circuited by the diode capacitances that at frequencies above 50 GHz present a reactance less than 10 Ω.

Fig. 5 also show that for mixer 2 the bandwidth it is flat over a band of nearly 2 GHz.

V. Conclusions

We have developed a sophisticated model for millimeter wave mixers, featuring a new, particularly accurate linear analysis portion that takes into account the terminations of all relevant mixing products over a wide band. The good degree of agreement between experimental and theoretical data confirms the effectiveness of the model.

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A Design Method for Lumped Broad-Band MMIC Matching Networks with Semiuniform Frequency-Dependent Losses

Quinan Lin, Hamid Aghvami, and Ian Robertson

***Abstract*— A new design method for broad-band MMIC matching networks, which consist of lumped inductors and capacitors, is presented in this paper. Based on the fictitious transformation between the lossy network and the lossless one [6] the transducer power gain (TPG) of the lossy network is calculated using impedance matrix and transmission parameter matrix methods. Then, on the basis of the result of TPG optimization, the matching network can be synthesized. Since only the complexity of the matching network needs to be specified, so the advantage of the "real frequency technique" [5] is retained. In this design procedure the frequency-dependent losses of lumped elements can be considered. Thus the actual gain response tends to coincide with the desired performance more than when the losses of elements are neglected. An example is given to show the application of the new method to broad-band GaAs FET amplifier design.**

I. Introduction

For the synthesis of matching networks for broad-band GaAs MESFET amplifiers, the elements of the network have generally been considered as lossless in the previously reported techniques. In order to occupy less GaAs chip area, the matching networks for MMICs are usually constructed by lumped or semi-lumped reactance elements. Because these elements have small size and operate in the microwave frequency range, neglecting the losses would cause rather evident errors in design [1]. Therefore it is necessary in practice to improve the synthesis procedure of the matching networks. Some approaches to the synthesis of broad-band matching networks with losses have been reported in the past few years. Based on Andersen's transform [2], Youla's analytical theory of broad-band matching can be extended to the lossy network of the semiuniform type (i.e. all inductors have one quality factor Q; all capacitors have another Q [3]). In the design of microwave broad-band amplifiers, if the unilateral model of the GaAs FET is used and the losses of the circuit elements are assumed frequency independent, the arbitrary non-uniform lossy network can be synthesized [4]. By use of the fictitious transformation between the lossy network and the lossless one and using the "real frequency technique" [5] the lumped matching network with semi-uniform frequency-dependent losses can be designed [6] and a lossy commensurate line network can be synthesized [7].

In this paper, we present a new synthesis method for the lumped broad-band matching network with seminuniform frequency-dependent losses. According to the fictitious transformation, the impedance parameters of the lossy network can be obtained from the lossless one. Then the transducer power gain of the lossy network is calculated using transmission parameter matrices and optimized using the least-square method. On the basis of the result of TPG optimization, the synthesis of the matching network can be implemented in a fictitious transformation domain. It is more direct

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than the reported technique [6], because the inverse calculation of scattering matrix between lossy and lossless networks is avoided. In this design procedure only the complexity of the matching network needs to be chosen, so the advantage of the "real frequency technique" is retained.

II. The Fictitious Lossless Network

In general the impedances of lumped microwave lossy elements can be described by the use of an equivalent series circuit [1], [8]. Using the unit normalized circuit, the normalized impedance of the inductor and capacitor may be written respectively as:

$$Z_L = \left(\frac{\omega^{1/2}}{Q_l} + j\omega\right) \cdot L = \alpha(\omega) \cdot L \tag{1a}$$

$$Z_c = \left[\left(\frac{\omega^{1/2}}{Q_c} + \frac{1}{Q_d\omega}\right) - \frac{j}{\omega}\right] \cdot \frac{1}{C} = \beta(\omega) \cdot \frac{1}{C}. \tag{1b}$$

Where Q_l is the quality factor of the inductors at the reference angular frequency of normalization, ω_o. Q_c and Q_d are the quality factors of the capacitor at ω_o and are determined separately by skin and dielectric loss, respectively.

