Design of Broad-Band Matching Network With Lossy Junctions Using the Real-Frequency Technique

Eric Kerherve, Pierre Jarry, and Pierre-Marie Martin

***Abstract*—A computer-aided design (CAD) procedure based on the real-frequency technique (RFT) is introduced for treating the matching of an arbitrary load to a complex generator. In this paper, the method has been applied to the design of interstage matching networks for microwave active circuits. The RFT provides several advantages over most of the usual techniques. It requires neither any transistor model because it directly includes measuring scattering- and noise-parameter data, nor a predetermined matching-circuit topology is necessary. A low-noise amplifier (LNA) design proceeding directly from experimental data is presented. Moreover, a new idea for treating the broad-band matching problem leads to the use of an equalizer topology containing cascaded transmission lines with lossy junctions. Thus, gain–flatness and stability are satisfied by designing the input and the output matching circuits by the line-segment technique. An example is presented for the matching of a 0.1–5-GHz amplifier.**

***Index Terms*—Broad-band matching, hybrid realizations, lossy junctions, low-noise amplifier, real frequency.**

I. Introduction

IN THE CASE of multistage microwave active-circuit design, the computer-aided design (CAD) programs must solve the classical matching problem: construct a lossless matching network between a generator and a load such that the transfer of power is maximized over a prescribed frequency band (see Fig. 1). In the case of a single matching problem, the source impedance is taken as a resistance $(Z_G = R_G)$ and the load has reactive and resistive components $(Z_L = R_L + jX_L)$. However, the most general case is the double matching problem where source and load impedances are complex.

In the case of the active two-port problem, the input and output of an active device are to be respectively matched to a generator and a load. In addition, matching problems of this sort may be complicated by other design considerations such as obtaining maximum transducer power gain (TPG) or minimum noise figure. A typical example of this kind is the design of a low-noise amplifier (LNA). Authors such as Carlin [1], Komiak, [2], Amstutz [3], [4], and Yarman [5],

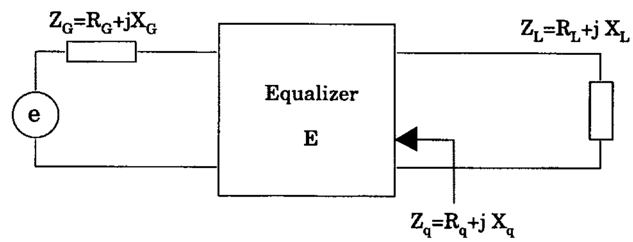

Fig. 1. Matching network with complex source and load terminations.

[6] have shown that the analytic theory of single and double matching problems is limited to simple problems (matching problems in which the generator and load networks include, at most, one reactive element). If the number of reactive elements increases on the load side, either the theory becomes inaccessible or the resulting gain performances turn out to be suboptimal and equalizers become unnecessarily complicated or sometimes completely unrealizable. On the other hand, in designing an active circuit, if the optimum topology is previously unknown, the task of designing matching networks of the circuit becomes difficult with the commercially available computer programs such as HP-EESOF, Supercompact, MDS, etc. The optimization can prove time-consuming, tedious, and without any guarantee regarding the convergence of the final result. It is even worse, the greater the number of parameters to be optimized and the wider the bandwidth. Carlin [1] was the first to propose a numerical approach called the real-frequency technique (RFT) to overcome the limitations of the analytical methods. Unlike the analytic and conventional CAD optimization techniques, this method utilizes the measured data obtained from the physical devices to be matched and requires no knowledge of an algebraic form of the transfer function or circuit topology. Carlin and Yarman [5] have presented the theoretical basis of the well-known method when it is applied to amplifier design. Authors have given examples which illustrate the merits of this synthesis technique for matching applied to broad-band microwave amplifier [2]–[9] and to microwave active filter [11]. A previous amplifier design [10] based on the RFT has included the noise figure over a frequency range of 100 MHZ to 6 GHz in the objective function. The continuity of this work using a modified RFT leads us to present a new circuit realization: a three-stage 14–14.5-GHz LNA. This paper also presents a new idea for treating the broad-band matching of a microwave amplifier.

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This idea leads to an alternative synthesis method based on the RFT. Lossy junctions are introduced between the cascaded transmission line of the equalizers [12] to offset the power gain at low frequencies. In this resistive matching case, the modified RFT directly synthesizes resistances and UE's. Both gain–flatness and stability are satisfied. Firstly, we only validate the new concept by designing a 0.1–5-GHz microwave amplifier. Secondly, the technique has been extended to design a two-stage 0.1–9-GHz broad-band amplifier.

