is situated $5\Delta d$ beneath the main line a and $1\Delta d$ beneath the secondary line b. If the plane of observation is nearer the main line a, the edge effect along the main line a will appear clearly. In Fig. 8(c), the main wave has passed over the crossing region. The wave reflected from the crossing appears on the left part of the secondary line b. The wave has the opposite phase because of the reflection by the conductor, but its magnitude is very small. The characteristics of these figures relate to the Poynting vector in Fig. 4 and the magnetic field in Fig. 6. The spatial spread of the distribution of the electric field E_y is not as wide as that of the magnetic field.

For 45 Degrees Crossing Angle: Fig. 9 shows the instantaneous time variation of the electric field E_y for the crossing angle of 45 degrees. Fig. 9(a) is the same as Fig. 8(a) because the wave has not reached the crossing area in both cases. Comparing Fig. 9(b) and (c) with Fig. 8(b) and (c), it is observed that the coupling characteristics of the 45 degrees case is more complicated than that of 90 degrees. This is caused by the different coupling process of each part of the crowd to the secondary line in space and time.

IV. Conclusions

In this paper the coupling between two crossing lines is described qualitatively. It is noted that the spatial spread of the fields of the propagating wave about the line causes complicated coupling characteristics. This also has importance in the analysis of the coupling characteristics of other configurations of lines, such as parallel lines. The quantitative evaluation of various coupling parameters is being studied now, and the results will be reported in a later paper. The importance of the Poynting vector corresponding to the power flow in the coupling characteristics should be investigated more rigorously. The coupling of more complex line configurations may be treated by the present method.

References

[1] D. V. Giri, S.-K. Chang, and F. M. Tesche, "A coupling model for a pair of skewed transmission lines," *IEEE Trans. Electromagn. Compat.*, vol. EMC-22, pp. 20–28, Feb. 1980.

[2] T. G. Bryant and J. A. Weiss, "Parameters of microstrip transmission lines and of coupled pairs of microstrip lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 1021–1027, Dec. 1968.

[3] C. R. Paul and A. E. Feather, "Computation of the transmission line inductance and capacitance matrices from the generalized capacitance matrix," *IEEE Trans. Electromagn. Compat.*, vol. EMC-18, pp. 175–182, Nov. 1976.

[4] A. K. Agrawal, H. J. Price, and S. H. Gurbaxani, "Transient response of multiconductor transmission lines excited by a nonuniform electromagnetic field," *IEEE Trans. Electromagn. Compat.*, vol. EMC-22, pp. 119–129, May 1980.

[5] Y. Kami and R. Sato, "Equivalent circuit for transmission lines excited by external electromagnetic waves," *Trans. IECE Japan*, vol. J66-B, pp. 687–688, May 1983.

[6] N. Schibuya, H. Takagi, K. Kumamoto, T. Homma, and K. Ito, "Crosstalk noise analysis of wiring on the printed circuit board," *Trans. IECE Japan*, vol. J68-B, pp. 1068–1076, Sept. 1985.

[7] S. Koike, N. Yoshida and I. Fukai, "Transient analysis of microstrip gap in three-dimensional space," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 726–730, Aug. 1985.

[8] S. Koike, N. Yoshida, and I. Fukai, "Transient analysis of directional coupler using coupled microstrip-slot-line in three-dimensional space," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 353–357, Mar. 86.

[9] N. Yoshida and I. Fukai, "Transient analysis of a stripline having a corner in three-dimensional space," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 491–498, May 1984.

[10] S. Koike, N. Yoshida, and I. Fukai, "Transient analysis of microstrip gap," *Trans. IECE Japan*, vol. J67-B, pp. 662–669, June 1984.

[11] S. Koide, N. Yoshida, and I. Fukai, "Transient analysis of coupled microstrip-slot-line," *Trans. IECE Japan*, vol. J68-B, pp. 811–818, July 1985.

