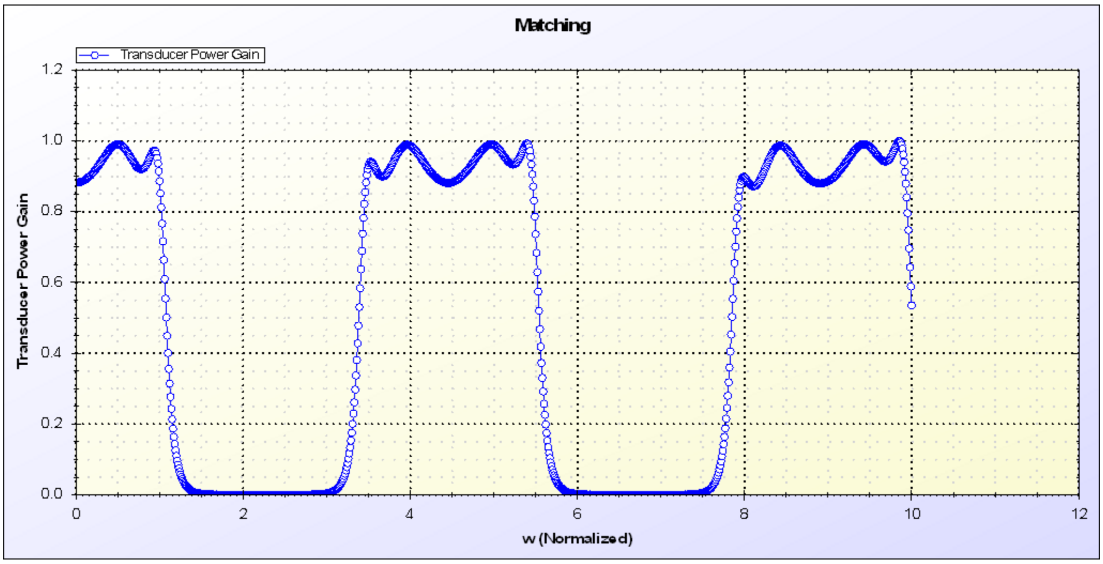


Figure 7.5 Single stage distributed matching transducer power gain

In this design, we get an average gain and minimum gain as 0.9434 and 0.8781, respectively

The output of the detailed report page is depicted in the Figure 7.6

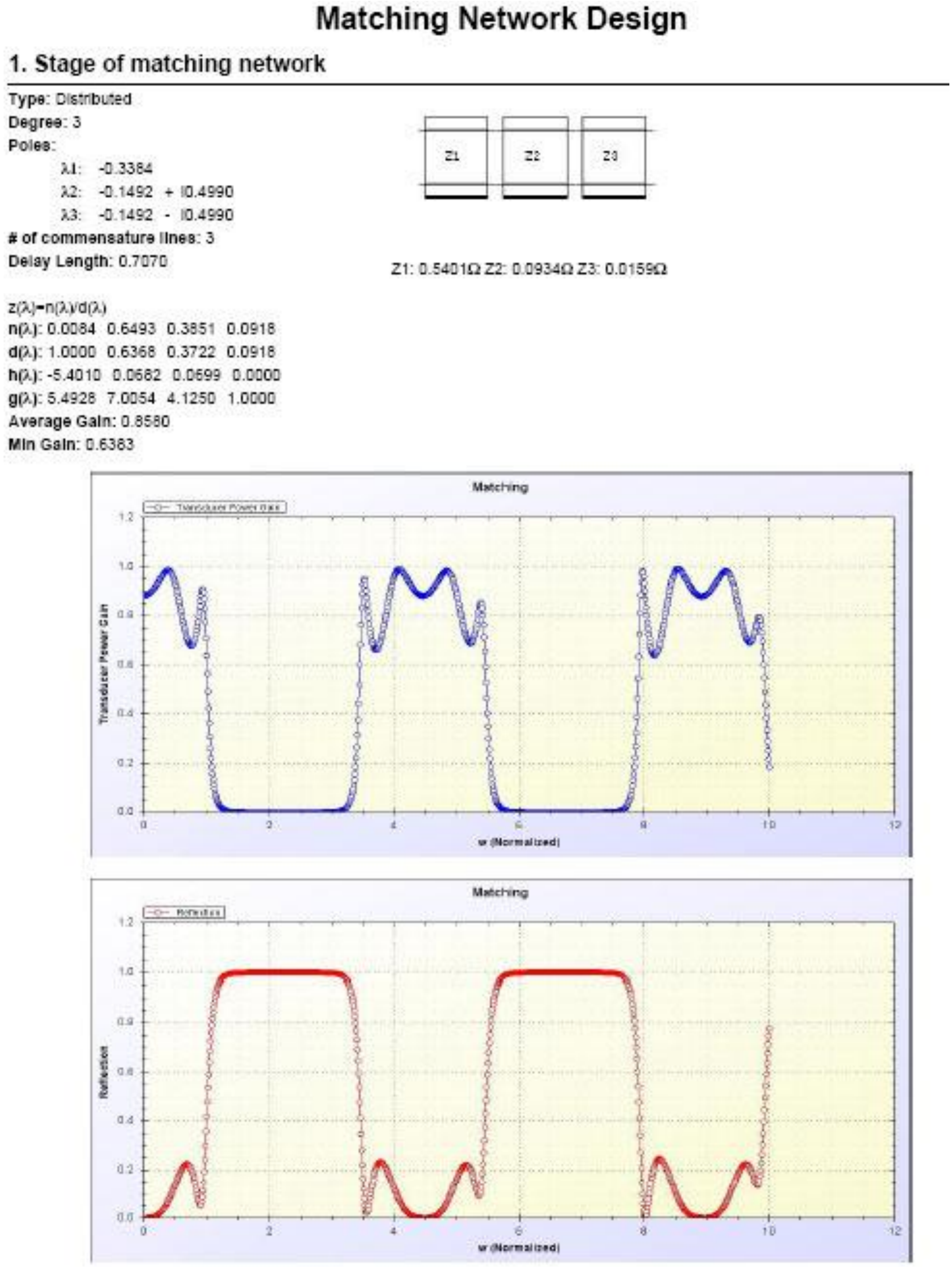


Figure 7.6 Output of the matching report page

7.1.3 Lumped Matching Example with Finite-Zeros.

In this example involves the construction of matching network for a broadband UHF antenna over the normalized frequency band of 0.575 Hz to 1.75 Hz. The antenna impedance is given as measured data and is listed as follows.

Frequency	**R_L**	**X_L**
0.575	12.00	6.00
0.6	7.00	-6.50
0.75	1.30	-1.70
1.0	0.93	-0.38
1.125	1.07	-0.25
1.25	1.17	-0.30
1.375	1.07	-0.38
1.5	0.93	-0.34
1.625	0.83	-0.31
1.75	0.72	-0.26

Step 1: We chose Z_G as 1Ω, complexity of the matching network, n, as 5, number of transmission zeros at DC, m_1, as 1 and number of transmission zeros at finite frequency of 2.5 Hz, m_2, as 1.

Step 2: *Optimization of the transducer power gain.*

After optimization process what we get is as follows.

$$Z_2(p) = \frac{214.6807p^4 + 672.1010p^3 + 1227.1724p^2 + 1931.2233p}{p^5 + 3.1307p^4 + 206.7612p^3 + 407.9940p^2 + 1141.8518p + 222.4079}$$

and

$$p_0 = -0.2087$$
$$p_1 = -0.9030 + j2.1321$$
$$p_2 = -0.9030 - j2.1321$$
$$p_3 = -0.5580 + j14.0877$$
$$p_4 = -0.5580 - j14.0877$$

Step 3: Finally $Z_2(\lambda)$ is synthesized and the values of the lumped components are found as follows.

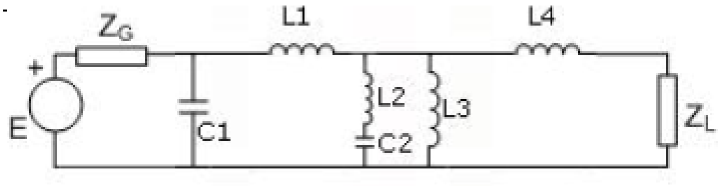


Figure 7.7 Matching example with finite zeros

C_1=0.0047F, C_2=0.2126F, L_1=0.9992H, L_2=0.7524H, L_3=7.6841H, L_4=0.0763H

Below, transducer power gain is depicted.

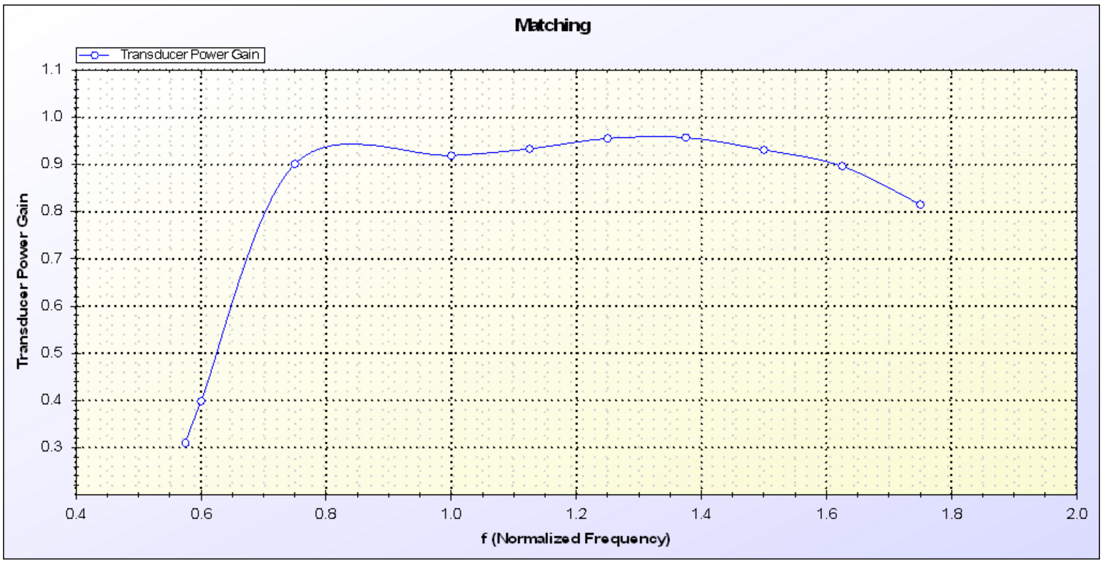


Figure 7.8 Transducer power gain of the matching example with finite zeros

Some can change the bandwidth of the network by changing the location of the finite-zero. Below, the transducer power gain of the same problem with no finite zero is depicted (All finite zeros are at infinity).

