Characterization of GaAs FET's in Terms of Noise, Gain, and Scattering Parameters Through a Noise Parameter Test Set

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Abstract —**A method for the complete characterization of GaAs FET's in terms of noise parameters (F_o, Γ_{on}, R_n), gain parameters (G_{ao}, Γ_{og}, R_g), and of those scattering parameters ($S_{11}, S_{22}, |S_{12}|, |S_{21}|, \angle S_{12}S_{21}$) that are needed for low-noise microwave amplifier design is presented. The instrumentation employed, i.e., a noise-figure measuring system equipped with a vectorial reflectometer, as well as the time consumption, are the same required for the determination of noise parameters only through conventional methods. The measuring setup and the experimental procedure are described in detail. Considerations about the computer-aided data processing technique are also provided. As an experimental result, the characterization of a sample device versus frequency (4–12 GHz) and drain current is reported. A comparison between the scattering parameters provided by the method and those measured by means of a network analyzer is also included.**

I. Introduction

OPTIMIZATION OF noise figure, gain, and input and output VSWR in designing wide-band, low-noise MESFET amplifiers requires a complete characterization of the device in terms of noise, gain, and scattering parameters versus frequency and drain current.

Scattering parameters are usually measured through a (automatic) network analyzer.

Noise and gain parameters cannot be measured through an instrument, but their determination requires time-consuming experimental and data-processing procedures.

To determine noise parameters, it is necessary to perform measurements of the device noise figure $F(\Gamma_s)$ for some (redundant, i.e., more than four, for accuracy) values of the input termination reflection coefficient Γ_s. Solving then the set of equations derived from the following relationship (or an equivalent one):

$$F(\Gamma_s) = F_o + 4N_n \frac{|\Gamma_s - \Gamma_{on}|^2}{(1-|\Gamma_s|^2)(1-|\Gamma_{on}|^2)} \tag{1}$$

the four noise parameters F_o, Γ_{on} (magnitude and phase), and N_n are obtained.

In order to evaluate the device noise figure $F(\Gamma_s)$, it is necessary to account for the noise contribution of the measuring stages following the device under test (DUT), making use of the well-known Friis formula

$$F_m(\Gamma_s) = F(\Gamma_s) + \frac{F_r(S'_{22})-1}{G_a(\Gamma_s)} \tag{2}$$

where $F_m(\Gamma_s)$ is the measured noise figure, $G_a(\Gamma_s)$ is the DUT available power gain, and $F_r(S'_{22})$ is the noise figure of the measuring stages when input is terminated on the DUT output reflection coefficient $S'_{22}(\Gamma_s)$. If an output matching network is employed, S'_{22} can be tuned to zero for each Γ_s; then $F_r(S'_{22})$ reduces to the constant value $F_r(0)$.

The gain $G_a(\Gamma_s)$ can be computed through the scattering parameters or measured as a power ratio by means either of a gain measuring system or the same instrumentation used for noise measurements [1].[1] After measuring some (redundant) values of $G_a(\Gamma_s)$ the gain parameters G_{ao}, Γ_{og} (magnitude and phase), and N_g defined by

$$\frac{1}{G_a(\Gamma_s)} = \frac{1}{G_{ao}} + 4N_g \frac{|\Gamma_s - \Gamma_{og}|^2}{(1-|\Gamma_s|^2)(1-|\Gamma_{og}|^2)} \tag{3}$$

can be derived through the same (computer-aided) data processing procedure as above for the noise parameters.

The described procedure, now commonly applied, refers to the state-of-the-art in the field of device characterization as assessed in 1969, when Lane first proposed to substitute with a computer-aided data processing technique the graphic procedure established ten years before by the IRE Standards [2], [3]. This procedure is, however, time-consuming and requires different measuring systems for the determination of all the parameter sets.