A Method for Measurement of Losses in the Noise-Matching Microwave Network While Measuring Transistor Noise Parameters

GIOVANNI MARTINES AND MARIO SANNINO

***Abstract* — A new method for measuring the loss of a tuner network used as the noise-source admittance transformer in a noise parameter test set is presented. Since the method is based on noise figure measurements, the tuner losses are determined on-line while performing measurements for determining transistor noise parameters.**

Experiments carried out on a coaxial slide-screw tuner by means of a computer-assisted measurement setup are reported.

I. Introduction

Modern techniques for the determination of the four noise parameters which characterize the noise behavior of microwave transistors, i.e., minimum noise figure, optimum reflection coefficient of the input termination (magnitude and phase), and noise resistance, require measurements of the noise figure $F(\Gamma_s)$ of the transistor under test for some (redundant, i.e., more than four) values of the reflection coefficients Γ_s of its input termination. These measurements are made in conjunction with a computer-aided data-processing procedure based on the least-squares method [1]. This procedure furnishes good results, provided that, in selecting the values Γ_s, well-established criteria are followed [2], [3].

The same procedure yields the four available gain parameters of the transistor under test after carrying out measurements of transistor power gain $G_a(\Gamma_s)$ for some (redundant) Γ_s through a gain-measuring setup [1]. These parameters and the scattering parameters can also be derived when determining transistor noise parameters if a proper experimental procedure for measuring $F(\Gamma_s)$ is followed [4]–[7].

The essential component of the noise-parameter-measuring setup is a coaxial or waveguide double-stub or slide-screw tuner inserted between the noise source and the device under test (DUT), in order to transform the noise source reflection coefficient Γ_{ns} to the desired value Γ_s of the DUT input termination. The DUT output noise powers are then detected, through a receiver, by a meter which measures the noises figure F_m of the whole measuring setup. This figure is given by Friis's formula for cascaded networks, i.e.

$$F_m(\Gamma_{ns}) = \alpha_{\Gamma_s}(\Gamma_{ns})\left(F(\Gamma_s) + \frac{F_r(S'_{22}) - 1}{G_a(\Gamma_s)}\right) \tag{1}$$

where α_{Γ_s} represents the tuner loss, which depends on the tuner configuration; F, G_a, and S'_{22} are the noise figure, the available power gain, and the output reflection coefficient of the DUT, respectively; and F_r is the receiver noise figure.

From (1), it appears that in order to determine $F(\Gamma_s)$ from the measured noise figure F_m, previous measurements of the tuner loss $\alpha_{\Gamma_s}(\Gamma_{ns})$ are needed. Very little has appeared so far in the literature concerning this measurement.

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Alternative measuring methods based on the use of a low-loss tuner as a reflecting input termination (e.g., sliding short) have been proposed [8], [9] and applied recently [10] to the characterization of the noise behavior of receivers. In current practice, however, the former methods are preferred for characterizing microwave transistors in terms of noise parameters, especially when there is interest in the *direct* determination of the optimum parameters using only the input-tuning procedure.

On the problem of determining the losses α_{Γ_s} of the tuner versus setting, a recent paper of Strid [11] is of interest. This suggests a unique procedure to determine the losses of a tuner by characterizing it by its scattering parameters. For any Γ_s, the loss α_{Γ_s} is then computed as the inverse of the tuner available power gain. Unfortunately, this method is very time consuming and also not convenient from an experimental viewpoint, particularly when the tuner is used in a transistor noise parameter test set. This is because it requires

a) carrying out two experimental procedures at different times by means of two different measuring systems, one to measure noise figures $F_m(\Gamma_s)$, the other to measure the tuner S-parameters; and
b) setting and resetting the tuner accurately in performing the above experimental procedures for each Γ_s.

The above drawbacks are avoided by the method proposed in this paper, which allows tuner losses to be measured with the same setup used for noise figure measurements. The tuner losses α_{Γ_s} can be determined on-line while performing the measurements of $F_m(\Gamma_s)$ intended to derive noise (gain and scattering) parameters of transistors. The method has been devised in order to be easily embodied in a system for computer-assisted measurements for transistor characterization.

II. Analysis of the Method

To illustrate the principle of the method, let us first refer to a setup dedicated to measurements of tuner loss only. In Section III its application using a noise parameter test set is discussed.

