P-5

LARGE-SIGNAL NARROW BAND QUASI-BLACK-BOX MODELLING OF MICROWAVE TRANSISTORS

F.FILICORI, A.MAMBRIONI, V.A.MONACO

Dipartimento di Elettronica Informatica e Sistemistica
Università di Bologna, Italy

SUMMARY

A method is proposed for the large-signal narrow-band characterization of microwave active devices. It does not require a detailed knowledge of the device internal structure, but only some measurements of small signal S parameters effected under different bias conditions and some other simple large-signal measurements to be effected by a standard network analyzer.

INTRODUCTION

In the computer-aided design of microwave small -signal amplifiers, active device are usually characterized by a black-box mathematical model, such as its scattering matrix, whose elements can be measured by an automating network analyzer for the given bias conditions and at many different frequencies (1) In large-signal operation, the identification of the model of an active device is usually effected on a physical basis which requires a detailed knowledge of the internal structure and difficult measurement procedures for the determination of the values of its characteristic parameters.
A purely mathematical approach, which does not depend on the actual physical structure considered would be preferable also in non-linear operation, especially in those cases in which the circuit designer, as often happens, cannot have detailed information on the manufacturing processes of commercially available devices. The lack of general methods for the black--box modelling of non-linear dynamic 2-ports makes however this approach difficult to apply, since in practice only simplified and partial characterization procedures are at present being used (2) - (5).

In order to overcome these limitations, a different approach, based on a quasi-black-box model, is proposed here for the modelling of the large-signal narrow-band performance of microwave transistors.

THE NARROW-BAND QUASI-BLACK-BOX MODEL.

In order to define the transistor model it is convenient to considere separately its active part (i.e. the "intrinsic transistor") which is mainly non-linear, form the remaining linear subnetwork.

The fundamental hypothesis on which the proposed method is based consists of assuming that, in relatively narrow-band and moderalety non-linear operation, the behaviour of the intrinsic transistor can be described in terms of fundamental harmonic components of voltages and currents at its input and output ports by the equations:

$$1)\quad \begin{aligned} I_i &= G_i(V_i) + G_t(V_o) \\ I_o &= G_f(V_i) + G_o(V_o) \end{aligned}$$

which are characterized by the four complex non-linear functions G, each one being dependent on only one complex variable.

Equations 1 represent a reasonable approximation of the intrinsic behaviour of various types of transistors and in particular they give a considerably accurate description of the Ebers-Moll model of a bipolar transistor. As far as other kinds of devices and, in particular, field-effect transistors are concerned, eqns 1 are not so immediately justifiable but in several papers (6) approximations which can lead to eqns 1 have been adopted in order to evaluate distortion and, in general, the behaviour of the device in non-linear operation.

In order to obtain the more convenient form of the mathematical model proposed, we observe that under the hypothesis of time-invariant operation of the device, a delay of the variable controlling any

0149-645X/86/0000-0393$01.00 © 1986 IEEE

function G must give only an identical delay without any other variation to the same function; thus the ratio

$$W(|V|) = G(V)/V$$

must necessarily be a function only of the module of the controlling variable V. Therefore eqns.1 may be conveniently put in the form:

$$2) \quad \begin{aligned} I_i &= W_i(|V_i|)\, V_i + W_r(|V_o|)\, V_o \\ I_o &= W_f(|V_i|)\, V_i + W_o(|V_o|)\, V_o \end{aligned}$$

and by separating the linear components:

$$3) \quad W^{\ell} \overset{\Delta}{=} \lim_{|V|\to 0} W(|V|)$$

from those which are purely non-linear:

$$4) \quad W^N(|V|) = W(|V|) - W^{\ell},$$

eqns 2 take the matrix form:

$$5) \quad \underline{I} = \underline{I}^{\ell} + \underline{I}^{N} = \underline{W}^{\ell}\,\underline{V} + \underline{W}^{N}\,\underline{V}$$

It should be considered that at microwave frequencies all the parasitic elements associated both to the device and to the package are very important, and also that all these elements are reasonably linear and can be represented by the above-mentioned remaining linear subnetwork.