In this paper, a method is presented which allows the simultaneous determination of the noise and gain parameters and of those scattering parameters that are needed for the design and analysis of microwave amplifiers ($S_{11}, S_{22}, |S_{12}|, |S_{21}|, \angle S_{12}S_{21}$) by means of a single measuring system. The instrumentation employed, i.e., a noise characterization setup equipped with a vectorial reflectom-

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[1] Commercial instruments for the simultaneous measurement of noise figure and gain of a device driven by a noise source are also available (e.g., AILTECH mod.7380 and Hewlett-Packard mod.8970). These instruments are very useful for measurements on matched devices; they are not convenient, however, for transistor characterization because in this case the use of a matching network at the DUT output port is required, which in turn implies time-consuming tuning adjustments as Γ_s varies and increased risk of oscillations.

eter, and the time-consumption are the same required for the determination of noise parameters only through conventional methods. The chosen test-set topology and the particular data processing procedures devised, together with the fact that measurements of signal sensitive parameters are performed at noise level, assure good repeatibility and accuracy. A further advantage of the proposed measuring system is that the possibility of oscillation build-up during characterization of potentially unstable devices is strongly reduced.

The theoretical analysis of the method, the measuring setup, the step-by-step experimental procedure, and the (computer-aided) data processing technique are fully discussed. Measurement procedures for testing the accuracy of the results obtained are also described.

As an experimental result, the complete characterization of a packaged GaAs MESFET versus frequency (4–12 GHz) and drain current (5–30 percent I_{DSS}) is reported. A comparison between the scattering parameters computed through the proposed method and the ones measured by a network analyzer is also included.

II. Analysis of the Method

The method presented here is the improvement of a method for the simultaneous determination of noise and gain parameters through noise-figure measurements only, already successfully applied for the characterization of bipolar transistors up to 4 GHz [4], [5].

The principle of the method can be discussed referring to the simplified block diagram shown in Fig. 1.

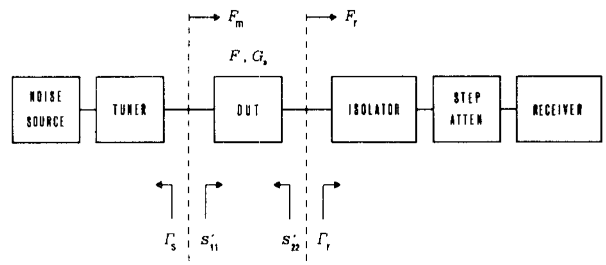

Fig. 1. Simplified block diagram illustrating the measurement principle.

As compared with conventional measuring systems, it can be observed that a matching network (tuner) at the output of the DUT is not used because a) it requires seeking for a careful tuning in order to maintain the matching for each Γ_s, and b) it may cause device oscillation (also outside of the measuring band) which falsifies the measurements. Instead of a tuner, an isolator and a step attenuator are used. The isolator allows the separation of the DUT from the stages following it, allowing us to use, for evaluating $F_r(S'_{22})$, the expression

$$F_r(S'_{22}) = F_r(0)\frac{|1-S'_{22}\Gamma_r|^2}{1-|S'_{22}|^2} \tag{4}$$

which permits the computation of F_r from the measured values of S'_{22}, $F_r(0)$, and Γ_r. The step attenuator is employed in order to easily obtain several different values of F_r for the same value of Γ_s.

Once a set of measurements of F_m is performed for a fixed value of Γ_s and for several (redundant, i.e., more than two) values of F_r, $F(\Gamma_s)$ and $G_a(\Gamma_s)$ are derived from (2). By repeating this measurement cycle for some (redundant, i.e., more than four) values of Γ_s, the noise and gain parameters of the DUT are derived from (1) and (3), respectively.

From the above measurements, the scattering parameters S_{11}, $|S_{21}|$, $|S_{12}|$, $\angle S_{12}S_{21}$ may also be derived by computation.

The DUT output reflection coefficient S'_{22} has been measured for each Γ_s to compute F_r. From the value of S_{22}, obtained by measuring S'_{22} for $\Gamma_s = 0$, and the relationship of the device gain under Z_0-terminated (usually 50 Ω) input conditions $G_a(0)$, $|S_{21}|$ given by

$$|S_{21}|^2 = G_a(0)\left(1-|S_{22}|^2\right) \tag{5}$$

is derived. From the relationship that expresses the dependence of $G_a(\Gamma_s)$ from the input termination reflection coefficient in terms of the scattering parameters we have

$$|1-S_{11}\Gamma_s|^2 = \frac{|S_{21}|^2\left(1-|\Gamma_s|^2\right)}{\left(1-|S'_{22}|^2\right)G_a(\Gamma_s)} \tag{6}$$

which allows the calculation, through proper algorithms, of the magnitude and phase of S_{11}.

