A New non linear optimization for microwave active circuits using a selective invariant transform

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Abstract- The main objective of this paper is to show one way of optimization, analysis and design of microwave active circuits, based on the transformation of the reflection coefficient for an amplifier or a microwave oscillator. This improved method is based on a non linear optimization of the circuit. This way consists on introducing a transformation invariant which allows to test the appropriate transistor for a given frequency bandwidth.

The transformation invariant is called coefficient *N*, and it determines very quickly, the possible modifications of the active multiport network by external circuits connected to it. This procedure is applied by looking at the microwave transistor as a three- port or as an equivalent two-port. The present method appears to be accurate and fast in the design and optimization of a microwave oscillator or an amplifier based on H.F transistors.

***Index Terms*. Microwave oscillator, Optimization, Computer aided design, H.F transistor.**

I. INTRODUCTION

The last four decades have been marked by rapid developments in RF and microwave circuits and by their wide application in a variety of commercial and military systems. Examples of systems in which such integrated circuits have been applied range from cellular telephones and microwave links on the commercial side to missile and electronic warfare systems on the military side, with many dual-use types of applications such as in radar and navigation systems. During the past three decades, several robust optimization techniques have been developed. These techniques supplied designers with strong and reliable tools necessary for the complex and demanding needs of modern circuit design. They utilize the circuit responses and possibly derivative information in the optimization loop.

Recently, commercial software packages, for example, have been developed that does solve Maxwell's equations for circuits of arbitrary geometrical shapes. Such simulators are denoted as Electromagnetic methods. The active elements are represented by non linear one-port devices such as diodes and feedback transistors [1].

For nearly a half a century computer-aided design (CAD) of electronic circuits have evolved from a set of special purpose, rudimentary simulators and techniques to a variety of highly flexible and interactive, general purpose software systems, with amazing visualization capabilities [2].

The big progress in the microwave semiconductor components technology allows more easily the realization of microwave sources with some transistors that are easily available.

By taking into consideration the elementary circuits theory and by applying the condition of quasi-linear oscillation, we can design tuned electrically oscillators.

The choice of a transistor is a very important parameter. In this paper we suggest a new technique that consists in optimizing and choosing the most appropriate high parameter for the large band microwave oscillator circuit [3].

The method consists for example the optimization of microwave sources.

It is based on the introduction of a parameter of homographic transformation that we call it coefficient of level *N*.

This criterion is simple and very quick. It is based on the coefficient of level *N*, on the modification *M* and on the load coefficient *W*.

The N coefficient is a transformation invariant by any active multi-port [4].

II. NECESSITY OF THE NO LINEAR OPTIMIZATION

When we analyze an active circuit, many natural limitations and some other technological constraints make the practical realization very difficult.
This is due the fact that analysis is always subject to compromise between conditions sometimes contradictory.
We can classify these conditions as follows:

1- It is almost always difficult to obtain values of reflection coefficients greater than one when the frequency increases.
2- If we adjust correctly the varactor, we can modify the variation of the area of the reflection coefficient phase, by coupling more or less the varactor and the circuit that is connected to it.
3- We observe a transistor reflectance decrease when we increase the frequency, although the variation of the variation of the diapason increases with the bandwidth.
4- When the coupling coefficient increases, we observe some losses in the reflection coefficient. In almost all cases, the absolute value of this coefficient decreases to stop the oscillation process or the amplification process.

III. ACTIVE CIRCUIT REPRESENTATION

The most important component of the microwave amplifier or the microwave oscillator is the active circuit. It can be designed by using the elementary circuits theory and by applying the quasi-linear oscillation condition.
A simple representation of the active circuit is given in Fig. 1 [5].