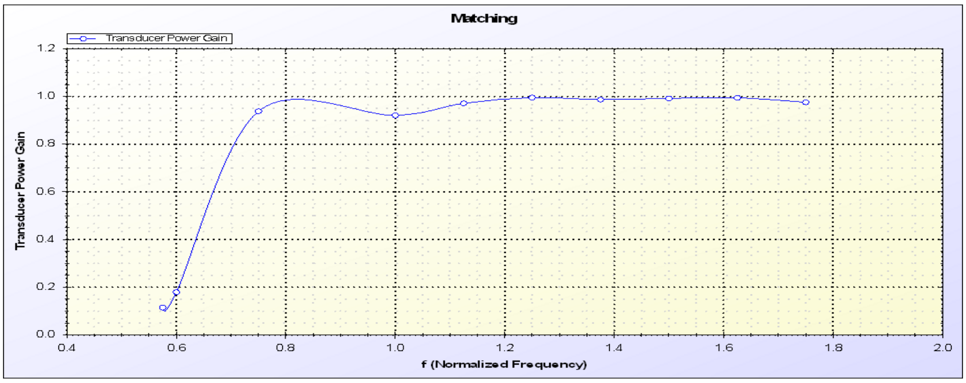


Figure 7.9 Antenna matching example with no finite zeros

The circuit and component values corresponding to antenna matching example with no finite zeros is depicted in the Figure 7.10.

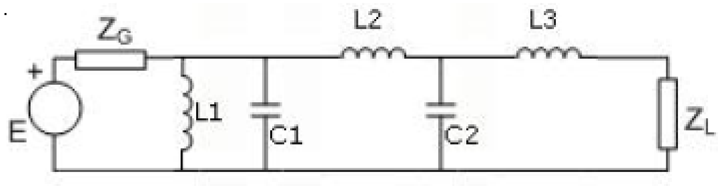


Figure 7.10 Matching example with no finite zeros

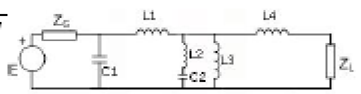
C_1=0.0821F, C_2=0.3737F, L_1=1.8288H, L_2=1.0563H, L_3=0.6769H

Finally, the output of the report page of the example with finite zeros at frequency 2.5 Hz is depicted in the Figure 7.11.

Matching Network Design

General Information about Matching Network

	Generator Impedance		Load Impedance	
f	Real	Imaginary	Real	Imaginary
0.575	1.0000	0.0000	12.0000	6.0000
0.6	1.0000	0.0000	7.0000	-6.5000
0.75	1.0000	0.0000	1.3000	-1.7000
1	1.0000	0.0000	0.9300	-0.3600
1.125	1.0000	0.0000	1.0700	-0.2500
1.25	1.0000	0.0000	1.1700	-0.3000
1.375	1.0000	0.0000	1.0700	-0.3600
1.5	1.0000	0.0000	0.9300	-0.3400
1.625	1.0000	0.0000	0.8300	-0.3100
1.75	1.0000	0.0000	0.7200	-0.2600



C1: 0.0047F C2: 0.2126F
L1: 0.9992H L2: 0.7524H L3: 7.6841H L4: 0.0763H

Type: Lumped,Highpass or Bandpass

Degree: 5

Poles:

P1: -0.2067
P2: -0.9030 + I2.1321
P3: -0.5580 + I14.0877
P4: -0.9030 - I2.1321
P5: -0.5580 - I14.0877

z(p)=n(p)/d(p)
n(p): 0.0000 214.6807 672.1010 1.227.1724 1.931.2233 0.0000
d(p): 1.0000 3.1307 206.7612 407.9940 1.141.8518 222.4079
h(p): -0.0022 0.4756 1.0461 1.8416 1.7746 -0.5000
g(p): 0.0022 0.4897 1.9758 3.6761 6.9086 0.5000
Average Gain: 0.8018
Min Gain: 0.3100

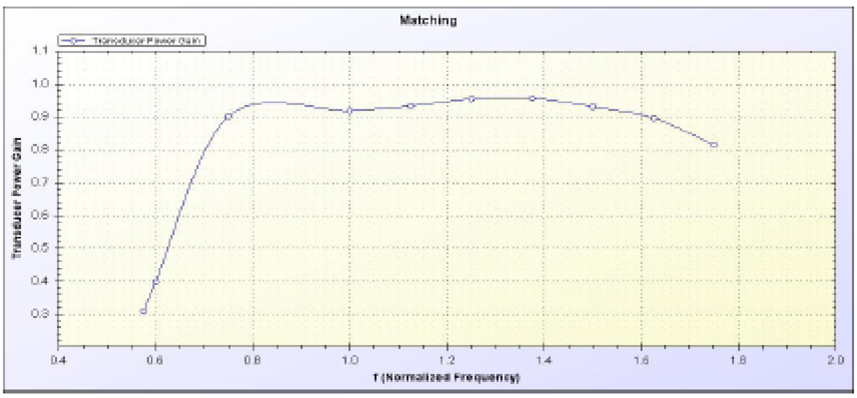


Figure 7.11 Output of the report page of the antenna matching example

7.1.4 Lumped Matching Example with Zeros at DC

In this example, the generator impedance Z_G consists of a resistance 1Ω parellel with the serial connection of a 2F capacitor and an inductor which is 1 Henry. The load impedance Z_L is given by a parallel resistor, R_L, and inductor, L_L, where R_L=1Ω and L_{L1}=4H in series with an inductance L_{L2} which has the value of .75H

The frequency band of the double matching problem is taken over B: $0.5 \leq w \leq 1$.

Step 1: *Topology of the network and*

Complexity of the network is taken as five and all the transmission zeros are set at DC, m_1=5

Step 2: *Optimization of the transducer power gain.*

In the optimization, the flat-gain level T0=1 is picked over the frequency band $0.5 \leq w \leq 1$, and Z_2(p) is computed as follows.

$$Z_2(p) = \frac{0.0757p^5 + 0.4428p^4 + 1.0316p^3 + 0.1785p^2 + 0.3377p}{p^5 + 2.4731p^4 + 0.8391^3 + 1.7455p^2 + 0.1454p + 0.2751}$$

By using $Z_2(p)$, we can find the optimized poles as below.

$p_0 = -2.4220$

$p_1 = -0.0020 + j0.4880$

$p_2 = -0.0020 - j0.4880$

$p_3 = -0.0236 + j0.6902$

$p_4 = -0.0236 - j0.6902$

Step 3: Finally $Z_2(p)$ is synthesized and the values of the capacitors and inductors are found as follows.

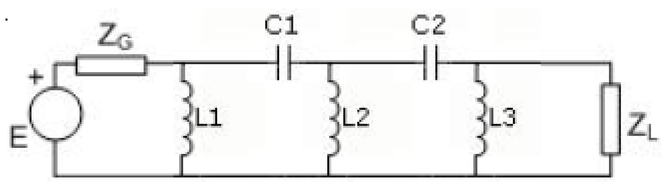


Figure 7.12 Lumped matching example with zeros at DC

L1=1.2276H, L2=0.1458H, L3=0.0400H, C1=2.6805F, C2=14.3415F

The transducer power gain of the matching network is depicted in the Figure 7.13.

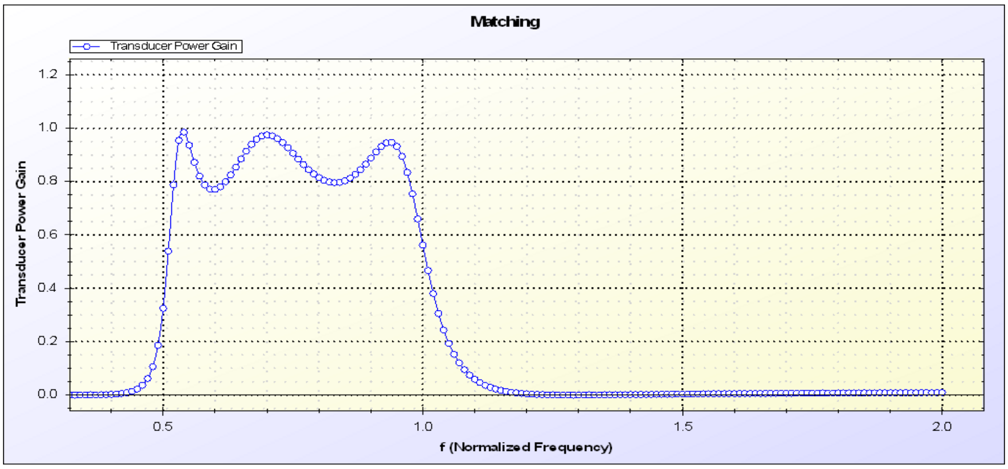


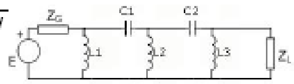
Figure 7.13 Transducer power gain of lumped matching example with zeros at DC

The report page of the matching problem is depicted in the Figure 7.14.