For example, by putting $x \equiv \mathrm{Re}\{S_{11}\}$ and $y \equiv \mathrm{Im}\{S_{11}\}$ from (6), we get

$$x^2 + y^2 + ax + by + c = 0 \tag{7}$$

where

$$a = -\frac{2\cos \angle\Gamma_s}{|\Gamma_s|} \qquad b = \frac{2\sin \angle\Gamma_s}{|\Gamma_s|}.$$

$$c = \frac{1}{|\Gamma_s|^2}\left(1-\frac{|S_{21}|^2\left(1-|\Gamma_s|^2\right)}{G_a(\Gamma_s)\left(1-|S'_{22}(\Gamma_s)|^2\right)}\right). \tag{8}$$

The set of equations obtained from (7) for some (redundant, i.e., more than two) values of Γ_s can be solved by means of the least-squares method and a (computer-aided) successive approximation procedure.

From the computed value of S_{11} and the relationship

$$S_{12}S_{21} = \frac{1}{\Gamma_s}\left(S'_{22}-S_{22}\right)\left(1-S_{11}\Gamma_s\right) \tag{9}$$

obtained from the expression of $S'_{22}(\Gamma_s)$, the product $S_{12}S_{21}$ is then computed (magnitude and phase).

It is noteworthy that the computed parameters (S_{11}, $|S_{12}|$, $|S_{21}|$, and $\angle S_{12}S_{21}$) are derived with high accuracy due to the redundancy in the processed data; only S_{22} is directly measured. In addition, the former set of parameters are determined at noise level, thus eliminating nonlinearity effects.

III. Measuring Setup and Experimental Procedure

The detailed block diagram of the measuring system used for both C- and X-band measurements is shown in Fig. 2.

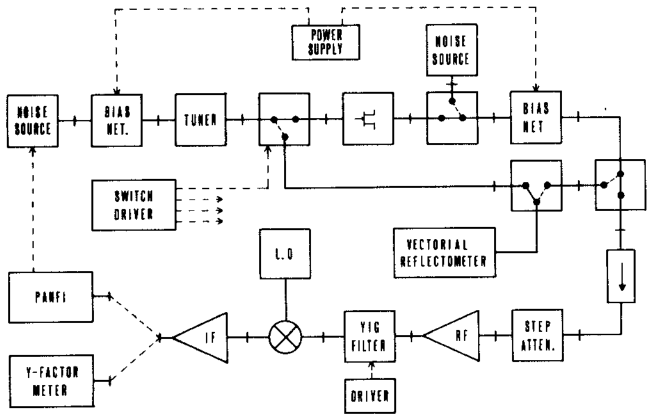

Fig. 2. Block diagram of the proposed noise, gain, and scattering parameter measuring system.

The instrumentation required is the same as for the determination of the noise parameters only through conventional methods. In order to go through the different phases (calibration, measurement, test) of the experimental procedure without manual reconfiguration of the system, some coaxial switches are employed, thus allowing better repeatibility and accuracy. A calibrated solid-state noise source is driven ON–OFF by the precision automatic noise-figure indicator (PANFI by AILTECH). Alternatively, a more accurate, but time-consuming, Y-factor meter can be employed to measure noise figure with the noise source switched ON–OFF manually. Since the source employed is well matched, ON–OFF mismatch effects are neglected here; in any case, either they can easily be taken into consideration by means of a proper data processing procedure [6] or strongly reduced by inserting an isolator at the noise source output and accounting for the reduction of the ENR due to the isolator losses [7].

The set of different values of the source reflection coefficient Γ_s is obtained by means of a coaxial slide-screw tuner. Criteria for properly selecting Γ_s values and evaluating the tuner losses are discussed in the following section. The measurements of Γ_s and $S'_{22}(\Gamma_s)$ are performed on-line by means of a vectorial reflectometer; their actual values at the input and output reference planes of the transistor are derived by computation after analytical modeling of the test fixture-switch cascade.

The measurement of F_r is performed by connecting the noise source in front of the output bias-network; a second noise source may be employed for convenience. The different values of F_r are obtained through a high-repeatibility step-attenuator. As previously stated, the isolator is inserted in order to maintain a 50-Ω match to the stages following it, thus simplifying the computation of the "receiver" noise figure $F_r(S'_{22})$. Should the device break into oscillation outside the measuring band where the isolator may no longer provide sufficient loading, a wide-band pad can be inserted in front of the isolator as a stabilizing load.