Section II presents a summary of the RFT theoretical formulation. In Section III, the technique has been successfully modified and then incorporated into a CAD program to realize the matching networks of an LNA. Section IV presents a new formalism to design broad-band microwave amplifier and gives the associated synthesis. In Section V, two broad-band amplifiers were designed and built using the new technique and the laboratory measurements are presented.

II. Real-Frequency Matching via Scattering Approach

The numerical approach called the real-frequency or line-segment technique was introduced [2] to overcome the limitations of the analytical methods when applied to the single-matching problem. Using only measured two-port active device data, the Carlin method consists of generating a positive real (PR) impedance $Z_q = R_q(\omega) + jX_q(\omega)$ looking into a resistively terminated lossless matching network. This impedance is assumed to be a minimum reactance function so as to be able to determine $X_q(\omega)$ uniquely from $R_q(\omega)$ by a Hilbert transformation. In this manner, the TPG function of Z_q and Z_L has only one unknown R_q, which is computed by using a set of line segments to approximate the desired TPG bandpass response. $Z_q(\omega)$ is approximated by a realizable rational function (describing a ladder network, for example) which fits the computed data. Finally, Z_q is synthesized using the Darlington procedure as a lossless two-port with a resistive termination. Despite several attempts, it has not proved convenient to apply this method to the double-matching problem. The potential power of the RFT led to the development of a new numerical synthesis procedure by Yarman and Carlin [5], which has all the merits of the line-segment technique. In double-matching, the final result of the new procedure is an optimized, physically realizable, and unit-normalized reflection coefficient $e_{11}(p)$ where $p = \sigma + j\omega$, which describes the equalizer alone. The equalizer is placed between a complex source Γ_G and complex load Γ_L, as shown in Fig. 2. If $e_{11}(p)$ is appropriately determined, then the equalizer E may be synthesized using the Darlington theorem, which states that any bounded real (BR) reflection coefficient $e_{11}(p)$ is realizable as a lossless reciprocal two-port terminated in a pure resistance and a ladder-type network may be extracted.

This method has the further advantages of generality, being applicable to all matching problems, and universality, as it involves neither equalizer values nor a predefined equalizer topology. The simple formalism of this technique allows us, without complex calculations, to optimize many performance parameters of single and multistage microwave active circuits.

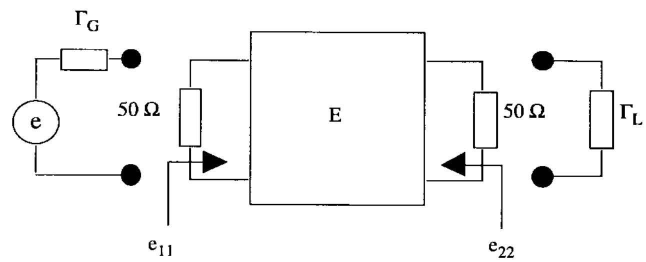

Fig. 2. RFT for double-matching problems.

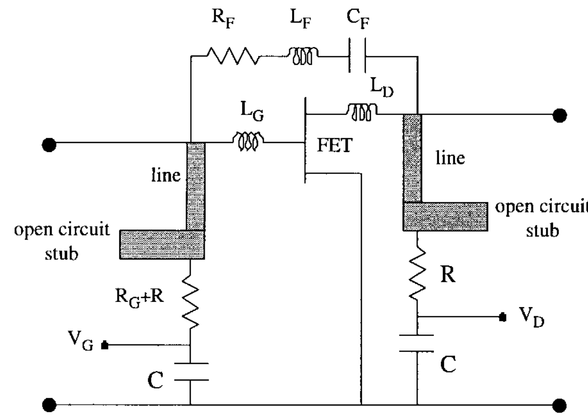

Fig. 3. TFB diagram.

III. Formalism to Narrow-Band Lossless Matching Problem

At microwave frequencies, many transistors are potentially unstable and are virtually impossible to match over large bandwidths. This is mainly due to the high input reflection coefficient and to the high gain of the device. Resistive feedback or lossy matching reduce terminal impedances and stabilize the transistor [13]. A solution integrating both methods has been adopted in our case. The transistor is thus embedded in a network including a parallel feedback loop (R_F, L_F, C_F), a drain and gate series inductances (L_D, L_G), and two bias circuits to form an elementary module termed the transistor feedback block (TFB) (Fig. 3).