Matching Network Design

General Information about Matching Network

	Generator Impedance		Load Impedance	
f	Real	Imaginary	Real	Imaginary
0.1	0.9611	-0.1934	0.1379	0.4198
0.29	0.7263	-0.4448	0.5737	0.7120
0.48	0.4462	-0.4971	0.7866	0.7697
0.67	0.2292	-0.4203	0.8778	0.8300
0.86	0.0947	-0.2928	0.9221	0.9130
1.05	0.0253	-0.1571	0.9464	1.0128
1.24	0.0010	-0.0312	0.9609	1.1237
1.43	0.0063	0.0789	0.9703	1.2421
1.62	0.0305	0.1719	0.9767	1.3657
1.81	0.0664	0.2490	0.9813	1.4930
2	0.1091	0.3118	0.9846	1.6231



C1: 2.6805F C2: 14.3415F
L1: 1.2276H L2: 0.1458H L3: 0.0400H

Type: Lumped,Highpass or Bandpass

Degree: 5

Poles:

P1: -2.4220
P2: -0.0020 + I0.4880
P3: -0.0236 + I0.6902
P4: -0.0020 - I0.4880
P5: -0.0236 - I0.6902

z(p)=n(p)/d(p)

n(p): 0.0757 0.4428 1.0316 0.1785 0.3377 0.0000

d(p): 1.0000 2.4731 0.8391 1.7455 0.1454 0.2751

h(p): -1.6800 -3.6902 0.3499 -2.8482 0.3496 -0.5000

g(p): 1.9551 5.2998 3.4001 3.4970 0.8781 0.5000

Average Gain: 0.8435

Min Gain: 0.3260

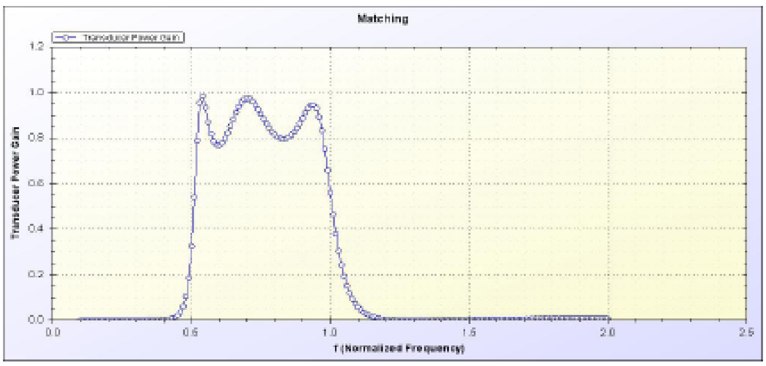


Figure 7.14 Output of the report page of the matching example with transmission zeros at DC

7.1.5 Mixed Matching Example with LPLU of Degree 5

In this example, degree is 5. So we need to split whole network in to two parts. You can find the result of the optimization routine below.

$P_0 = -0.6013$
$P_1 = -0.2211 + j1.4362$
$P_1 = -0.2211 - j1.4362$

$\lambda_0 = -0.7244 - j0.6421$
$\lambda_1 = -0.7244 - j0.6421$

$$Z_2(0,\lambda) = \frac{0.8780\lambda^2 + 2.4267\lambda + 0.9370}{\lambda^2 + 1.4487\lambda + 0.9370}$$

$$Z_2(p,0) = \frac{1.0938p^2 + 1.1413p + 1.2696}{p^3 + 1.0434p^2 + 2.3774p + 1.26960}$$

$$\Lambda_h = \begin{pmatrix} 0 & 0.5218 & -0.0651 \\ -0.4868 & 0.0345 & -1.2081 \\ 0.0198 & -1.7858 & 0 \\ -0.3938 & 0 & 0 \end{pmatrix}$$

$$\Lambda_g = \begin{pmatrix} 1 & 2.0680 & 1.0021 \\ 1.3858 & 3.1198 & 1.2081 \\ 0.8417 & 1.7858 & 0 \\ 0.3938 & 0 & 0 \end{pmatrix}$$

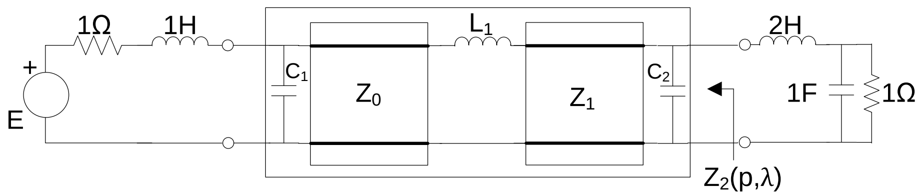


Figure 7.15 Mixed matching circuit with fifth order LPLU

Z_0=0.9624Ω, Z_1=0.7214Ω , C_1=0.9142F, C_2=0.9582F, L_1=0.8989H

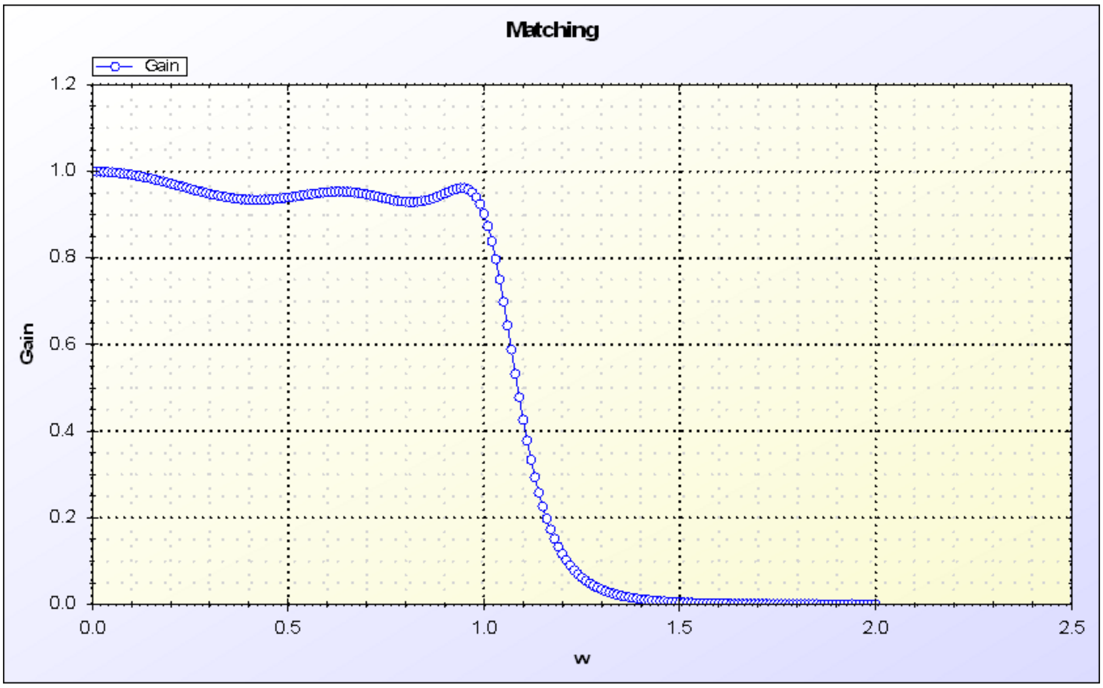


Figure 7.16 Transducer power gain of mixed matching network with fifth order LPLU

7.2 Amplifier Design Examples

In this section, we will investigate several types of amplifier examples like

- Single stage lumped amplifier example.
- Single stage distributed amplifier example.
- Single stage mixed amplifier example
- Single stage mixed amplifier example with LPLUs
- Multistage lumped amplifier example.
- Multistage distributed amplifier example.
- Multistage mixed amplifier example

7.2.1 Single-Stage Lumped Amplifier Example

This amplifier design example is taken from [6]. We get very close results. In this example Z_G and Z_L is taken as 1 Ω. HFET2001 is used as an active device. We examined amplifier examples within the frequency band of 6-16GHz. We took the scattering parameters of the transistor from the database.

Step 1: Stage of the amplifier is set as one

Step 2: The complexity of the backend and frontend equalizers is set.

The degree of frontend network and backhand network is taken as 3. According to degree of networks, initial parameters are given as follows.

FRONTEND

$\mathbf{P_0}$ = -15.3841
$\mathbf{P_1}$ =-0.5427 + j 0.7518
$\mathbf{P_2}$ =-0.5427 - j 0.7518

BACKEND

$\mathbf{P_0}$ = - 1.1748
$\mathbf{P_1}$ =-0.2813+ j 0.8107
$\mathbf{P_2}$ =-0.2813 - j 0.8107

Foster Section
Serial inductor : 3.23

These values are computed by h and g polynomials taken form [6]. Below, you can find the non-optimized results of the design by using poles above.