In the measuring system shown in Fig. 1, the noise source is alternatively switched ON and OFF to inject two different noise power levels to the tuner under test; the noise powers are then detected by a single sideband receiver equipped with an input isolator and a noise figure indicator.

Supposing that the noise source is a well-matched one, i.e., $\Gamma_{ns}=0$, the noise figure F_r of the tuner–receiver cascade (Fig. 1(b)) is given by

$$F_{r\alpha}(0)=\alpha_{\Gamma_s}F_r(\Gamma_s) \tag{2}$$

from which

$$\alpha_{\Gamma_s}|_{dB}=F_{r\alpha}(0)|_{dB}-F_r(\Gamma_s)|_{dB} \tag{3}$$

which furnishes the tuner loss α_{Γ_s} if $F_r(\Gamma_s)$ is known.

To compute $F_r(\Gamma_s)$, we need in general the four noise parameters of the receiver. In the case of an input-isolated receiver, however, we have, more simply, [6], [7], [12]

$$F_r(\Gamma_s)=\frac{F_r(0)}{1-|\Gamma_s|^2}. \tag{4}$$

The receiver noise figure $F_r(0)$ in the input-matched condition is then measured by directly connecting the source to the isolator, and $|\Gamma_s|$ is measured through a reflectometer. Relationship (4) is valid if $T_a\cong T_0$ and $\Gamma_r\cong 0$, where T_a and T_0 are the (isolator) ambient temperature and the reference temperature (290°K), respectively, and Γ_r is the input reflection coefficient of the receiver.

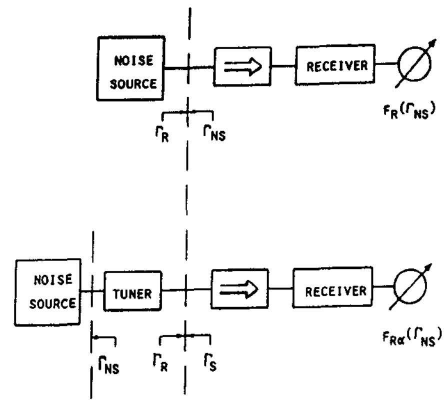

Fig. 1. The two different configurations of the measuring system for determination of tuner loss through noise figure measurements.

If these hypotheses do not hold, we have (see the Appendix)

$$F_r(\Gamma_s)=\frac{1}{\mu'}F_r(0)+\left(\frac{1}{\mu'}-1\right)\left(\frac{T_a}{T_0}-1\right) \tag{5}$$

where

$$\mu'=\frac{1-|\Gamma_s|^2}{|1-\Gamma_r\Gamma_s|^2}. \tag{6}$$

Let us consider finally the case in which the noise source is not well matched and consequently $F_{r\alpha}(0)$ and $F_r(0)$ cannot be measured. If Γ_{ns1} and Γ_{ns2} are the values of the *nominal* source reflection coefficient Γ_{ns} in the OFF and ON conditions, respectively, we can compute $F_{r\alpha}(0)$ and $F_r(0)$ by means of relationships similar to (5). Thus (see also the Appendix)

$$F_{r\alpha}(0)=\mu_2'F_{r\alpha}(\Gamma_{ns})+(\mu_2'-1)\left(\frac{T_a}{T_0}-1\right) \tag{7}$$

and

$$F_r(0)=\mu_2'F_r(\Gamma_{ns})+(\mu_2'-1)\left(\frac{T_a}{T_0}-1\right) \tag{8}$$

with

$$\mu_2'=\frac{1-|\Gamma_{ns2}|^2}{|1-\Gamma_{ns2}\Gamma_r|^2}. \tag{9}$$

From (7), (8), (5), and (2), the tuner loss α_{Γ_s} is determined by computer-aided processing of measured noise figures $F_{r\alpha}(\Gamma_{ns})$ and $F_r(\Gamma_{ns})$.