The multistage synthesis method outlined in the following section is applied, with the TFB replacing the transistor as the active element. The extraction of lumped-element topologies of the ladder form using the Darlington procedure [14] has been described in [11]. We shall detail the synthesis procedure based on the RFT using the Richards' transformation $t = j\Omega$ for the extraction of the distributed commensurate transmission lines. In the case of the double matching problem, it has been shown [15] that the scattering parameters of an equalizer E can be completely determined from the numerator polynomial $h(t)$ of the input reflection $e_{11}(t)$. The insertion gain of a cascaded lossless transmission lines may be described [16] as a ratio of two even-power polynomials at t:

$$e_{12}(t)e_{12}(-t) = \frac{(1-t^2)^n}{P_n(-t^2)} \leq 1 \tag{1}$$

where $P_n(-t^2)$ is an even-power polynomial function of degree $2n$ in t. E is assumed to be a ladder network, thus

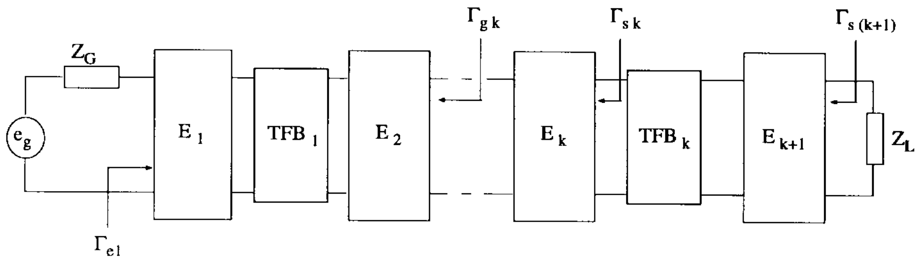

Fig. 4. Multistage microwave amplifier.

the scattering parameters are given as follows (Belevitch representation):

$$e_{11}(t) = \frac{h(t)}{g(t)} = \frac{h_0 + h_1 t + \cdots + h_n t^n}{g_0 + g_1 t + \cdots + g_n t^n} \tag{2}$$

$$e_{12}(t) = e_{21}(t) = \frac{(1-t^2)^{n-q/2}}{g(t)} \tag{3}$$

$$e_{22}(t) = -\frac{h(t)}{g(t)} \tag{4}$$

where n is the number of cascaded unit elements (UE's) of the matching network and q is the number of open circuit shunt stubs in a low-pass topology. The polynomial $g(t)$ is generated as a Hurtwitz polynomial by the factorization of

$$|e_{11}(t)|^2 + |e_{12}(t)|^2 = 1. \tag{5}$$

Thus, the physical realizability of the matching network is already built into the procedure. To obtain the scattering parameters of E it is, therefore, sufficient to generate the Hurwitz denominator polynomial $g(t)$ from $h(t)$. We are now in a position to obtain the parameters of the equalizer. In [17], the RFT algorithm has been detailed in the case of distributed UE matching-network design. Knowing $e_{ij}(t)$, we compute the TPG, input and output VSWR's, and noise figure, and we create the objective function including the desired parameters of the amplifier.

A. A Sequential Procedure for Designing a LNA

Referring to Fig. 4, the TPG for the first k cascaded stages active circuit is computed with (6)

$$T_k = T_{k-1} G_k \tag{6}$$

$$G_k = \frac{|e_{21k}|^2 |S_{21k}|^2}{|1 - e_{11k}\Gamma_{gk}|^2 |1 - S_{11k}\Gamma_{sk}|^2} \tag{7}$$

T_{k-1} is the gain of the first $(k-1)$ stages with normalized resistive terminations, $(e_{ij})_k$ are the scattering parameters of the kth equalizer E_k, and $(S_{ij})_k$ are the scattering parameters of the kth transistor with bias networks. The design of a multistage microwave active circuit is realized as the following procedure. First, we only synthesize the input equalizer E_1 by optimizing (8) (we only consider the first transistor with normalized resistive terminations [Fig. 4]). Next, the interstage equalizer E_2 is synthesized with the second transistor stage in place and we optimize (9). This last step is repeated for each k successive transistor stage. Once the kth TPG has been computed, the output equalizer E_{k+1} (Fig. 4) is added between the last transistor output and load impedance Γ_L and synthesized by optimizing the overall gain $T(\omega)$ defined by (10):

$$T_1 = (1 - |\Gamma_g|^2) G_1 \tag{8}$$

$$T_2 = T_1 G_2 \tag{9}$$

$$T = (T_1 T_2 \cdots G_{k+1}). \tag{10}$$

Using the same method, we are able to define the noise figure and the VSWR's of the multistage microwave amplifier.