The YIG-tuned filter eliminates the effect of the image frequency arising from down-conversion. As for any noise measuring system, the use of the filter may be avoided since it has been shown that the image-frequency effects can be fully accounted for by computation [8], [9]. To this end, it is sufficient, if the intermediate frequency is not too high, to measure the source reflection coefficient also at the image frequency. In this case, however, the number of measurements to be performed will grow too much, so that the elimination of the filter is suggested only when the experimental procedure is executed by means of a computer-controlled setup.

The measuring system of Fig. 2 has been designed in view of automation. By replacing all the manual instruments (power supply, switch driver, step attenuator, YIG-filter deriver, etc.) with their programmable counterparts, measurements, data acquisition and processing versus frequency and bias condition can be performed, for each value of Γ_s, under the control of a desk-top computer. For a full automatic characterization of a device, however, a computer-driven tuner is also necessary [10]. Unfortunately, a satisfactory solution to this problem has not been proposed so far.

For each frequency point, the suggested measuring procedure is the following:

1) By adjusting the tuner, obtain a value of Γ_{si} ($i = 1, 2, \cdots, p$) and measure it through the reflectometer.

2) Measure the corresponding value of $F_{mj}(\Gamma_{si})$ ($j = 1, 2, \cdots, q$) for each of the q values of attenuation inserted by the step attenuator.

3) For the same set of attenuations, measure the values of the receiver noise figure $F_{rj}(0)$ under the 50-Ω input-terminated condition.

4) Measure the device output reflection coefficient $S'_{22i} \equiv S'_{22}(\Gamma_{si})$ and compute $F_{rj}(S'_{22i})$ from (4) and the previously measured value of the receiver input reflection coefficient Γ_r.

From the above collected data, $F(\Gamma_{si})$ and $G_a(\Gamma_{si})$ are computed through (2).

5) Repeat steps 2) and 4) for the other bias conditions (when required).

6) Repeat steps 1), 2), 4), and 5) for the next Γ_s.

From the p values of $F(\Gamma_{si})$ and $G_a(\Gamma_{si})$ so determined, the device noise and gain parameters are derived from (1) and (3).

7) By repeating steps 2), 4), and 5) under the 50-Ω input-terminated condition, compute $F(0)$ and $G_a(0)$.

From the measured values of S'_{22i} and S_{22} (steps 4) and 7)) and the computed set of $G_a(\Gamma_{si})$ and $G_a(0)$, all the remaining scattering parameters are obtained through (5)–(9), as shown in the previous section.

In addition, comparison between the values of $F(0)$ and $G_a(0)$ obtained from the measurements of the $F_{mj}(0)$, and the ones computed from the noise and gain parameters already derived, represents a check of the accuracy of the experimental procedure carried out, including the evaluation of the tuner losses.

IV. Experimental Verifications

After the effectiveness of the method has been examined through computer-aided simulation, experiments have been carried out on low-noise transistors. The experimental verifications reported here concern the complete characterization of a GaAs FET (NE24483 by Nippon Electric Company; common source configuration; $V_{DS} = 3$ V) in terms of noise, gain, and scattering parameters versus frequency (4–12 GHz) and drain current ($I_{DS} = 5$, 10, 15, and 30-percent I_{DSS}).

The two noise-measuring setups used for C- and X-band characterization correspond to the same block diagram of Fig. 2. Only octave-bandwidth-limited components (isolator, RF amplifier, etc.) have been substituted. For the 8-GHz frequency, which is common to both the instrumentation benches, the measuring procedure was executed twice in order to check for repeatibility.

Measurements of the noise figure $F_{mj}(\Gamma_{si})$ have been performed for twelve different values of Γ_s and for 0, 3, 6, and 10 dB of inserted attenuation in the C-band. The values of the attenuation steps depend on the device gain. High values of gain require high values of attenuation in order to obtain noticeable increments of the measured noise figure, which is necessary for good accuracy. On the other hand, low values of device gain require smaller values of maximum inserted attenuation to avoid an exceedingly high "receiver" noise contribution, which again reduces accuracy. For this reason, the 10-dB attenuation step has not been used in X-band measurements, where the transistor gain is lower.