By putting together all the linear elements both relative to the package and to the intrinsic transistor, the device may be represented as shown in fig.1. The NLIT accounts only for the strictly non-linear part of the intrinsic transistor and the ALN is the associated linear network containing all the linear elements and having four ports: 1 and 2 are the input and output ports of the device while the other two are the internal ones (not accessible), to which the NLIT is connected.

The behaviour of ALN is completely described by a system of linear equations between the eight electrical variables at its four ports.

The connection between the two subnetworks imposes the constraints $I_3 = I_i^N$, $I_4 = I_0^N$, $V_3 = V_i$ and $V_4 = V_o$ which, taking into account that in small signal operation $I_i^N = I_o^N = 0$ (the functions W_N being null by definition due to eqns.3 and 4), allow for the determination of voltages V_i and V_o applied to NLIT in terms of the wave variables at the input and output ports:

$$6) \quad \begin{aligned} V_i &= V_3 = k_{A1}(a_1 + k_1 b_1) \\ V_o &= V_4 = k_{A2}(a_2 + k_2 b_2) \end{aligned}$$

where the parameters k_{A1}, k_{A2}, k_1, k_2 depend only on the subnetwork ALN.

In large-signal operation, the currents I_3 and I_4 at the ports of the NLIT are not zero and can be determined in terms of the voltages $V_3 = V_i$, $V_4 = V_o$. To this end, considering only moderately non-linear operation, it is possible, in the computation of the voltages V_i and V_O, to neglect the terms due to the same non-linearities, that is, to utilize expressions 6 relative to small-signal operation thus obtaining the narrow-band mildly non-linear mathematical model of the device defined by the vector equation:

$$7) \quad \underline{b} = \underline{S}\,\underline{a} + \underline{F}(\underline{a} + \underline{k}\underline{b})$$

between the column vectors $\underline{b}$ and $\underline{a}$ of reflected and incident wave-variables, where $\underline{S}$ is the scattering matrix of the device with respect to its external ports in small-signal operation for a given frequency and assigned bias conditions; $\underline{F}$ is a matrix whose four elements are complex non-linear functions of a real variable

$$8) \quad F_{ij} = F_{ij}(|a_j + k_j b_j|) \qquad i,j = 1, 2$$

which describe the behaviour of the non-linear intrinsic transistor (NLIT), and $\underline{k}$ a diagonal matrix whose elements k_1 and k_2 are complex parameters depending on the associated linear network (ALN).

MODEL IDENTIFICATION

The measurement of the S parameters of the device for a given frequency and assigned bias does not present any problem since it may be effected by a standard network analyzer. Instead parameters k_1 and k_2, due to the non-accessibility of ports 3 and 4 of the ALN, cannot be directly measured. It is therefore necessary to adopt an indirect identification procedure.

In particular, according to the method suggested in (10), the parameter k_2 can be determined by using the formula:

$$9) \quad k_2 = \frac{\displaystyle\sum_{j=1}^{n}\left(\frac{1}{S_{21}^{(j)}} - \frac{1}{n}\sum_{i=1}^{n}\frac{1}{S_{21}^{(i)}}\right)\cdot\left(\frac{S_{22}^{(j)}}{S_{21}^{(j)}} - \frac{1}{n}\sum_{i=1}^{n}\frac{S_{22}^{(i)}}{S_{21}^{(i)}}\right)^{*}}{\displaystyle\sum_{j=1}^{n}\left|\frac{S_{22}^{(j)}}{S_{21}^{(j)}} - \frac{1}{n}\sum_{i=1}^{n}\frac{S_{22}^{(i)}}{S_{21}^{(i)}}\right|^{2}}$$

where * indicates the complex conjugate; and $S^{(j)}$ the scattering parameters measured for a set of n suita bly chosen bias points; in the case of bipolar transistors this set of bias points is generated by changing only the base-collector d.c. voltage.

The validity of this method for the evaluation of k_2 has been verified with reference to a MSC 3000 transistor for which a physical model was already available (2) and whose accuracy had been proved in several applications (8). In order to determine the value of k_1, a procedure similar to that defined for k_2 can be used once the port numbers 1 and 2 have been reversed.