B. Synthesis with Distributed Commensurate Transmission Lines

The method used to calculate the characteristic impedances of cascaded commensurate lines (specified for simplicity in the UE formulation), is now briefly outlined. The Richards' theorem states, in a manner analogous to the Darlington theorem [14], that the driving point impedance $Z_0(t)$ of a UE of characteristic impedance z_1 and terminated with a rational PR impedance $Z_1(t)$ is both rational and PR [Fig. 5(a)].

Extended to a cascade of n UE's [Fig. 5(b)], we obtain

$$Z_1(t) = Z_0(1)\,\frac{Z(t) - tZ(1)}{Z(1) - tZ(t)} \tag{11}$$

since $Z_0(1) = z_1$. Then we apply the generalized form of (11) to deal with the jth equalizer, j varying from 2 to n as follows:

$$Z_{j-1}(t) = \frac{V_{j-1}(t)}{U_{j-1}(t)} = z_{j-1}\,\frac{Z_{j-2}(t) - tz_{j-1}}{z_{j-1} - tZ_{j-2}(t)}. \tag{12}$$

Each time we extract a UE from the polynomial ratio, the order of Z_{j-1} decreases. In practice, this is ensured by dividing the numerator and denominator of (12) by $(1-t^2)$. Thus,

$$Z_{j-1}(t) = \frac{V_{j-1}(t)/(1-t^2)}{U_{j-1}(t)/(1-t^2)}. \tag{13}$$

Finally,

$$z_j = Z_{j-1}. \tag{14}$$

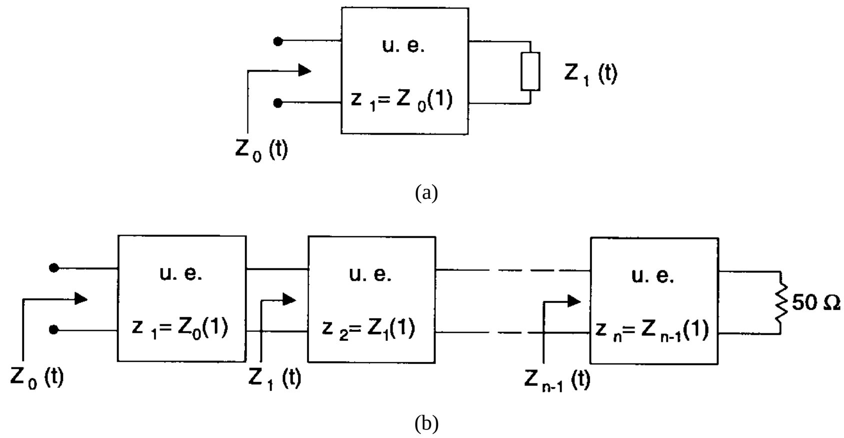

Fig. 5. (a) Diagram for derivation of the Richards' theorem. (b) Richards' theorem applied to a cascade of UE.

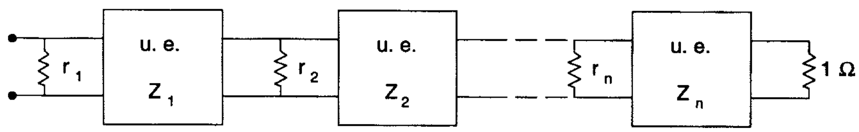

Fig. 6. n cascaded transmission-line network with lossy junctions.

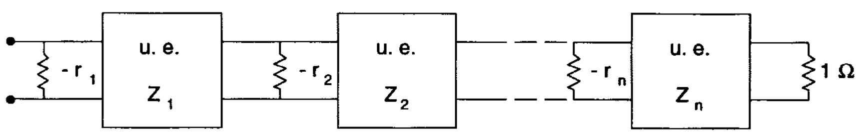

Fig. 7. Image network $\overline{N}$.

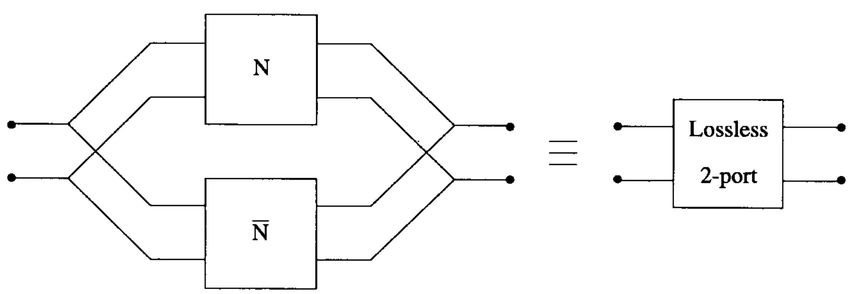

Fig. 8. Parallel connection of real and image networks.