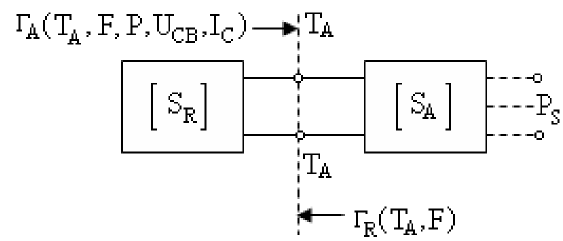

Fig.1. Circuit of the electrically tuned active circuit

The active circuit is the H.F transistor. In this circuit and at the level of reference plane that is chosen, we have a port by which the transistor decreases the losses of the connected circuit. The reflection coefficient Γ_A has to be greater than one.
It is a non linear coefficient [6] that depends on:

1- The source (V_{cb}, I_c).
2- The generated power by the circuit (Ps)
3- The frequency f.
4- The value of Γ for small signals denoted Γ_{A0}.

The adjustment circuit is defined by its coefficient Γ_R that depends on the frequency.
The oscillation condition can be written as follows:

$$\Gamma_A(V_{CB}, I_C, T_A, P, f).\Gamma_R(T_A, f) = 1 \qquad (1)$$

By introducing the retreat function $R\ (P_S)$ we can establish the oscillation as follows:

$$|\Gamma_A|.|\Gamma_R| = \frac{1}{|R(P_S)|} \qquad (2)$$

Introducing the phase's condition:

$$\theta_{A0} + \theta_R = -\varphi(R(P_S)) \qquad (3)$$

We have:

$$R(P_S) = \frac{\Gamma_A}{\Gamma_{A0}} \qquad (4)$$

$$R(P_S) = \left|R(P_{(S)})\right|.e^{j\phi_R} \quad (5)$$

$$\Gamma_R = \left|\Gamma_R\right|.e^{j\theta_R} \quad (6)$$

$$\Gamma_{A0} = \left|\Gamma_{A0}\right|.e^{j\theta_{A0}} \quad (7)$$

P_S is the output power.

These conditions are always verified at the stable point of oscillation [7],[8],[9].
The stability factor should be less than unity for any possibility of oscillation. If this condition is not satisfied, either the common terminal should be changed or positive feedback should be added. Next, the passive terminations must be added to resonate the input and output ports at the frequency of oscillation.
The microwave transistors used germanium and the silicon. The choice of transistors is between the silicon BJT and the gallium arsenide MESFET. A comparison of these devices permits to say that Silicon has another important advantage, a lower flicker corner frequency, which is important for oscillator applications [10].

IV. SCALE COEFFICIENT AND ACTIVE CIRCUIT MODIFICATION

The relation that shows this modification and that links the two reflection coefficient depends on the homographic transformation between the two parameters newly introduced and that are the modification coefficient and the load coefficient [11].

The modification is the variation of a circuit parameter in a chosen direction. To understand this, we compare many values of level coefficients that are obtained for either the same structure of the circuit, or for many structure based on different transistor *(FET* or *Bipolar*). We consider the active circuit as a two-port between the T_A and T_M planes. T_A is the connection plane showed in Fig. 2.

The relation between Γ_A and Γ_M:

$$\Gamma_A = S_{AA} + \frac{S_{AM}.S_{MA}\Gamma_A}{1 - S_{MM}\Gamma_M} \quad (8)$$

Another possible written form is:

$$\Gamma_A = S_{AA}.(1 + N.M) \quad (9)$$

Where:

$$N = \frac{S_{AM}.S_{MA}}{S_{AA}.S_{MM}} \quad (10)$$

is the level coefficient and

$$M = \frac{W}{W-1} \quad (11)$$

is the load coefficient and

$$W = S_{MM}.\Gamma_M \quad (12)$$

is the modification coefficient

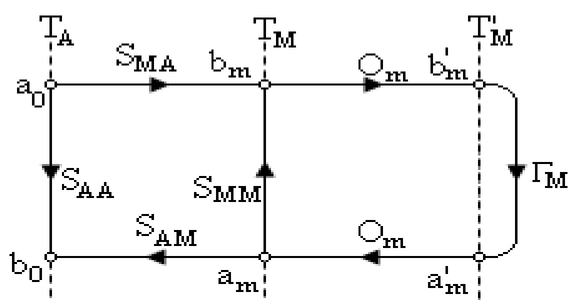

Fig.2. Fluency graph issued from the transformation of Γ_M to Γ_A

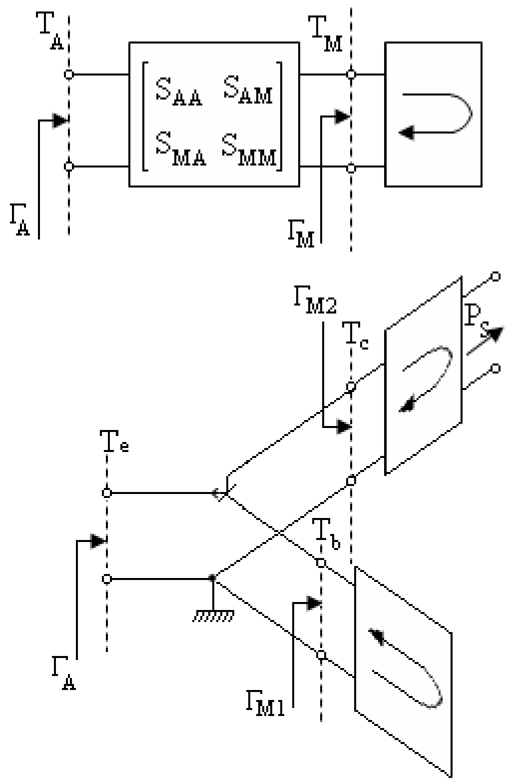

Fig.3. Active circuit modification by external circuits

This technique uses the transistor as the three-port or an equivalent two-port and using the modification theory represented in Fig. 3. [12], [13].

The main principle of this technique is to consider the transistor as an equivalent two or three-port. And by using the modification theory based on the introduction of the level's coefficient *N*, we can show that some configurations are more appropriate than some others. We present also the Smith chart that shows the transformation *M* and the modification and reflection coefficients shape in the active plane of the oscillator and this over a large wide band of frequencies.

To understand the modification of the transistor, we compare the three measured scale coefficients N_{eb}, N_{ec}, N_{bc} for the same transistor taken as a three-port *(HP 35820A).*

The three-port matrices are useful for converting CE to CB and CC configurations. This is most easily done using the three-port *Y* parameters, since the three-port and two-port *Y* parameters are measured to be the same. If we arbitrarily label the emitter port 1, the collector port 2, and the emitter port 3, we may easily show the common-emitter.

In Fig. 4, the best possibility to modify this transistor is the one taken between the base and the emitter.

By comparing the bipolar transistor reflectance taken as an equivalent two-port, we remark that the biggest value of this reflectance is issued from the B-C configuration in Fig. 5.

An additional advantage of the transfer matrix is revealed in cascaded structures, where the matrix of the cascaded connection can be determined by simply multiplying the transfer matrices of its components. [14], [15].

The disadvantage of using wave matrices is that there is no information about input structure and processes because the matrix describes characteristics of outside parameters only.

A frequency resonance condition simply requires the circuit imaginary term be zero.

If the impedance resonance is on the left-hand real axis, this is a series resonance; that is, at frequencies above resonance the impedance is inductive and below resonance the impedance is capacitive. If the impedance resonance is on the right-hand real axis, the resonance is a parallel resonance; that is, at frequencies above resonance the impedance is capacitive and below resonance the impedance is inductive.

These circuits are investigated in the steady state, and a frequency domain is applied. However, some questions for the best configuration are better answered by a time domain analysis. The investigated circuit is then regarded as being quasi-linear because the time domain investigation of non linear circuits cannot normally be carried out with analytical functions. [16], [17], [18].

