OPTIMIZATION AND STABILITY ANALYSIS TECHNIQUES IN MICROWAVE GaAs FET OSCILLATORS FOR MMIC APPLICATIONS

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ABSTRACT

This paper tries to give some answers about the analysis and optimization techniques in dealing with microwave GaAs FET oscillators for MMIC applications. The methods treated here are based on multiharmonic formulation in order to take into account the best approach to the real waveform. Some experimental results are presented in order to validate the general approach.

INTRODUCTION

During recent years, the high frequency non linear circuits have obtained great importance in communication systems and furthermore, there has been considerable development of these techniques for MMIC applications. Fortunately there are several commercial programs for developing microwave nonlinear functions that aid strongly to the designer. But all of these software programs are not able to optimize a nonliner circuit and so to obtain the better performances. Furthermore, intrinsic problems such as large signal stability, quality of oscillation, noise, etc, are not treatet at all.

Taking into account the above considerations, we have developed a complete analysis and optimization program based upon the harmonic balance technique lied to the conjugate gradient method, by using appropriate GaAs-MESFET models.

In order to validate the theoretical model and the simulated optimization, we have measured a MGF-1802 GaAs-MESFET transistor, and these results have been used to design and construct a X-band varactor tuned oscillator circuit. We have observed a good agreement with the experimental results.

The harmonic optimization procedure is quite simple: given a nonlinear active device, the external voltages and currents can be optimized independently of the embedding circuit [2], [3]. The synthesis of the linear circuit may be performed a posteriori.

HARMONIC OPTIMIZATION

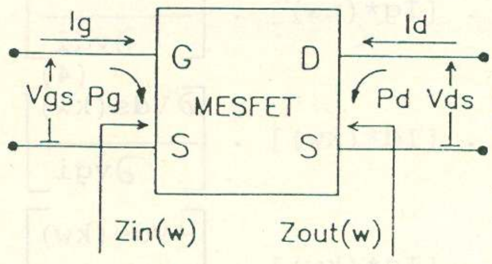

Fig.1

Figure 1 represents the nonlinear two-port device. Knowing the internal functions given by [1], [2], we can obtain the external voltages and currents given by (1). Therefore, we have 2(2N+1) independent variables, where N is the highest order of the Fourier development.

$$vgs(t) = Re\sum_{k=o}^{N} [Zpg(kw)Igk+Zg(kw)Igintk+Vfk-Zs(kw)Isintk].exp(jkwt) \quad (1-a)$$

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$$vds(t) = Re\sum_{k=o}^{N} [Zpd(kw)Idk+Zd(kw)Idintk+Vdik-Zs(kw)Isintk].exp(jkwt) \quad (1-b)$$

$$ig(t)=Re\sum_{k=o}^{N} \frac{[Zpgs(kw)Igintk+Zg(kw)Igintk+ Vfk-Zs(kw)Isintk]}{Zpgs(kw)}.exp(jkwt) \quad (1-c)$$

$$id(t)=Re\sum_{k=o}^{N} \frac{[Zpds(kw)Idintk+Zd(kw)Idintk+Vdik-Zs(kw)Isintk]}{Zpds(kw)}.exp(jkwt) \quad (1-d)$$

Note that the equations (1) present a problem when the package capacitors Cpgs and Cpds are not implemented. In this case:

$$\lim_{Zpgs \to \infty} [ig(t)] = igint(t)$$
$$\lim_{Zpds \to \infty} [id(t)] = idint(t) \quad (2)$$

The problem is to know the power and the power gradient at the gate and drain ports in order to optimize the fundamental added power and the harmonic content functions. The power expressions are:

$$Pg(nw) = 1/2.Re[Vgsn.Ign^*] = 1/2.[Vgsnr.Ignr+Vgsni.Igni]$$
$$Pd(nw) = 1/2.Re[Vdsn.Idn^*] = 1/2.[Vdsnr.Idnr+Vdsni.Idni] \quad (3)$$

where Pg(nw) and Pd(nw) are the n-harmonic dissipated power.

