OPTIMIZATION OF GAIN, VSWR AND NOISE OF THE BROADBAND MULTISTAGE MICROWAVE MMIC AMPLIFIER BY THE REAL FREQUENCY METHOD. SYNTHESIS IN LUMPED AND DISTRIBUTED ELEMENTS

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Abstract

Up until now the simplified "real frequency" technique has been applied to the design of broadband multistage microwave amplifiers where gain and VSWR are optimized. In this paper we extend the method to the optimization of the noise figure in parallel with the gain and VSWR. The synthesis of the networks is carried out in two ways, with lumped elements and distributed elements. Then we give several examples of design ; ultra wide band GaAs monolithic amplifier, transimpedance amplifier for optical receivers and low noise ultra wide band amplifier.

I - Introduction

In the design of broad-band multistage microwave amplifiers the double-matching real frequency method [1] has been used favourably. It is a powerful synthesis tool developed to design lossless impedance matching networks. This method utilizes directly the measured data obtained from the generator and the load networks (scattering parameters of the FET devices). Neither an a priori choice of an equalizer topology, nor an analytic form of the system transfer function, is assumed. No transformers are used in the synthetized solutions. The optimum solution of the problem is garanteed as the method is associated with a strong optimization process, such as the J. MORE routine [2] which is an improvement on the levenberg-Marquardt technique. It should be noted that the nonunilateral behavior of the transistor is taken into account.

In very high data rate fiber-optic systems, such as an optical receiver, it is necessary to have a bandwidth spanning usually three decades. But at low frequencies, microwave GaAs FETs are unstable and virtually impossible to match over large bandwidths. It is thus necessary to use either lossy matching or RL feedback to reduce terminal impedances and stabilize the transistor [3]. The GaAs MESFET is thus embedded in a network including a parallel feedback loop, a drain series inductance and a gate shunt resistance to form an elementary module termed Transistor Feedback Block (TFB). Then we extract the scattering parameters of this TFB and use the simplified real frequency method. With this approach we have designed and built a 6 MHz-6 GHz microwave amplifier in hybrid and monolithic technology (PLESSEY foundry, GB) for a wide-band fiber ptic receiver at 4.8 Gbit/s (fig. 1 and fig. 2) [3] [4]. We explicit now the simplified real frequency method; the initial version and the extension of the technique [3, 4].

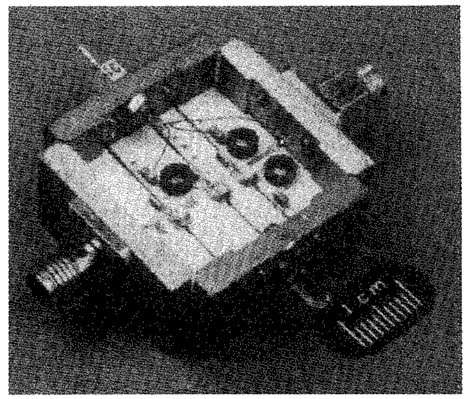

Fig. 1 - Hybrid three-stage 6MHz-6GHz amplifier - LANNION (F)

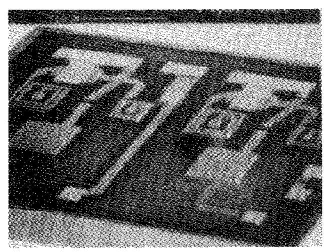

Fig. 2 - Monolithic two-stage 6MHz-7GHz amplifier - PLESSEY (GB)

II - The simplified real frequency technique

Initially, the simplified real frequency technique was introduced by YARMAN and CARLIN [1]. It is an optimum approach to broadband matching. In this section we show how the simple formalism of this method allow us, without complex calculations, to optimize many characteristics of a multi-stage microwave amplifier.

