EFFICIENT NUMERICAL METHOD TO THE DESIGN OF MICROWAVE ACTIVE CIRCUITS

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ABSTRACT

The real frequency technique with the Marquardt-More optimization algorithm is probably one of the best microwave active circuit synthesis method available for the design of amplifiers and active filters. This numerical approach was introduced to overcome the limitations of the analytical methods, when applied to the single or double matching problem.

INTRODUCTION

The present classical synthesis techniques applied to the design of microwave active circuits, give no indication *a priori* on the values of the constituent elements of interstage equalizers. Therefore, optimization can prove long, tedious and without any guarantee as regards the convergence of the final result (this optimization is even longer, the greater the number of parameters to be optimized and the longer the bandwidth length). For the real frequency method, neither an *a priori* choice of matching network topology nor an analytical form of the system transfer function is necessary. Moreover, most of the usual techniques utilizes a model of the 2-port active device contrary to the presented procedure which directly includes the measured scattering parameters.

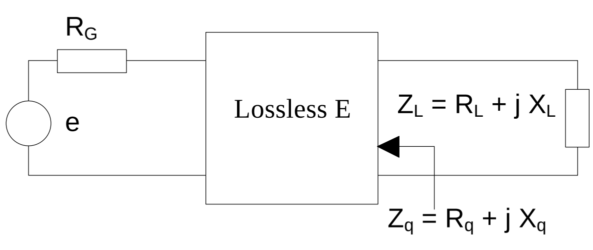

Figure 1: Real frequency technique for single-matching problem.

In the version described in this paper, the optimization process is applied simultaneously to gain, input and output VSWRs, noise figure and group delay of multistage microwave active circuits. The « FREELCD » software uses the real frequency method with the Marquardt-More optimization algorithm and is applied to the design of low-noise amplifier and bandpass filter.

I. THE REAL FREQUENCY METHOD

The numerical approach called the real frequency or line-segment technique was introduced [1] in 1977 to overcome the limitations of the analytical methods, when applied to the single-matching problem. Using only measured 2-port active device data, the Carlin method consists in 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 transducer power gain 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 transducer power gain 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 2-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 real frequency technique led to the development of a new

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numerical synthesis procedure by Yarman and Carlin [2] in 1982, 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, unit-normalized reflection coefficient $e_{11}(p)_{(p=\sigma+j\omega)}$, which describes the equalizer alone. The equalizer is placed between a complex source Γ_G and complex load Γ_L.

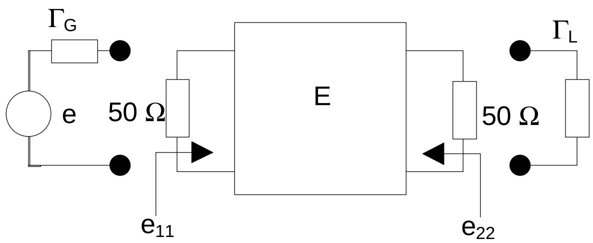

Figure 2**:** Real frequency technique for double-matching problem.

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 2-port terminated in a pure resistance [2] 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.

A. Formalism [3]

In the case of the double matching problem, the scattering parameters of an equalizer *E*, can be completely determined from the numerator polynomial h(p) of the input reflection $e_{11}(p)$. *E* is assumed to be a ladder network, thus the scattering parameters are given as following (Belevitch representation)

$$e_{11}(p) = \frac{h(p)}{g(p)} = \frac{h_0 + h_1 p + \ldots + h_n p^n}{g_0 + g_1 p + \ldots + g_n p^n} \qquad (1)$$

$$e_{12}(p) = e_{21}(p) = \frac{f(p)}{g(p)} \qquad (2)$$

$$e_{22}(p) = -\frac{(-1)^k h(-p)}{g(p)} \qquad (3)$$

where *n* is the number of passive elements of the matching network *E* and k is an integer which specifies the order of the transmission zeros (k < n) at the origin. The polynomial h(p) is chosen as the unknown and assumes the equalizer to be lossless. For the amplifier design, the equalizers *E* have all been assumed to be a minimum phase structure with transmission zero only at $\omega = \infty$, $\omega = 0$ (i. e. $f(p) = \pm p^k$). For the active filter design, we consider the general form of f(p) given as follows

$$f(p) = \pm p^k \prod_{i=1}^{m} (p^2 + \omega_i^2) \qquad (4)$$

where m is the number of the finite attenuation poles. The filter will be low-pass if k = 0, bandpass if 1 < k+2m < n or high-pass if k+2m = *n* . The polynomial g(p) is generated from the Hurwitz factorization

$$|e_{11}(p)|^2 + |e_{12}(p)|^2 = 1 \qquad (5)$$

$$g(p)g(-p) = h(p)h(-p) + f(p)f(-p) \qquad (6)$$

To obtain the scattering parameters of *E*, it is therefore sufficient to generate the Hurwitz denominator polynomial g(p) from h(p).

