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MODERN APPROACHES TO BROADBAND MATCHING PROBLEMS

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MODERN APPROACHES TO BROADBAND MATCHING PROBLEMS†

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ABSTRACT

The purpose of this paper is to review from a practical viewpoint analytic gain-bandwidth theory and commercially available CAD techniques as applied to matching problems. Then, modern approaches, so-called real frequency techniques, to gain bandwidth problems are summarized to provide necessary back ground to the reader for easy access to the relevent literature.

† Part of this paper is a summary of the talk presented at the Workshop on Broadband Matching and Design of Microwave Amplifiers, sponsored by the 1983 IEEE International Microwave Symposium, May 30, Boston.

I. INTRODUCTION

During the Workshop on broadband matching and design of microwave amplifiers, sponsored by the 1983 IEEE Microwave Symposium in Boston, U.S.A. [1], importance of the new approaches to broadband matching problems was affirmed by the speakers and the audiance. This renewed interest in matching problems stems from recent developments in solid-state device technology, which in turn activated research and development efforts in both commercial and military communication systems. The classical analytical approaches to broadband matching problems had limited applicability to the more demanding and complicated matching problems which were outcomes of modern technology. Particularly in commercial satellite and wideband military communication systems, need for new broadband matching techniques become apparent.

In order to meet the needs of the industry, several computer packages had already been developed to handle matching and amplifer design problems. These computer-aided design (CAD) packages, with both analysis and optimization capabilities, have been quite useful to microwave engineers. On the other hand, they are all based upon brute force methods, and as such suffer from certain deficiencies in handling complicated microwave design problems.

In the workshop mentioned above, analytic methods, commercially available computer packages, device modelling techniques, and new procedures for designing broadband matching networks and microwave amplifiers were discussed in detail. The so-called real frequency techniques for broadbanding were qualitatively shown to offer superior design performance over other available techniques.

In this review paper, practical aspects of analytic methods and commercially available CAD techniques are discussed first. Then, in the words of Dr.Levy, "Modern Synthesis" approaches to brocdband matching problems are scanned, with references to the relevent literature. It should be noted that details of the new techniques or application examples are not given in this paper. Our wish is to collect comparative theoretical and practical examples in a different paper.

II. MATCHING PROBLEMS

The classical matching problem is one of constructing a lossless matching network or equalizer (E) between a generator and a load such that the transfer of power is maximized over a prescribed frequency band. Performance of the entire system is usually measured with the Transducer Power Gain (TPG), which is defined as the ratio of the power delivered to load (P_L) to the available power from the generator (P_A) (Fig.1). We study matching problems under the following headings:

a) Single Matching: In this classical matching problem, the generator network is given as a pure resistance in series with an ideal voltage source, whereas the load is arbitrary (Fig.1a).

b) Double Matching: This is the power transfer problem from a complex generator to a complex load (Fig.1b).

c) Active Two-Port problem: In this case, the input and output of an active device is to be simultaneously matched to a generator and a load (Fig. 1c).

Thus, one may be involved with single and/or double matchings. In addition, matching problems of this sort may be complicated by other desing considerations such as obtaining maximum power delivery from the active device, good noise figure, and efficiency from the overall system, etc. A typical example of this kind is the design of a microwave amplifier.

Analytic theory of single and double matching problems are extensively studied [2] - [8] in the literature. Analytic theory is essential in understanding the gain-bandwidth limitations of the given loads to be matched. Explicit formulas are available for certain simple cases [9] - [13]. On the other hand, application of the analytic theory is limited to simple problems. By simple, we mean those problems of single or double matching in which the generator and load networks include at most one reactive element, namely either a capacitor or an inductor (Fig.2). For simple problems, low-pass equi-ripple prototype designs yield reasonable solutions given a fixed complexity for the matching networks. On the other hand, if the number of reactive elements increases on the load side, either the theory becomes inaccesible, or if it is still capable of handling the problem, the resulting gain performances turn out to be sub-optimal, equalizer structures become unnecessarily complicated, and sometimes, even altogether unrealizable [14], [15].

Commercially available computer programs such as Super Compact [16], Cosmic [17], ANA [18], etc. are very good in analyzing and optimizing given circuit structures, but they do not include network sythesis procedures in a literal sense. In other words, in designing a matching network or amplifier, a topology for the matching network should be supplied to a commercially available package. Furthermore, subject to the optimization goals, element values of the chosen topology should be sufficiently close to the final result. Thus, these CAD packages work as fine trimming tools on the element values when the circuit is synthesized in advance. If the problem is well known, that is, if the design engineer knows the optimum topology, with element values close enough to the final design, then commercial programs yield reasonable solutions. Especially, if the frequency band is narrow (e.g. less than one octave), the choice of circuit topology for the lossless matching network and initialization of the element values are not difficult tasks. Usually, a simple two-element low-pass L-C Ladder is a practical solution. However, if the optimum topology is unknown or the problem calls for substantial bandwidth, the task of designing matching networks or microwave amplifiers becomes difficult, and it is not easy to use the above-mentioned CAD packages because of the following reasons:

-For highly complicated broadband single or double matching problems, the choice of circuit topology is unclear. Even if the topology is known, with three or more elements, it is not straight forward to come up with approximate element values to initialize the nonlinear optimization problem.

-The TPG is a highly nonlinear function of the unknown element values of the circuit topology. Therefore, it is not easy to know whether local or global maximum is reached in the search algorithm.

Let us now briefly investigate the degree of the non-linearity of TPG in terms of the element values of the chosen topology. For this purpose, let us take a look at a lowpass L-C filter problem for which we know that equiripple functions yield optimum solutions for a fixed number of elements. For this problem, the TPG can be explicitly written in terms of the element values of the Ladder.

We describe the degree of nonlinearity of the TPG in terms of the order of multiplications of the unknowns (i.e. the elements of the ladder network), and we let 0 denote the Least Degree of Nonlinearity. Then the degree of nonlinearity $D\{\cdot\}$ of the inverse gain function. $(1/T(\omega))$ may be described by

$$D\{\frac{1}{T(\omega)}\} \sim 0^{2n} + 0^{2(n-1)} + \ldots + 0^2 + 0 \qquad (1)$$

As can be seen from the above relation, maximization of the TPG or minimization of $1/T(\omega)$ constitutes a highly nonlinear optimization problem in terms of the unknown element values. If the problem is one of single or double matching, then the optimization of $T(\omega)$ becomes much more difficult.