a - Formalism

It has been shown [5] that the scattering parameters of lossless reciprocal two-port E, or equalizer, can be completely determined from the numerator polynomial h(p) of the input reflexion $e_{11}(p)$. E is assumed to be a ladder network. Then the scattering parameters are given as (Belevitch representation) :

$$e_{11}(p) = h(p) \,/\, g(p)$$
$$e_{12}(p) = e_{21}(p) = \pm\, p^r \,/\, g(p) \qquad (1)$$
$$e_{22}(p) = -(-1)^r\, h(-p) \,/\, g(p)$$

where $r > 0$ is an integer and specifies the order of the zeros of transmission. The polynomial $h(p) = h_0 + h_1\, p + \dots + h_n p^n$ is chosen as unknown and the polynomial g(p) is generated from the Hurwitz factorisation of

$$g(p)\, g(-p) = h(p)\, h(-p) + (-1)^r\, p^{2r} \quad \text{(E assumed lossless)} \quad (2)$$

With these expressions we are able to describe all the characteristics of a multistage amplifier (i.e, gain, $VSWR_S$ and noise figure) using the measured scattering parameters and noise data of the FET.

b - Transducer power gain

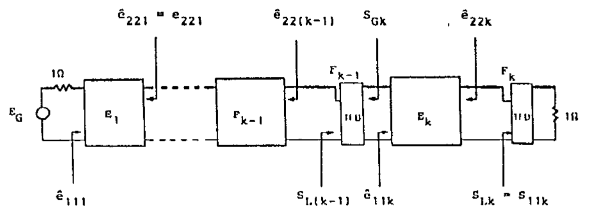

Fig. 3 - Computation steps for designing a broad-band multistage amplifier

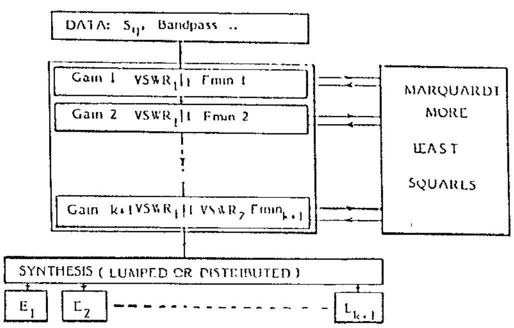

Fig. 4 - Synopsis of CAD procedure FREEL

Refering to Fig. 3 for the first k cascaded amplifier stages, the transducer power gain (TPG) is given by [1] :

$$T_k(\omega) = T_{k-1}(\omega)\, \frac{|e_{21_k}|^2 |S_{21_k}|^2}{|1 - e_{11_k} S_{G_k}|^2 |1 - \hat{e}_{22_k} S_{11_k}|^2} \qquad (3)$$

i.e. $\qquad T_k(\omega) = T_{k-1}(\omega).E_k(\omega) \qquad (4)$

where T_{k-1} : is the TPG of the first (k-1) stages with normalized resistive terminations,

$(e_{ij})_k$: scattering parameters of the kth equalizer E_k

$(S_{ij})_k$: scattering parameters of the kh TFB.

The overall gain $T(\omega)$ is defined after the final equalizer E_{k+1} has been added :

$$T(\omega) = (T_1 \cdot T_2 \dots T_k)\, E_{k+1}(\omega) \qquad (5)$$

c - Input and output VSWR

Using the same method, we are able to define the input and output VSWR. For the first k cascaded stages, the input and output VSWR are given by [3,4] :

$$R_{ink} = \frac{1 + |\hat{e}_{11\,1}|}{1 - |\hat{e}_{11\,1}|}\Big|_k \qquad (6)$$

$$R_{outk} = \frac{1 + |S_{22\,k}|}{1 - |S_{22\,k}|} \qquad (7)$$

d - Noise figure

Up until now the simplified real frequency method has been applied to the optimization of gain and VSWR characteristics. Now we show how this technique can take into account the noise parameters of the FET and optimize the noise figure of each stage. So, refering to Fig. 3, the noise figure of the amplifier at stage k is given as :

$$F_k = F_{k-1} + \frac{1}{T_{ak-1}} \left[F_{mink} + R_{nk} G_{ok} \frac{|\hat{e}_{22k} - \hat{e}_{22optk}|^2}{|1 - \hat{e}_{22optk}|^2 (1 - |\hat{e}_{22k}|)^2} - 1\right] \qquad [8]$$

where :

