A SIMPLIFIED "REAL FREQUENCY" TECHNIQUE APPLIABLE TO BROADBAND MULTISTAGE MICROWAVE AMPLIFIERS*

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ABSTRACT

A computer-aided design procedure, a simplified "Real Frequency" technique, applicable to broadband, multistage FET amplifiers is presented. The design procedure requires no decisions to be made in advance as to algebraic form of transfer function, or circuit topology. Furthermore it is more efficient, accurate and complete than currently-available CAD methods.

I. Introduction

In the design of broadband, multistage microwave amplifiers the fundamental problem is to realize lossless interstage equalizers as well as front-end and back-end equalizers so that the transfer of power from source to load is maximized over a prescribed frequency band. Here, the major problem is the "Double Matching", power transfer from a complex generator to an arbitrary load (Figure 1). In this paper, we will first introduce a new computer-aided design procedure, a simplified "Real Frequency" technique for double matching problems, then we will extend the technique to the design of broadband, multistage FET amplifiers.

II. Simplified "Real Frequency" Technique For Double Matching Problems

(Real Frequency Matching via scattering approach):

The method described here is an alternate version of the "Direct Computational" technique [3] which is the generalized form of "Real Frequency Matching" introduced by Carlin [1,2]. This new method has all the merits of [1,2,3]. That is, when it is implemented, there is no need to choose an algebraic transfer function class which is a measure of performance; nor is it necessary to assume an equalizer topology in advance; no equivalent circuit need be determined for the devices; the real frequency experimental data for the devices to be broadband is processed directly. The final result of the procedure is an optimized physically realizable unit normalized reflection coefficient (e_{11}) which describes the equalizer alone. As contrasted with [1,2,3], in the new design procedure, numerical evaluation of the Hilbert transformation is eliminated since the use of minimum reactance or minimum susceptance immittance function to describe the lossless matching network is omitted. Thus the computational efficiency is significantly improved. Furthermore, just as in the other versions of the "Real Frequency" method, the result and equalizers are generally simpler with superior gain properties as compared to structures obtained by the analytic procedure [4].

The heart of the technique resides in the generation of the unit normalized scattering parameters [$\{e_{ij}\}$, i=1,2] of the lossless equalizer E from the unknown - but initialized - numerator polynomial h(s) of the unit normalized reflection coefficient $e_{11}(s) = h(s)/g(s)$. For the sake of simplicity, E is assumed to be a minimum phase structure with transmission zero only at $\omega = \infty$, $\omega = 0$. Employing the well known Belevitch [5] representation, the scattering parameters $\{e_{ij}\}$ are written as

$$e_{11}(s) = h(s)/g(s), \quad e_{12}(s) = e_{21}(s) = \pm s^k/g(s)$$

and (1)

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

where $k \geq 0$ is an integer. By losslessness

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

The Hurwitz polynomial g(s) is explicitly computed during each iteration of the optimization routine until the final improved set of coefficients is found. Thus, bounded-real (BR) scattering parameters $\{e_{ij}\}$ of the lossless equalizer E are generated from the initialized numerator polynomial $h(s) = h_0 + h_1 s + \ldots + h_n s^n$ of $e_{11}(s) = h(s)/g(s)$. Hence, following the above procedure the transducer power gain $T(\omega^2)$ of the doubly terminated structure (Figure 1) can be generated as a function of e_{ij} and the complex terminations (see (3) below). The unknown real coefficients $h_0, h_1, \ldots, h_n$ of h(s) are optimized using a nonlinear optimization routine such that $T(\omega^2)$ is maximized over a specified pass band. Here, it is important to note that there is no restriction other than reality imposed on the unknown coefficients (h_i, i=0,.,n). Realizibility is simply achieved by choosing the denominator g(s) as a Hurwitz polynomial. Therefore, any unconstrained, straightforward optimization routine can be used to determine the unknown coefficients h_i.

