Optimum Noise-Source Reflection-Coefficient Design with Feedback Amplifiers

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***Abstract*—The issue of designing a low-noise microwave feedback amplifier for a given optimum noise-source coefficient $\Gamma_{S_{\rm opt}}$ is addressed and a set of original formulas is presented. These expressions define a new procedure which does not rely on computer optimization in order to get the required noise performance of the low-noise amplifier stage. The technique permits the design of a circuit which is simultaneously noise and power matched at its input port without an input matching circuit. This method can be used to screen devices for an optimum noise performance and it provides the essential mathematical tool for designing the core of a feedback amplifier.**

***Index Terms*—Amplifier noise, circuit noise, feedback amplifiers, feedback circuits, microwave amplifiers, noise.**

I. Introduction

The design of low-noise amplifiers has been investigated widely [1]–[3]; feedback is often cited as the method to move the optimum noise reflection coefficient $\Gamma_{S_{\rm opt}}$ on the Smith chart. Feedback amplifiers have been analyzed in the past [4]–[8]. Parallel feedback [9] has been shown to allow wider band response [10], [11] as well as to improve input $|\Gamma_{\rm in}|$ and output $|\Gamma_{\rm out}|$ return losses [12]; series feedback has been experimentally demonstrated to provide low input return loss and $\Gamma_{\rm in} \simeq (\Gamma_{S_{\rm opt}})^\star$ simultaneously [13], [14]. Today, computer optimization is applied to low-noise amplifiers in order to determine the series feedback value [15].

This paper develops some expressions for the noise parameters of the feedback amplifier and then addresses the issue of designing for either a specified value of $\Gamma_{S_{\rm opt}}$ or $|\Gamma_{S_{\rm opt}}| < \epsilon$. The aim is to achieve $\Gamma_{\rm in} = (\Gamma_{S_{\rm opt}})^\star = 0$ for a microwave amplifier without an input matching circuit. According to the correlation matrix noise theory [16], the transmission representation matrix $\mathbf{C}_n$ of the cascaded circuits is

$$\mathbf{C}_n = \mathbf{C}_M + \mathbf{T}_M \mathbf{C}_A \mathbf{T}_M{}^+ \tag{1}$$

where the subscript M refers to the input matching circuit, A to the following amplifier, $\mathbf{C}$'s are correlation matrices, $\mathbf{T}_M$ is the matching circuit transmission matrix, and $^+$ is the Hermitian conjugate operation. The stages driven by the amplifier are neglected in (1) on the basis that the amplifier gain can reduce their noise contribution [17].

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Equation (1) demonstrates that the elements of the matrix $\mathbf{C}_n$ are nonlinear combinations of the signal and the noise parameters of the cascaded stages. Direct control of $\mathbf{C}_n$ is therefore very difficult. The design is simplified by removing the input matching network: (1) then simply becomes $\mathbf{C}_n = \mathbf{C}_A$.

II. Expressions for the Design

The equations involving the noise parameters are written as functions of the elements of $\mathbf{C}_A$:

$$\mathbf{C}_A = \begin{bmatrix} R_n^A & \rho_o^{A\star} \\ \rho_o^A & g_n^A \end{bmatrix}, \qquad \text{where } \rho_o^A = \rho_n^A \sqrt{g_n^A R_n^A}.$$

The term $4kT_o\Delta f$ has been dropped.

A. Expression for a Given $\Gamma_{S_{\rm opt}}^A$

Suppose that an optimum source reflection coefficient $\Gamma_{S_{\rm opt}}^A$ has to be achieved. According to [18]

$$\sqrt{(G_c^A)^2 + \frac{G_n^A}{R_N^A}} - jB_c^A = Y_{S_{\rm opt}}^A \tag{2}$$

where $Y_{S_{\rm opt}}^A = G_{S_{\rm opt}}^A + jB_{S_{\rm opt}}^A$ is the admittance which corresponds to $\Gamma_{S_{\rm opt}}^A$, $Y_c^A = G_c^A + jB_c^A$ is the correlation admittance of the stage, G_n^A and R_n^A are its uncorrelated noise conductance and resistance. After rewriting

$$\begin{aligned} Y_c^A &= \rho_n^A \sqrt{\frac{g_n^A}{R_n^A}} \\ &= \rho_n^A \frac{\sqrt{g_n^A R_n^A}}{R_n^A} \\ &= \frac{\rho_o^A}{R_n^A} \end{aligned} \tag{3}$$

$$G_n^A = g_n^A - |Y_c^A|^2 R_n^A \tag{4}$$

and substituting (3) and (4) into (2), the system

$$\Im m[\rho_o^A] = -B_{S_{\rm opt}}^A R_n^A \tag{5a}$$

$$g_n^A = |Y_{S_{\rm opt}}^A|^2 R_n^A \tag{5b}$$

is obtained.