The characterization of the model also requires the identification of the four complex functions F which may however be effected on the basis of conventional wave variable measurements at the external ports of the device in large-signal operation. Considering for simplicity "quasi-unilateral" devices, the determination of the functions F_{ij} can be effected with a standard network analyzer by measuring,

10) $$\phi_{i2}(|a_2|) = \left.\frac{b_i}{a_2}\right|_{a_1=0} \qquad \phi_{i1}(|a_1|) = \left.\frac{b_i}{a_1}\right|_{a_2=0}$$

with i = 1,2 and by computing

$$F_{i2}(|x_2|) = \frac{\phi_{i2}(|a_2|) - S_{i2}}{1 + k_2\phi_{22}(|a_2|)}$$

11)

$$F_{i1}(|a_1|) = \phi_{i1} - S_{i1} - F_{i2}(|k_2\phi_{21}\ a_1|)k_2\phi_{21}$$

with $x_2 = a_2 + k_2 b_2 = a_2\left[1 + k_2\phi_{22}(|a_2|)\right]$

The model is completely identified when the non-linear functions have been determined for a set of discrete values of $|x_2|$ and $|a_1|$

ANALYSIS OF A LARGE-SIGNAL AMPLIFIER BY THE PROPOSED NON-LINEAR MODEL.

The analysis of active circuits in moderately non-linear operation and with narrow band signal is greatly simplified by employing the quasi-black-box model here described. With reference to the circuit represented in fig.2, the analysis consists in solving the systems of equations obtained by considering the model eqns.7 together with the linear source and load equations at the input and output ports of the active device:

$$\underline{a} = \underline{\rho}\,\underline{b} + \underline{\Gamma}$$

where $\underline{\rho}$ is the diagonal matrix of source and load reflectivities and $\underline{\Gamma}$ the vector of the forcing terms.

Owing to the non-linearity, the solution of this system involves the adoption of an iterative technique. Since the non-linear level is not too high, it is possible to adopt a simple iterative sequence whose n^{th} iteration is defined by:

$$\underline{b}^{(n+1)} = \left[\underline{1} - \underline{S}\underline{\rho} - \underline{F}^{(n)}\ (\underline{\rho}+\underline{k})\right]^{-1}\ (\underline{S} + \underline{F}^{(n)})\ \underline{\Gamma}$$

where the values of functions F_{ij} have approximated by their values at the previous iteration. This iterative formula has shown good convergence properties and my be initialized by starting with $F^{(o)}=0$, which correspondes to small-signal operation.

VALIDATION OF THE PROPOSED METHOD AND EXPERIMENTAL RESULTS

In order to verify the validity of the mathematical model proposed and of the procedure for determinating its parameters, the middle power bipolar transistor MSC 3000, whose parameters relative to an accurate physical model were already known (2), has been considered.

The scattering parameter, in small-signal operation at 1.6 GHz and relative to I_{co} = 30 mA with different values of V_{CBO}, have first been measured and the complex parameter k_2 has been determined by applying expression 9 with I_{co} = 30 mA and V_{CBO}=28 V as reference bias condition.

By means of a standard microwave network analyzer the non-linear functions ϕ_{ij} defined by expressions 10 have then been measured and used to compute according to eqns.11 the non-linear functions F_{ij} of the model; then, by following the procedure described in the previous section, the analysis was performed of two amplifiers characterized by two different values of the source and load reflection coefficients ρ_S, ρ_L. The computed values of the tranducer gain and the phase characteristic vs. input power are plotted in fig.3 where the corresponding measured values are also shown. The generally good agreement between measured and computed quantities confirm the validity of the model also in strongly non-linear operation.

In order to verify the validity of our model for simulating FET's, the data reported in (7) by Tucker and relative to a G_aA_s FET were used. These data were to so-called "large-signal S-parameters" corresponding to the ϕ's in eqns.10 by which, through the expressions 11 and 12, the non-linear functions F_{ij} of the model were computed.

In this way the large-signal FET amplifier described in (7) was analyzed using the tecnique proposed in this paper. The good agreement between the values of transducer gain obtained by using our model and those measured and computed by Tucker using a different approach (see fig.4) confirm the validity of the proposed method also in the case of FET'S.

CONCLUSION

A method has been proposed for characterizing an active device in large-signal operation based on the identification of a non-linear model which does not require a detailed knowledge of the internal structure of the device by only some measurements of small signal S parameters effected under different bias conditions and some other simple large--signal measurements to be effected by a standard network analyzer.