Approximately, all inductors have one quality factor Q_l, all capacitors have the same Q_c and the same Q_d.

Using the transformation

$$\tilde{Z} = \frac{Z(\omega)}{\sqrt{\alpha(\omega) \cdot \beta(\omega)}} \tag{2}$$

the impedance of the inductors and capacitors can be written respectively as follows:

$$\tilde{Z}_L = \frac{Z_L}{\sqrt{(\alpha(\omega)\beta(\omega)}} = \sqrt{\frac{\alpha(\omega)}{\beta(\omega)}} \cdot L \tag{3a}$$

$$\tilde{Z}_C = \frac{Z_c}{\sqrt{\alpha(\omega)\beta(\omega)}} = \sqrt{\frac{\beta(\omega)}{\alpha(\omega)}} \cdot \frac{1}{C}. \tag{3b}$$

If we let

$$\lambda = \sqrt{\frac{\alpha(\omega)}{\beta(\omega)}} \tag{4}$$

then (3a) and (3b) can be written as

$$\tilde{Z}_L = \lambda L \tag{3a}$$

$$\tilde{Z}_C = \frac{1}{\lambda C} \tag{3b}$$

It is clear that (3a) and (3b) are similar to the operational impedances of a lossless network. Thus it can be seen when using the transformation (2) that a semiuniform lossy network becomes a fictitious lossless network. λ is the fictitious complex frequency. Obviously, if the impedance matrix $\tilde{Z}(\lambda)$ of the fictitious lossless network is determined, then the corresponding impedance matrix $Z(\omega)$ of the semiuniform lossy network can be obtained by inverse transformation, i.e.

$$Z(\omega) = \sqrt{\alpha(\omega)\beta(\omega)}\tilde{Z}(\lambda)|_{\lambda=\sqrt{\alpha(\omega)/\beta(\omega)}} \tag{5}$$

III. The Transducer Power Gain of the Lossy Network

It is well known that the transducer power gain of two-port networks in cascade can be calculated conveniently by the multiplications of their transmission parameter matrices. For example, the total transmission parameter matrix of a two-stage amplifier is given by:

$$T = T^{(1)} \cdot T^{(2)} \cdot T^{(3)} \cdot T^{(4)} \cdot T^{(5)} \tag{6}$$

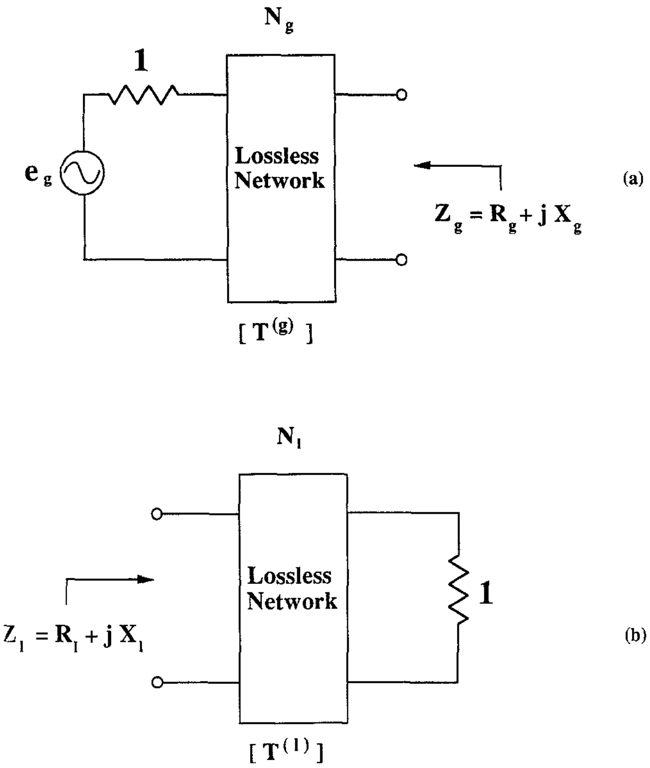

Fig. 1. Equivalent circuit for (a) complex generator and (b) complex load.