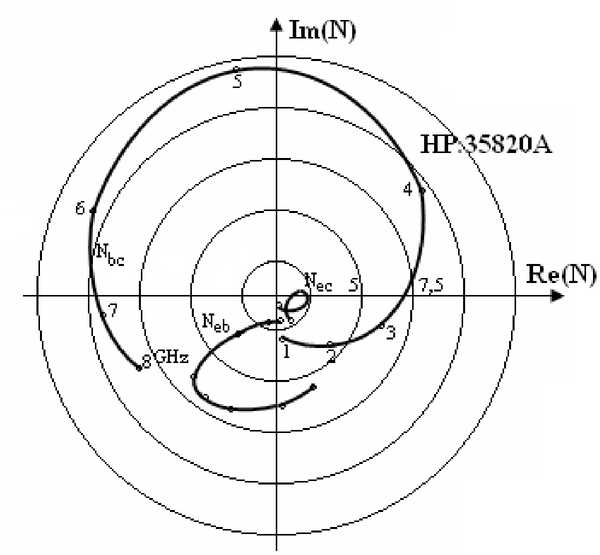

Fig. 4. Comparison between 3 level coefficients of transistor HP 35830A

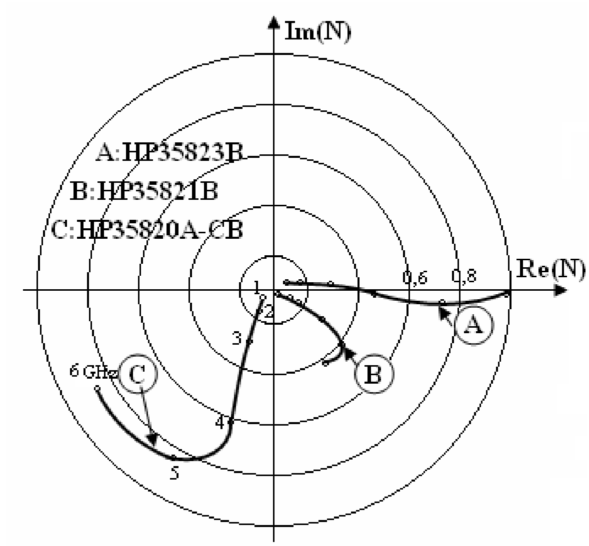

Fig. 5. Comparison between bipolar scale coefficients

V. MEASURMENT AND TEST

The elements of a scattering matrix are relatively easy to measure experimentally. *S*-parameters are related to the power incident on and waves reflected from network ports. In the microwave range, besides frequency we can measure only power and reflection coefficients (module and phase) [19].

These measurements are equivalent to measurements of incident and reflected waves. The *S* parameters related to the terminal voltages and currents cannot be measured because appropriate microwave test equipment does not exist, and also because broadband short and open circuits are very difficult to achieve [20], [21].

The principle of conservation of energy for a given device is easily expressed through the unitary condition of the scattering matrix; it is not easy to verify this condition when *Z* and *Y* matrices are used.

When reference planes of networks are changed, only the phases of scattering matrix elements are altered. By contrast, *Z* and *Y* elements change both in module and phase.

An additional advantage of the transfer matrix is revealed in cascaded structures, where the matrix of the cascaded connection can be determined by simply multiplying the transfer matrices of its components.

The disadvantage of using wave matrices is that there is no information about input structure and processes because the matrix describes characteristics of outside parameters only.