The described procedure has been applied to the design of 14–14.5-GHz transmission-line LNA's. The obtained results are given in Section V.

IV. Formalism to Matching With Lossy Junctions

A. Design Theory

For the broad-band amplifier, the matching-network topology is modified introducing lossy junctions in cascaded transmission line N network (Fig. 6). The mathematics concept [12] termed *image network* is applied to the *real network* (Fig. 7). The parallel connection of these networks results in a purely reactive two-port (Fig. 8).

Based on (15), where I is the identity matrix, S and $\overline{S}$ the scattering matrix of N and $\overline{N}$ networks, (16) is obtained

$$I - S\overline{S}^* = 0 \tag{15}$$

$$S_{11}\overline{S_{11}}^* + S_{21}\overline{S_{21}}^* = 1. \tag{16}$$

Thus, the scattering-parameter TPG are now given by [19] as follows:

$$e_{11}(t) = \frac{h(t)}{g(t)} \tag{17}$$

$$e_{12}(t) = e_{21}(t) = \frac{(1-t^2)^{n/2}}{g(t)} \tag{18}$$

$$e_{22}(t) = -\frac{\overline{h(-t)}}{g(t)} \tag{19}$$

where $h(t) = \alpha(t) + \beta(t)$ and $\overline{h(t)} = \alpha(t) - \beta(t)$. $\alpha(t)$ is an even polynomial and $\beta(t)$ is an odd polynomial. The polynomial $g(t)$ is obtained by the factorization of (20) or (21). $\overline{e_{ij}(t)}$ are the S-parameters of the image TPG:

$$e_{11}\overline{e_{11}}^* + e_{21}\overline{e_{21}}^* = 1 \tag{20}$$

$$g(t)\overline{g(-t)} = h(t)\overline{h(-t)} + (1-t^2)^n. \tag{21}$$

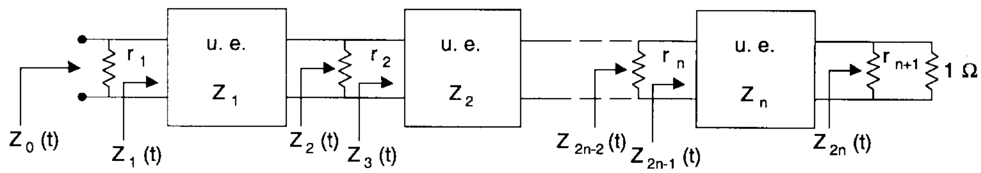

Fig. 9. Richards' theorem applied to a cascade of UE with lossy junctions.

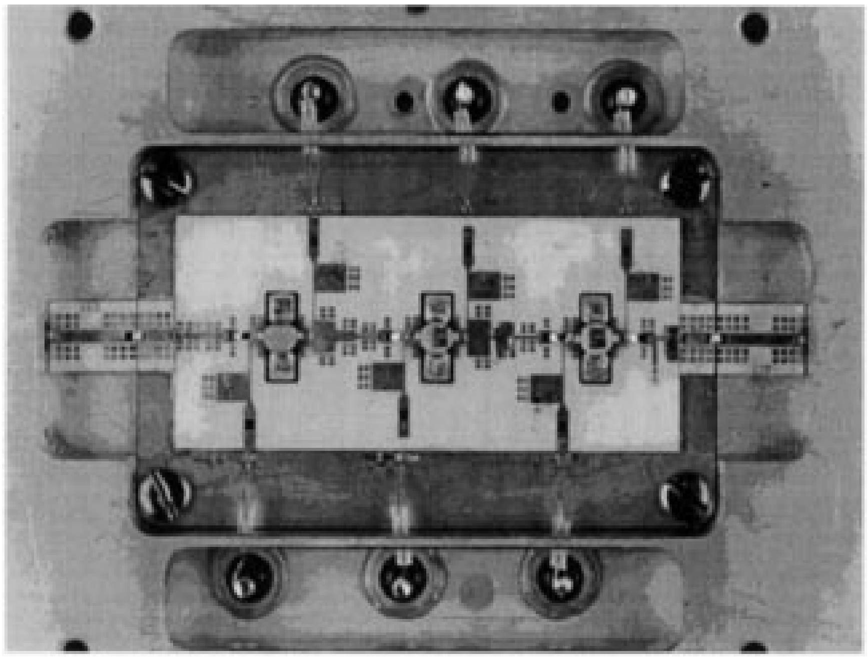

Fig. 10. Hybrid three-stage 14–14.5-GHz LNA.