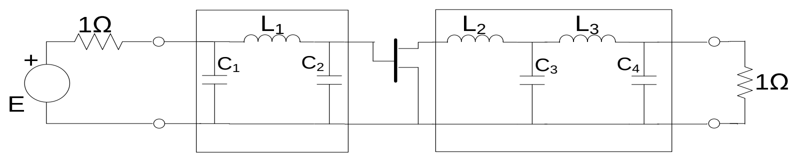


Figure 7.17 Single stage lumped amplifier circuit

C_1=1.2663F, L_1=0.8360H, C_2=0.1295F, L_2=3.23H, C_3=1.0395F, L_3=1.9308H, C_4=0.5741F

You can see the transducer power gain below

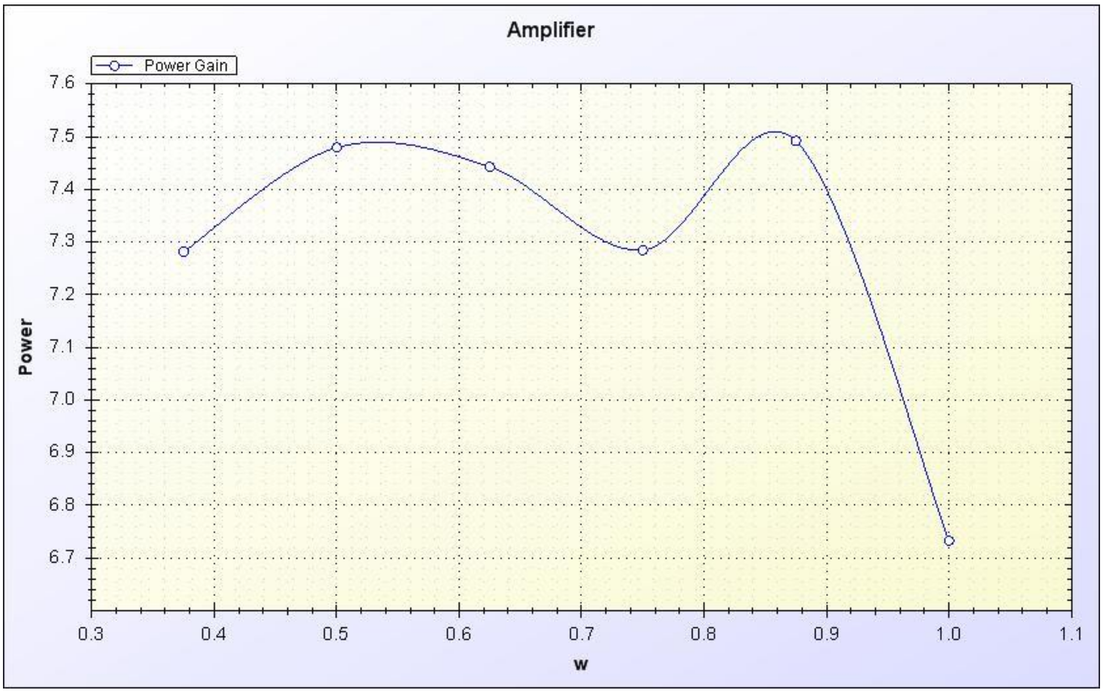


Figure 7.18 Transducer power gain of single stage lumped amplifier network with no optimization

We get very close transducer power gain to the example in [6].

Step 3: *Optimization*

When we input all required parameters, we run the optimization routine to make transducer power gain to converge desired level. The steps to be needed to compute transducer power gain is mentioned in 4.2.1. If we take the desired gain level 10 and perfect match, after optimization we get

FRONTEND

$$Z_2(p) = \frac{0.7079p^2+14.8437p+14.8721}{p^3+20.970p^2+21.7194p+14.8721}$$

$$P_0 = -19.9171$$

$$P_1 = -0.5265 + j\,0.6852$$

$$P_2 = -0.5265 - j\,0.6852$$

BACKEND

$$Z_2(p) = \frac{2.9937p^4+14.2506p^3+10.9204p^2+13.6662p+2.8689}{p^3+4.7602p^2+3.2914p+2.8689}$$

$$P_0 = -4.1316$$

$$P_1 = -0.3143 + j\,0.7717$$

$$P_2 = -0.3143 - j\,0.7717$$

Step 4: *Synthesis*

After synthesis we get the following component result

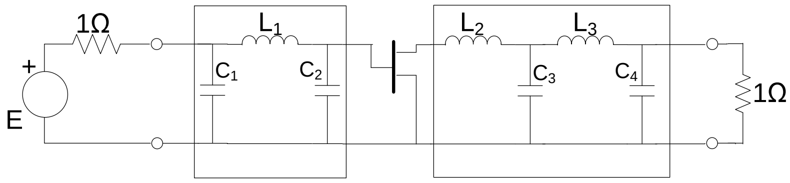


Figure 7.19 Single stage lumped amplifier circuit

C_1=1.4127, L_1=0.9981H, C_2=0.0477F, L_2=2.9937H, C_3=0.9373F, L_3=1.7704H, C_4=0.210F

Transducer power gain of the system is depicted as follows.

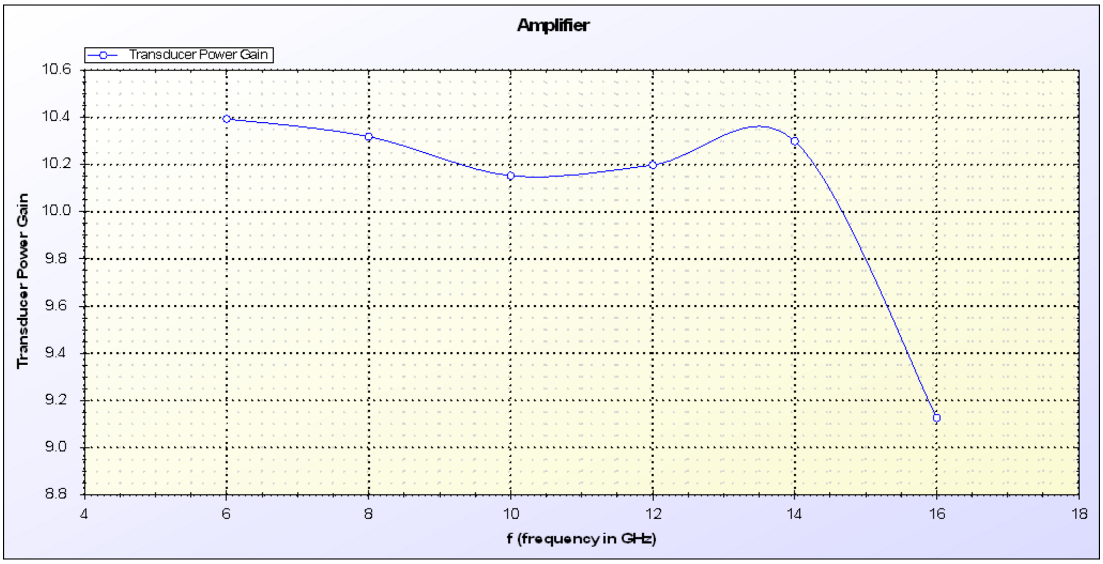


Figure 7.20 Transducer power gain of single stage lumped amplifier network with optimization and perfect match

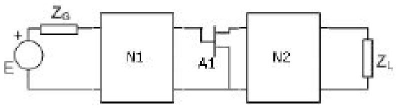
Average gain = 9.8818dB

Minimum gain in band = 9.7258 dB

The output of the report page is depicted as follows;

Amplifier Design

General Information about Amplifier Design



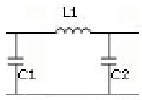
1. Transistor Parameters A1 (HFET2001)

f(GHz)	A11M	A11P	A12M	A12P	A21M	A21P	A22M	A22P
6	0.88	-65	0.05	60	2	125	0.71	-22
8	0.83	-85	0.06	53	1.81	109	0.68	-30
10	0.79	-101	0.06	51	1.64	95	0.66	-37
12	0.76	-113	0.06	52	1.48	84	0.66	-43
14	0.73	-126	0.06	54	1.39	73	0.64	-48
16	0.71	-141	0.07	55	1.32	61	0.63	-56

Figure 7.21 Output of the first page of amplifier report

N1

Type: Lumped
Degree: 3
Poles:
- **P1:** -19.9171
- **P2:** -0.5265 + i0.6852
- **P3:** -0.5265 - i0.6852



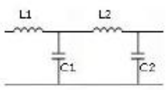
z(p)=n(p)/d(p)
n(p): 0.0000 0.7079 14.8437 14.8721
d(p): 1.0000 20.9701 21.7194 14.8721
h(p): –0.0336 –0.6812 –0.2312 0.0000
g(p): 0.0336 0.7288 1.2293 1.0000

C1: 1.4127F L1: 0.9981H C2: 0.0477F

Figure 7.22 Output of the second page of the amplifier report

N2

Type: Lumped
Degree: 3
Poles:
- **P1:** -4.1316
- **P2:** -0.3143 + i0.7717
- **P3:** -0.3143 - i0.7717



z(p)=n(p)/d(p)
n(p): 2.9937 14.2506 10.9204 13.6662 2.8686
d(p): 0.0000 1.0000 4.7602 3.2914 2.8686
h(p): 0.5218 2.3096 1.0737 1.8083 0.0000
g(p): 0.5218 2.6582 2.7332 2.9557 1.0000
Average Gain: 10.1018
Min Gain: 9.1262

L1: 2.9937H C1: 0.9373F L2: 1.7704H C2: 0.2101F

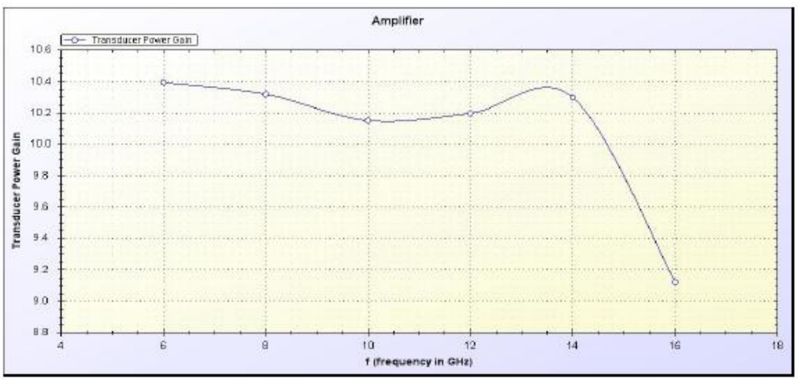


Figure 7.23 Output of the last page of the amplifier report

7.2.2 Single Stage Distributed Amplifier Example

In this example Z_G and Z_L and active device are same with the previous example.

Initial steps are similar with the previous example.

Step 1: Stage of the amplifier is set as one.

Step 2: The complexity of the backend and frontend equalizers is set. The degree of frontend network and backhand network is taken as 3. According to degree of networks, initial parameters are given as follows.

FRONTEND

$\lambda_0 = -15.3841$

$\lambda_1 = -0.5427 + j\,0.7518$

$\lambda_2 = -0.5427 - j\,0.7518$

Number of transmission line = 3

Commensurate delay=0.707

BACKEND

$\lambda_0 = -1.1748$

$\lambda_1 = -0.2813 + j\,0.8107$

$\lambda_2 = -0.2813 - j\,0.8107$

Number of transmission line = 3

Commensurate delay=0.707

Step 3: *Optimization*

When we input all required parameters, we run the optimization routine to make transducer power gain to converge desired level. In this case, as we did in distributed matching problem, we must change p $(p = \sigma + j\omega)$ with $(\lambda = \Sigma + j\omega)$ and we must use modified version of the f(p).