If the degree of the nonlinearity $D\{\frac{1}{T(\omega)}\}$ is transformed to $\sim[0^2 + 0]$ by defining a different set of unknown parameters, then the numerical problem would become one of quadratic optimization, the solution of which is well established. The real frequency techniques reviewed in this article have successfully achieved this goal.

Before we proceed with real frequency techniques, let us briefly illustrate the difficulties of both analytic gain-bandwidth theory and commercially available CAD techniques with the help of the following example. In Fig.3a, measured impedance data for a high-Q monopole antenna is depicted. Here, the problem is to design a lossless matching network over more than 3 octave bandwidth to transfer maximum power from a resistive generator to the antenna over more than 3-octave bandwidth. Thus, the problem is one of broadband single matching. First of all the following questions should be asked.

- What is the best performance that can be achieved? In other words, we wish to know the gain-bandwidth limitations of the load (highest flat gain level which is physically realizable).

- What should be the practical equalizer structure to reach the best performance?

In applying analytic gain-bandwidth theory, the following path is followed:

a) Measured data is modelled in as simple a manner as possible to be handled by the theory. Fig.3b shows a possible model which fits the measured data.

b) A proper transfer function is selected with unknown parameters in which the load model is embedded. But, to the best our knowledge, there is no transfer function available in the literature which can readily include the load shown in Fig.3b. Therefore, theory is inaccessible in this case.

c) Unknown parameters of the transfer function is computed to optimize the gain and satisfy the complicated gain-bandwidth equations. Then, the transfer function is synthesized, yielding the desired matching network. In the present problem, however, we are not able to follow these steps.

An attempt can be made to solve it by employing commercially available CAD packages. In this case, the major difficulty is to come up with the optimum circuit topology. Even if the circuit topology is known, it is not clear how to initialize the highly nonlinear optimization algorithm.

As can be seen from the above discussion, conventional techniques are not capable of handling this problem. However, the modern approaches reviewed in this paper can readily be employed to solve this kind of single and double matching problem.

III. NEW APPROACHES TO GAIN BANDWIDTH PROBLEMS

III.1. Line Segment Technique

In 1977 a new numerical approach known as the "real frequency" technique was introduced by Carlin for the solution of single matching problems [18]. The real frequency technique utilizes measured data, by passing analytic theory. Neither the equalizer topology nor the analytic form of a transfer function are assummed. They are the result of the proposed design method. Measured data obtained from the devices to be matched is directly processed. The heart of Carlin's approach resides in the generation of the positive real (PR) input impedance $Z_q = R_q(\omega) + jX_q(\omega)$ looking into a lossless matching network with resistive termination. Let the measured load impedance be $Z_L(j\omega) = R_L(\omega)+jX_L(\omega)$; then the transducer gain $T(\omega)$ is given by

$$T(\omega) = 4 \frac{R_q(\omega)\ R_L(\omega)}{[R_q(\omega) + R_L(\omega)]^2 + [X_q(\omega) + X_L(\omega)]^2} \quad . \tag{2}$$

Once the impedance of the device $Z_L(j\omega)$ is specified, $T(\omega)$ depends only on $R_q(\omega)$ and $X_q(\omega)$. Close examination of Eq(2) reveals that the unknown functions $R_q(\omega)$ and $X_q(\omega)$ are related to $\frac{1}{T(\omega)}$ in a quadratic manner. If we assume that the unknown impedance $Z_q(j\omega)$ is a minimum-reactance function, then $X_q(\omega)$ can be uniquely determined from $R_q(\omega)$ by a Hilbert transformation. In Carlin's approach, the matching problem is solved in two steps. First, $R_q(\omega)$ is described as a set of linear combinations of unknown line segments and, assumming $Z_q(j\omega)$ is minimum reactance, $X_q(\omega)$ in also expressed in terms of the same unknown line segments (Fig.4)†. These straight lines are then computed in such a way that TPG is optimized over the band of operation. The second step is to approximate $Z_q(j\omega)$ by a realizable rational function which fits the computed data (R_q, X_q), Finally, Z_q is synthesized, using Darlington's procedure, as a lossless two-part with resistive termination [19].

In most practical cases, first we choose an analytic form for $R_q(\omega)$ which corresponds to a ladder network. That is,

$$R_q(\omega) = \frac{A_K\ \omega^{2k}}{B_o + B_1\omega^2 + \ldots + B_n\omega^{2n}} \tag{3}$$

where k and n are positive integers ($k \leq n$) fixing the complexity of the equalizer. The coefficients A_k and B_i ($i = 0,1,\ldots,n$) are computed to fit the real part data, obtained from the first step of the technique, by linear regression. After wards, $Z_q(s)$ ($s = \sigma + j\omega$) is generated from (3) using the Bode or Gewertz procedure [20]. It has been shown that the line segment technique yields superior design performances over the analytic and other CAD approaches [14], [21].

$\dagger$ In a similar manner, $Y_q = \frac{1}{Z_q} = G_q + jB_q$ can be employed as a minimum-susceptance in put admittance and B_q computed from G_q by means of a Hilbert transformation. In this case design procedure remains unchanged.

Almost optimum circuit topology is resolved. Furthermore, gain-bandwidth limitations of a given load may be determined by means of computer experiments [15].

On the other hand, the following may be regarded as disadvantages of this technique.

The design procedure involves two basic computation steps (line segment computations and approximation problem with the evaluation of Hilbert transformations), which may be Laborious and expensive. Even though we no longer have to choose a circuit topology, decisions have still to be made whether the input immitance (Z_q or Y_q) is to be a minimum-reactance or a minimum-susceptance function. If we do not wish to restrict the design, some reactive elements can be extracted from the equalizer leaving a minimum-reactance or minimum-susceptance input immitance. Althoush this process improves the flexibility of the technique, one has to decide what to extract (capacitance or inductance) and how to extract (series or parallel) by trial and error, which in turn increases the computation time.

Despite the said drawbacks of this technique, it is reasonably satisfactory for single matching problems. Later, the line segment technique was extended to handle double matching problems as well, but the computational efficiency of the technique turned out to be poor [22]. A recently proposed double matching technique, the so-called "direct computational technique" overcomes some of the difficulties of the line segment techniqu [8].