The equations (1) show that the expressions (3) are functions of each harmonic component of the independent variables, and every harmonic power will need 2(2N+1) derivatives to solve the optimization problem by using the conjugated gradient method. Knowing the Jacobian matrix defined by [2] and the expressions (1), and taking into account the row vectors given in [2], we can obtain two row vectors for the power gradient at gate port and the equivalent two row vectors at drain port given by (4).

$$\left[\frac{\partial Pg(nw)}{\partial vgi}\right] = 1/2 \;.\; [Vgs(nw)] \;.\; \left[\frac{\partial Ig^*(nw)}{\partial vgi}\right] + 1/2 \;.\; [Ig^*(nw)] \;.\; \left[\frac{\partial Vgs(nw)}{\partial vgi}\right]$$

$$\left[\frac{\partial Pg(nw)}{\partial vdi}\right] = 1/2 \;.\; [Vgs(nw)] \;.\; \left[\frac{\partial Ig^*(nw)}{\partial vdi}\right] + 1/2 \;.\; [Ig^*(nw)] \;.\; \left[\frac{\partial Vgs(nw)}{\partial vdi}\right] \quad (4)$$

$$\left[\frac{\partial Pd(kw)}{\partial vgi}\right] = 1/2 \;.\; [Vds(kw)] \;.\; \left[\frac{\partial Id^*(kw)}{\partial vgi}\right] + 1/2 \;.\; [Id^*(kw)] \;.\; \left[\frac{\partial Vds(kw)}{\partial vgi}\right]$$

$$\left[\frac{\partial Pd(kw)}{\partial vdi}\right] = 1/2 \;.\; [Vds(kw)] \;.\; \left[\frac{\partial Id^*(kw)}{\partial vdi}\right] + 1/2 \;.\; [Id^*(kw)] \;.\; \left[\frac{\partial Vds(kw)}{\partial vdi}\right]$$

Typically, the main function to be optimized in amplifier and oscillator circuits is the maximum added power [2], [3], [4] at the fundamental. However, the harmonic content is an effect that degrades the microwave behavior. We will take into account N harmonics in order to minimize the harmonic content:

$$Padd = - Pd(w) - P(w) = maximum \quad (5)$$

with

$$\sum_{n=2}^{N} Padd(nw) = \sum_{n=2}^{N} - [Pd(nw) + Pg(nw)] = minimum \qquad (6)$$

and the gradient is given by (7).

$$\left[\frac{Padd(w)}{vgi}\right] = - \left[\frac{Pd(w)}{vgi}\right] - \left[\frac{Pg(w)}{vgi}\right]$$

$$\left[\frac{Padd(w)}{vdi}\right] = - \left[\frac{Pd(w)}{vdi}\right] - \left[\frac{Pd(w)}{vdi}\right]$$

$$\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots \qquad (7)$$

$$\left[\frac{Padd(nw)}{vgi}\right] = - \left[\frac{Pd(nw)}{vgi}\right] - \left[\frac{Pg(nw)}{vgi}\right]$$

$$\left[\frac{Padd(nw)}{vdi}\right] = - \left[\frac{Pd(nw)}{vdi}\right] - \left[\frac{Pg(nw)}{vdi}\right]$$

When we considere N harmonics we have 4N+2 independent variables and the expresions (7) are row vectors of 2N+1 elements.

At the last, the constraints are a very important aspect in optimization process if one want to increase the speed of the calculation. This work employs three different groups of constrains [2]: the dc balance for each bias point, the stability equations in the circuit and the non power dissipation concept for superior harmonics.

Note that this formulation is valid, not only for first harmonic oscillators and amplifiers, but also for multipliers and dividers. In these last cases, we will consider different harmonic number at the gate and drain ports in the equation (5).

These concepts will be applied to optimize a varactor tuned oscillator that use a comercially available varactor diode which limit the electronic tuning. Figure 2 shows a typical equivalent circuit of a varactor diode which has an impedance Zv and the total circuit to be optimized is given in figure 3.

Taking into account the nonlinear model proposed by [1] and figure 3, the expressions (1) we can write:

$$Vv(t) = Re \sum_{k=0}^{N} [Igk.Zv(kw)+Idk.Zv(kw)] = Re \sum_{k=0}^{N} Vvk.exp(jkwt) \qquad (8)$$

On the other hand, known the voltage at the varactor and the drain current, we can calculate the load impedance:

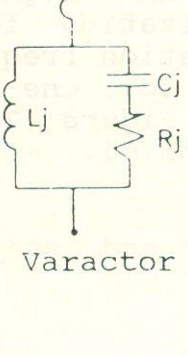

Fig.2 - Varactor diode

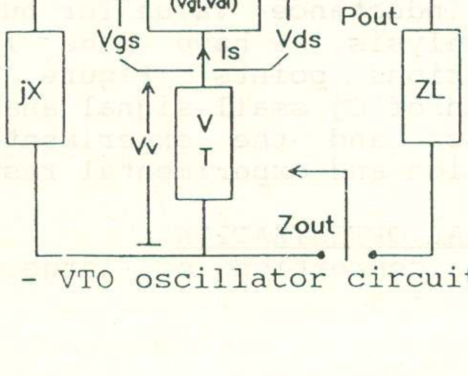

Fig.3 - VTO oscillator circuit

$$ZL = - \sum_{k=0}^{N} \frac{Vvk + Vdsk}{Idk} \qquad (9)$$

In this case, the problem is to know the optimum load impedance at each frequency, for a given value of Lg, in order to obtain the maximum output power at the desired harmonic along with, a minimum harmonic content. We can write:

$$Pout(nw) = - 1/2 \; . \; Re[VdsnIdn^* + Zv(nw) IgnIdn^* + Zv(nw).Idn.Idn^*] \qquad (10)$$

The method used here needs the knowledge of the power gradient in analytical form given by (4) and (7), we can define two row matrix for the harmonic n.

$$\left[\frac{Pout(nw)}{vgi}\right] = - 1/2.Re \; [Id^*(nw)].[\left[\frac{Vds(nw)}{vgi}\right] + Zv(n\;). \left[\frac{Ig(nw)}{vgi}\right] +$$

$$+ Zv(nw). \left[\frac{Id(nw)}{vgi}\right]] + [[Vds(nw)] + Zv(nw) \; [Ig(nw)]$$

$$+ Zv(nw)[Id(nw)]]. \left[\frac{Id^*(nw)}{vgi}\right] \qquad (11)$$

We can make the derivative expression with respect to vdi in analogous form.

The voltages and currents of (11) are rows vectors given by [2] of 2N+1 elements where only the element of order n is different of zero and the derivatives with respect to Vgi and Vdi are matrix of (2N+1)*(2N+1) elements with known gradient. Also, the equation (11) shows that it is necessary to know the harmonic Jacobians of the gate and drain currents and the voltage Vds in analytical form.

Once that we know the output power expressions and the analytical gradient, we can calculate for each frequency the optimization functions, they will be:

$$M - Pout(w) = minimum \quad and \quad \sum_{n=2}^{N} Pout(nw) = minimum \qquad (12)$$

where M is a upper bound of the output power.

A very important aspect in optimization process, in order to increase the speed of the simulation, is the calculation of the constraints. The constraints used in this paper are given by[2].

The experimental results and the nonlinear simulations are a good agreement. The device used is a commercial MGF-1802 GaAs-MESFET transistor and the equivalent circuit model is given in [1]. The varactor diode exibits a junction capacitance ratio of aproximately ten. This fact will limit the bandwidth between 6 to 9 GHz. In the following step, we have made a small signal analysis of Zout of the VTO structure (Fig. 3) for several values of Lg. The better inductance value for our pourposes is Lg = 1nH. Right after this analysis, we have made large signal optimization for each oscillations points. Figure 6 shows the oscillation frequencies function of Cj small signal analysis, optimization for one and two harmonics and the experimental verifications. Figure 7 shows simulation and experimental results of Pout function Cj.

GRAPHICAL OPTIMIZATION

The conception of large signal amplifiers and oscillators operating under nonlinear saturated conditions needs sophisticated computer programs in order to treat all the nonlinear problems of optimization[5]. However, we always need a simple and accurate starting point if we want to arrive to the correct final solution.

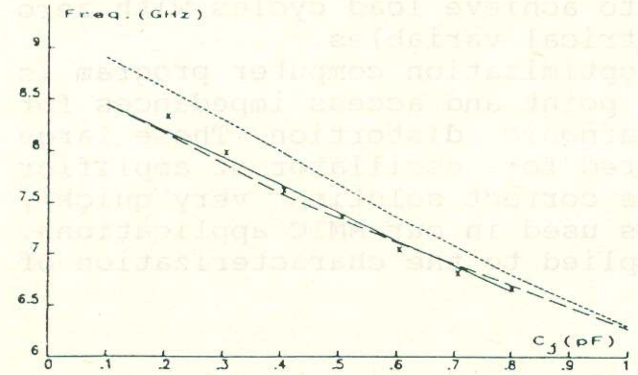

Figure 6 - Frequency function of C_j. .... Small signal results. ---- 1 harmonic optimization. —— 2 harmonic optimization. xxxx experimental results.