In the case of a cascaded transmission line with lossy junctions, the RFT algorithm is similar to the one used in Section III. Thus, to construct the real S-parameters from $h(t)$ and $g(t)$ using (21), we firstly generate the polynomial

$$G(t) = g(t)\overline{g(-t)} = G_0 + G_1 t + G_2 t^2 + \cdots + G_{2n} t^{2n} \quad (22)$$

where

$$\begin{aligned} G_0 &= \alpha_0^2 - \beta_0^2 + 1 \\ G_1 &= 2\alpha_0\beta_1 - 2\alpha_1\beta_0 \\ &\vdots \\ G_k &= G_{2i} \\ &= (-1)^i(\alpha_i^2 - \beta_i^2 + C_n^i) \\ &\quad + 2\sum_{j=1}^{i} [(-1)^{j-1}(\alpha_{j-1}\alpha_{2i-j+1} - \beta_{j-1}\beta_{2i-j+1})], \\ &\qquad \text{odd } k \\ G_{k+1} &= G_{2i+1} \\ &= 2\sum_{j=1}^{i+1} [(-1)^{j-1}(\alpha_{j-1}\beta_{2i-j+2} - \alpha_{2i-j+2}\beta_{j-1})], \\ &\qquad \text{even } (k+1) \\ &\vdots \\ G_{2n} &= (-1)^n(1 + \alpha_n^2 - \beta_n^2). \end{aligned} \quad (23)$$

In a second step, we find the roots of $G(t)$. For a TPG with only transmission lines, the roots are conjugate and symmetrical to the imaginary axis and in the case of resistive equalizers, the roots are still conjugate, but are no longer symmetrical. Finally, we choose the left half-plane of $G(t)$ and form the polynomial $g(t)$. Knowing $e_{ij}(t)$, we compute the TPG, input and output VSWR's, and noise figure, and we create the objective function U.

B. Synthesis with Distributed Commensurate Transmission Lines and Lossy Junctions

In this case, the Richards' theorem is used to calculate (r_i, z_i) (Fig. 9).

We have

$$Z_0(t) = \frac{r_1 Z_1(t)}{r_1 + Z_1(t)}. \quad (24)$$

And we apply the Richards' theorem

$$Z_1(t) = z_1 \frac{Z_2(t) + t z_1}{z_1 + t Z_2(t)} \quad (25)$$

$$z_1 = Z_1. \quad (26)$$

We deduce r_1 from (27) and z_1 from (28):

$$r_1 = \frac{2Z_0(1)Z_0(-1)}{Z_0(1) + Z_0(-1)} \quad (27)$$

$$z_1 = \frac{2Z_0(1)Z_0(-1)}{Z_0(1) - Z_0(-1)}. \quad (28)$$

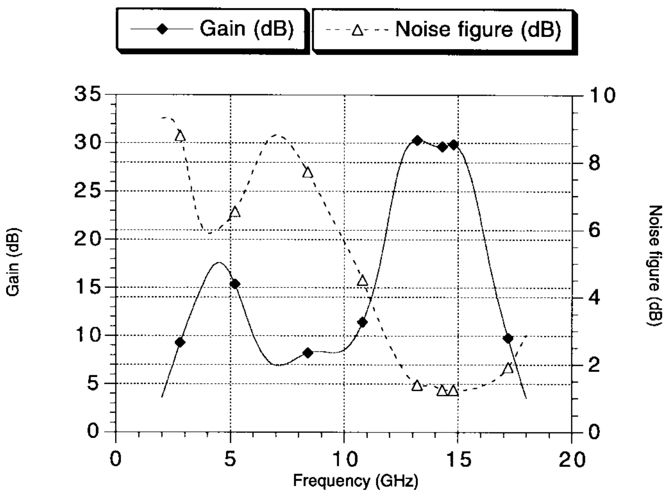

Fig. 11. Measured gain and noise figure performances of a 14–14.5-GHz LNA.