III.2. Direct Computational Technique

The basic idea employed in the direct computation method is similar to that of the line segment technique. That is, the lossless matelning network is described by its driving point inpedance Z_q with a resistive termination (Fig.5). The core of the new method resides in an expression for the overall TPG of the doubly terminated system in terms of the unknown even part $R_q(-s^2)$ of the input impedance Z_q

$$Z_q = \frac{N(s)}{D(s)} = \frac{a_0 + a_1 s + \dots + a_m S^m}{1 + b_1 s + \dots + b_n s^n} = R_q(-s^2) + \text{Odd}Z_q(s) \qquad (4)$$

where Odd $Z_q(s)$ is the odd part of $Z_q(s)$. In this procedure, Z_q is of course assumed to be a minimum-reactance function. The real part $R_q(\omega^2)$ is specified by Eq(4) and TPG is generated by explicit factorization of $R_q(-s^2)$:

$$R_q(-s^2) = \frac{A(s)\ A(-s)}{D(s)\ D(-s)} = F(s)\ F(-s) \tag{5}$$

The terms A(s) A(-s) and D(s) D(-s) denote spectral factorizations of the numerator and denominator polynomials of $R_q(-s^2)$, and F(s) is defined as

$$F(s) \triangleq \frac{A(-s)}{D(s)} \quad . \tag{6}$$

On the jω axis,

$$F(j\omega) = \sqrt{R_q(\omega)}\ e^{j\phi_F(\omega)} \quad .$$

Then the TPG is expressed in terms of the generator impedance Z_G, load impedance Z_L, and the unknown real part $R_q(\omega^2)$ of Z_q

$$T(\omega) = \frac{(1-|g_{22}|^2)(1-|S_{in}|^2)}{|1-S_{in}\ g_{22}|^2} \tag{7}$$

where

$$S_{in}(j\omega) = e^{2j\phi_F(\omega)}\ \left(\frac{Z_L - Z_q(-j\omega)}{Z_L + Z_q}\right) \tag{8}$$

and

$$g_{22} = \frac{Z_{G-1}}{Z_{G+1}}\ , \quad 2q = R_q + jX_q, \quad X_q = H(R_q)$$

In (8), Z_q can be easily generated from $R_q(\omega^2)$ by employing the Gewertz procedure. For most practical cases, it is useful to choose $R_q(\omega^2)$ as in (3) to construct ladder networks.

Once the TPG is generated, it is maximized to determine the unknown coefficients A_k and B_i of $R_q(\omega^2)$ (see Eq(3)). In this design method, the line segment approach is simply omitted. Thus the computational efficiency is improved. This new double matching technique has all the merits of the line segment technique. However, decisions have again to be made whether to make the input impedance Z_q minimum reactance or minimum susceptance, etc. For interested readers, we leave the details of this double matching technique to the references [8] and [22].

III.3. Simplified Real Frequency Technique, "A Scattering Approach"

The simplified real frequency technique (SRFT) is also a CAD procedure for double matching problems. It possesses all the out standing merits of the line segment and direct computational techniques. Moreover, it does not require any a priori decisions, and it eliminates the need for numerical evaluation of the Hilbert transformations. The numerical set-up problem is well-defined for the optimization of TPG. Hence, it is faster than the other existing CAD techniques [21], [23], [24], [25] and easier to use. It is also naturally suited to design broadband microwave amplifiers. Employing SRFT, several matching networks and amplifiers have been designed and built; laboratory performance measurements exhibit good agreement with theoretical computations [21], [23], [25].

The basis for the new scattering approach to be described in the following is to deal with both the optimization of the system transducer gain between complex generator and load and the realization of the equalizer directly in terms of the unit normalized reflection factor of the equalizer alone, $e_{11}(s)$ (Fig.6). If $e_{11}(s)$ is appropriately determined, then the equalizer E may be synthesized using the Darlington theorem which states that any bounded reall (BR) reflection coefficient $e_{11}(s)$ is realizable as a lossless reciprocal two-port terminated in a pure resistance [26]. For simplicity, E is assummed to be a minimum phase structure with transmission zeros only at $\omega=\infty$, $\omega=0$. This is a convenient assumption since it assures realization without coupled coils, except possibly for an impedance level transformer. The algorithm of SRFT to be described is generally applicable to all matching problems as it involves neither equalizer element values nor equalizer topology.

Suppose $e_{11}(s)$ is given as

$$e_{11}(s) \triangleq \frac{h(s)}{g(s)} = \frac{h_0 + h_1 s + \dots + h_n s^n}{g_0 + g_1 s + \dots + g_n s^n} \qquad (9)$$

where n specifies the number of reactive elements in E. Then, employing the well known Belevitch representation [27] the real normalized scattering parameters of E are given as

$$e_{11}(s) = \frac{h(s)}{g(s)}$$

$$e_{12}(s) = e_{21}(s) = \pm \frac{s^k}{g(s)} \tag{10}$$

$$e_{22}(s) = -(-1)^k \frac{h(-s)}{g(s)}$$

where $k \geq 0$ is an integer and specifies the order of the transmission zeros. Since the matching network is lossless, it follows that

$$g(s)\ g(-s) = h(s)\ h(-s) + (-1)^k\ s^{2k} \tag{11}$$

In the iterative approach presented below, the coefficients of the numerator polynomial h(s) are chosen as unknowns. In order to construct the scattering parameters of E, it is sufficient to generate the Hurwitz denominator polynomial g(s) from h(s). It can be readily shown that once the coefficients of h(s) are initialized at the start of the optimization process and the complexity of the equalizer E is specified (i.e. n and k are fixed), g(s) can be generated as a Hurwitz polynomial by explicit factorization of (11). Thus, the physical realizibility of the scattering parameters $[\{e_{ij}\},\ i,j = 1,2]$ is already built into the procedure. It should be noted that in choosing the polynomial h(s) and integer k we can not allow $h(0) = 0$ and $k \neq 0$ simultaneously, since this violates the losslessness criterion of (11). The transducer power gain of the dorbly terminated structure of Fig.6 may now be constructed in terms of the e_{ij} and the given complex terminations as follows:

$$T(\omega) = T_g \frac{|e_{21}|^2 |\ell_{21}|^2}{|1-e_{11}S_G|^2 |1-\hat{e}_{22}S_L|^2} \tag{12}$$

where

$$T_g = 1 - |S_G|^2, \quad |L_{21}|^2 = 1 - |S_L|^2 \quad ,$$

$$\hat{e}_{22} = e_{22} + \frac{e_{21}^2 S_G}{1 - e_{11} S_G}$$

and

$$S_G = \frac{1 - Z_G}{1 + Z_G} \quad \text{and} \quad S_L = \frac{1 - Z_L}{1 + Z_L}$$