Compared with gas-discharge noise sources, modern solid-state sources offer differences between mismatches in ON and OFF conditions which are usually neglected. In this case, obviously, the above relationships hold with $\Gamma_{ns1}\cong\Gamma_{ns2}\cong\Gamma_{ns}$. On the other hand, different ON–OFF mismatches, i.e., $\Gamma_{ns1}\neq\Gamma_{ns2}$, are not acceptable when characterizing transistors through a noise parameter test set, because this implies that the value Γ_s is ambiguous while measuring $F(\Gamma_s)$ of (1).

For this reason, the mismatch difference is to be reduced to a negligible value. To this end, an isolator (or attenuator) can be connected to the source output, provided that the attenuation so introduced is accurately measured in order to determine the actual noise levels furnished by the equivalent noise source so obtained [13], [14].

The method presented here is suitable for this, using the same setup shown in Fig. 1 where the two-port under test is now the isolator–(or attenuator)–tuner cascade.

III. Application of the Method while Performing Computer-Controlled Measurements of the Noise (Gain and Scattering) Parameters of Microwave Transistors

The method proposed requires that the measurements be preceded by a calibration procedure for both the receiver and the noise source employed.

The calibration procedure is (for each frequency):

a) measure the receiver input reflection coefficient Γ_r through a vectorial reflectometer;
b) measure the noise source reflection coefficient Γ_{ns}; if the hypothesis $\Gamma_{ns} \cong \Gamma_{ns1} \cong \Gamma_{ns2}$ does not hold, Γ_{ns2} must be measured; in the case of transistor characterization, however, the ambiguity must be reduced, as stated above, through an isolator or an attenuator;
c) measure the noise figure $F_r(\Gamma_{ns})$ using the setup of Fig. 1(a).

The measurements to carry out on the tuner under test for each setting and frequency are:

a) measure the noise figure $F_{r\alpha}(\Gamma_{ns})$ using the setup of Fig. 1(b);
b) measure the reflection coefficient Γ_s of the tuner output.

From this it appears that, as previously stated, the method has been conceived with the objective of conveniently solving the problem of measuring the loss of a tuner when it is used as a source admittance transformer in a transistor noise figure measuring system. Actually, we note that the measurement a) is performed through noise figure measurements and step b) is already part of the experimental procedure for the determination of the noise (and gain) parameters of the transistor. In other words, the main advantage of the method is that it requires measurements of noise figure and reflection coefficient, which can be added to those already required by experimental procedures for computer-aided determination of transistor noise and gain parameters [1], [4], [5]. No further experimental effort is needed, the tuner loss being taken into account when carrying out computer-aided data processing.

The method can be easily implemented in a system for computer-assisted measurements. To this end, the setup of Fig. 1 is assembled by employing computer-controlled instruments and computer-driven microwave switches to realize the instrument connections. This can be accomplished using a test setup for complete characterization of transistors shown in Fig. 2, where the computer-assisted version of the system described in [6] is reported.

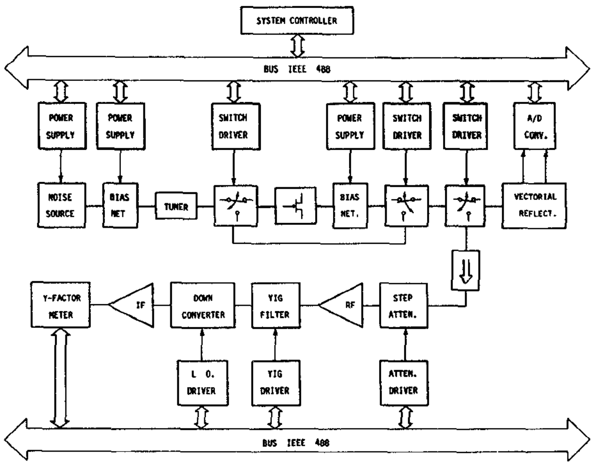

Fig. 2. Implementation of the tuner loss measuring method in the computer-controlled version of the transistor noise, gain, and scattering parameter measuring system proposed in [6].

IV. Experimental Results

The losses of the coaxial slide-screw tuner have been measured through a computer-assisted implementation of the measuring system of Fig. 1.