System (5) can be solved for two unknowns. Equation (5a) states that real optimum source reflection coefficients (e.g., $\Gamma_{S_{\rm opt}}^A = 0$) require $\Im m[\rho_o^A] = 0$.

B. Expression for $|\Gamma^A_{S_{\rm opt}}| \le \epsilon$

Suppose the goal is

$$|\Gamma^A_{S_{\rm opt}}| \le \epsilon \le 1. \tag{6}$$

Equation (6) defines a circle on the normalized admittance plane $y^A_{S_{\rm opt}} = Y^A_{S_{\rm opt}}/Y_o$:

$$|y^A_{S_{\rm opt}} - C_\epsilon| \le R_\epsilon \tag{7}$$

where

$$C_\epsilon = \frac{1+\epsilon^2}{1-\epsilon^2}$$

and

$$R_\epsilon = \frac{2\epsilon}{1-\epsilon^2}.$$

If (2) is substituted into (7), and (3) and (4) are used, the resulting general expression is

$$(g^A_n Z_o)^2 + f^2_\epsilon (R^A_n Y_o)^2 - 2h_\epsilon (g^A_n Z_o)(R^A_n Y_o) + 4C^2_\epsilon \Im m[\rho^A_o]^2 = 0 \tag{8}$$

where

$$f_\epsilon = C^2_\epsilon - \eta R^2_\epsilon$$

and

$$h_\epsilon = C^2_\epsilon + \eta R^2_\epsilon$$

η is a parameter, ranging from 0 to 1, which transforms the inequality (7) into an equation ($R^2_\epsilon \to \eta R^2_\epsilon$) and is useful for software implementation. One unknown can solve (8).

III. Expansion for the Feedback Amplifier

Expressions (5) and (8) will be expanded as functions of the noisy feedback elements $Z_s = R_s + jX_s$ and $Y_p = G_p + jB_p$ of a feedback amplifier (see Fig. 1). R^A_n, g^A_n, and $\rho^A_o = \rho^A_n \sqrt{g^A_n R^A_n}$ have been derived in [19] as functions of the feedback elements.

A. Expansion for a Given $\Gamma^A_{S_{\rm opt}}$

Substituting R^A_n, g^A_n, and ρ^A_o into (5a) and (5b) results in the system:

$$\begin{aligned} &kA_{11} g_p x^2_s + kA_{11} g_p r^2_s \\ &\quad + kA_{10} r^2_s + kA_{10} x^2_s + (kA_{21} + D_{rg}) r_s g_p \\ &\quad + (kA_{31} + D_{xg}) g_p x_s + D_{xb} x_s b_p \\ &\quad + D_{rb} r_s b_p + (kA_{20} + D_r) r_s + (kA_{30} + D_x) x_s \\ &\quad + (kA_{41} + D_g) g_p + D_b b_p \\ &\quad + (kr_{t_n} + D_o) = 0 \end{aligned} \tag{9a}$$

$$\begin{aligned} &qA_{11} g_p x^2_s + qA_{11} g_p r^2_s - B_{11} r_s b^2_p - B_{11} r_s g^2_p \\ &\quad + qA_{10} r^2_s - B_{10} g^2_p + qA_{10} x^2_s - B_{10} b^2_p \\ &\quad + (qA_{21} - B_{21}) r_s g_p - B_{31} r_s b_p \\ &\quad + qA_{31} g_p x_s + (qA_{20} - B_{41}) r_s + qA_{30} x_s \\ &\quad + (qA_{41} - B_{20}) g_p - B_{30} b_p \\ &\quad + (qr_{t_n} - g_{t_n}) = 0 \end{aligned} \tag{9b}$$