are the reflection coefficients of the generator and the load networks, respectively. It should be noted that in generating the Hurwitz denominator polynomial g(s) from the initialized coefficients of h(s), one first computers g(s) g(-s) as in (11):

$$g(s)\ g(-s) = G_0 + G_1\ S^2 + \ldots + G_n\ S^{2n} \tag{13}$$

where G_i's are given as follows:

$$\begin{aligned}
G_0 &= h_0^2 \\
G_1 &= -h_1^2 + 2h_2h_0 \\
&\vdots \\
G_i &= (-1)^i\ h_i^2 + 2(h_{2i}h_0 + \sum_{j=2}^{i} (-1)^{j-1} h_{j-1}\ h_{2j-i+1}) \\
&\vdots \\
G_k &= G_i \Big|_{i=k} + (-1)^k \\
&\vdots \\
G_n &= (-1)^n\ h_n^2 \quad .
\end{aligned} \tag{14}$$

Explicit factorization (13) then follows and the polynomial g(s) is formed from the left half plane zeros of g(s) g(-s). Once g(s) is formed, the scattering parameters of E are generated as in (10) and T(ω) constructed from (12). Once T(ω) is computed, the unknown coefficients h_i are determined by means of an optimization routine. Details of the numerical work can be found in [23] and [24]. It has been qualitatively shown that the least-square optimization technique provides satisfactory results [23], [24], [28]. However, any nonlinear optimization technique may be employed with the design method. It was also shown that the linear programming techniques of Dantzing [29] also provide very good results when used with the real frequency techniques [15].

The convergence behaviour of the simplified real frequency technique is excellent. Examination of (12) together with (10) indicates that TPG is almost inverse quadratic in the unknown coefficients h_i. To be more specific, $T(\omega)$ can be put in the following form

$$T(\omega) = \frac{W(\omega)}{h(j\omega)\ h(-j\omega) + \omega^{2k}} \tag{15}$$

where the term $W(\omega)$ can be regarded as a weighting function and is given by

$$W(\omega) = \frac{T_g |L_{21}|^2}{|1-e_{11}S_G|^2 |1-\hat{e}_{22}S_L|^2} \tag{16}$$

Despite the presence of the terms e_{11} and $\hat{e}_{22}$ in $W(\omega)$, optimization of the gain is heavily dominated by the quadratic term $\{h(j\omega)\ h(-j\omega) + \omega^{2k}\}$ in (15). It should be noticed that as $T(\omega)$ is maximized, $|e_{21}|^2$ is also maximized, which in turn reduces the amplitudes of e_{11} and e_{22} due to the losslessness of the equalizer E. Thus, in (15), $W(\omega)$ tends to depend only on the generator and load terms $T_g = 1 - |S_G|^2$ and $|\ell_{21}|^2 = 1 - |S_L|^2$, respectively. Therefore, the degree of the nonlinearity of $D\{\frac{1}{T(\omega)}\}$ tends to becomes quadratic. It is also worth mentioning that the numerical stability of the computer algorithm written for SRFT discussed above is excellent, since all the scattering parameters e_{ij} and reflection coefficients S_G and S_L are bounded real, i.e. $\{|e_{ij}|, |S_G|, |S_L|\} \leq 1$.

As is usually the case, an intelligent initial guess is important in efficiently running the program. Detailed initialization techniques may be found in the literative given in this paper. However, it has been experienced that for many practical problems, an ad hoc direct choice for the coefficients h_i (e.g. $h_i = 1$ or -1) provides satisfactory initialization to start the simplified real frequency technique algorithm.

IV. APPLICATIONS OF THE REAL FREQUENCY TECHNIQUES

So far, all of the real frequency techniques discussed in this paper have been applied to many theoretical and practical cases. Examples are scattered in the literature [8] [14] [15] [18] [21] [25].

Extension of these techniques to the design of broadband multistage microwave amplifiers have been carried out successfully [8], [23], [25], [31]. The new methods presented throughout this paper all result in lumped elements in the matching circuits. However, we can easily proceed directly to the realization of the equalizers made from transmission lines by using Richards' transformation for the frequency variable

$$\Omega = \tan \omega\tau \quad ,$$

where τ is the delay length of the commensurate transmission line [32]. All the approximations are now calculated in the transformed Ω domain, but otherwise the real frequency procedure remains basically unchanged [30]. An extended version of the SRFT with transmission line elements is given in [32].

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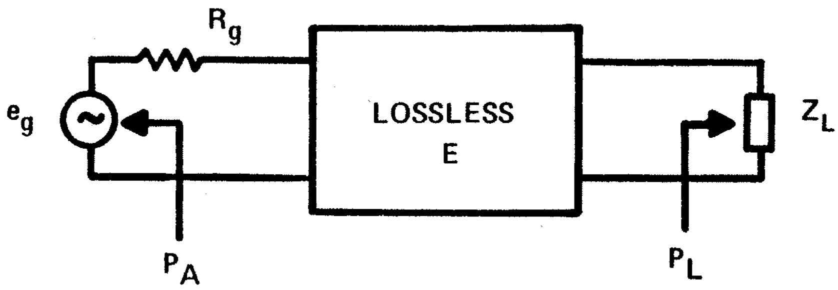
FIGURE CAPTIONS

Figure 1	Classical Broadband Matching Problem
1a	Single Matching
1b	Double Matching
1c	Active 2-port Problem
Figure 2	Analytic Theory and Simple Problems
Figure 3	Implementain of the analytic gain-band width theory
3a	Measured impedance data for a monopole antenna
3b	A possible model for the data shown in Fig.3a.
Figure 4	Real frequency technique for Single Matching Problems
Figure 5	Direct Computational Technique for double matching problems
Figure 6	Simplified Real Frequency Technique, "A scattering approach".