Where $T^{(2)}$ and $T^{(4)}$ is the transmission parameter matrix of the FET, and $T^{(1)}, T^{(3)}$ and $T^{(5)}$ are the transmission parameters of the input, interstage, and output matching networks, respectively. For double-matching problems, i.e. Z_g and Z_l are complex numbers, $Z_g = R_g + jX_g$ and $Z_l = R_l + jX_l$, the following method can be used: A passive one-port network can be represented as a lossless Darlington two-port network with a resistive termination [5], and so Z_g and Z_l may be represented as shown by Fig. 1(a) and 1(b) respectively. Assuming the lossless networks N_g and N_l are symmetrical, the impedance parameters of the lossless network $N_t(t = g, 1)$ can be expressed as follows:

$$\left.\begin{aligned} Z_{11}^{(t)} &= Z_{22}^{(t)} = j\left(\frac{X_t}{1 - R_t}\right) \\ Z_{12}^{(t)} &= Z_{21}^{(t)} = \pm j\left(\frac{\sqrt{R_t\{(1 - R_t)^2 + X_t^2\}}}{1 - R_t}\right) \end{aligned}\right\} \tag{7}$$

On the basis of the property of symmetrical lossless networks, and in accordance with the relationshp between the transmission parameter matrix and the impedance matrix, the transmission parameter matrix $T^{(t)}(t = g, 1)$ of network N_t may be described directly using R_t and X_t as

$$T^{(l)} = \frac{1}{2\sqrt{R_t\{(1 - R_t)^2 + X_t^2\}}} \cdot \begin{bmatrix} \pm 2X_t, \mp j(1 - R_t^2 - X_t^2) & \mp j\{(1 - R_t)^2 + X_t^2\} \\ \pm j\{(1 - R_t)^2 + X_t^2\} & \pm 2X_t \pm j(1 - R_t^2 - X_t^2) \end{bmatrix} \tag{8}$$

In this formula (8) the two groups of number may be selected arbitrarily. Thus, in the case of complex generator and load, the right side of (6) is multiplied at the left by $T^{(g)}$ and the right side by $T^{(l)}$ only. It should be noted that with this modification the transducer power gain of the overall amplifier (including complex generator and load) remains unchanged [9].

IV. The Synthesis of the Semiuniform Lossy Network

For convenience of design and manufacture, the lossless network is usually assumed to be a minimum phase structure with transmission zeros only at $S = 0$ and $S = \infty$ on the complex frequency domain. This assumption is advantageous to the design because it ensures that the realized network structure has no coupling coil, except perhaps there may be an ideal transformer. In this condition the scattering parameters may be expressed employing the well known Belevitch representation as follows [5]:

$$\left.\begin{aligned} S_{11}(s) &= \frac{h(s)}{g(s)} = \frac{h_0 + h_1 s + \cdots\cdots + h_n s^n}{g_0 + g_1 s + \cdots\cdots + g_n s^n} \\ S_{12}(s) &= S_{21}(s) = \frac{f(s)}{g(s)} = \pm\frac{s^k}{g(s)} \\ S_{22}(s) &= -(-1)^k \frac{h(-s)}{g(s)} \end{aligned}\right\} \quad (9)$$

moreover there is the relation

$$g(s)g(-s) = h(s)h(-s) + (-1)^k s^{2k} \quad (10)$$

where n specifies the maximum number of reactive elements in the network, integer $k (0 < k < n)$ specifies the order of the zero of transmission at $s = 0$.