F_{k-1} : is the noise figure of the (k-1) first stages

$F_{min\,k}$: minimum noise figure of the kth FET

$T_{a\,k-1}$: Available gain of the (k-1) first stages

R_{nk} : noise resistance of the kth FET

G_{ok} : normalized conductance

$\hat{e}_{22k}$: reflection coefficient at the output of the equalizer E_k to the left

$\hat{e}_{22\,opt\,k}$: optimum value of the reflection coefficient $\hat{e}_{22k}$

Consequently, the noise figure of each stage can be optimized with only the polynomial **h** as the gain and $VSWR_S$.

e - Optimization of gain, input VSWR and noise figure

Here, the optimization will be performed simultaneously upon the transducer power gain, the input VSWR and the noise figure. Relating to the gain, it has been shown [5] that the convergence behavior of the simplified real frequency technique is excellent (the degree of the nonlinearity of $1/T(\omega)$ tends to be quadratic).

At stage k (fig. 3), according to the least square criterion, the objective function can be written as :

$$E_k^2 = \sum_{j=1}^{m} W_1(T_k(\omega_j)/T0_k-1)^2 + W_2(R_{ink}(\omega_j)/R0_k-1)^2 + W_3(F_k(\omega_j/F0_k-1)^2 \quad (9)$$

where $T0_k$, $R0_k$ and $F0_k$ are the desired gain, input VSWR and noise figure to be approximated, m is the number of sampling frequencies over the bandpass, W_1, W_2 and W_3 are weighting functions. At each iteration, $h_j + \Delta h_j$ are corrected to minimize the objective function using the J. MORE routine which is an improvement of the Levenberg-Marquardt algorithm. Vector $\Delta \mathbf{h}$ is given as :

$$\Delta \mathbf{h} = -[\mathbf{J}^T\mathbf{J} + \alpha \mathbf{D}^T\mathbf{D}]^{-1} \mathbf{J}^T \mathbf{e}_o \quad (10)$$

where :

- $\mathbf{e}_o$: is the error vector
- $\mathbf{J}$: the Jacobian Matrix of $\mathbf{e}$ (where the elements are : $\partial e_j / \partial h_i$; j = 1... m ; i = 1... n)
- $\mathbf{D}$: a diagonal matrix which take into account the scaling of the problem
- α : the Levenberg-Marquardt parameter.

J. MORE introduced relationships between $\mathbf{J}$, $\mathbf{D}$ and α which permit rapid convergence, and this without initial guess on the vector $\mathbf{h}$.

f - Design with transmission line elements

The real frequency technique presents results in lumped elements in the matching circuits. However, it is able to proceed directly to the realization of the equalizers made from transmission lines by using Richards's transformation for the frequency variable

$$t = j\,\Omega = j \tan \omega \tau \quad (11)$$

where τ is the delay length of the commensurate transmission line. All the approximations are now calculated in the transformed Ω domain, but otherwise, the real frequency procedure remains basically unchanged [6]. Here the real normalized scattering parameters of E are given as :

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

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

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

where q is the number of parallel stubs.

Fig. 4 shows the different stages of the CAD procedure associated to the method namely FREEL. The results obtained are in concordance with the commercially available computer programs.