Then we apply the generalized form of (27) and (28) to deal with the jth equalizer, j varying from 2 to n:

$$r_{j-1} = \frac{2Z_{2j-2}(1)Z_{2j-2}(-1)}{Z_{2j-2}(1) + Z_{2j-2}(-1)} \tag{29}$$

$$z_{j-1} = \frac{2Z_{2j-2}(1)Z_{2j-2}(-1)}{Z_{2j-2}(1) - Z_{2j-2}(-1)}. \tag{30}$$

Finally,

$$r_{n+1} = \frac{Z_{2n}}{1 - Z_{2n}}. \tag{31}$$

This approach has been applied to a 100-MHz–5-GHz amplifier and to a 100-MHz–9-GHz multistage amplifier.

V. Applications

A CAD program based on the RFT and the Levenberg–Marquardt optimization algorithm is applied to synthesize the TPG's of an LNA and two broad-band amplifiers.

A. LNA

The proposed method can be applied using the Richards' transformation $t = j\Omega$ to distributed commensurate transmission-line extraction [16]. This approach has been used in a three-stage amplifier design. A hybrid 14–14.5-GHz LNA has been produced using three NEC24283A GaAs FET's with bias $V_{ds} = 2$ V and $I_{ds} = 10$ mA. Fig. 10 shows the hybrid implementation on a 0.635-mm-thick alumina substrate. The measured performances of this amplifier are shown in Fig. 11. In the band of interest, the gain is 29.7 dB with a good flatness ±0.1 dB. A noise figure less than 1.4 dB was obtained. This example shows that the matching equalizers obtained by the RFT are realizable.

B. Broad-Band Amplifier

The first example shows the use of the modified RFT to design a 100-MHz–5-GHz amplifier. This broad-band amplifier was realized using one NEC72084A GaAs FET's with bias $V_{\rm gs} = -1.25$ V, $V_{\rm ds} = 3$ V, and $I_{\rm ds} = 31.2$ mA. The gate and drain bias are the bias networks of the network analyzer. Fig. 12(a) shows the FET with lossy TPG's. The hybrid implementation [Fig. 12(b)] of the input and output equalizers was made on two separate 1-in^2 0.635-mm-thick alumina substrate.

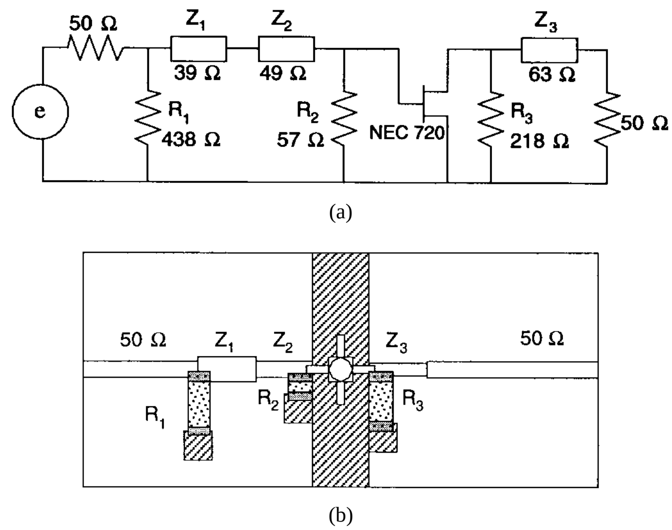

Fig. 12. (a) Schematic diagram. (b) Hybrid one-stage 100-MHz–5-GHz amplifier.

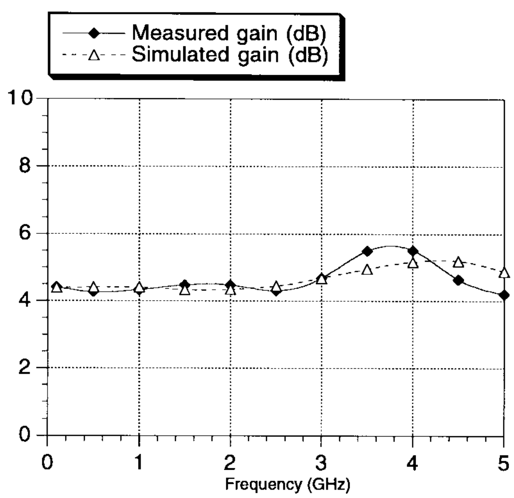

Fig. 13. Measured and simulated gain performances of 100-MHz–5-GHz amplifier.

The measured gain performance of this amplifier is shown in Fig. 13 together with the theoretical gain. The small-signal gain is 4.5 dB across the 100-MHz–5-GHz frequency range. Fig. 14 shows that input and output return loss are less than −5 dB. Thus, this example gives experimental results in good agreement with the theoretical calculations.