From the above it can be seen that when n and k are selected, we need only determine $h(s)$ and from it to produce $g(s)$, and the character of the matching network is defined totally. Through solving (10) and by choosing the LHP roots of $g(s) \cdot g(-s)$, $g(s)$ can be constructed in terms of these roots. The synthesis of lumped broad-band matching networks with semiuniform frequency-dependent losses can be accomplished from the fictitious lossless network. Based on the above approach, first n and k of each network are determined. In normal conditions k may be stipulated for $0 < k < n$ (i.e. passband type), when the resulting matching network contains an ideal transformer so as to absorb it by using impedance transformation with inductive and capacitive T and pi networks [10]. Then the initial value of coefficients of $h_i \; (i = 0, 1, \cdots\cdots, n)$ are given, thereby $h(\lambda)$ and $f(\lambda)$ are written as:

$$h(\lambda) = h_0 + h_1\lambda + \cdots\cdots + h_n\lambda^n \quad (11)$$

$$f(\lambda) = \lambda^k \quad (12)$$

and we generate the polynomial

$$\begin{aligned} g(\lambda)g(-\lambda) &= h(\lambda)(-\lambda) + (-1)^k\lambda^{2k} \\ &= A_0 + A_1\lambda^2 + \cdots\cdots + A_n\lambda^{2n} \end{aligned} \quad (13)$$

The Hurwitz polynomial $g(\lambda)$ in the λ plane will be obtained from solving (13). In order to avoid the transformation between lossy and lossless networks, which involves matrix inversion, the relationship between scattering parameters and impedance can be used to express the impedance matrix $\tilde{Z}(\lambda)$ from $h(\lambda)$, $f(\lambda)$ and $g(\lambda)$ as

$$\tilde{Z}(\lambda) = \frac{1}{g_0(\lambda) - h_0(\lambda)} \begin{bmatrix} g_e(\lambda) + h_e(\lambda) & f(\lambda) \\ f(\lambda) & g_e(\lambda) - h_e(\lambda) \end{bmatrix} \quad (14a)$$

(when $f(\lambda)$, is an even function)

$$\tilde{Z}(\lambda) = \frac{1}{g_e(\lambda) - h_e(\lambda)} \begin{bmatrix} g_0(\lambda) + h_0(\lambda) & f(\lambda) \\ f(\lambda) & g_0(\lambda) - h_0(\lambda) \end{bmatrix} \quad (14b)$$

(when $f(\lambda)$ is an odd function)

where subscripts "e" and "o" indicate that even and odd part of the polynomial of λ respectively. In terms of (5), taking inverse transformation the transmission parameter matrix $T(\omega)$ of each lossy matching network can be obtained from the following expression:

$$T = \frac{1}{2Z_{21}} \begin{bmatrix} |Z| + Z_{11} + Z_{22} + 1 & -(|Z|) - Z_{11} + Z_{22} - 1) \\ |Z| + Z_{11} - Z_{22} - 1 & -(|Z| - Z_{11} - Z_{22} + 1) \end{bmatrix} \quad (15)$$

where

$$|Z| = Z_{11}Z_{22} - Z_{12}Z_{21}$$

Thus the transducer power gain of the overall chain of networks can be calculated. According to the bandwidth requirement a normalized frequency $\omega_j (j = 1, 2, \cdots\cdots m)$ is chosen and coefficients h_i are taken as design variables, a simple approach to optimize the transducer power gain $G_T(\omega)$ may be formulated by using the least-square method. The objective function may be written as

$$\delta = \sum_{j=1}^{m} \{G_T(\omega) - G_0\}^2 \quad (16)$$

where G_0 is the desired flat gain level to be approximated in the least-square sense. An optmization program is used to minimize the error δ, and the final result of $h(\lambda)$ is obtained, and then $\tilde{Z}(\lambda)$ is determined. On the basis of $\tilde{Z}(\lambda)$ the synthesis of the fictitious lossless network is carried out in the λ domain, so the topology and normalized value of the elements are obtained. In other words, the topology and each element's value for the lossy network which can satisfy the desired performance are determined.

V. Design Example

To demonstrate the synthesis technique, a two-stage microwave integrated circuit broad-band amplifier is designed with the following specifications:

50 Ω input and output impedance.
CX571 GaAs FET device (made in China).
Bandwidth from 3.2 GHz to 4.4 GHz.
Quality factors of elements are given as $Q_l = 50, Q_c = 100$ and $Q_d = 1000$.
Desired Gain 22 dB.
Maximum complexity of the matching networks:
 Input: $n = 3, k = 0$;
 Interstage: $n = 3, k = 0$;
 Output: $n = 2, k = 0$.