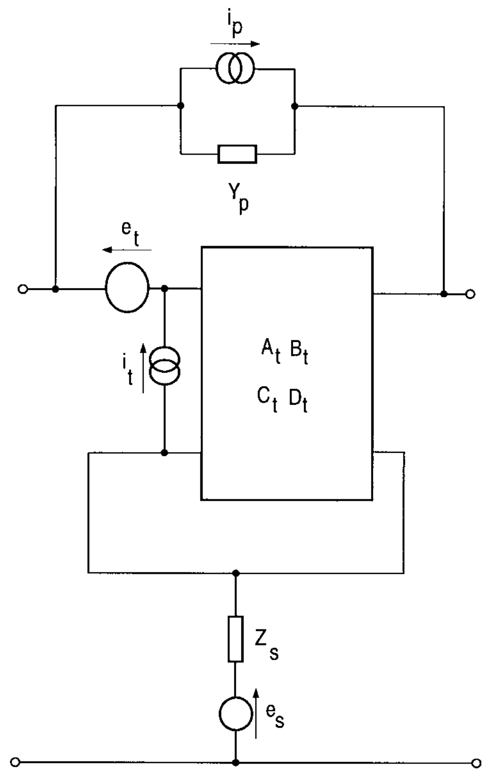

Fig. 1. Schematic of noisy two-port with series and parallel feedback.

The unknowns are $r_s = \Re e[Z_s/Z_o]$, $x_s = \Im m[Z_s/Z_o]$, $g_p = \Re e[Y_p/Y_o]$, and $b_p = \Im m[Y_p/Y_o]$. $Z_o = 1/Y_o$ is the characteristic impedance of the system. The coefficients of (9) are

$$\begin{aligned} k &= \Im m[Y^A_{S_{\rm opt}}] Z_o \\ q &= \left(\frac{|Y^A_{S_{\rm opt}}|}{Y_o}\right)^2 \\ \Delta &= 1 - a - d - (A_t D_t - B_t C_t) \\ a &= 1 - A_t \\ d &= 1 - D_t \\ \rho_{o_t} &= \rho_t \sqrt{R_t g_t} \\ A_{10} &= (g_t |a|^2 + R_t |C_t|^2 + 2\Re e[a\rho_{o_t} C^\star_t]) Z_o \\ A_{11} &= |\Delta|^2 \\ A_{20} &= |a|^2 + 2\Re e[a\rho_{o_t}] + 2R_t \Re e[C_t] \\ A_{21} &= 2\Re e[\Delta B^\star_t] Y_o \\ A_{30} &= -2\Im m[C_t R_t + a\rho_{o_t}] \\ A_{31} &= -2\Im m[\Delta B^\star_t] Y_o \\ A_{41} &= (|B_t| Y_o)^2 \\ r_{t_n} &= \frac{R_t}{Z_o} \\ B_{10} &= R_t |d|^2 + g_t |B_t|^2 + 2\Re e[d\rho^\star_{o_t} B^\star_t] Y_o \\ B_{11} &= |\Delta|^2 \\ B_{20} &= |d|^2 + 2\Re e[d\rho^\star_{o_t}] + 2g_t \Re e[B_t] \\ B_{21} &= 2\Re e[\Delta C^\star_t] Z_o \\ B_{30} &= -2\Im m[B_t g_t + d\rho^\star_{o_t}] \end{aligned}$$

$$
\begin{aligned}
B_{31} &= -2\Im m[\Delta C_t^\star] Z_o \\
B_{41} &= (|C_t| Z_o)^2 \\
g_{t_n} &= \frac{g_t}{Y_o} \\
c_1 &= g_t a^\star + \rho_{o_t} C_t^\star \\
c_2 &= g_t a^\star B_t + \rho_{o_t} C_t^\star B_t + \rho_{o_t}^\star a^\star d + R_t\, d C_t^\star \\
c_3 &= \rho_{o_t} B_t + R_t\, d \\
c_4 &= -d\Delta^\star \\
c_5 &= -a^\star \Delta \\
c_6 &= -B_t^\star\, d \\
c_7 &= -C_t a^\star \\
D_{rg} &= \Im m[c_2 + c_5 + c_4] \\
D_{xb} &= \Im m[c_2] \\
D_{rb} &= \Re e[c_2 + c_5] \\
D_{xg} &= -\Re e[c_2 + c_4] \\
D_r &= \Im m[(c_1 + c_7) Z_o] \\
D_x &= -\Re e[c_1 Z_o] \\
D_g &= \Im m[(c_3 + c_6) Y_o] \\
D_b &= \Re e[c_3 Y_o] \\
D_o &= \Im m[\rho_{o_t}]
\end{aligned}
$$

k and q specify $\Gamma_{S_{\rm opt}}^A$ to be achieved; $A_t B_t C_t D_t$ is the transmission matrix of the transistor, R_t, g_t, and ρ_t its noise parameters. The subscript t refers to the transistor of the feedback amplifier stage.