If we take the desired gain level 10 and perfect match, after optimization we get

FRONTEND

$$Z_2(\lambda) = \frac{14.8382\lambda^3 + 14.8985\lambda^2 + 11.7693\lambda + 3.8520}{\lambda^3 + 18.3987\lambda^2 + 14.6803\lambda + 3.8520}$$

$\lambda_0 = -17.5759$

$\lambda_1 = -0.4114 + j\,0.2234$

$\lambda_2 = -0.4114 - j\,0.2234$

BACKEND

$$Z_2(\lambda) = \frac{15.6868\lambda^3 + 27.3180\lambda^2 + 20.2731\lambda + 3.9607}{\lambda^3 + 7.7296\lambda^2 + 9.1684\lambda + 3.9607}$$

$\lambda_0 = -6.3922$

$\lambda_1 = -0.6687 + j\,0.4153$

$\lambda_2 = -0.6687 - j\,0.4153$

Step 4: *Synthesis*

After synthesis we get the following component result

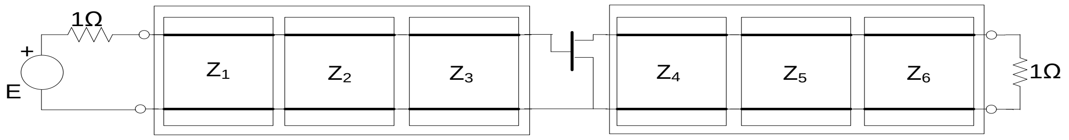


Figure 7.24 Single stage distributed amplifier circuit

Z_1=1.1958Ω, Z_2=3.2461 Ω, Z_3=10.4566 Ω, Z_4=3.0761 Ω, Z_5=10.5972 Ω, Z_6=13.6447 Ω

Transducer power gain of the system is depicted as follows.

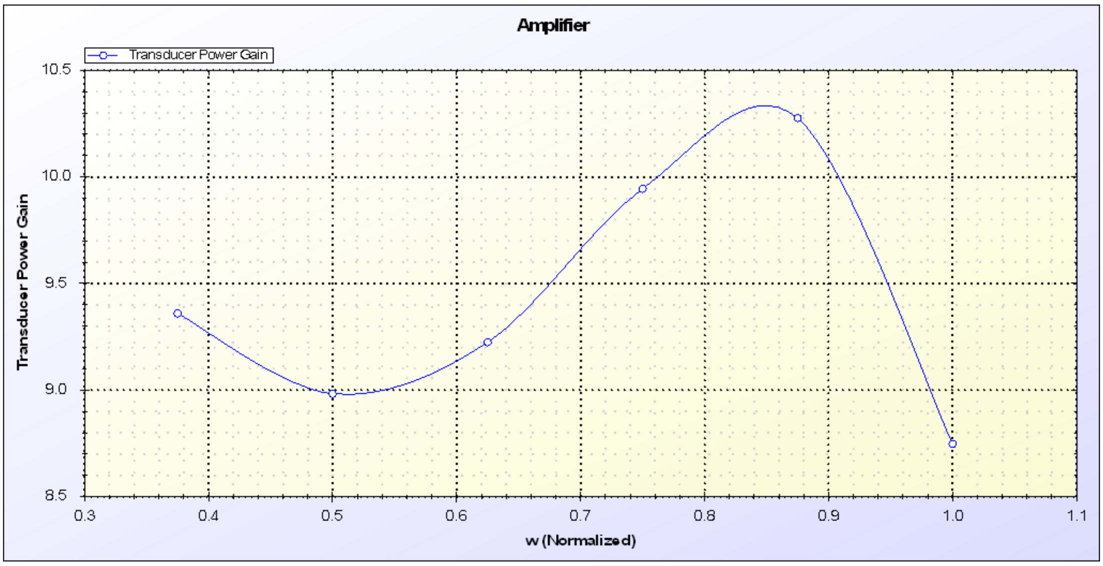


Figure 7.25 Transducer power gain of single stage amplifier network with optimization and perfect match

Average gain calculated as 8.8220 dB, and minimum gain in band width=8.7477dB.

Output of the report page is depicted as follows.

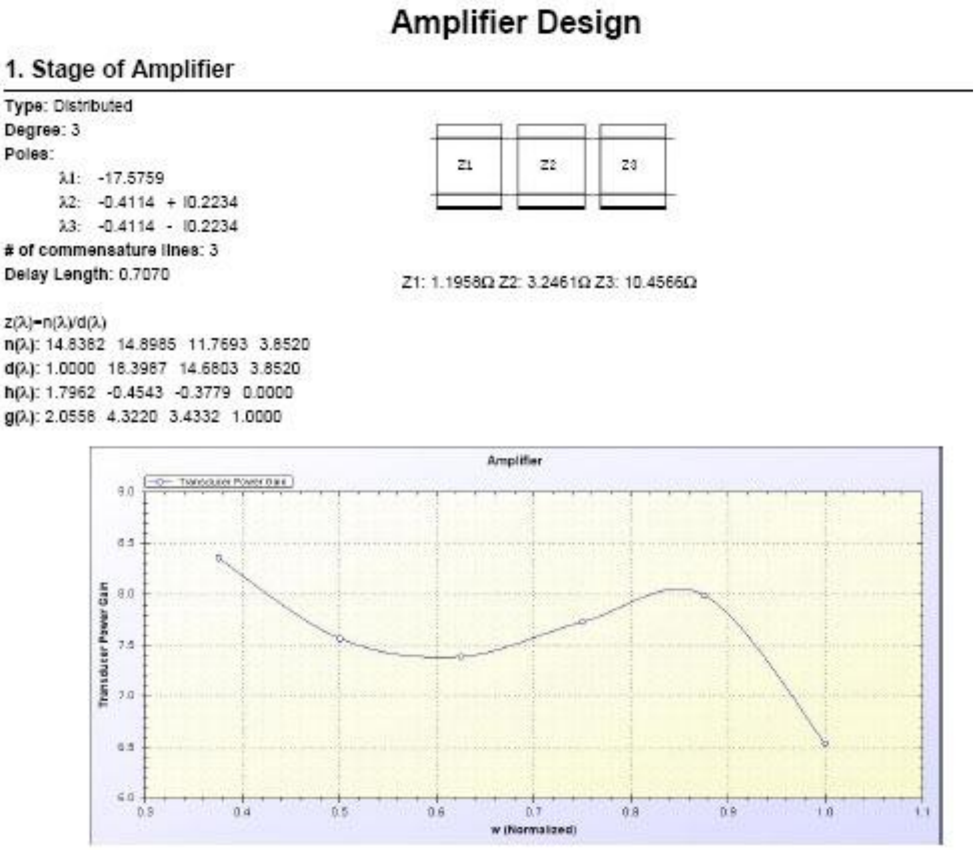


Figure 7.26 Output of the first page of the report page

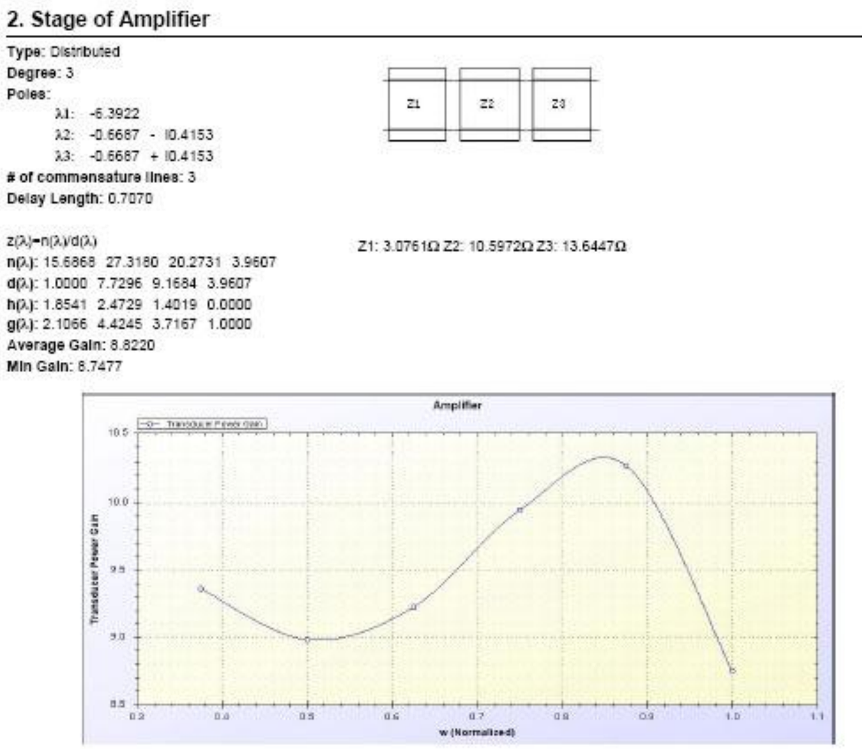


Figure 7.27 Output of the second page of the distributed amplifier design report page

7.2.3 Single-Stage Mixed Amplifier Example with Lumped Frontend and Distributed Backend Equalizers

In this example Z_G and Z_L and active device and initialization steps are same with the previous example.

Step 1: Stage of the amplifier is set as one.

Step 2: The complexity of the backend and frontend equalizers is set. The degree of frontend network and backhand network is taken as 3. According to degree of networks, initial parameters are given as follows.