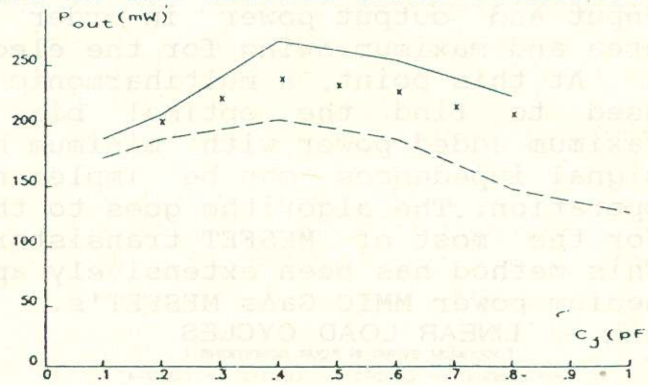

Figure 7 - Output power function of C_j. ---- 1 harmonic optimization. —— 2 harmonic optimization. xxxx experimental results.

Fortunately many solutions, most of them starting from linear knowledge of a particular topology, have been reported in the past. We present here a simple graphical method to obtain the optimum bias point and load impedances of a given FET to be applied to the design of optimal MMIC oscillators as well as amplifiers. The method is based on the draw of the load cycles relating the drain voltage and current.

In order to give a simplified approach to this method, we start saying that the drain current of a given FET, apart the intrinsic resistances, can be written as:

$$Ids(t) = F[Vgs(t-\tau), Vds(t)] \tag{13}$$

In the frequency domain, and under a linearized form, we can write:

$$Ids(w) = gm.\exp(-jw\tau).Vgs(w) + gd.Vds(w) \tag{14}$$

where gm and gd are the derivatives versus Vgs and Vds respectively. The output power for this formulation is:

$$Pout = 1/2.Real[Vds.Ids^*] \tag{15}$$

Taking the phase reference for Vds, Vgs should be delayed by an amount of ϕ degrees. In this case:

$$Pout = 1/2.Real[gm.Vgs.Vds.\exp(-jw\tau+j\phi) + gd.Vds2] \tag{16}$$

The output power will be a maximum if:

$$w\tau - \phi = (2K-1)\pi \qquad K=1,2,\ldots \tag{17}$$

or Vgs and Vds in opposition. From a graphical point of view that means that the curve representing the load cycle, in the Ids versus Vds plane, must be with zero surface.

With these ideas in mind, and supposing that we have an accurate nonlinear model for the transistor, we use a conventional linear program to obtain the source and load impedances for optimum gain conversion (MAG or MSG) in the broadband of interest. Because the limiting effects such as breakdown and gate conduction, the maximum swing for the Vgs and Vds voltages is known. For every bias point, and starting from the above linear impedances, we can use a nonlinear program to draw the true load cycles as a function of input and output power in order to achieve load cycles with zero area and maximum swing for the electrical variables.

At this point, a multiharmonic optimization computer program is used to find the optimal bias point and access impedances for maximum added power with minimum harmonic distortion. These large signal impedances can be implemented for oscillator or amplifier operation. The algorithm goes to the correct solution very quickly for the most of MESFET transistors used in our MMIC applications. This method has been extensively applied to the characterization of medium power MMIC GaAs MESFET's.

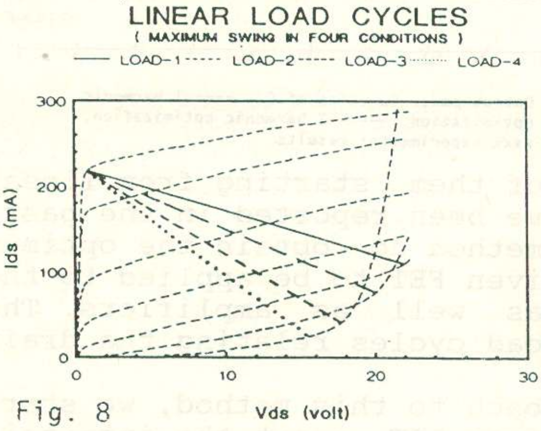

Fig. 8

Fig.(8) shows the linear load cycles for different conditions of load impedances, while Fig.(9) shows the same contours in nonlinear saturated conditions at a given frequency. Fig.(10) shows the added power versus input power for the four cases. It is easy to see that the four load cycles are optimum in zero area and maximum swing, but only one gives the maximum added power.