B. Expansion for $|\Gamma_{S_{\rm opt}}^A| \leq \epsilon$

The expansion is carried out for the particular case of a reactive feedback amplifier ($Z_s = jX_s$, $Y_p = 0$) because this configuration is widely used for achieving a simultaneous match between the input reflection coefficient $\Gamma_{\rm in}^A$ and the conjugate of the optimum source reflection coefficient $(\Gamma_{S_{\rm opt}}^A)^\star$. Thus, (8) is expanded into

$$\sum_{i=0}^{4} \alpha_i x_s^i = 0 \tag{10}$$

where

$$
\begin{aligned}
\alpha_4 &= f_\epsilon^2 A_{10}^2 \\
\alpha_3 &= 2 f_\epsilon^2 A_{10} A_{30} \\
\alpha_2 &= f_\epsilon^2 A_{30}^2 + 2 f_\epsilon^2 A_{10} r_{t_n} + 4 C_\epsilon^2 D_x^2 - 2 h_\epsilon g_{t_n} A_{10} \\
\alpha_1 &= 2 f_\epsilon^2 A_{30} r_{t_n} - 8 C_\epsilon^2 \rho_y D_x - 2 h_\epsilon g_{t_n} A_{30} \\
\alpha_0 &= f_\epsilon^2 r_{t_n}^2 + 4 C_\epsilon^2 \rho_y + g_{t_n}^2 - 2 h_\epsilon g_{t_n} r_{t_n} \\
\rho_y &= \Im m[\rho_{o_t}].
\end{aligned}
$$

The unknown is $x_s = \Re e[Z_s/Z_o]$. The α_i's are defined in terms of the coefficients of (9).

IV. Discussion

Some observations about the system (9) are as follows.

- The set of (9a) and (9b) allows determination of the values of the feedback elements for a circuit to provide a given $\Gamma_{S_{\rm opt}}^A$ at the design frequency. No control on other stage parameters is exerted by (9).
- System (9) is nonlinear.
- System (9) has more unknowns than equations.

An exact solution of (9) can be formally derived by setting to zero two of the four variables r_s, x_s, g_p, and b_p and then substituting one equation into the other. Table I shows the number N_s of expected solutions as a function of the unknowns chosen. Some of them may be physically meaningless—for example, a solution (x_s; b_p) can be complex. The desired pair of feedback elements may not exist or may not be achievable at certain frequencies. However, the procedure applied to a number of different commercially available MESFET's has always found a numerical solution for a given $\Gamma_{S_{\rm opt}}^A$. The solution involving a resistive element is expected to correspond to a higher minimum noise figure than the one which makes use of reactive elements only; nonetheless, Table III demonstrates a decrease in $F_{\rm min}$ can result.

TABLE I
NUMBER N_s OF SOLUTIONS OBTAINABLE FROM (9) AFTER SETTING A PAIR OF UNKNOWNS TO ZERO: THE SYMBOL X SHOWS THE CHOSEN UNKNOWNS

r_s	x_s	g_p	b_p	N_s
X	X	–	–	2
X	–	X	–	6
X	–	–	X	5
–	X	X	–	6
–	X	–	X	4
–	–	X	X	2

TABLE II
EXPECTED MINIMUM IN $|\Gamma_{S_{\rm opt}}^A|$ FOR THREE HEWLETT PACKARD MESFET'S AND A THIRD-DEGREE POLYNOMIAL LEAST SQUARES APPROXIMATION (N.A.: DATA BOOK PARAMETERS NOT AVAILABLE)

GHz	ATF21186	ATF35376	ATF10136
0.5	0.024	N.A.	N.A.
1.0	0.057	N.A.	0.296
2.0	0.213	0.418	0.079
4.0	0.452	0.080	0.276
6.0	0.639	0.144	0.360
8.0	0.800	0.271	0.251

TABLE III
SOME OF THE SOLUTIONS ACHIEVABLE WITH ATF21186 @ 1 GHz FOR $\Gamma_{S_{\rm opt}} = 0.1e^{j45^\circ}$ (THE FIRST ROW SHOWS THE DEVICE PERFORMANCE WITHOUT FEEDBACK)