B. Algorithm

The inputs of the algorithm are n, k, Γ_G, Γ_L and h_i (i=1,2…n). The computational steps are:

① Generate the polynomial

$$G(p^2) = g(p)g(-p) = G_0 + G_1(p^2) + \cdots + G_n(p^{2n}) \qquad (7)$$

where

$$G_0 = h_0^2$$

$$G_1 = -h_1^2 + 2h_0h_1$$

$$G_i = (-1)^i h_i^2 + 2\left(h_{2i}h_0 + \sum_{j=2}^{i} (-1)^{j-1} h_{j-1} h_{2i-j+1} \right)$$

$$G_n = (-1)^n h_n^2$$

The initial values of h_i (with i = 1,2 … n) are selected in a pseudorandom manner for several different cases from a wide range of values. The coefficients G_i , corresponding to selected h_i are then computed using Marquardt-More

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algorithm [4]. ② Find the roots of $G(p^2)$. ③ Choose the left half plane of $G(p^2)$ and form the polynomial g(p). ④ Construct the real normalized scattering parameters $e_{ij}(p)$ from h(p) and g(p) using eq.6. ⑤ Knowing e_{ij}, compute the transducer power gain, input and output VSWRs, noise figure and group delay from classic formulae [5]. ⑥ Create the multiobjective function with respect to the parameters desired

$$U=\sum_{j=1}^{m} W_1\left(\frac{G(\omega_j)}{G_0}-1\right)^2+W_2\left(\frac{R_{in}(\omega_j)}{R_{in0}}-1\right)^2 +W_3\left(\frac{R_{out}(\omega_j)}{R_{out0}}-1\right)^2+W_4\left(\frac{F(\omega_j)}{F_0}-1\right)^2+W_5\left(\frac{D(\omega_j)}{D_0}-1\right)^2 \quad (8)$$

where G_{0_k}, R_{in0_k}, R_{out0_k}, F_{0_k} and D_{0_k} are respectively the specified transducer power gain, input VSWR, output VSWR, noise figure and group delay at the k^{th} equalizer ; m is the number of sampling frequencies ; W_i are the appropriate weighting functions.

C. Flow chart of FREELCD program

The flow chart represents the manner in which FREELCD is organized. From the input data which include S-parameter data for each type of transistor used, amplifier specifications of bandwith, gain, input and output VSWRs, noise figure, group delay and number of elements for each equalizer, the program optimizes the vector h_i for each stage of the active circuit using the optimization routine.

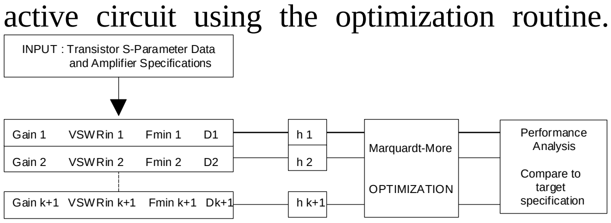

Figure 3**:** Flow chart of FREELCD program.

Once the vector h_i and the corresponding equalizer parameters e_{ij} are obtained for each stage, the overall circuit response is compared with the target specification. If the result complies with the specification, FREELCD carries on to the next stage. If not, a new set of starting values for h_i is selected in the random manner, and the optimization process repeated. Once the equalizer parameters are obtained, each stage may be synthesized from the parameter e_{ij} to a given low-pass, high-pass or bandpass specification with a lumped or distributed element topology. The manner in which the high-pass, bandpass or low-pass considerations impose following conditions on the polynomial *h(p)*

- low-pass $h_0 = 0$ and $h_n = g_n$,
- bandpass $h_0 = \pm 1$ and $h_n = g_n$,
- high-pass $h_0 = \pm 1$ and $h_n = 0$.

In all cases, the solution obtained with no approximation on transistor S_{12} is optimal. No improvement can be obtained in circuit performances by using a circuit optimization program.

III. APPLICATIONS

A. Bandpass filter design

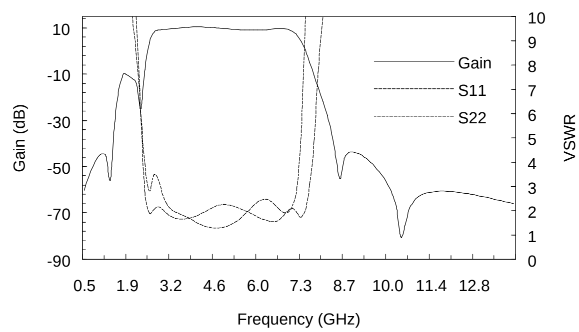

Figure 5**:** Gain and VSWRs of the bandpass filter.