FRONTEND	BACKEND
$p_0 = -15.3841$	$\lambda_0 = -1.1748$
$p_1 = -0.5427 + j\,0.7518$	$\lambda_1 = -0.2813 + j\,0.8107$
$p_2 = -0.5427 - j\,0.7518$	$\lambda_2 = -0.2813 - j\,0.8107$
	Number of transmission lines = 3
	Commensurate delay=0.707

Step 3: *Optimization*

When we input all required parameters, we run the optimization routine as we did in earlier examples. Everything is same, except, backend and frontend equalizers are in different types, one lumped and other distributed. When dealing with backend network we must change p $(p = \sigma + j\omega)$ with $(\lambda = \Sigma + j\omega)$ and we must use modified version of the f(p). If we take the desired gain level 10 and perfect match, after optimization we get

FRONTEND	BACKEND
$Z_2(p) = \frac{0.7128p^2+50.9996p+50.9384}{p^3+71.5441p^2+72.1702p+50.9384}$	$Z_2(\lambda) = \frac{7.2403\lambda^3+17.9193\lambda^2+12.8989\lambda+2.6908}{\lambda^3+4.5022\lambda^2+6.3612\lambda+2.6908}$
$p_0 = -70.5311$	$\lambda_0 = -0.7746$
$p_1 = -0.5065 + j\,0.6824$	$\lambda_1 = -1.8638 + j\,0.0032$
$p_2 = -0.5065 - j\,0.6824$	$\lambda_2 = -1.8638 - j\,0.0032$

Step 4: After synthesis of the components, we get lumped and distributed components as below.

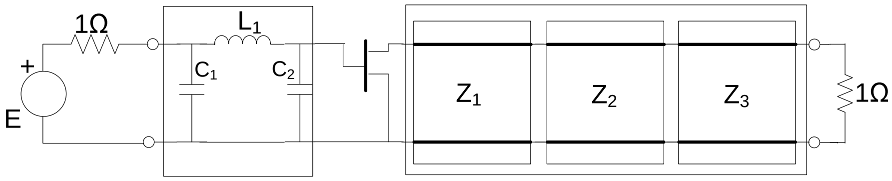


Figure 7.28 Single stage mixed amplifier network with lumped frontend, distributed backend equalizers

C_1=1.4028F, C_2=0.0140F, L_1=1.0012H, Z_1=2.7976Ω, Z_2=7.7092Ω, Z_3=7.4096Ω

Transducer power gain of the amplifier is depicted in the Figure 7.29

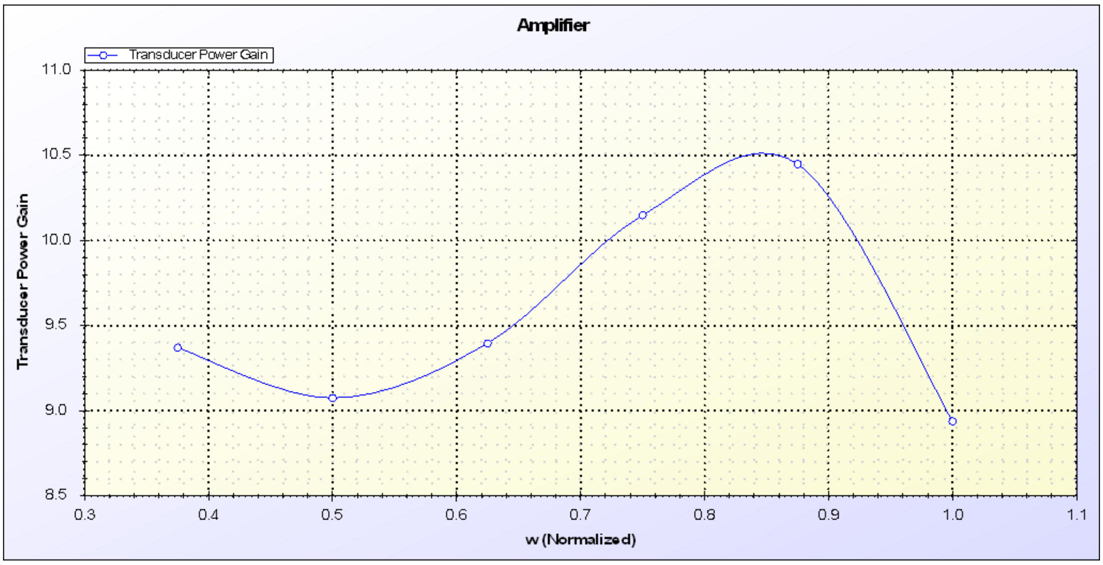


Figure 7.29 Single stage mixed amplifier network with lumped frontend, distributed backend equalizers

Average gain: 9.1168dB

Minimum gain: 8.9380dB

The output of the report page as follows.

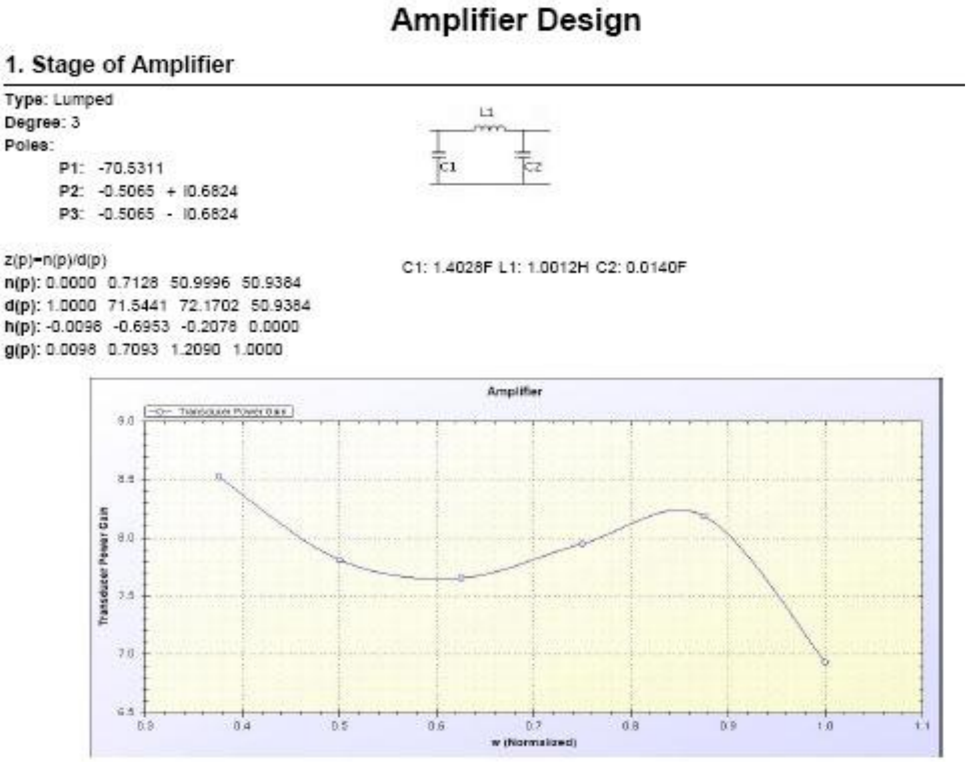


Figure 7.30 Output of the first page of the mixed amplifier design report

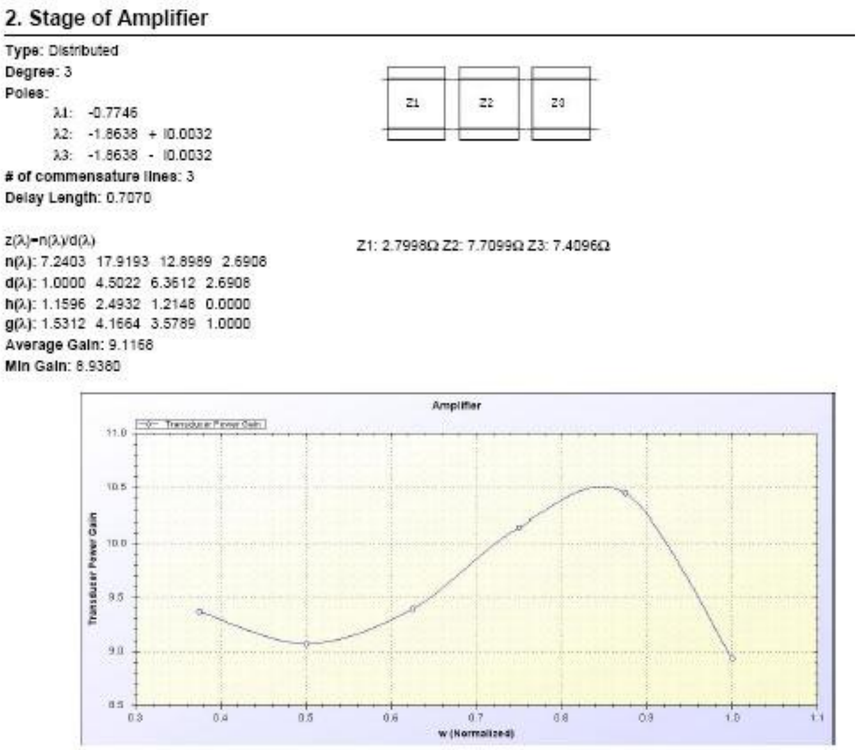


Figure 7.31 Output of the second page of the mixed amplifier design report

7.2.4 Single-Stage Mixed Amplifier Example with LPLUs

In this example, we try to use low pass ladders with unit elements during the design of backend and frontend equalizers. We mentioned how to design matching network with LPLUs before. So, it will not be explained in detail again. We took again Z_G=1Ω and Z_L=1Ω as generator and load impedances.

Step 1: The stage of the amplifier is set as 1.

Step 2: Degree of frontend and backend equalizers is set as 4 and 5, respectively. According to degree of the networks, initial parameters are given as follows.