R_s/Z_o	X_s/Z_o	G_p/Y_o	B_p/Y_o	F_{min} [dB]	G_{av} [dB]
0	0	0	0	0.55	15.1
0.0266	3.2737	0	0	0.46	4.9
0.5864	0	0.7551	0	17.5	-13.8
$1.6505\ 10^3$	0	0	6.1704	0.01	0.0
0	0.3492	0.2466	0	3.49	4.5
0	3.1565	0	-0.0890	0.50	9.4
0	0	0.3219	-0.2325	4.57	5.0

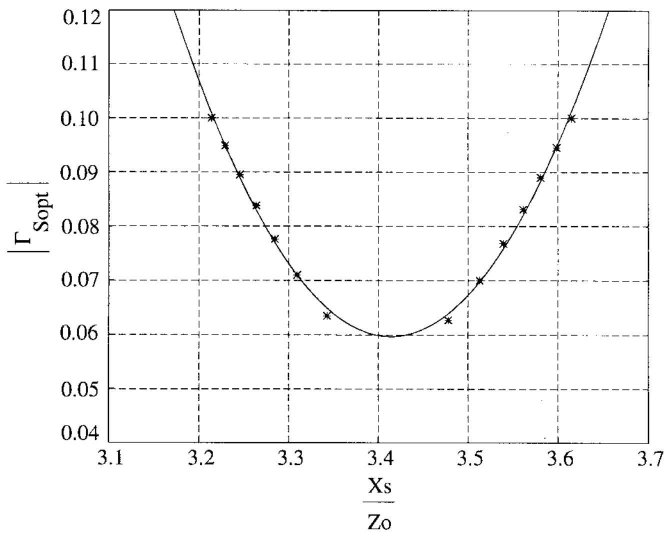

Fig. 2. Computed $\Gamma^A_{S_{\rm opt}}$ (*) versus normalized series feedback x_s with $\epsilon = 0.1$ and $Y = ax_s^2 + bx_s + c$ (solid line; $a = 1.0303$, $b = -7.0350$, $c = 12.0685$) for a Hewlett Packard ATF21186 GaAs MESFET at 1 GHz: $|\Gamma^A_{S_{\rm opt}}|_{\min} = 0.0596$ @ $x_s = -b/2a = 3.4141$.

TABLE IV
AMPLIFIER DESIGN VALUES FOR AN HP ATF21186 AT 1 GHz, $\epsilon = 0.1$

x_s	3.43	–
L_s	27.29	nH
$\mid\Gamma^A_{S_{opt}}\mid$	0.06	–
$\angle\Gamma^A_{S_{opt}}$	-164.81	deg
F^A_{min}	0.39	dB
R^A_n	1.18	Ω
$\mid\Gamma^{SSNM}_L\mid$	0.91	–
$\angle\Gamma^{SSNM}_L$	9.05	deg
G^A_{av}	4.80	dB
G^A_T	4.73	dB

The procedure for designing either $\Gamma^A_{S_{\rm opt}}$ or $|\Gamma^A_{S_{\rm opt}}| \le \epsilon$ is outlined below.

1) Choose a pair of unknowns and solve system (9) for the given $\Gamma^A_{S_{\rm opt}}$ or solve (10) for the given ϵ.
2) For each acceptable solution work out the signal and noise parameters.
3) Calculate the value of the load which allows to get the input reflection coefficient $\Gamma^A_{\rm in} = (\Gamma^A_{S_{\rm opt}})^\star$, where $\star$ is the conjugate operation; this particular load is

$$\Gamma_L^{\rm SSNM} = \frac{S_{11} - \Gamma^\star_{S_{\rm opt}}}{\Delta - S_{22}\Gamma^\star_{S_{\rm opt}}}$$

where Δ is the determinant of the scattering matrix of the stage—transistor plus feedback elements. SSNM is the acronym for simultaneously signal and noise matched.
4) Find the transducer power gain G_T when $\Gamma_L^{\rm SSNM}$ loads the output along with other signal and noise parameters as desired.
5) If the required circuit performance is not satisfied, rerun this procedure with a different set of unknowns.

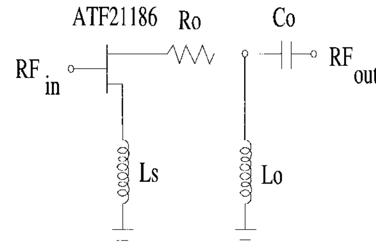

Fig. 3. Circuit layout for frequency domain simulation: $R_o = 10\ \Omega$, $L_o = 27.61$ nH, $C_o = 0.69$ pF for the output matching circuit.