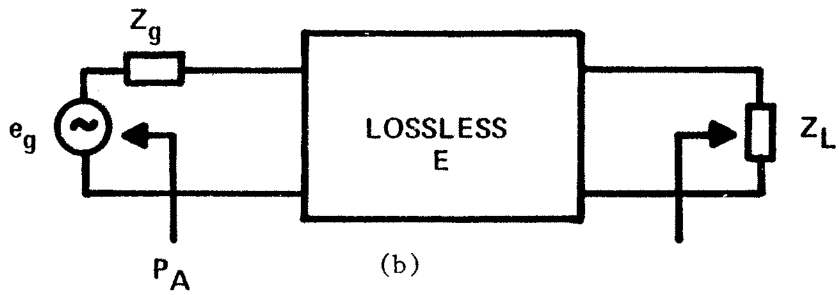
CLASSICAL BROADBAND MATCHING PROBLEMS

- SINGLE MATCHING



(a)

- DOUBLE MATCHING



(b)

GOAL: MAXIMIZE POWER TRANSFER FROM GENERATOR TO LOAD

$$T(\omega) = \frac{PL}{PA}$$

IS OPTIMIZED OVER PASSBAND

FIGURE-1.

ACTIVE TWO-PORT PROBLEM

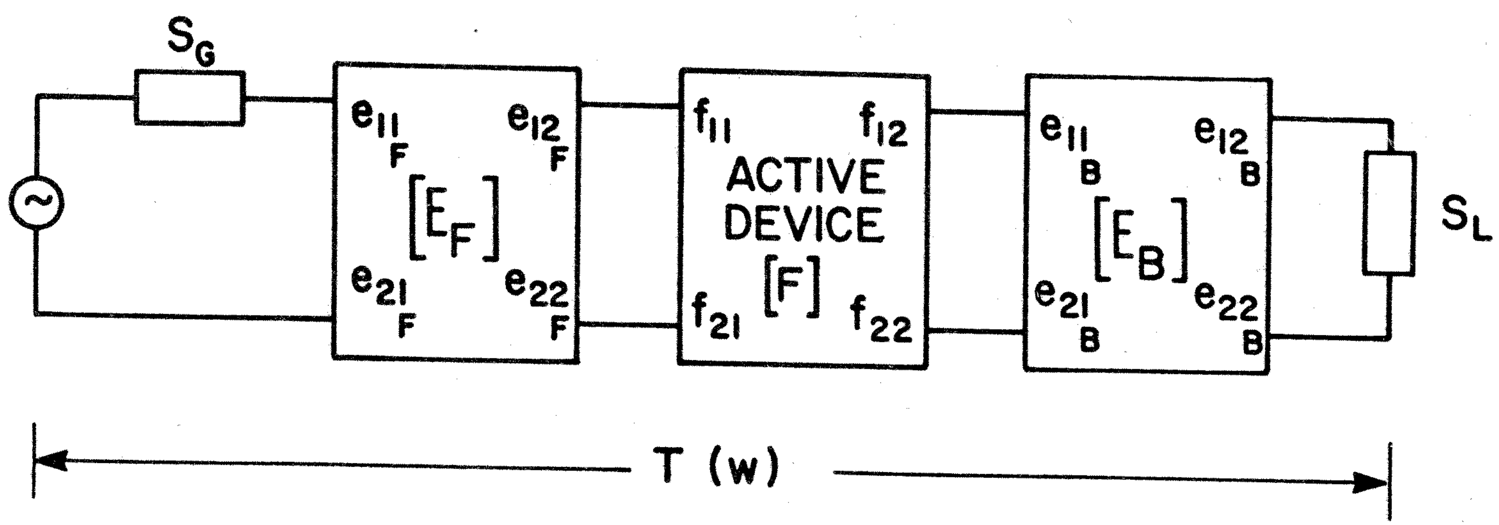
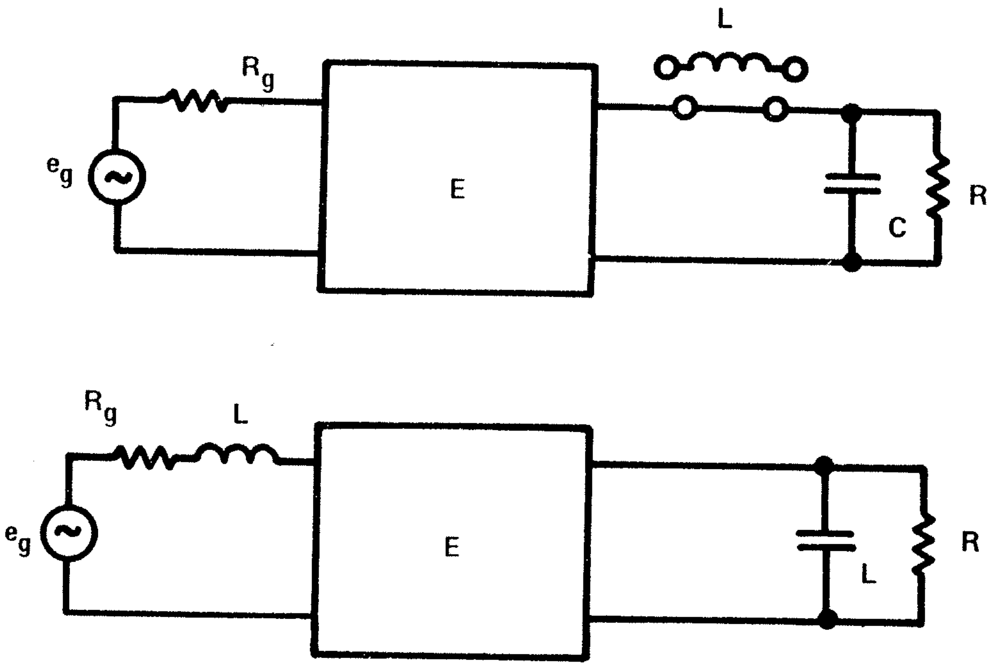


FIGURE-1c.