Improper selection of the set of Γ_s may cause serious problems in determining device parameters, as for any noise measuring system. This topic was discussed theoretically elsewhere, where some practical selection criteria to follow in order to reduce error sensitivity are also given [11], [12].

In our case, after the overall system optimum noise source admittance Γ_{om} has been determined for each frequency, values of Γ_s have been realized in its neighborhood a) by acting on the screw for a fixed carriage position and b) by sliding the carriage for a fixed screw position. Since the tuner employed exhibits negligible variation of the loss versus configuration, provided that deep insertion of the screw is avoided, full characterization of the tuner [13] is not necessary, and the loss of the bias network-tuner cascade is measured for one configuration only (50-Ω match). The influence of the switch and transistor test-fixture input line are accounted for by modeling as low-loss line.

Measurements of $F_m(\Gamma_s)$ are performed through the automatic noise-figure meter. Since measurements of $F_{rj}(0)$ are performed one time only for each frequency, the more accurate Y-factor meter is employed.

Experimental data have been processed by means of an HP 9835 desk-top computer. In order to reduce computational run-time, (1) and (3) have been rewritten in linearized form by introducing indirect noise and gain parameters related to conventional ones through relationships [5]. A successive approximation technique such as the one proposed in [14] has not been employed in this case, because no significant improvement was obtained. The device noise and gain parameters versus frequency, for the fixed drain current value $I_{DS} = 15$-percent I_{DSS} (where optimum noise performances are expected), are presented in Fig. 3. The associated noise figure $F(\Gamma_{og})$ and available gain $G_a(\Gamma_{on})$, i.e., the noise figure corresponding to the optimum source reflection coefficient for maximum available gain and the power gain corresponding to the optimum source reflection coefficient for minimum noise figure, respectively, are also shown in Fig. 3(a). For a better

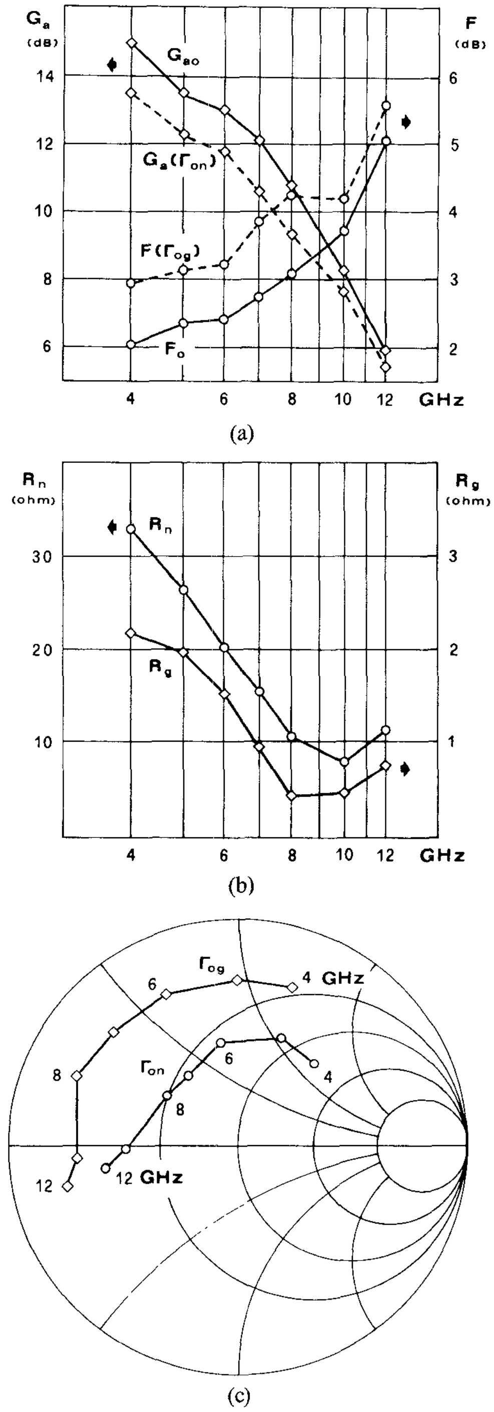

Fig. 3. Noise and available gain parameters for the NE24483 FET versus frequency ($I_{DS} = 15$-percent I_{DSS}, $V_{DS} = 3$ V, $T_{amb} = 25°$C). (a) Minimum noise figure F_0 and associated available gain $G_a(\Gamma_{on})$, and maximum available gain G_{ao} and associated noise figure $F(\Gamma_{og})$. (b) Equivalent noise R_n and gain R_g resistances. (c) Optimum source reflection coefficient for minimum noise figure Γ_{on} and maximum available gain Γ_{og}.