FRONTEND

$$p_0 = -0.2813 + j\,0.8107$$

$$p_1 = -0.2813 - j\,0.8107$$

$$\lambda_0 = -0.5427 + j\,0.7518$$

$$\lambda_1 = -0.5427 - j\,0.7518$$

Commensurate delay=0.707

BACKEND

$$p_0 = -1.2813$$

$$p_1 = -0.2813 - j\,0.8107$$

$$p_2 = -0.2813 - j\,0.8107$$

$$\lambda_0 = -0.5427 + j\,0.7518$$

$$\lambda_1 = -0.5427 - j\,0.7518$$

Commensurate delay=0.707

Step 3: Overall transducer of the amplifier is calculated sequentially. In each iteration, LPLU networks are split into only lumped and only distributed sections. We try to converge to gain level 10 and under perfect matching we get the result as follows

FRONTEND

$$p_0 = -5.1544 + j\,12.0805$$

$$p_1 = -5.1544 - j\,12.0805$$

$$\lambda_0 = -0.4260 + j\,0.3905$$

$$\lambda_1 = -0.4260 - j\,0.3905$$

Commensurate delay=0.707

BACKEND

$$p_0 = -1.7369$$

$$p_1 = -0.0583 - j\,13.0926$$

$$p_2 = -0.0583 - j\,13.0926$$

$$\lambda_0 = -1.3360 + j\,0.0123$$

$$\lambda_1 = -1.3360 - j\,0.0123$$

Commensurate delay=0.707

$$Z_2(0,\lambda) = \frac{0.1116\lambda^2 + 0.6976\lambda + 0.3340}{\lambda^2 + 0.8521\lambda + 0.3340}$$

$$Z_2(p,0) = \frac{16.7338p + 172.5054}{p^2 + 10.3038p + 172.5054}$$

$$\Lambda_h = \begin{pmatrix} 0 & -0.2312 & -1.3300 \\ 0.0186 & -0.1242 & -0.0692 \\ -0.0029 & -0.0019 & 0 \end{pmatrix}$$

$$\Lambda_g = \begin{pmatrix} 1 & 2.3198 & 1.6640 \\ 0.0784 & 0.1861 & 0.0692 \\ 0.0029 & 0.0019 & 0 \end{pmatrix}$$

$$Z_2(0,\lambda) = \frac{3.1860\lambda^2 + 5.1809\lambda + 1.7849}{\lambda^2 + 2.6720\lambda + 1.7849}$$

$$Z_2(p,0) = \frac{27.0875p^2+50.2090p+297.7424}{p^3+1.85364p^2+171.6225p+297.7424}$$

$$\Lambda_h = \begin{pmatrix} 0 & 0.7028 & 0.6123 \\ -0.2039 & 0.7650 & -0.2276 \\ 0.0424 & -0.0563 & 0 \\ -0.0017 & 0 & 0 \end{pmatrix}$$

$$\Lambda_g = \begin{pmatrix} 1 & 2.1998 & 1.1726 \\ 0.3725 & 0.9628 & 0.2276 \\ 0.0486 & 0.0563 & 0 \\ 0.0017 & 0 & 0 \end{pmatrix}$$

Step 4: After synthesis, we get the following component values.

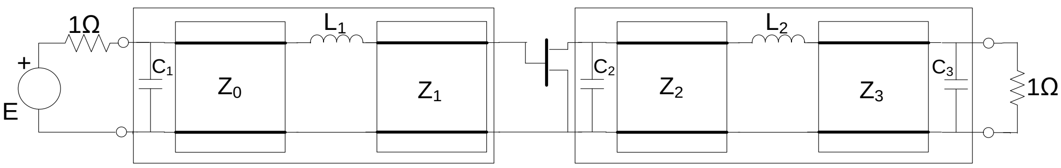


Figure 7.32 Single stage mixed amplifier network with LPLUs

C_1=0.0598F, C_2=0.374F, C_3=0.5397F, L_1=0.0970H, L_2=0.1689H, Z_0=0.5229Ω, Z_1=0.1747Ω, Z_2=1.8603Ω, Z_3=3.3203Ω

You can see the transducer power gain in Figure 7.33

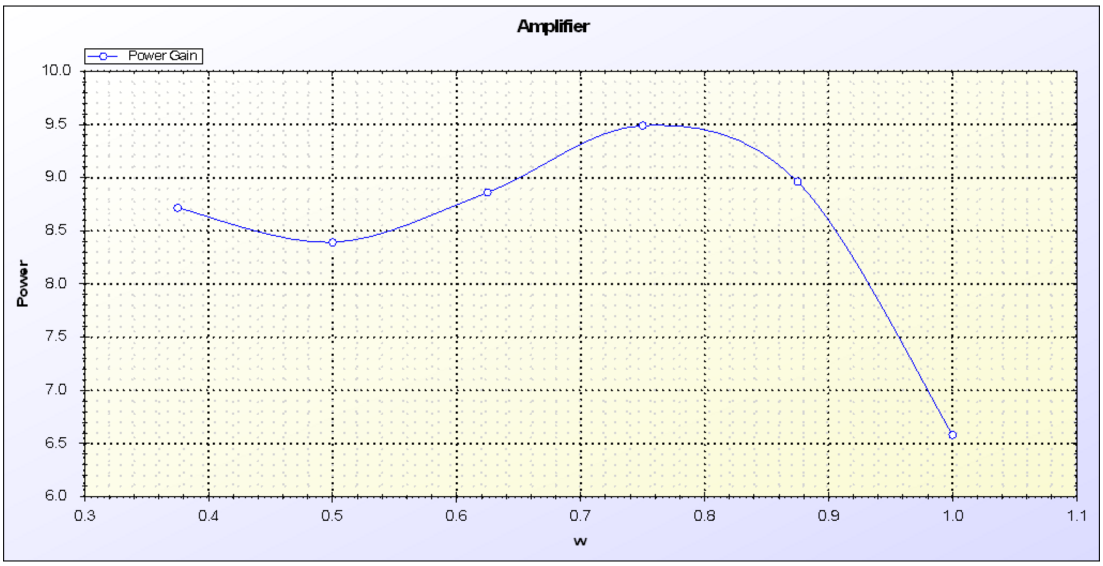


Figure 7.33 Transducer power gain of a single stage mixed amplifier network with LPLUs

7.2.5 Multistage Mixed Amplifier Example

In this example, we try to design a multi-stage amplifier with lumped or distributed equalizers. In this design, two active devices, HFET2001, were used as in the Figure 7.34. We took again Z_G=1Ω and Z_L=1Ω as generator and load impedances.

Step 1: The stage of the amplifier is set as 2.

Step 2: Degree of equalizers is set as 3 and type of the networks is set. According to degree of the networks, initial parameters are given as follows.

N_1

Type: Distributed
Degree: 3

$\lambda_0 = -1$
$\lambda_1 = -1 + j\,1$
$\lambda_2 = -1 - j\,1$

Number of transmission lines: 3
Commensurate delay=0.707

N_2

Type: Lumped
Degree: 3

$p_0 = -1$
$p_1 = -1 + j\,1$
$p_2 = -1 - j\,1$

N_3

Type: Lumped
Degree: 3

$p_0 = -1$
$p_1 = -1 + j\,1$
$p_2 = -1 - j\,1$

Step 3: Overall transducer of the amplifier is calculated sequentially as mentioned in previous sections. We try to converge to gain level 25. Under perfect matching we get the result as follows

N_1

$\lambda_0 = -0.3498$
$\lambda_1 = -1.3654 + j\,0.0260$
$\lambda_2 = -1.3654 - j\,0.0260$

Commensurate delay=0.707

$$Z_{2_1}(\lambda) = \frac{0.4258\lambda^3 + 1.2788\lambda^2 + 1.4614\lambda + 0.6525}{\lambda^3 + 3.0807\lambda^2 + 2.8205\lambda + 0.6525}$$

N_2

$p_0 = -21.7693$
$p_1 = -0.3852 + j\,1.2639$
$p_2 = -0.3852 - j\,1.2639$

$$Z_{2_2}(p) = \frac{2.2584p^2 + 50.9032p + 38.0056}{p^3 + 22.5396p^2 + 18.5148p + 38.0056}$$

N_3

$p_0 = -18.2562$
$p_1 = -0.4061 + j\,1.0377$
$p_2 = -0.4061 - j\,1.0377$

$$Z_{2_3}(p) = \frac{1.5234p^2 + 29.0488p + 22.6692}{p^3 + 19.0684p^2 + 16.0695p + 22.6692}$$

Step 4: After synthesis, we get the following component values.