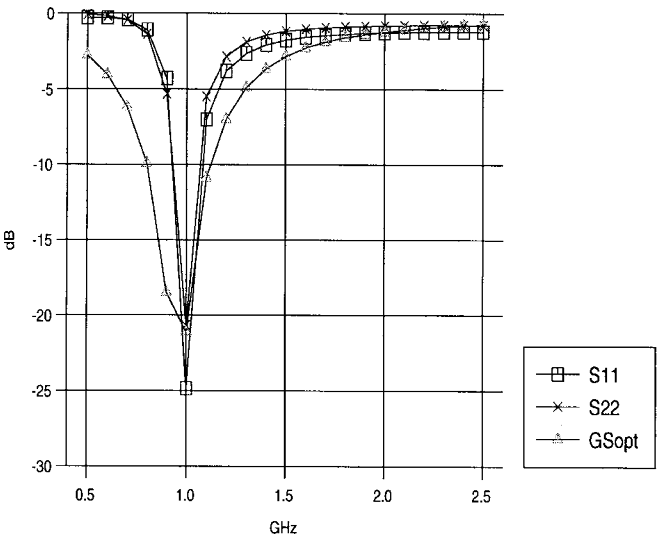

Fig. 4. Frequency dependence of the input return loss (S_{11}), the output return loss (S_{22}), and the optimum noise reflection coefficient ($GS_{\rm opt}$) for the designed circuit.

When a series reactive feedback is considered, $\Gamma^A_{S_{\rm opt}}$ has a minimum in magnitude. Equation (10) suggests a way to find this minimum. A least squares method may successfully be applied in order to evaluate this minimum (see Fig. 2). If $\epsilon < |\Gamma^A_{S_{\rm opt}}|_{\min}$, a different device must be selected (see Table II); the input insertion loss when the simultaneous match $\Gamma_{\rm in} = (\Gamma_{S_{\rm opt}})^\star$ is achieved cannot be better than $|\Gamma^A_{S_{opt}}|_{\min}$.

This procedure has been applied to a Hewlett Packard ATF21186 low-noise GaAs MESFET [20]. Table IV collects the design results for the circuit shown in Fig. 3. Finally, a simulation in the frequency domain has been carried out as shown in Figs. 4–6.

The simulation at the design frequency gives the same response as the calculations described above. The device is inherently unstable and this stability is usually further degraded by the calculated feedback elements. Both resistive and reactive components have to be properly added to the circuit in order to control the input and output return loss and restrain the amplifier from oscillating. Since this will affect $\Gamma^A_{S_{\rm opt}}$, the number of circuit components should be kept as small as possible and should preferably be added after the transistor.

The output stage has the main task of providing the necessary $\Gamma_L^{\rm SSNM}$ at its input port when loaded at its output by 50

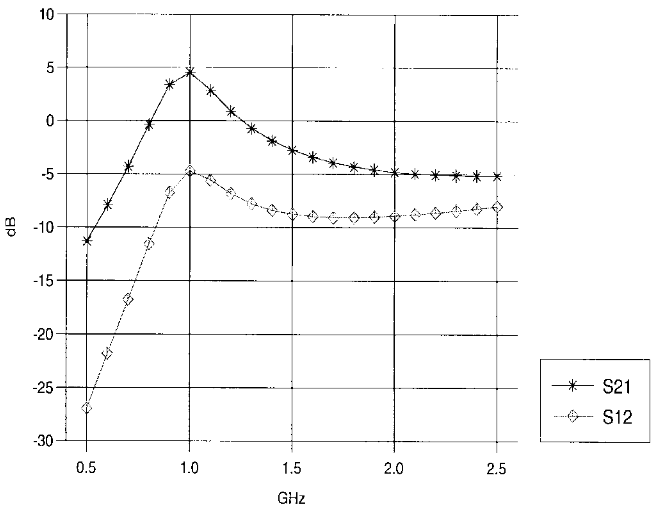

Fig. 5. Frequency dependence of the forward (S_{21}) and the inverse (S_{12}) transmission coefficients for the designed circuit.