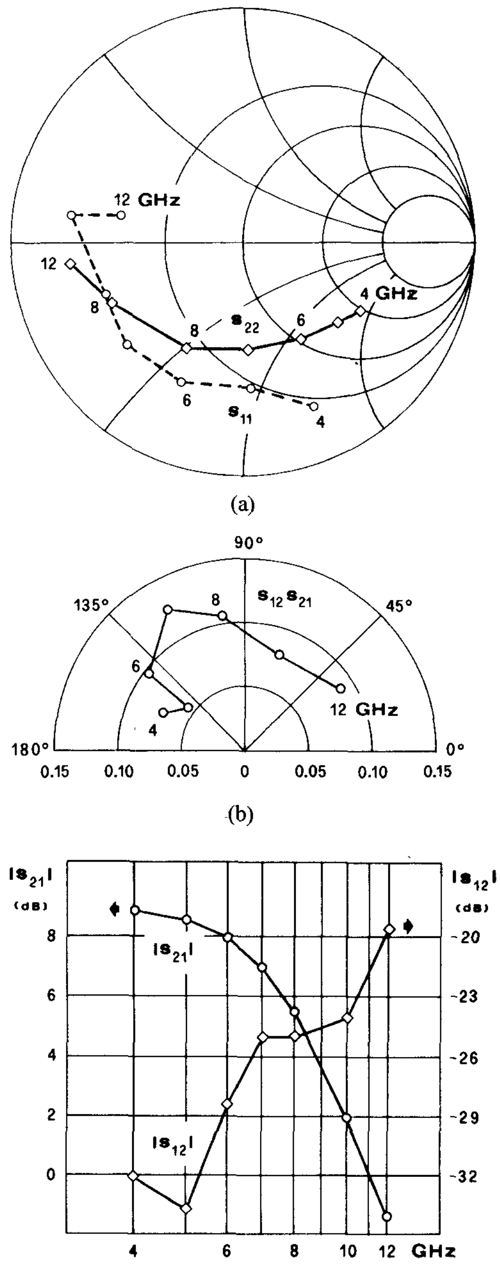

Fig. 4. NE24483 scattering parameters versus frequency as provided by the method (I_{DS} = 15-percent I_{DSS}, $V_{DS} = 3$ V, $T_{amb} = 25°C$). (a) S_{11} and S_{22}; (b) $S_{12}S_{21}$; (c) $|S_{21}|$ and $|S_{12}|$.

technical understanding, instead of the "terminal invariant" [15] parameters N_n and N_g, the more commonly employed equivalent noise and gain resistances R_n and R_g are reported in Fig. 3(b). The scattering parameters are given in Fig. 4. The comparison between the scattering parameters computed through the method presented here and the ones measured by means of a network analyzer is reported in Table I. Good agreement can be observed, except at 12 GHz where the increase of the measured noise-figure values reduces the accuracy of the automatic noise-figure indicator employed. The dependance of some device parameters on drain current for fixed frequency values is shown in Fig. 5. Comparison between measured and computed values of the device noise figure and avail-