As instrument system controller (via HP-IB), an HP 9836 computer with a HP 6942A multiprogrammer, is used. To switch ON and OFF the solid-state noise source HP 346B and to actuate the microwave switches and the receiver programmable step attenuator HP 8494H, the multiprogrammer is equipped with relay card HP 69730A. Narrow-band (spot) noise figures are measured with the known Y-factor method by a HP 436A power meter with a HP 8484A (0.1 nW–0.01 mW) probe through a 30-MHz filter (1-MHz bandwidth). Reflection coefficient measurements are carried out by using as the vectorial reflectometer a network analyzer in the reflection mode only and an A/D converter card HP 69751A of the multiprogrammer. The tuner tested is the 2–18-GHz slide double-screw Maury Microwave 2640D.

The tuner losses have been measured at 8, 10, and 12 GHz for four different spacings of one screw from the inner conductor of the tuner and for 57 positions of the carriage, as shown in Fig. 3. The other screw (the left and larger) was fully extracted. The relevant values of the VSWR are reported in Table I.

V. Conclusions

The model of the tuner tested through the method presented here is the same as that previously characterized by Strid through S-parameter measurements [11]; qualitative comparisons between the two methods show that the results are in good agreement. This does not hold, however, when measuring the loss of a tuner employed as a noise source admittance transformer in a noise parameter test set. The main advantage of the method proposed here is that it requires measurements which are part of the experimental procedure already required for characterization of noisy two-ports.

This is valid in particular for determining noise, gain, and scattering parameters with the noise parameter test set recently proposed [6], [7]. It is thus becoming an accurate and convenient method for the complete characterization of microwave transistors.

Appendix

Computation of the Noise Figure $F_r(\Gamma_s)$ of a Receiver with Isolated Input for any Input Termination Reflection Coefficient Γ_s as Function of the Noise Figure $F_r(0)$ in Input Matched Condition

The noise temperature (noise power) of a receiver with an input isolator depends on the effective noise temperature of the isolator, which varies as Γ_s varies, and on the noise temperature (constant versus Γ_s) of the stages following the isolator. The isolator noise temperature is given by $T_a(\alpha_i(\Gamma_s)-1)$ [14], with $\alpha_i(\Gamma_s)=1/G_{ai}(\Gamma_s)$, where α_i and G_{ai} are the attenuation and the

TABLE I
APPROXIMATE VALUES OF THE VSWR OF THE TUNER OUTPUT VERSUS SETTING AND FREQUENCY

	8 GHz	10 GHz	12 GHz
P1	7.14	8.33	9.67
P2	3.76	4.31	4.74
P3	2.09	2.25	2.43
P4	1.38	1.44	1.51

available power gain, respectively. Thus, the receiver noise temperature is given by

$$T_{er}(\Gamma_s) = T_a(\alpha_t(\Gamma_s)-1)+\alpha_t(\Gamma_s)\tau = \alpha_t(\Gamma_s)(T_a+\tau)-T_a \tag{A1}$$

where [12]

$$\alpha_t(\Gamma_s) = \alpha_t(0)\frac{|1-\Gamma_r\Gamma_s|^2}{1-|\Gamma_s|^2} \tag{A2}$$

as can be derived from the relationship of the available gain

$$G_{at}(\Gamma_s) = \frac{|S_{21}|^2(1-|\Gamma_s|^2)}{|1-S_{11}\Gamma_s|^2(1-|S'_{22}(\Gamma_s)|^2)}. \tag{A3}$$

Recalling that $S_{12}=0$ for an isolator and, consequently, $S_{11}=\Gamma_r$ and $S'_{22}(\Gamma_s)=S_{22}$, the measured Y-factor is given by [13]

$$Y=\frac{\mu_2[T_2+T_e(\Gamma_{s2})]}{\mu_1[T_1+T_e(\Gamma_{s1})]} \tag{A4}$$

where Γ_{s2} and Γ_{s1} are the reflection coefficients of the isolator input termination when the noise source is in the ON and OFF conditions, respectively, and Γ_2 and Γ_1 are the corresponding values of the mismatch factor given by

$$\mu=\frac{(1-|\Gamma_s|^2)(1-|\Gamma_r|^2)}{|1-\Gamma_s\Gamma_r|^2}. \tag{A5}$$