The amplifier design generated by the synthesis procedure is shown in Fig. 2 and its calculated and measured gain performance is shown in Fig. 3.

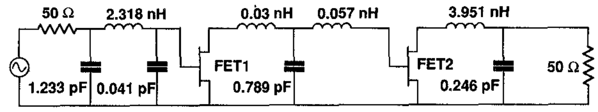

Fig. 2. Circuit diagram of the synthesized two-stage amplifier.

VI. Conclusions

The design of broad-band matching networks is one of the main problems in the realization of broad-band microwave amplifiers. In this paper we have presented a technique for the synthesis of lossy matching networks by using a fictitious transformation between the lossy and lossless networks. Since the synthesis of the network is carried out in the fictitious transformation domain, so it is no more difficult than the existing methods. However, because the loss of the network can be considered, the design accuracy is improved.

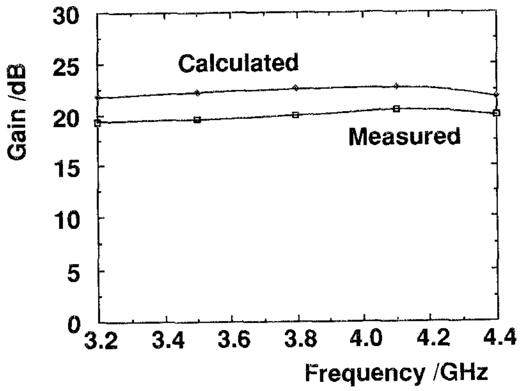

Fig. 3. Performance of the amplifier.

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X-Band Low Phase Distortion MMIC Power Limiter

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Abstract—**This paper describes the design and performance of a 8 GHz MMIC MESFET power limiter. This limiter incorporates a special gate biasing scheme and makes use of appropriate load conditions which reduce the unexpected phase variations experienced by the signal through the device.**

Measured performances (phase variations less than 8° over a 22 dB input power range) are found to be in agreement with the theoretical ones obtained from large signal simulations.

I. INTRODUCTION

Microwave power limiters are required for signal processing under constant amplitude conditions, for example to suppress AM components before demodulation of a microwave FM signal or to insure the incoming signal to a receiver does not exceed a given threshold. So far, this function has been fulfilled with p-i-n diodes, but this is not suitable for building MMIC since these diodes are not yet available from commercial foundries. Then GaAs FET limiters have been introduced [1] additionally providing a good input-output isolation.

The basic idea is to utilize the output power saturation of a class A GaAs FET amplifier. The FET is operated beyond its compression point so that the output RF signal cannot exceed its saturation level whatever the input signal amplitude. Over an input power dynamic range, output power variations may be reduced by cascading several FET stages [2].

The aforementioned applications require a low AM to PM conversion. Unfortunately the phase shift introduced by the limiter depends on the amplitude of the RF signal. Similarly the maximum range $\Delta\Phi$ of the observed phase shift variations (phase distortion) may greatly vary from 20° on some devices to a mere 1° on others, for a given power range.

The analysis of this phenomenon has been already reported [3] and the first section of this paper summarizes the most significant results. Then the second section describes a large signal model which yields input-output phase variations in good qualitative agreement with the experiment. Finally these two sections support the design of a low phase distortion MMIC power limiter whose characteristics are measured.

II. PHASE SHIFT DISTORTION ORIGINS

The results concerning the phase shift distortion origins are presented [3]. It has been shown that phase shift variations are mostly related to large input capacitance variations at moderate power levels and that at higher levels they are related to the forward-biased gate during a fraction of the RF cycle.

We therefore proposed the addition of an external series resistor, Rg, in the gate bias supply circuit [4]. This resistor shortens the fraction of the RF cycle during which the gate is forward biased and

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