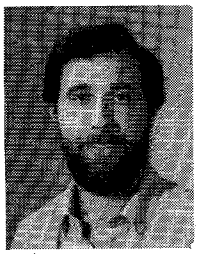
[11] M. Sannino, "On the determination of device noise and gain parameters," *Proc. IEEE* (*Letters*), vol. 67, pp. 1364–1366, Sept. 1979.

[12] 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.

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

[14] M. Mitama and H. Katoh, "An improved computational method for noise parameter measurements," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 612–615, June 1979.

[15] J. Lange, "Noise characterization of linear two-ports in terms of invariant parameters," *IEEE J. Solid-State Circuits*, vol. SC-2, pp. 37–40, June 1967.



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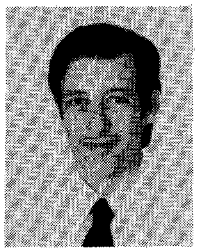

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Stability Margins in Microwave Amplifiers

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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short produces capacitive or inductive reactances ranging in value from 0 to ∞ ohms at any given frequency, i.e., it covers only the perimeter of the Smith Chart from short to open-circuit. Such terminations are those most likely to cause oscillation because, ideally, they add no damping. Experience has not shown instability to occur in amplifiers under this test if $K > 1$ over the entire frequency range of interest. However, this test does not measure feedback and, indeed, there is only one case known to the author of published [1] values of feedback computed from the circuit diagram of a microwave amplifier.

The purpose of this paper is to provide designers of microwave amplifiers and systems with a new method of stability assessment based upon the total feedback present in a given physical amplifier or circuit diagram. The significant features are as follows.

1) A theory, proven at frequencies below 1 GHz, is extended to at least 18 GHz.

2) State-of-the-art commercially available automatic network analyzers (ANA), transistor test fixtures, and precision microwave probes enable accurate and repeatable measurements to be made on transistors and circuits.

3) Theory and measurements combine, resulting in a more meaningful criterion of stability based on the total amount of feedback actually present.

To obtain the truest assessment of feedback it is necessary to use measured S-parameters which take account of strays and those largely indefinable departures from theoretical predictions encountered at microwave frequencies. With some sacrifice in realism, such as might be appropriate at the design or feasibility stages, circuit models of active and passive components, tracks, etc., can be used in place of measurements on an actual amplifier. Data from either of these sources is combined with the author's recent developments of H. W. Bode's feedback theory to compute a transfer function analogous to loop-gain employed at lower frequencies.

The benefits claimed for negative feedback have been listed [2] but cannot be treated quantitatively without a sound measure of feedback such as the present paper gives. It appears that the amounts of feedback obtainable at microwave frequencies are relatively modest, so one would expect the benefits to be likewise, but this has not yet been investigated.

The new method is outlined briefly in Section II, using a simple but commonly used example of a feedback amplifier. A broad-band representation for a transistor is introduced which is applicable to 3-terminal devices, including pre-matched FET's (i.e., those in which several matching elements are included within the package) at one extreme, to chips at the other. A sequence of five steps leads from measured S-parameters to computed feedback, and the Nyquist criterion for self-oscillation. The detailed results of applying the method to an amplifier over the band 2 to 18 GHz forms Section III. A number of related points of interest are described in Section IV, and a brief discussion of recent stability criteria including Rollett's is given in the Appendix.

II. The Embedding Network Method

Recently, a key contribution to the circuit designer's art has been the publication [3] of an "in depth presentation of a new technique for assessing amplifier stability margins." It has been called the embedding network method because attention is focused on one transistor and the remainder of the amplifier in which this active device is embedded. Developments of H. W. Bode's feedback theory are combined with S-parameter measurements made by automated network analyzers. These provide a firm theoretical foundation second-to-none, and the most practical and accurate measurements in the frequency bands of interest (e.g., 2–18 GHz).

Microwave amplifier designers are familiar with ANA's such as the Hewlett-Packard 8409 B/C, but are probably not familiar with the theory mentioned earlier. It provides a rigorous definition of feedback in terms of an output-to-input transfer function called return ratio (RR). The feedback (analog of $1\text{-}\mu\beta$ in conventional terms) is then $1 + RR$. The transfer function can be chosen to give the total (internal, local, and loop) feedback around the chosen transistor; specifically, the feedback around the controlled source representing the nonreciprocal gain (analog of g_m). Instead of using the well-known forms of hybrid-pi circuit models—which have their drawbacks [3]—a simpler representation is used comprising a parallel connection of a 'pi' of admittances and a 2-port voltage-controlled current source, all of which are frequency-dependent quantities uniquely defined by formulas [4] in terms of measured S-parameters. The representation is shown within the dashed-line box in the amplifier circuit diagram of Fig. 1.

The small-signal admittances y_a, y_b, y_c, and y_m are computed from S-parameters measured in a new accurate microwave-transistor test fixture [5]. The transfer function of interest is that which gives the return ratio for the transadmittance $y_m = y_{21} - y_{12}$. It is well known that a physically meaningful quantity like noise figure of linear noisy 2-ports can be computed from two fictitious noise generators and a noise-free 2-port. In a like manner, the physically meaningful feedback can be evaluated by computing the ratio of the returned voltage V' (that across y_a) to a fictitious independent voltage V (instead of the gate–source voltage) controlling the dependent source of transadmittance y_m. By definition, the return ratio for y_m is $-V'/V$, and is analogous to a measurable loop (voltage) gain $\mu\beta$.