By means of a simple numerical analysis algorithm, which can be run on a personal computer, the model can be used to predict the amplifier response in terms of transducer gain, maximum output power, AM-PM conversion and other distortion parameters.

REFERENCES

(1) V.A.Monaco, et al., IEEE Trans. MTT-22, March 1974

(2) G.P.Bava et al., Rep.3 and 4 ESTEC contract 3334/NL/AK

(3) C.Rauscher et al., IEEE Trans. MTT-28, oct. 1980

(4) S.R.Mazumder et al., IEEE Trans., MTT-26, June 1978

(5) F.Sechi, IEEE Trans., MTT-28, Nov.1980

(6) G.P.Bava et al., Electronics Letters, Vol.18, 18 Feb.1982

(7) R.S.Tucker, IEEE Trans., MTT-29, Aug.1981

(8) F.Filicori et al., IEEE Trans. MTT-27, Dec.1979

(9) O.Shimbo, IEEE Proc., February 1971

(10) F.Filicori et al., ALTA FREQUENZA, Vol.55, n.1 January 1986.

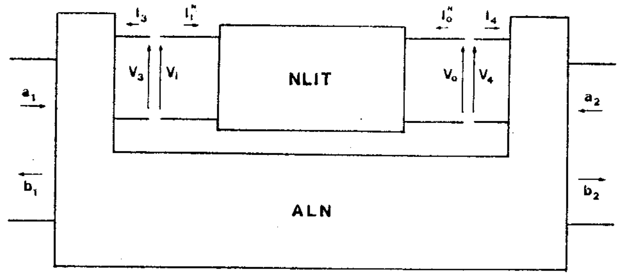

Fig.1 Decomposition of the active device into two subnetworks: the Non-Linear Intrinsic Transistor and the Associated Linear Network.

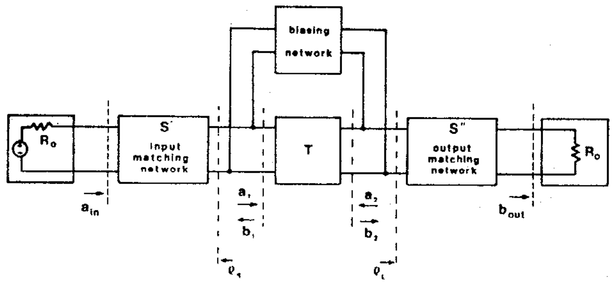

Fig.2 Circuit configuration of the analyzed amplifier.

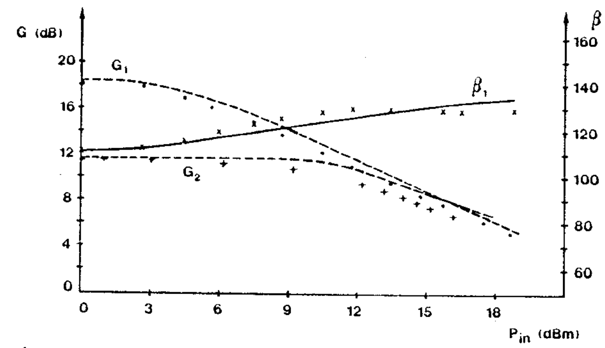

Fig.3 Computed transducer gain G(---) for two source/load conditions compared with the measured values (.)(+); phase response β computed (—) and measured (x).

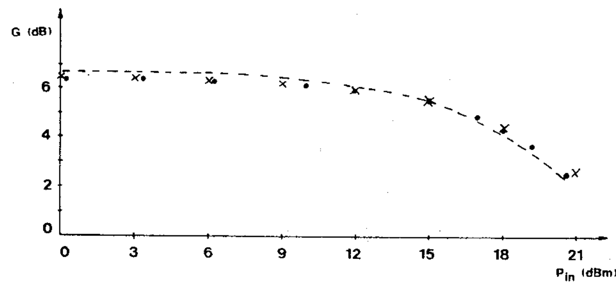

Fig.4 Curve of the transducer gain vs input power for the FET amplifier computed by Tucker (+), compared with the measured values (·) and with those computed by the quasi-black-box model (--).