From (A4) and (A5), we have

$$Y=\frac{\mu_2(T_2-T_a)+\alpha_t(0)(T_a+\tau)(1-|\Gamma_r|^2)}{\mu_1(T_1-T_a)+\alpha_t(0)(T_a+\tau)(1-|\Gamma_r|^2)}. \tag{A6}$$

From (A6) and (A1), the effective noise temperature of the receiver in the input matched condition is derived in the form

$$T_{er}(0)=\alpha_t(0)(T_a-\tau)-T_a = \frac{\mu'_2T_2-Y\mu'_1T_1}{Y-1}-\frac{(\mu'_2-1)-Y(\mu'_1-1)}{Y-1}T_a \tag{A7}$$

where μ'_1 and μ'_2 are given by

$$\mu'_1=\frac{1-|\Gamma_{s1}|^2}{|1-\Gamma_{s1}\Gamma_r|^2} \tag{A8}$$

$$\mu'_2=\frac{1-|\Gamma_{s2}|^2}{|1-\Gamma_{s2}\Gamma_r|^2}. \tag{A9}$$

Recalling that

$$F=\frac{T_e}{T_0}-1 \tag{A10}$$

from (A7) we obtain the receiver noise figure

$$F_r(0)=\frac{\mu'_2T_2-Y\mu'_1T_1}{(Y-1)}-\frac{(\mu'_2-1)-Y(\mu'_1-1)}{(Y-1)T_0}T_a+1. \tag{A11}$$