III. Design of Multistage FET Amplifiers

Referring to Figure 2c for any cascaded (k) equalizer - FET stages, the transducer power gain $T_k(\omega^2)$ is given by

$$T_k(\omega^2) = T_{(k-1)} \left[\frac{|e_{21_k}|^2 \quad |\ell_{21_k}|^2}{|1 - e_{11_k} S_{G_k}|^2 \; |1 - \hat{e}_{22_k} S_{\ell_k}|^2} \right] = T_{(k-1)} \cdot E_k(\omega^2) \quad, k \geq 1 \tag{3}$$

where $T_{(k-1)}$ is the transducer power gain of the first (k-1) equalizer - FET stages with resistive termination, $E_k(\omega)$ is the term in brackets [ ], e_{ij_k} are the unit normalized scattering parameters of the (k)th equalizer E_k, S_{G_k} is the unit normalized reflection coefficient measured at port G_K to left, $\hat{e}_{22_k}$ and S_{ℓ_k}

*Sponsored under Joint Services Contract F49620-81-C-0082

 0149-645X/82/0000-0529 $00.75

are the unit normalized reflection coefficients measured at port L_K to left and right, respectively. It is straightforward to generate $\hat{e}_{22_k}$ and S_{ℓ_k} using the scattering parameters of the FET (F_k) and the information obtained from the previous stages [7]. Utilizing the simplified real frequency technique above, a multistage microwave FET amplifier may be designed as follows:

1 - Construct the front-end equalizer $E_F = \{e_{ij_F}\}$ for the first stage such that $T_1(\omega^2) = T_o \cdot E_1(\omega^2)$ is optimized (Fig. 2a), where $T_o = 1 - |S_G|^2$, and $S_G = \dfrac{Z_G - 1}{Z_G + 1}$ which is the known reflection coefficient of the source network.

2 - Second equalizer - FET stage cascaded with the first one. The interstage equalizer $E_2 = \{e_{ij_2}\}$ is computed so that $T_2(\omega^2) = T_1 \cdot E_2(\omega^2)$ is optimized. Notice that, in $T_2(\omega^2)$, $T_1(\omega^2)$ is regarded as a weighting function (Figure 2b).

3 - Repeat Step 2 for designing the cascaded stages (Figure 2c).

4 - Finally the back-end equalizer $E_B = \{e_{ij_b}\}$ is determined for given complex load S_L $(j\omega)$ so that over all transducer power gain $T(\omega^2)$,

$$T(\omega^2) = (T_1 \cdot T_2 \cdot\cdots\cdot T_k) \cdot E_B(\omega^2) \qquad (4)$$

is optimized. In (4) the term $(T_1 \cdot T_2 \cdot \ldots \cdot T_k)$ is received from the previous stages and acts as a weighting factor on $E_B(\omega^2)$ which is a function of the back-end equalizer (Figure 2d), and given by replacing "k" with "b" in $E_k(\omega^2)$ of (3). In this case,

$S_{\ell_b} = S_L = \dfrac{Z_L - 1}{Z_L + 1}$, $|\ell_{21b}|^2 = 1 - |S_L|^2$, which is the known reflection coefficient of the load.

In the course of the above design process, the gain taper of each FET is compensated at the port connected to the preceding equalizer in the cascade. The term $E_B(\omega^2)$ in (4) only provides impedance matching. It should be noted that (4) completely specifies the gain with all equalizers and FETs in place, and the non-unilateral behavior of the FETs taken into account. However, the design procedure can be iterated one or more times in order to improve the maximum flat gain level.

The design technique that has been discussed is applicable to optimizing a variety of objective functions as in [2]. Thus it can be used for maximizing the minimum passband gain, or for minimizing maximum noise figure, or high gain flat noise figure design, etc.