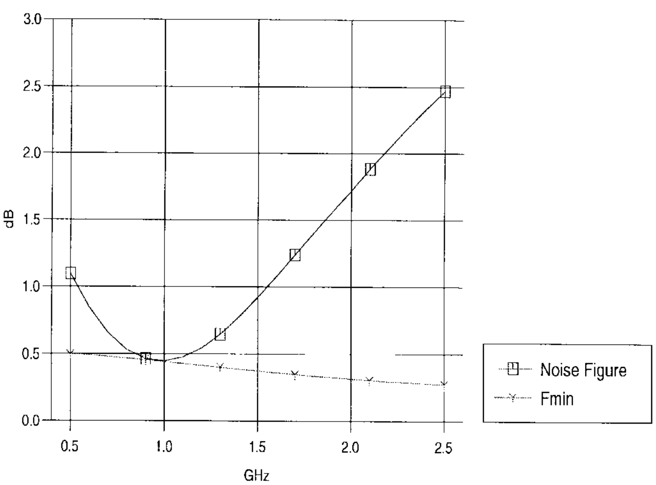

Fig. 6. Frequency dependence of the noise figure and the minimum noise figure ($F_{\min}$) for the designed circuit.

Ω. A resistor R_o can improve the stability without affecting the noise performance of the stage (see Fig. 3). Usually, the calculated $\Gamma_L^{\rm SSNM}$ is close to $\Gamma_L = (\Gamma_{\rm out})^\star$, the load at which the output port is power matched. Thus, the transducer power gain G_T^A of the feedback amplifier is close to its available power gain G_{av}^A.

The design must be considered as a starting point for a subsequent optimization. The optimization is required because this design does not take into account every physical component or the parasitic elements of the complete circuit. Frequency dependent elements have to be added to the network in order to improve its stability. Transmission lines to the active device input port are particularly important [21] because they have a large effect on the noise parameters. The design seems to be sensitive to these elements even if the input line is very short. However, the optimization at the design frequency is able to achieve the required $\Gamma_{S_{\rm opt}}$.

The authors are not aware of any other analytical technique to directly control $\Gamma_{S_{\rm opt}}$. These expressions are valid for either active and passive linear two-ports with feedback elements.

V. Conclusion

Original expressions for designing either a given $\Gamma_{S_{\rm opt}}$ or $|\Gamma_{S_{\rm opt}}| \leq \epsilon$ are derived and applied to a feedback amplifier. These formulas allow the design of a circuit simultaneously matched at its input port $\Gamma_{\rm in} = (\Gamma_{S_{\rm opt}})^\star = 0$ without the need of an input matching circuit. When a reactive series feedback is used, the procedure can select the most suitable device since a minimum value of $|\Gamma_{S_{\rm opt}}|$ as a function of the feedback exists. These equations apply to any linear noisy two-ports with feedback elements.

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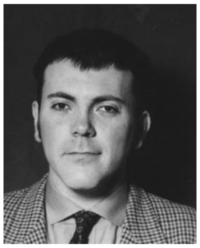

Luciano Boglione (S'96) was born in Turin, Italy. He received the Laurea degree in electronic engineering from the Politecnico di Torino, Turin, Italy, in 1991. He is currently pursuing his Ph.D. degree from The University of Leeds, Leeds, U.K.

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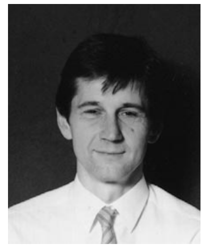
He holds the Chair in high-frequency measurements in the Department of Electronic and Electrical Engineering, The University of Leeds, where he has been a Faculty Member since 1974 as well as being a Consultant to the Hewlett-Packard Company, Santa Rosa, CA, since 1981. He is jointly responsible for the activities of the Microwave and Terahertz Technology Research Group which has over 40 active researchers, a strong graduate program, and has made contributions to microwave passive and active device research. His personal research interests are in microwave network measurements, calibration and error correction, microwave and millimeter-wave circuits, and large-signal and nonlinear characterization.

Dr. Pollard is a Chartered Engineer and a member of the Institution of Electrical Engineers (IEE), U.K. He is serving his second term as an elected member of the IEEE MTT-S Administrative Committee, where he is 1997 Vice President.



Vasil Postoyalko (S'82–M'82) was born in Leeds, U.K., in 1956. He received the B.Sc. degree in mathematics and Ph.D. degree in electronic engineering and applied mathematics from The University of Leeds, Leeds, U.K., in 1978 and 1985, respectively.

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