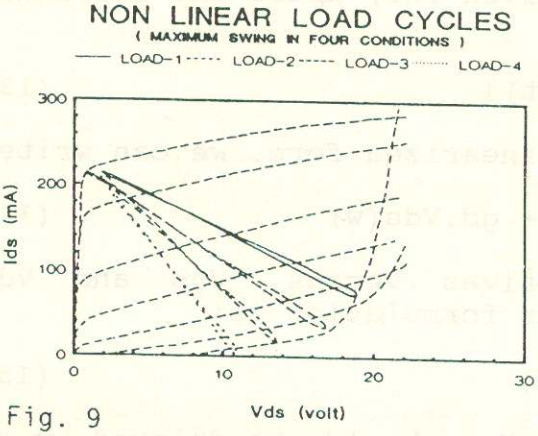

Fig. 9

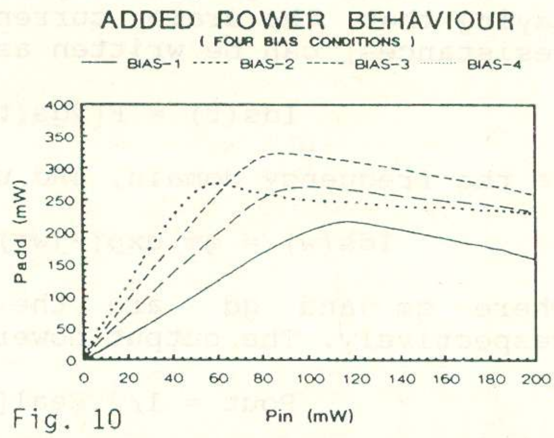

Fig. 10

NONLINEAR STABILITY

Most of the negative resistance oscillators appearing in the bibliography [7] are referred to oscillators using two terminal active devices, such as Gunn or Impatt diodes, or sometimes to three terminals devices but converted into two terminals when considering only one opening port. When trying to determine the stability of the latter, the above conversion often produces different results, even contradictory, depending on the considered opening port.

But In an N-port active device, N independent temporal variables should be considered, being an appropiate choice the controlling voltages of the active device. Between the N+1 external terminals of the circuit, connected to generator resistors, and ground, N+1 opening ports may be defined, though only N are necessary, satisfying Kurokawa oscillation condition [6] at each of the M harmonic frequencies considered. NxM transfer functions, equal to zero, are thus defined, by using Y-formalism:

$$H^i_k(w, V^1_1,...,V^1_M,...,V^N_1,...,V^N_M, \phi^1_1,...,\phi^1_M,...,\phi^N_1,...,\phi^N_M)= 0 \quad (18)$$

with i = 1...N and k = 1...M. The superindexes indicate the port number and the subindexes, the harmonic number. Small perturbations about the operating point are now considered,

so the defined transfer functions become: $H^i_k = H^i_{k0} + dH^i_k$

Since the perturbations are small quantities, the new transfer functions can be expanded in a Taylor series about the operating point, considering only the first order term. The resulting equations can be expressed matrixially:

$$\left[V^i_k \frac{\partial H^1_m}{\partial V^i_k} \right] \frac{\delta \vec{V}^i_k}{V^i_k} + \left[\frac{\partial H^1_m}{\partial \phi^i_k} \right] \delta \vec{\phi}^i_k + \left[\frac{dH^1_m}{dw_1} \right] \delta\omega_1 = 0 \quad (19)$$

where $\delta V^i_k/V^i_k$ and $\delta\phi^i_k$ are the perturbations vectors.

Since pertubations are small and all the frequencies are harmonically related [1]:

$$\delta w^i_k = k\,\delta w^i_1 \quad \text{or} \quad \delta\dot{\phi}^i_k - j\,\delta\dot{v}^i_k = k\,(\,\delta\dot{\phi}^i_1 - j\,\delta\dot{v}^i_1\,) \qquad k = 1,M$$

As well, after the quasistatic assumption, the instantaneous frequency does not depend on the measuring point of the circuit.

$$\delta w^i_k = \delta w^1_k \quad \text{or} \quad \delta\dot{\phi}^i_k - j\,\delta\dot{v}^i_k = (\delta\dot{\phi}^1_1 - j\,\delta\dot{v}^1_1\,) \qquad i = 1,N$$

Taking into account these relations we obtain an equations system which can be easily solved for one of the unknowns as a function of another, so an exclusive relation between δv and its time derivative can be obtained. Such a relation will be of the kind: $\delta\dot{v}^1_1 + S\delta v^1_1 = 0$. The stability condition will be S>0

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