III - Examples of design

1) MMIC Amplifier 50 Ω - 50 Ω. 1 MHz - 10 GHz

We have already designed and built an amplifier 50 Ω - 50 Ω 6 MHz - 6 GHz [3], [4] for an optical receiver at 4.8 Gbit/s. For a similar project at the CNET of LANNION (France) we have designed a 50 Ω - 50 Ω, 1 MHz - 10 GHz for an optical receiver at 9.8 Gbit/s. It is a three stage monolithic amplifier which will be carried out at the GaAs ANADIGICS foundry (USA). In this circuit, diodes permit decoupling bias at low frequencies (fig. 5).

2) Transimpedance amplifier

An optical receiver consists of a photodetector and a low-noise amplifier (transimpedance). A photodiode is assumed to be a high impedance current source with a shunt capacitance. The transimpedance amplifier is placed before the post-amplifier (ex : amplifier 1 MHz - 10 GHz design above). We give an example of transimpedance design (200 Ω - 50 Ω; 0.3 - 2.5 GHz) which will be carried out in hybrid technology with the NEC 32000 transistor (fig. 6). Gain, input VSWR and noise figure are optimized simultaneously. Noise figure is less than 2 over 1 GHz to 2.5 GHz.

3) Ultra-wide band low noise amplifier

The latest developement in GaAsFET technology (low noise HEMT transistor) permits a higher break frequency amplifier, fig. 7 shows the design of an ultra-wide band (0.1 GHz - 30 GHz) low-noise amplifier (transistor FHX02X Fujitsu) with distributed elements. We obtained a good flatness of the gain and $VSWR_S$ less than 2.

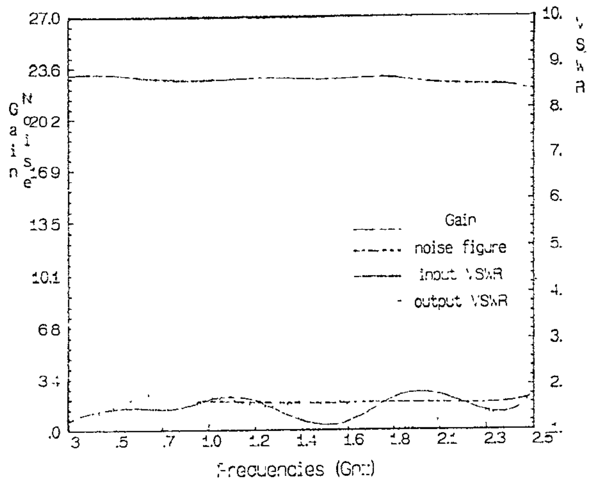

Fig. 6 - 0.3 - 2.5 GHz Transimpedance

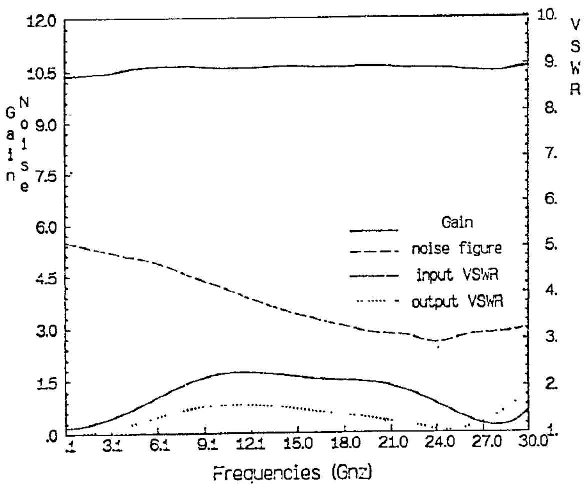

Fig. 7 - 0.1 - 30 GHz two-stage amplifier

IV - Conclusion

The double-matching real frequency method described in this paper is particularly useful in practical design problems such as broad-band microwave amplifiers. After the VSWR optimization, we have included noise in the method, and optimized it using the same routine. Loads and devices need only be specified by empirical data, and since an a priori choice of an equalizer topology is not assumed, the technique has wide flexibility in its range of application. Furthermore this technique can be applied to other devices where broad-band matching is important, such as antennas or active filters.

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