ANALYTIC THEORY

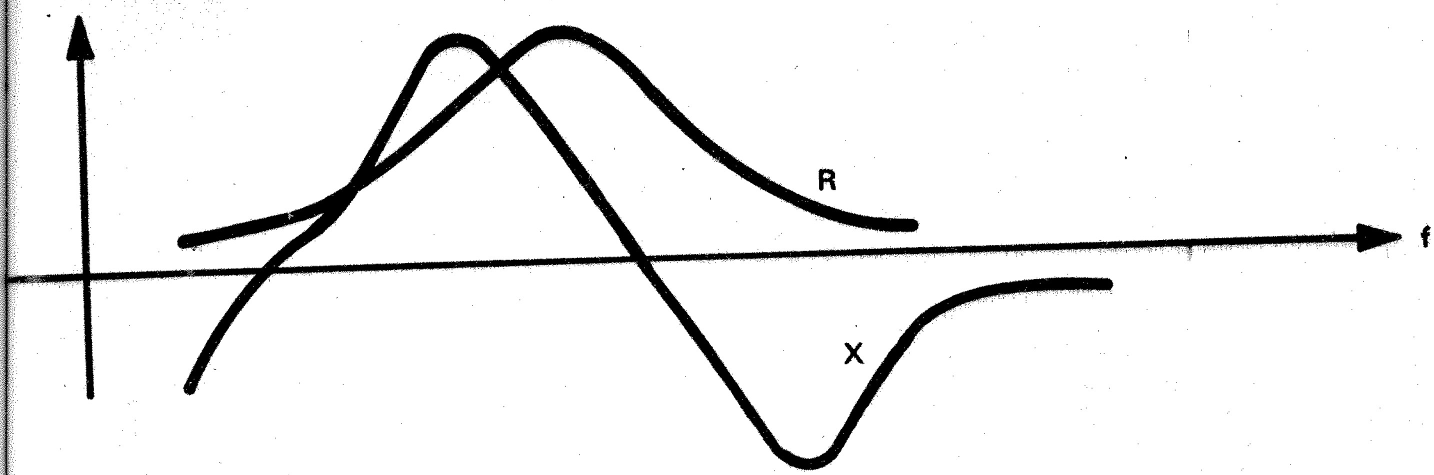
- ESSENTIAL TO UNDERSTAND GAIN BANDWIDTH PROBLEMS
- CAPABLE OF SOLVING SIMPLE PROBLEMS. EXPLICIT FORMULAS ARE AVAILABLE



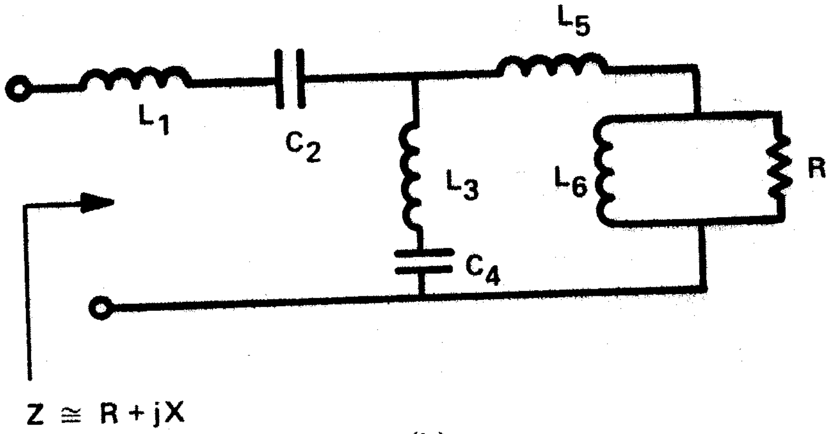
- FOR MORE COMPLICATED PROBLEMS THEORY IS INACCESSIBLE

FIGURE-2.

PROCEDURE



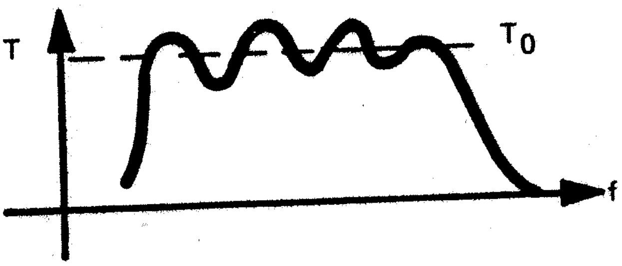
(a)



(b)

MODEL GENERATOR & LOAD NETWORKS

2.



CHOOSE A TRANSFER FUNCTION

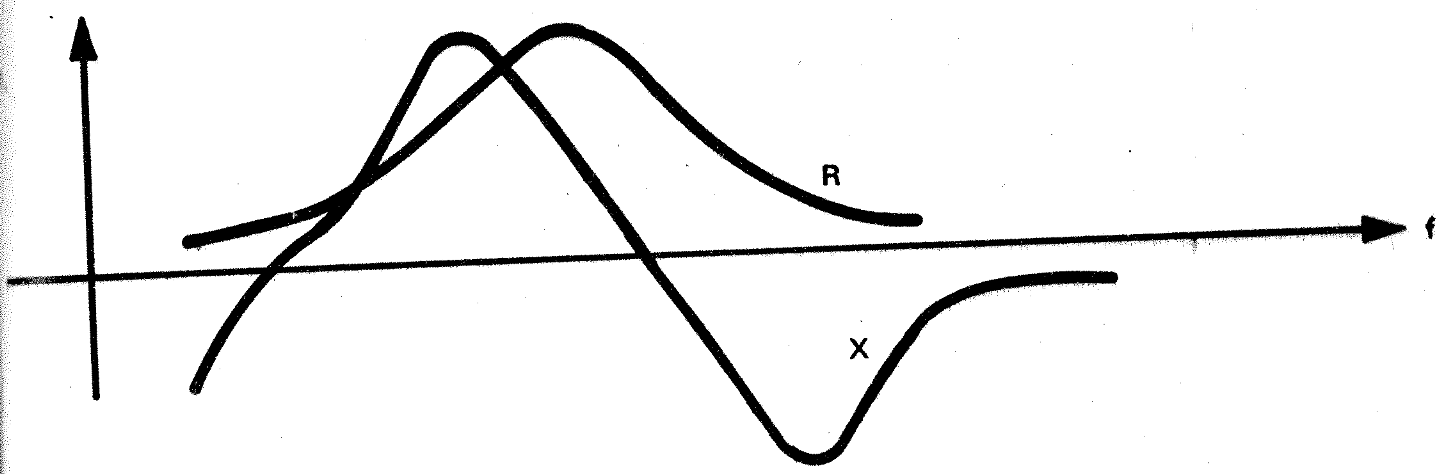
$T(w, x_1, x_2, x_3, x_4, x_5, x_6, x_7 \ldots)$

3. SOLVE GAIN BANDWIDTH EQS

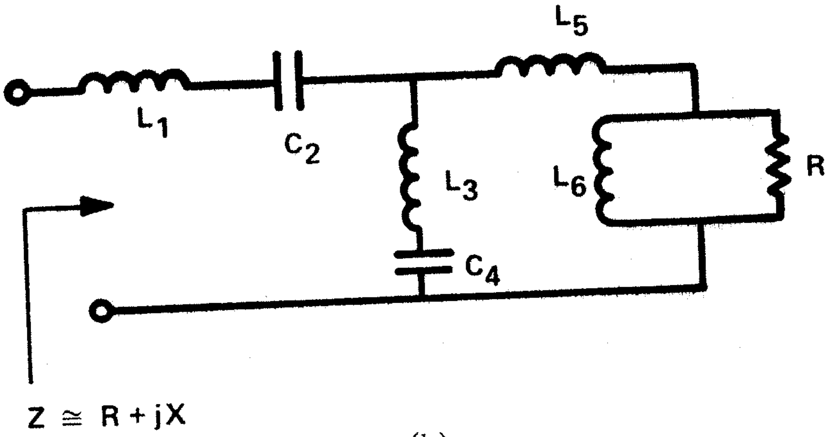
4. SYNTHESIZE THE MATCHING NETWORK

FIGURE-3.