For the second example, we will only demonstrate that the synthesis procedure is applicable to multistage broad-band amplifier design. A two-stage amplifier has been optimized and simulated by the modified RFT over the 100-MHz–9-GHz frequency band. Fig. 15 shows the schematic diagram using two NEC71083 GaAs FET's with bias $V_{\rm gs} = 0$ V, $V_{\rm ds} = 3$ V, and $I_{\rm ds} = 40$ mA. The simulated gain performance (Fig. 16)

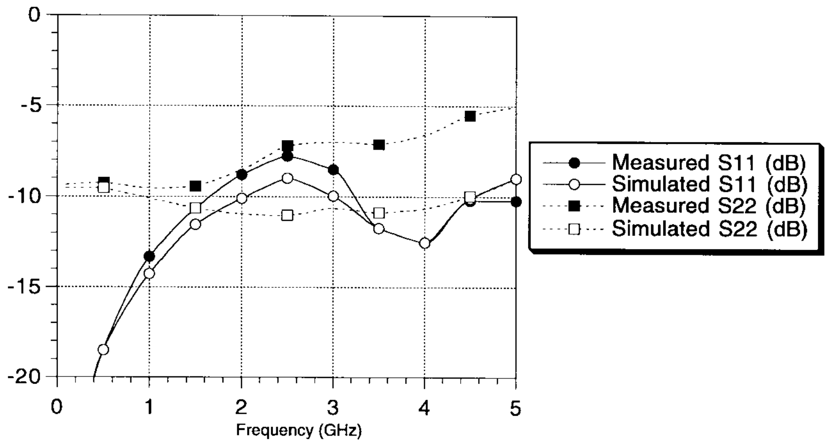

Fig. 14. Measured and simulated return-loss performances of 100-MHz–5-GHz amplifier.

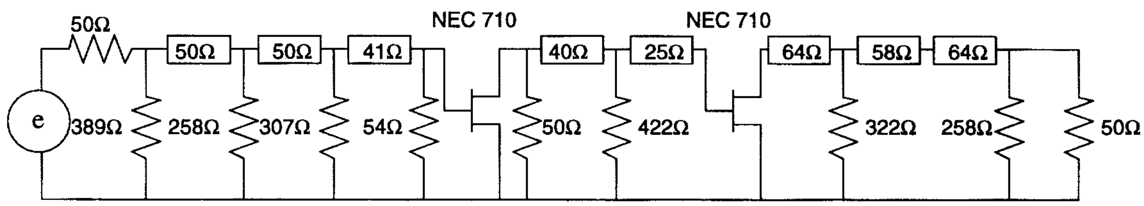

Fig. 15. Schematic diagram of the two-stage 100-MHz–9-GHz amplifier.

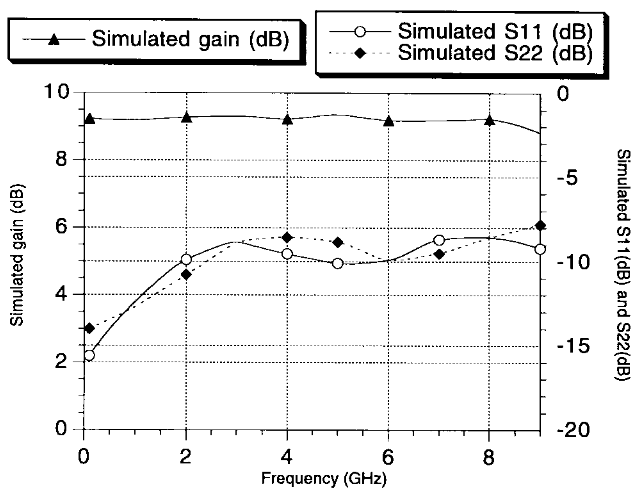

Fig. 16. Simulated gain and return loss performances of 100-MHz–9-GHz amplifier.

is 9, 1 dB with ± 0.2-dB flatness. The return losses are less than -7.8 dB over the prescribed frequency band.

The two examples validate the formalism for broad-band amplifiers using lossy matching networks.

VI. Conclusion

The proposed method allows designing microwave active circuits. It does not require any two-port active model because it directly includes the S-parameters. Moreover, the stability and the feasibility of computed equalizers are ensured by the Hurtwitz factorization. In this method, the RFT is modified such that the design of LNA's with distributed commensurate transmission-line matching networks and the design of a broad-band amplifier with resistive matching networks are possible. The computer program includes performance specifications on TPG, noise figure, and input and output VSWR's. The validity and advantages of the proposed method have been demonstrated by three design examples.

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