This example demonstrates the use of the proposed program to design a 3-7 GHz 4-finite transmission zeros filter with lumped elements. The extraction of lumped element topologies of the ladder form using the Darlington procedure has been applied to obtain the constituent elements of two equalizers. The transducer power gain in the passband is 10.2±0.2 dB and input and output VSWRs are respectively less than 1.9 and 1.7.

B. Low-noise amplifier design

The proposed method can be applied using the Richards transformation $t=j\Omega$ to distributed commensurate transmission line extraction.

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This approach has been used to a 3-stage low-noise amplifier design. The hybrid implementation of the amplifier fabricated on a 0.635 mm thick alumina substrate is shown in Figure 6. The measured transducer power gain is 32±0.16 dB. The noise figure measured on the passband is less than 1.5 dB. The input and output return loss are less than -12.0 dB.

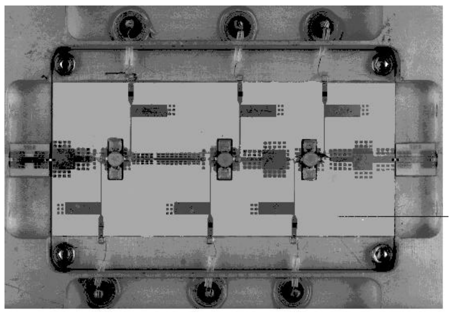

Figure 6**:** Photography of the low-noise amplifier.

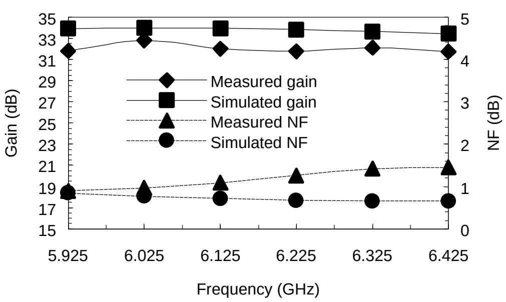

Figure 7**:** Gain and noise figure of the amplifier.

IV. CONCLUSION

The real frequency technique associated with More optimization algorithm is an original method to design microwave active circuits. This technique does not require any 2-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.

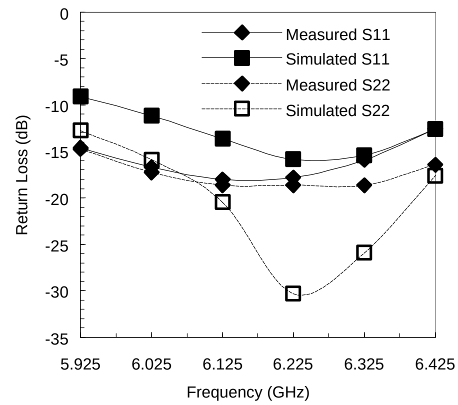

Figure 8**:** Return loss of the amplifier.

The validity and advantages of the proposed method have been demonstrated by two design examples. This procedure can be applied to design power or broadband amplifiers.

REFERENCES

[1] H. J. Carlin and J. J. Komiak, « A new method of broadband equalization applied to microwave amplifiers », *IEEE Trans. Microwave Theory Tech*., Vol. MTT-27, pp.93-99, Feb. 1979.

[2] B. S. Yarman and H. J. Carlin, "A simplified real frequency technique applied to broad-band multistage microwave amplifiers", *IEEE Trans. Microwave Theory Tech.*, Vol. MTT-30, pp. 2216-2222, Dec. 1982.

[3] P. Jarry, E. Kerherve, H. El Hendaoui and H. Tertuliano, « CAD of microwave integrated active filters by the real frequency-computer program FREELCD », *Annals of Telecommunications*, pp. 191-205, May-June 1996.

[4 J. J. More, "The Levenberg-Marquardt algorithm : implementation and theory", *Lecture notes in mathematics 630*, Springer Verlag, 1978.

[5] R. Soares, A. Perennec, A. Olomo-Ngongo, P. Jarry, « Application of a simplified real-frequency synthesis method to distributed-element amplifier design », *International Journal of Microwave and Millimeter-Wave Computer-Aided Engineering*, Vol.1, No. 4, pp. 365-378, 1991.

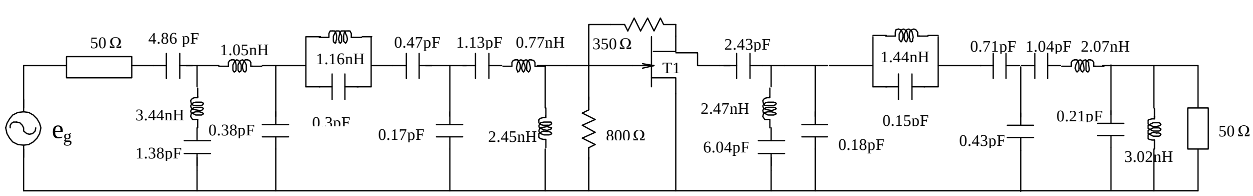

Figure 4**:** Schematic diagram of the bandpass filter.