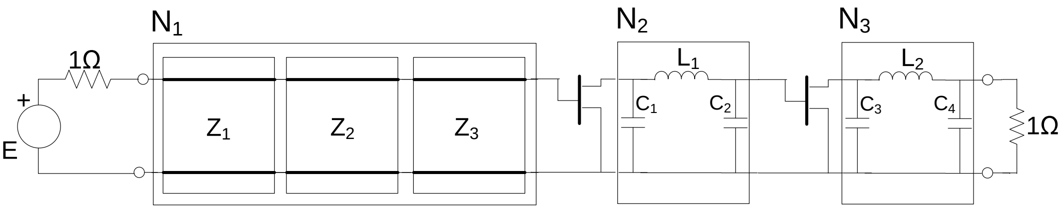


Figure 7.34 Multi stage mixed amplifier network

C_1=0.4428F, C_2=0.0444F, C_3=0.6564F, C_4=0.0524F, L_1=1.3394H, L_2=1.2814H, Z_1=0.5055Ω, Z_2=0.3376Ω, Z_3=0.4357Ω

You can see the transducer power gain of the overall network in the Figure 7.35

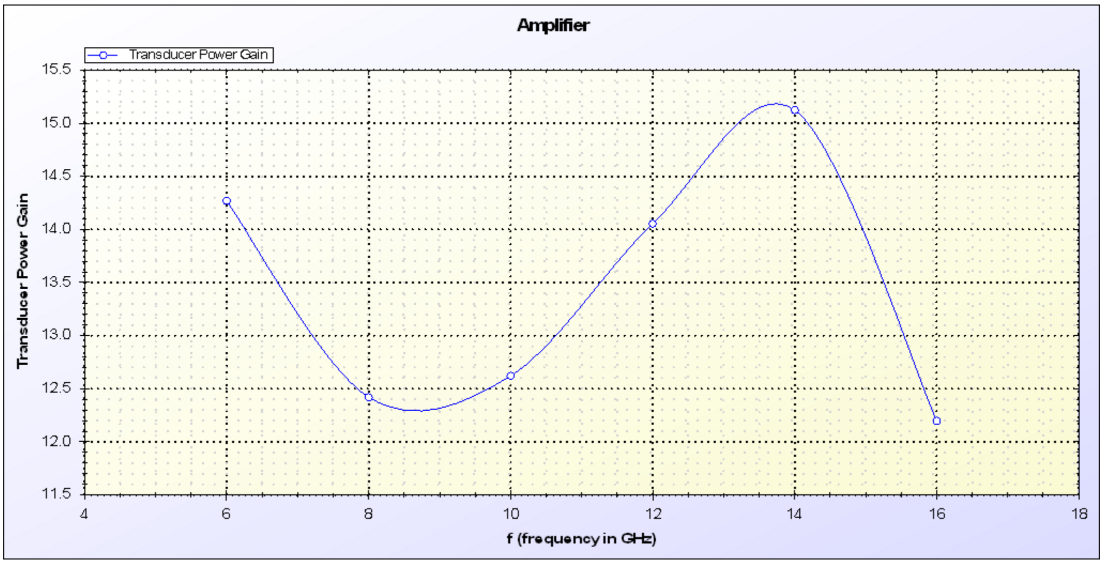
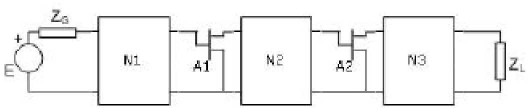


Figure 7.35 Transducer power gain of a multistage mixed amplifier network

The output of the report page depicted in the figures;

Amplifier Design

General Information about Amplifier Design



1. Transistor Parameters A1 (HFET2001)

f(GHz)	A11M	A11P	A12M	A12P	A21M	A21P	A22M	A22P
6	0.88	-65	0.05	60	2	125	0.71	-22
8	0.83	-85	0.06	53	1.81	109	0.68	-30
10	0.79	-101	0.06	51	1.64	95	0.66	-37
12	0.76	-113	0.06	52	1.48	84	0.66	-43
14	0.73	-126	0.06	54	1.39	73	0.64	-48
16	0.71	-141	0.07	55	1.32	61	0.63	-56

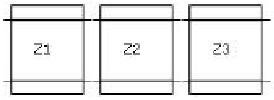
2. Transistor Parameters A2 (HFET2001)

f(GHz)	A11M	A11P	A12M	A12P	A21M	A21P	A22M	A22P
6	0.88	-65	0.05	60	2	125	0.71	-22
8	0.83	-85	0.06	53	1.81	109	0.68	-30
10	0.79	-101	0.06	51	1.64	95	0.66	-37
12	0.76	-113	0.06	52	1.48	84	0.66	-43
14	0.73	-126	0.06	54	1.39	73	0.64	-48
16	0.71	-141	0.07	55	1.32	61	0.63	-56

Figure 7.36 The first page of the report of multistage amplifier

N1

Type: Distributed
Degree: 3
Poles:
λ1: -0.3498
λ2: -1.3654 - i0.0260
λ3: -1.3654 + i0.0260
# of commensature lines: 3
Delay Length: 0.7070



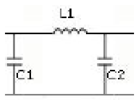
z(λ)=n(λ)/d(λ)
n(λ): 0.4258 1.2788 1.4614 0.6525
d(λ): 1.0000 3.0807 2.8205 0.6525
h(λ): –0.4400 –1.3808 –1.0415 0.0000
g(λ): 1.0925 3.3406 3.2811 1.0000

Z1: 0.5055Ω Z2: 0.3376Ω Z3: 0.4357Ω

Figure 7.37 The second page of the report of the multistage amplifier

N2

Type: Lumped
Degree: 3
Poles:
P1: -21.7693
P2: -0.3852 + i1.2639
P3: -0.3852 - i1.2639



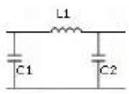
z(p)=n(p)/d(p)
n(p): 0.0000 2.2584 50.9032 38.0056
d(p): 1.0000 22.5396 18.5148 38.0056
h(p): –0.0132 –0.2668 0.4261 0.0000
g(p): 0.0132 0.3262 0.9133 1.0000

C1: 0.4426F L1: 1.3394H C2: 0.0444F

Figure 7.38 The third page of the report of the multistage amplifier

N3

Type: Lumped
Degree: 3
Poles:
P1: -18.2562
P2: -0.4061 + i1.0377
P3: -0.4061 - i1.0377



z(p)=n(p)/d(p)
n(p): 0.0000 1.5234 29.0488 22.6692
d(p): 1.0000 19.0684 16.0695 22.6692
h(p): –0.0221 –0.3870 0.2863 0.0000
g(p): 0.0221 0.4542 0.9951 1.0000
Average Gain: 13.5867
Min Gain: 12.1973

C1: 0.6564F L1: 1.2814H C2: 0.0524F

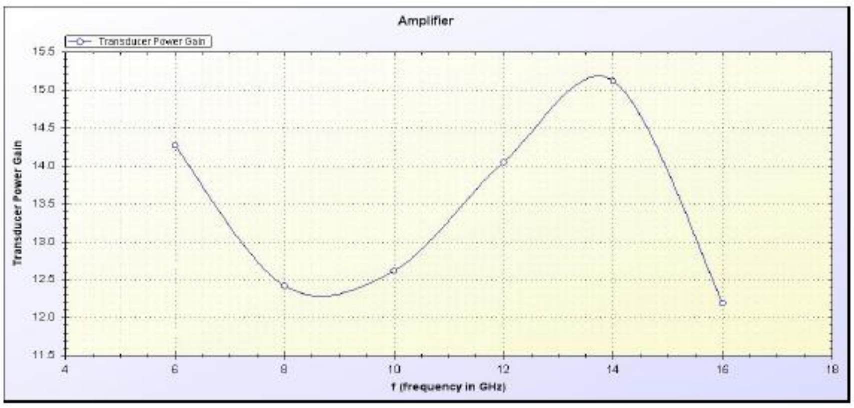


Figure 7.39 The last page of the report of the multistage amplifier

Chapter 8
Conclusion

In this work, real frequency design of broadband matching networks is studied. Among various real frequency technique algorithms, impedance based parametric approach and scattering based design technique for broadband matching are examined in detail. These approaches are combined to implement a standalone tool for the construction of matching networks with mixed lumped and distributed elements. Thus a generalized multivariable formulation for the mixed type of structures is obtained to end up with an integrated design tool for different matching network design problems.

In the course of proposed design tool for multivariable mixed element networks, once the complexity and location of the transmission zeros of the network is decided, we first split the network into only-lumped and only-distributed parts. Then by using parametric approach the impedance functions describing the lumped and distributed sub networks are obtained in an explicit manner in terms of the chosen unknown pole locations of the networks. Then, in terms of the impedance functions, we generate the canonic polynomials, $h(p,0), g(p,0)$ and $h(0,\lambda), g(0,\lambda)$ which fully describe the scattering functions of the network to be designed. These polynomials are then used as the first column and first row of the two-variable polynomials $h(p,\lambda)$ and $g(p,\lambda)$ that characterize the mixed network. Once the network characterization is obtained, transducer power gain of the system is optimized via a nonlinear search routine to end up with the required matching network impedance parameters. The resultant impedance functions are then synthesized to obtain the element values. In this approach, only the degrees and the poles of the lumped and distributed sections are assumed as the descriptive variable parameters. In the developed design tool, no attention is paid to the initialization of these parameters, where add-hoc initialization is assumed. The initial value generation is left to the practical engineering know-how. However, it has been exercised by several examples that, with an intelligent use of the design tool, the designer can come up with reasonable design solutions after a number of sequential design processes.

Along with the design of matching networks, the amplifier design problem which inherently involves the front-end, back-end and inter-stage matching networks is also studied and the application of the tool to the design of multistage amplifiers is implemented.

The developed design tool is capable of solving lumped, distributed, mixed matching and amplifier design problems of low-pass, high-pass or band-pass type. The user interface of the program is self explanatory and informs the user about the load, generator data input and all required parameter initialization requirements. For amplifier design the selection of the transistor type from a database or the entry of the numerical transistor scattering data is also included. Although the stability issues of the transistors are discussed in the text, the program ensures realizable designs for only stable gain block case. Stability features of the transistors require further elaboration, but can easily be incorporated to the design software.

In the design tool developed, modules that are responsible of managing numerical calculations were written in MATLAB. Because of the huge choice of graphical user interface components and flexibility in implementing and design of software, we preferred to integrate MATLAB modules with the .NET environment.

The integrated multivariable tool is a complete design tool which can analyze a given matching or amplifier network, try to make it more efficient by using optimization and prepare a design report in pdf form, that informs user about the transducer power gain, characterization of equalizer networks, circuit topology and component values. As far as the network topologies are concerned a variety of choices are supported; lumped networks, distributed networks, mixed lumped-distributed networks, active matching problems for multistage amplifiers with lumped, distributed or mixed type equalizers.

As far as the performance of the developed code is concerned, it might be faster if some basic numerical calculations run in .NET layer instead of MATLAB layer, however, because of the cost of implementing an optimization routine and complex numbers library in .NET, we preferred to use of the self modules which were already written in MATLAB. For further improvement of the tool, user interface of the program may also be enhanced by the means of human computer interaction during later developments.

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Curriculum Vitae

Hayri Şimşek was born on 20 July 1981, in İstanbul. He received his BS degree in Electronics Engineering and Computer Science and Engineering in 2005 and M.S. degree in 2009 in Electronics Engineering both from Işık University. He worked as a research assistant at the Department of Electronics Engineering of Işık University from 2006 to 2009. His research interest is designing microwave matching and amplifier networks.