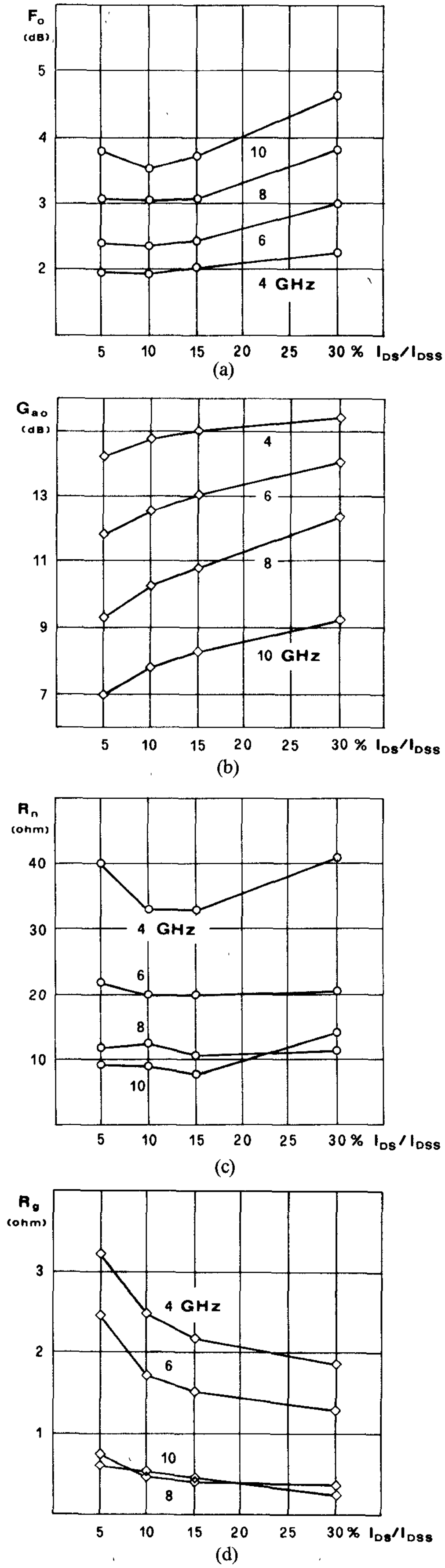

Fig. 5. Dependance on the drain current of (a) the minimum noise figure F_o, (b) the maximum available power gain G_{ao}, (c) the equivalent noise resistance R_n, and (d) the equivalent gain resistance R_g of the NE24483 for fixed frequency values ($V_{DS} = 3$ V, $T_{amb} = 25°C$).

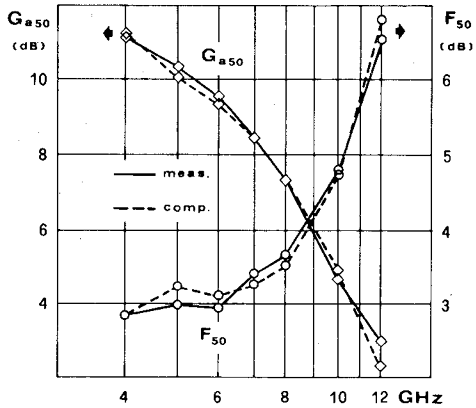

Fig. 6. Measured and computed noise figure F_{50} and available gain G_{a50} of the NE24483 under 50-Ω input-terminated condition ($V_{DS} = 3$ V, I_{DS} = 15-percent I_{DSS}, $T_{amb} = 25°C$).

TABLE I
COMPARISON BETWEEN THE SCATTERING PARAMETERS COMPUTED THROUGH THE METHOD HERE PRESENTED AND THE ONES MEASURED BY MEANS OF A NETWORK ANALYZER (REPORTED IN BRACKETS)

GHz	$\lvert s_{11}\rvert$	$\angle s_{11}$	$\lvert s_{21}\rvert$	$\lvert s_{12}\rvert$	$\angle s_{12}s_{21}$
4	.77 (.77)	−66° (−61°)	2.70 (2.52)	.025 (.029)	156° (157°)
5	.63 (.63)	−87° (−87°)	2.67 (2.63	.021 (.021)	144° (154°)
6	.66 (.66)	−115° (−116°)	2.52 (2.62)	.037 (.035)	142° (147°)
7	.67 (.67)	−139° (−139°)	2.24 (2.41)	.056 (.048)	119° (121°)
8	.64 (.66)	−160° (−159°)	1.90 (2.10)	.056 (.048)	100° (106°)
10	.75 (.73)	171° (171°)	1.26 (1.45)	.063 (.055)	70° (65°)
12	.55 (.73)	167° (151°)	0.90 (1.05)	.096 (.062)	33° (32°)

ble gain for a 50-Ω input termination, F_{50} and G_{50}, respectively, is also given in Fig. 6.