PROCEDURE



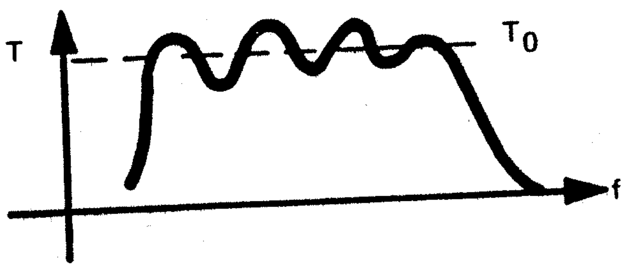
(a)



(b)

MODEL GENERATOR & LOAD NETWORKS

2.



CHOOSE A TRANSFER FUNCTION

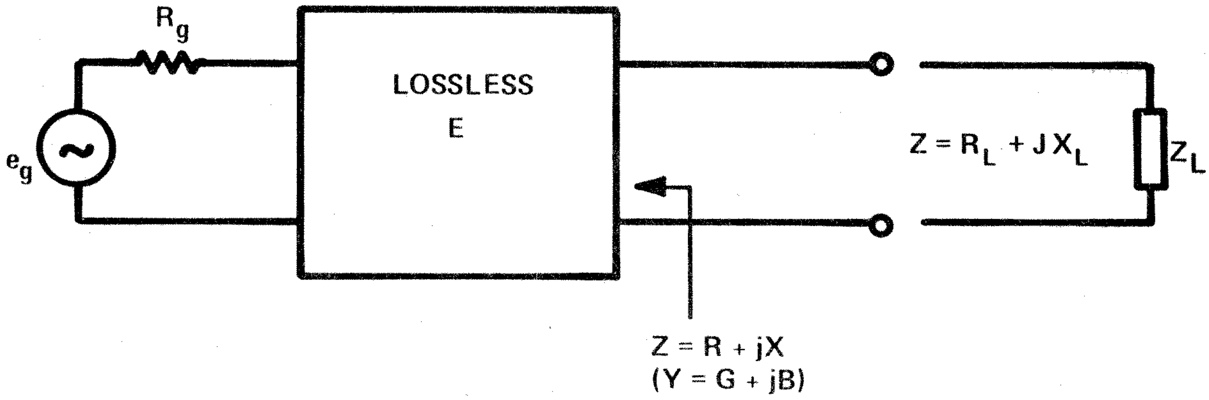
$T(w, x_1, x_2, x_3, x_4, x_5, x_6, x_7 \ldots)$

3. SOLVE GAIN BANDWIDTH EQS

4. SYNTHESIZE THE MATCHING NETWORK

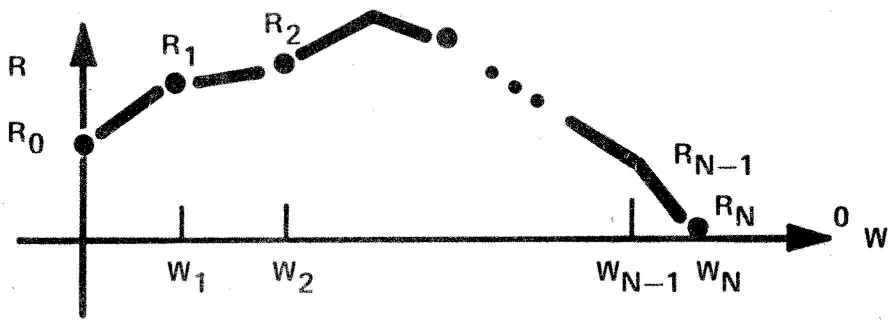
FIGURE-3.

REAL FREQUENCY TECHNIQUE
FOR
SINGLE MATCHING PROBLEMS
(CARLIN – 1977)



Z IS MINIMUM REACTANCE
(or Y IS MINIMUM SUSEPTANCE)

X = H(R)



$$R_q = \Sigma\, a_i(w) R_i \geqslant 0$$

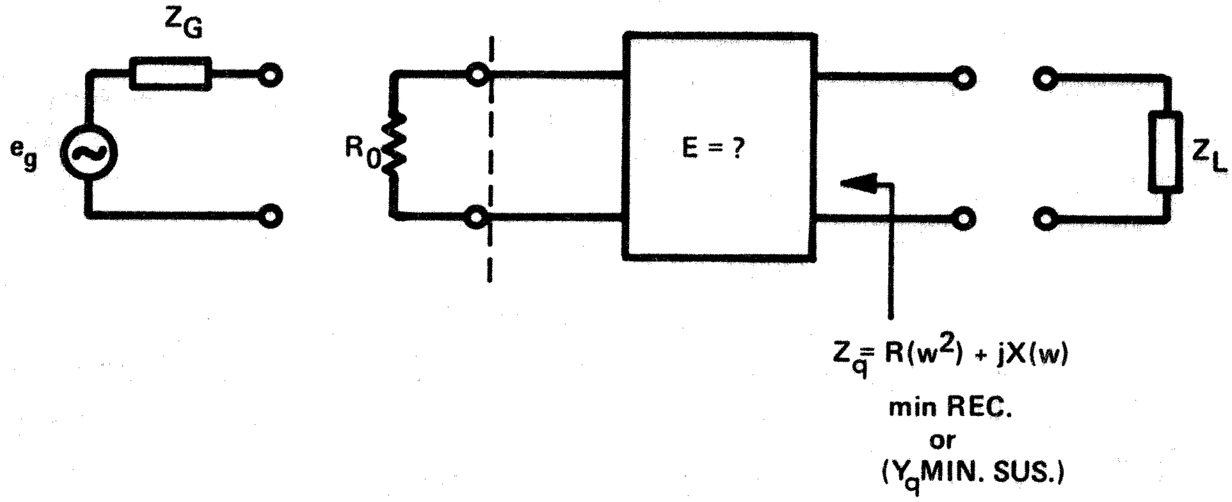
$$X_q = H(R) = \Sigma\, b\, R$$

$$T(w) = \frac{4\, R_q R_L}{(R_q + R_L)^2 + (X_q + X_L)^2}$$

FIGURE-4.