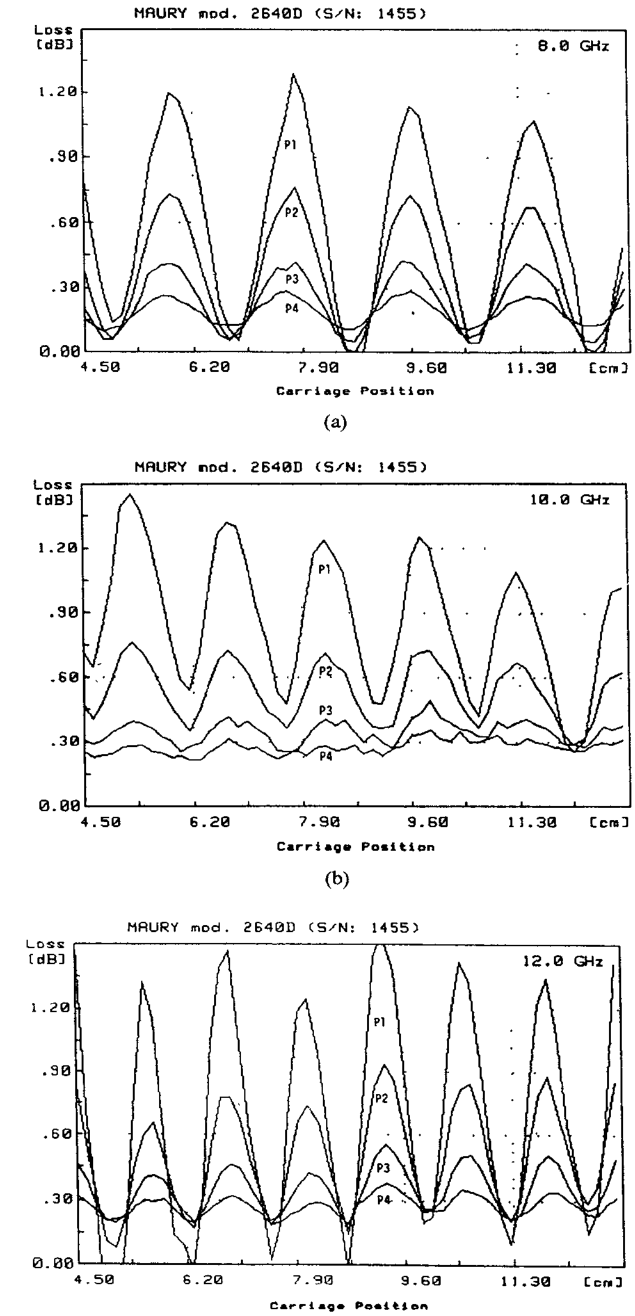

Fig. 3. Loss of the Maury Microwave 2640D tuner versus carriage position for four spacings P_1–P_4 of one screw from the inner conductor with the other screw fully extracted. here $P_1=0.050$, $P_2=0.150$, $P_3=0.350$ and $P_4=0.650$ in. The frequencies are (a) 8.0 GHz, (b) 10.0 GHz, and (c) 12.0 GHz.

If $T_1=T_a$, (A11) becomes

$$F_r(0)=\frac{\mu'_2T_2-(\mu'_2-1)T_a-YT_a}{(Y-1)T_0}+1 \tag{A12}$$

which, after some algebra, can be written

$$F_r(0)=\mu'_2F_r(\Gamma_s)+(\mu'_2-1)\left(\frac{T_a}{T_0}-1\right) \tag{A13}$$

where

$$F_r(\Gamma_s)=\frac{T_2-YT_1}{(Y-1)T_0}+1 \tag{A14}$$

represents the receiver noise figure corresponding to the *nominal* value Γ_s of the input termination reflection coefficient, whose actual value is Γ_{s1} or Γ_{s2}.

References

[1] R. Q. Lane, "The determination of device noise parameters," *Proc. IEEE*, vol. 57, pp. 1461–1462, Aug. 1969.

[2] M. Sannino, "On the determination of device noise and gain parameters," *Proc. IEEE (Lett.)*, vol. 67, pp. 1364–1366, Sept. 1979; also in *Low-noise Microwave Transistors and Amplifiers*, H. Fukui, Ed. New York: IEEE Press, 1981.

[3] G. Caruso and M. Sannino, "Computer-aided determination of microwave two-port noise parameters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 639–642, Sept. 1978.

[4] M. Sannino, "Simultaneous determination of device noise and gain parameters through noise measurements only," *Proc. IEEE (Lett.)*, vol. 68, pp. 1343–1345, Oct. 1980.

[5] G. Martines and M Sannino, "Determination of microwave transistor noise and gain parameters through noise-figure measurements only," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, pp. 1255–1259, Aug. 1982.

[6] E. Calandra, G. Martines, and M. Sannino, "Characterization of GaAs FET's in terms of noise, gain, and scattering parameters through a noise parameter test set," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 231–237, Mar. 1984.

[7] G. Martines and M. Sannino, "Simultaneous determination of transistor noise, gain, and scattering parameters for amplifier design through noise figure measurements only," *IEEE Trans. Instrum. Meas.*, vol. IM-34, pp. 89–91, Mar. 1985.

[8] G. F. Engen, "A new method of characterizing amplifier noise performance," *IEEE Trans. Instrum. Meas.*, vol. IM-19, pp. 344–349, Nov. 1970.

[9] V. Adamian and A. Uhlir, "A novel procedure for receiver noise characterization," *IEEE Trans. Instrum. Meas.*, vol. IM-22, pp. 181–182, June 1973.

[10] V. Adamian and A. Uhlir, "Simplified noise evaluation of microwave receivers," *IEEE Trans. Instrum. Meas.*, vol. IM -33, pp. 136–140, June 1984.

[11] E. W. Strid, "Measurements of losses in noise matching networks," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 247–253, Mar 1981.

[12] M. Sannino, "Comments on 'Simultaneous determination of device noise and gain parameters through noise measurements only'," reply of the author," *Proc. IEEE (Lett.)*, vol. 70, pp. 100–101, Jan. 1982.

[13] G. Mamola and M. Sannino, "Source mismatch effects on measurements of linear two-port noise temperatures," *IEEE Trans. Instrum. Meas.*, vol. IM-24, pp. 239–242, Sept. 1975.

[14] G. Mamola and M. Sannino, "Source mismatch effects on noise measurements and their reduction," *Alta Frequenza*, vol. 44, pp. 233–239, May 1975.