As an example of the typical uncertainty obtained in parameter evaluation following the suggested experimental procedure, worst-case deviations of F_o and G_{ao} as function of the number (n) of input terminations Γ_{si} are presented in Fig. 7. The computation was carried out by processing n values of Γ_s out of $p(=12)$. The reduced spreading in the determination of noise with respect to gain parameters can be observed.

V. CONCLUSIONS

A method for the simultaneous determination of noise, gain, and scattering parameters of microwave transistors through a noise parameter test set is presented. Instrumentation and time-consumption are the same required by conventional methods for the determination of noise parameters only. Theory of the method, measuring setup, experimental procedure for measuring and testing, and data-processing techniques are described. A computer-controlled version of the measuring system is also suggested. As experimental verification, the complete characterization of a GaAs FET versus frequency (4–12 GHz) and drain current is reported.

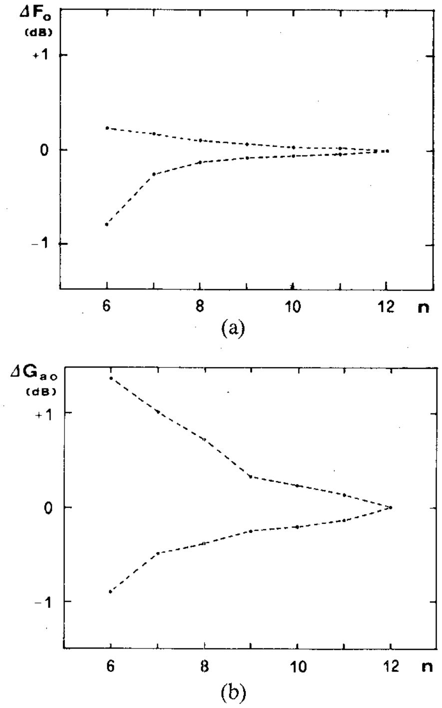

Fig. 7. Decrease of the uncertainty in the determination of (a) the minimum noise figure F_o and (b) the maximum available gain G_{ao} of the NE24483, as a function of the number of input terminations selected in the data processing procedure: worst-case deviation for n out of twelve measurement points.

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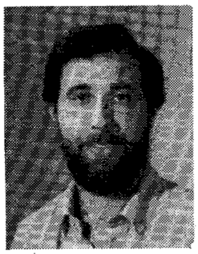

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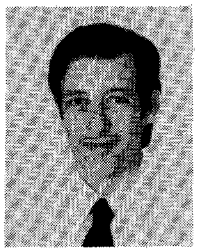

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Stability Margins in Microwave Amplifiers

DOUGLAS J. H. MACLEAN

Abstract —**Shunt feedback around single GaAs MESFET's is becoming more widespread to ease matching to 50-Ω terminations and improve gain flatness. The most accurate and meaningful method of assessing feedback, intentional or unintentional, is described. A simple sequence of steps leads from measured S-parameters to a plot of return ratio and Nyquist's criterion of stability. An amplifier using an accurately measured NE 70083 FET is analyzed to illustrate the method, and to present graphs of frequency-dependent admittances of a broad-band representation for transistors which is simpler than hybrid-pi models, and valid over the entire 2 to 18-GHz measured frequency range. The return ratio quantifies the total feedback present, thus enabling the most realistic stability margins to be found, and the benefits of feedback on performance to be quantified.**

I. Introduction

THE DESIGN, modeling, and realization of microwave amplifiers for operational systems are topics full of theoretical and practical limitations. By their very nature, all microwave transistors are active devices with feedback, while the distributed nature of the associated circuit may well give rise to feedback even where none is intended. The corresponding theoretical limitation has arisen from the absence, until very recently, of a feedback theory applicable at the frequencies of interest.

At present, Rollett's stability factor is commonly used to determine whether a given amplifier is absolutely or conditionally stable when viewed as a 2-port between arbitrary passive terminations. This criterion can be applied to a circuit diagram or to a physical amplifier, and any values of the factor less than one denote conditional stability at the frequencies concerned. In practice, an amplifier under test first has its 2-port S-parameter measured, then the measurements are error-corrected and the stability factor K computed. The absolute stability signified by the necessary but not sufficient condition $K > 1$ can be verified experimentally by using a sliding short at one port, then on the other, and finally at both ports simultaneously. The sliding

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