DIRECT COMPUTATIONAL TECHNIQUE
FOR
DOUBLE MATCHING (INTERSTAGE) PROBLEM

- LINE SEGMENT TECHNIQUE IS SIMPLY OMITTED



FOR LADDER STRUCTURES

$$R_q(w^2) = \frac{A_0 w^{2k}}{1 + B_1 w^2 + \dots + B_n w^{2n}} = \frac{AA_*}{DD_*}$$

CHOOSE A_0 & B_i AS UNKNOWNS

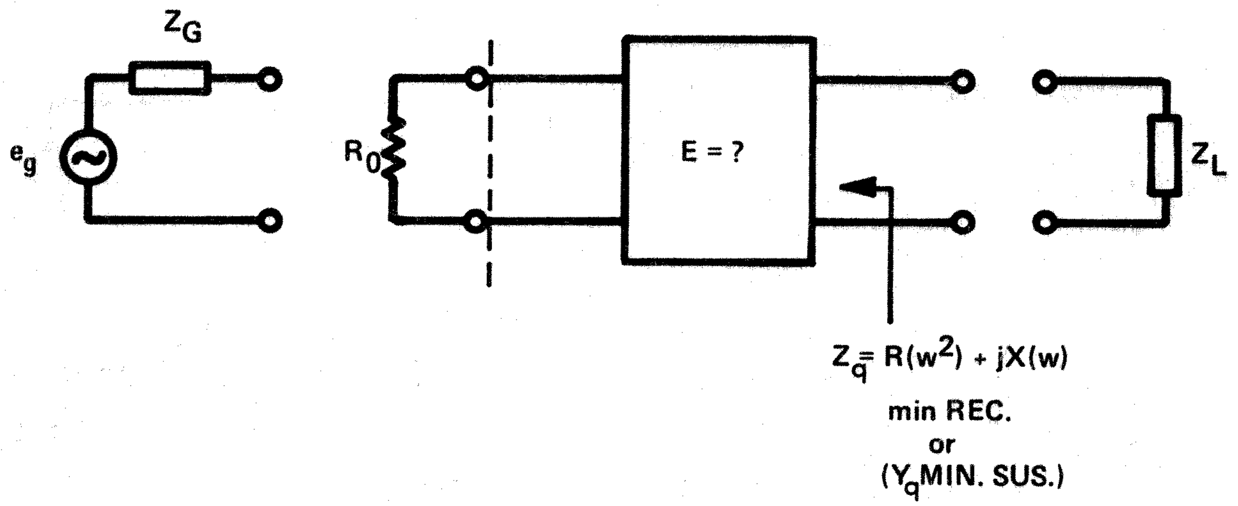
GENERATE GAIN IN TERMS OF $R(w^2)$ USING HILBERT TRANSFORMATION AND EXPLICIT FACTORIZATION TECHNIQUES.

DETERMINE A_0 & B_i TO OPTIMIZE GAIN: T (w, Z_G, Z_L, R)
k and n FIXED IN ADVANCE

FIGURE-5.

DIRECT COMPUTATIONAL TECHNIQUE
FOR
DOUBLE MATCHING (INTERSTAGE) PROBLEM

- LINE SEGMENT TECHNIQUE IS SIMPLY OMITTED



FOR LADDER STRUCTURES

$$R_q(w^2) = \frac{A_0 w^{2k}}{1 + B_1 w^2 + \ldots + B_n w^{2n}} = \frac{AA_*}{DD_*}$$

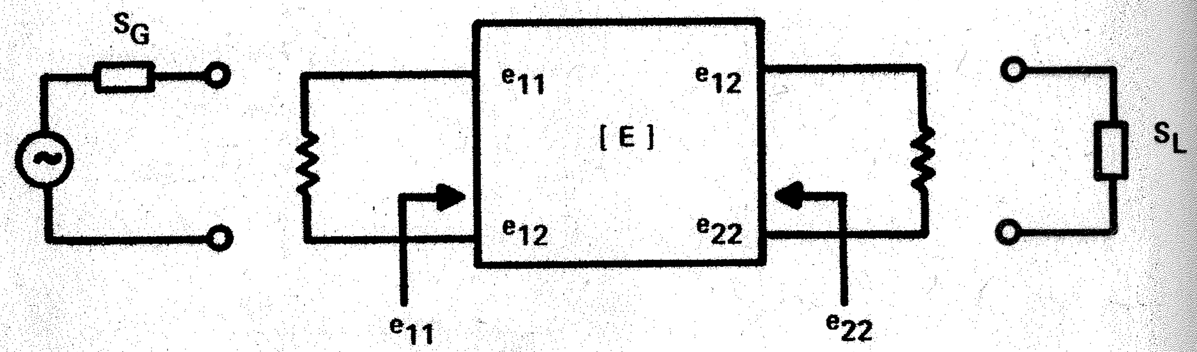
CHOOSE A_0 & B_i AS UNKNOWNS

GENERATE GAIN IN TERMS OF $R(w^2)$ USING HILBERT TRANSFORMATION AND EXPLICIT FACTORIZATION TECHNIQUES.

DETERMINE A_0 & B_i TO OPTIMIZE GAIN: $T(w, Z_G, Z_L, R)$
k and n FIXED IN ADVANCE

FIGURE-5.

SIMPLIFIED REAL FREQUENCY TECHNIQUE
"A SCATTERING APPROACH"



[E] IS DESCRIBED IN TERMS OF e_{ij}.

$$e_{11} = \frac{H(S)}{G(S)} = \frac{H_0 + H_1S + \ldots + H_nS^n}{G_0 + G_1S + \ldots + G_nS^n}$$

COMPLETE SCATTERING PARAMETERS OF [E] CAN BE DETERMINED FROM H(S). THUS GAIN IS A FUNCTION OF H(jw). THEN CHOOSE $H_0, H_1, \ldots H_n$ AS UNKNOWNS AND COMPUTE THEM TO OPTIMIZE THE GAIN:

$$T(w, S_G, S_L, H(jw))$$

FIGURE-6.