EXAMPLE:

It is desired to design two-stage FET amplifiers using Mitsubishi MGF 2124 FETs for the frequency band of 11.7 GHz < f < 12.2 GHz. FETs are assumed to be identical and measured scattering parameters are given in reference [6]. Following the design procedure introduced above, front-end (E_F), interstage (E_2) and back-end (E_B) equalizers were computed sequentially. The transducer power gain was found to be 7 ± 0.5 dB. The complete design is depicted in Figure 3. Here it is important to note that this design is based upon roughly-measured scattering parameters of FETs with test fixtures. Therefore, it also includes the loss of the test fixtures which results in reduction of gain. However, this initial design may be improved by iterating the above procedure and using more accurate scattering parameters for FETs.

Acknowledgment

Helpful discussions with Dr. Klatskin and Dr. Huang are gratefully acknowledged.

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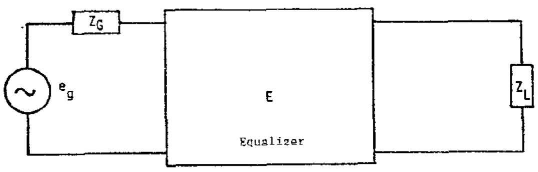


Figure 1 Double Matching Problem

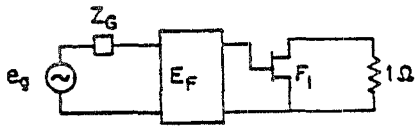


Figure 2a: Step 1 - Design of the Front end Equalizer E_F

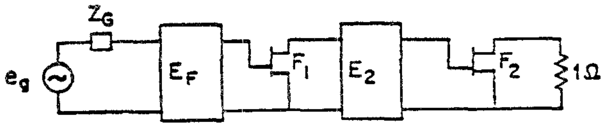


Figure 2b: Step 2 - Design of interstage equalizer E_2

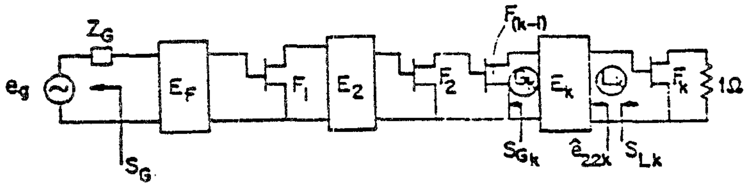


Figure 2c: Step k - Repeat Step 2 k times

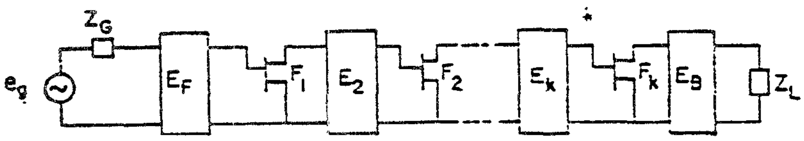
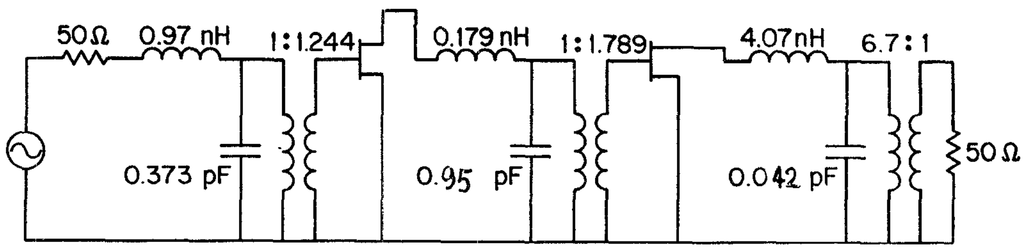


Figure 2d: The last step - Design of Back end equalizer E_B

Figure 2 Computation steps for
Designing Broadband Multistage Microwave FET Amplifiers



11.7 GHz $\leqslant$ f $\leqslant$ 12.2 GHz
Figure 3. Design of Two-Stage FET Amplifiers for X-Band