Extending the foregoing small-signal model for nonlinear operation can be done in a variety of ways. The most obvious is to base a new circuit model on large-signal *S* parameters. This can easily be accomplished by measuring device *S* parameters at elevated power levels. The power level is usually chosen to correspond to the level encountered in the final circuit application. However, there are some negative aspects of using large-signal *S* parameters. First, as we have learned, *S* parameters are defined in a linear *n*-port system with constant load impedance. Under large-signal conditions, the microwave *n*-port is not linear and large-signal *S* parameters cannot predict device performance for load impedances other than the one used during measurement [22], [23]. Also, the value obtained for *S*22 is not the conjugate of the optimum load impedance. Having established the oscillation conditions, we now apply the equivalent-circuit elements and the nonlinear equations from the dc (*I–V)* simulation to determine the oscillator properties under steady-state oscillation conditions for the permissible range of load terminating conditions. The result of the nonlinear analysis is shown in Figure 4.
Shown are closed constant-output power contours (in dBm) as a function of load (drain) terminating conditions. Also shown are

intersecting loci of constant-frequency contours. For example, the 7.5-GHz contour shows the predicted power output at various terminating admittance levels. The power levels indicated are in the range obtained experimentally, as shown by the measured data [25],[26].

To alleviate the limitations noted above, the active elements must be characterized as a function of terminating impedance as well as a function of drive level. This technique, commonly called "load pull," has the advantage that the FET is operated. It is evident from the discussion above that a numerical model with the efficiency of a frequency-domain linear solution is required for efficient interactive nonlinear design. Several such modeling approaches, commonly referred to as "harmonic balance" methods, are solved in the frequency domain but employ time-domain descriptions for the active-element nonlinearities. Thus nonlinear behavior of the total microwave circuit is obtained. Development of a CAD model of this type is relatively straightforward and begins with the linear FET model described previously [27], [28], [29].

Two-port oscillator design may be summarized as follows:

1. Select a microwave transistor with level coefficient, sufficient gain and output power capability for the frequency of operation. This may be based on oscillator data sheets, amplifier performance, or *S*-parameter calculation.
2. Select a topology that gives a correct *N* at the operating frequency.
3. Select an output load matching circuit that gives $|S11|>1$ over the desired frequency range. In the simplest case this could be a 50 Ω load.
4. Resonate the input port with a lossless termination so that $\Gamma_A = 1$. The value of $S22$ will be greater than unity with the input properly resonated.

In all cases the transistor delivers power to a load and the input of the transistor [30], [31].

Practical considerations of realizability and dc biasing will determine the best design.

For both bipolar and FET oscillators, a common topology is common base or common gate, since a common-lead inductance can be used to raise $S22$ to a large value, usually greater than unity even with a 50 Ω generator resistor. However, it is not necessary for the transistor $S22$ to be greater than unity, since the 50 Ω generator is not present in the oscillator design. The requirement for oscillation is $S11|< 1$; then resonating the input with a lossless termination will provide that $|S22|>1$. [32], [33].

A simple example will clarify the design procedure. A common-base bipolar transistor (H35830A) was selected to design a fixed-tuned oscillator at 5 GHz. [34], [35],[36].

VI. CONCLUSION

Another critical point in the algorithm is the choice of the first guess to choose *N*. A well chosen first guess of the best value of *N* will considerably ease the convergence of the algorithm to the correct solution. If the circuit is mildly nonlinear, the linear solution, obtained for a low-level input, will probably be a good first guess for a given *N* and *M*. If the circuit is driven into strong non-linearity, a continuation method will probably be the best approach. The level coefficient of the input signal is first reduced to a quasi-linear excitation and a mildly nonlinear analysis is performed; then, the input level is increased stepwise, using the result of the previous step as a first guess. In most cases the intermediate results will also be of practical interest, as in the case of a power amplifier or an oscillator driven from small-signal level into compression. Most commercially available CAD software automatically enforces this method when convergence becomes difficult or when it is not reached at all. The tests results show that this technique permits to choose the appropriate HF transistor and to be an efficient selection criterion to optimize active microwave structures, because the choice of the required transistor is based essentially on the analysis and on comparison of the level coefficients of different useful transistors. The methods produce comparable results but are implemented quite differently.

The presented circuits may thus be regarded as examples which can further be utilized for the treatment of many similar circuits.

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