The most realistic predictions of stability margins are found when return ratios are computed from data banks obtained from S-parameter measurements on the amplifier under test. Unlike a single set of the four parameters S_{11}, S_{12}, S_{21}, and S_{22} at all frequencies of interest which enable the stability factor K to be calculated, the return ratio requires measurements on the transistor to be used in the amplifier and on its actual embedding network. For the circuit of Fig. 1, the embedding network can be fully characterized by one set of S-parameter measurements taking the input port between gate pad and ground and output between drain pad and ground. If the common

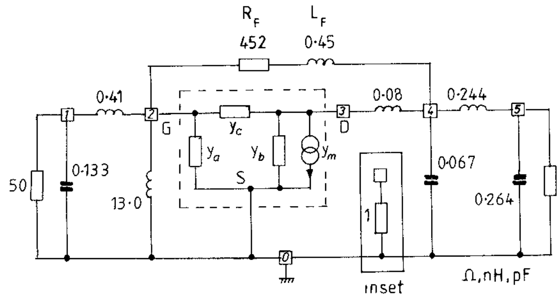

Fig. 1. Simple feedback amplifier using a NE 70083 GaAs MESFET in grounded source configuration.

terminal of the transistor is lifted off ground by inserting a component between source and ground, the resulting embedding network requires [3] three, not one, sets of S-parameter measurements to describe it fully.

The sequence of steps required for the calculation of return ratio using a nodal analysis type of program can be summarized as follows.

1) The primary data consists of a) S-parameters of the actual transistor to be used in the amplifier under test, b) S-parameters of the remainder of the amplifier as seen from the missing transistor. These two 2-ports can easily be identified in Fig. 1 as that within the box and that outside the box.

2) The conversion from error-corrected S-parameters to secondary data comprising files of 2-port admittances using standard formulas [4]. It is convenient to denote these files of admittances by $[TR]$ and $[EN]$, for the transistor and embedding network, respectively.

3) The third step separates the reciprocal and nonreciprocal properties of the transistor by forming

$$[TR]=\begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix}=\begin{bmatrix} y_{11} & y_{12} \\ y_{12} & y_{22} \end{bmatrix}+\begin{bmatrix} 0 & 0 \\ y_{21}-y_{12} & 0 \end{bmatrix} \tag{1}$$

which can be denoted by

$$[TR]=[RTR]+[DS] \tag{2}$$

where RTR stands for reciprocal transistor and DS for dependent source. The relationships between the 2-port admittances in (1) and the admittances in Fig. 1 are simply

$$y_a=y_{11}+y_{12} \quad y_c=-y_{12}$$
$$y_m=y_{21}-y_{12} \quad y_b=y_{22}+y_{12}. \tag{3}$$

4) So far, the only three nodes of interest are the gate, the drain, and the grounded source. An additional node and unit admittance connecting it to ground are introduced as shown in the insert in Fig. 1. The voltage V across this branch controls the dependent source simply by re-assigning the input of the 2×2 admittance file $[DS]$ as this new node rather than the gate node, while leaving the output as the drain.

5) Finally, the return ratio is computed as the voltage ratio $-V'/V$, where V' is the response between gate and ground to the independent voltage V. Evidently, signals can be returned from drain to gate through y_c or through external couplings such as R_F and L_F, and these contributions can be computed separately if desired.

The most accurate answers are obtained when the primary data is taken from measurements on the specific transistor and embedding network of interest, but certainly not from manufacturer's data sheets and simple circuit diagrams. Typical transistor data is commonly obtained by averaging over a few samples, resulting in S-parameters which are fictitious in that they do not correspond to any one physical device. Moreover, they represent historical data which may differ significantly from subsequent batches. Accurate and repeatable measurements of FET's up to 18 GHz are now practical [5]. The measurement of embedding networks is more difficult, but recent work [6] suggests that suitable probes and circuit constructions can be used to facilitate measurements. This is very desirable at frequencies (above say 5 GHz) where the field is no longer simply TEM and where several modes may be present.

The total feedback expressed as a return ratio is plotted, usually in polar form, then examined in terms of the well-known Nyquist criterion for stability against self-oscillation [3].

III. A Broad-Band Feedback Amplifier

For simplicity, the embedding network is taken to be adequately represented by the lumped elements shown in Fig. 1. By lumped element is meant a component whose maximum dimension is, at most, a tenth of a wavelength at the highest operating frequency of interest. Since λ (mm) = $300/f$ (GHz), a component's largest dimension must be smaller than 1.5 mm if 20 GHz is the largest frequency. The range of values of inductance and capacitance shown can be realized by lumped components.

A packaged transistor has been chosen as the active device for several reasons.

1) They can be pre-tested and screened to desired quality levels [2].

2) This form is much more easily handled than a chip [2].

3) S-parameters can be measured on a device under suitable bias conditions and that device then used in the amplifier under test. This cannot be done with a chip transistor.

4) A newly available test fixture [5] for packaged transistors can provide results of an accuracy and repeatability not previously obtainable.

The first published results using this new jig refer to a 0.5-μm recessed gate FET type NE 70083 biased at 3 V and 10 mA. With two corrections, these results have been used as the primary data for the evaluation of feedback. The high quality of this data is evident from plots of the S-parameters (magnitude and phase) but perhaps more strikingly in the admittances derived from them. Another relevant consideration is that this is a commercially available device rather than a laboratory prototype.

Programs based on nodal analysis require that S-parameters be transformed into admittances. For stability investigations, these can best be taken as the branch admittances of the broad-band representation (Fig. 1) of the NE 70083. The real (conductance) and imaginary (susceptance) com-

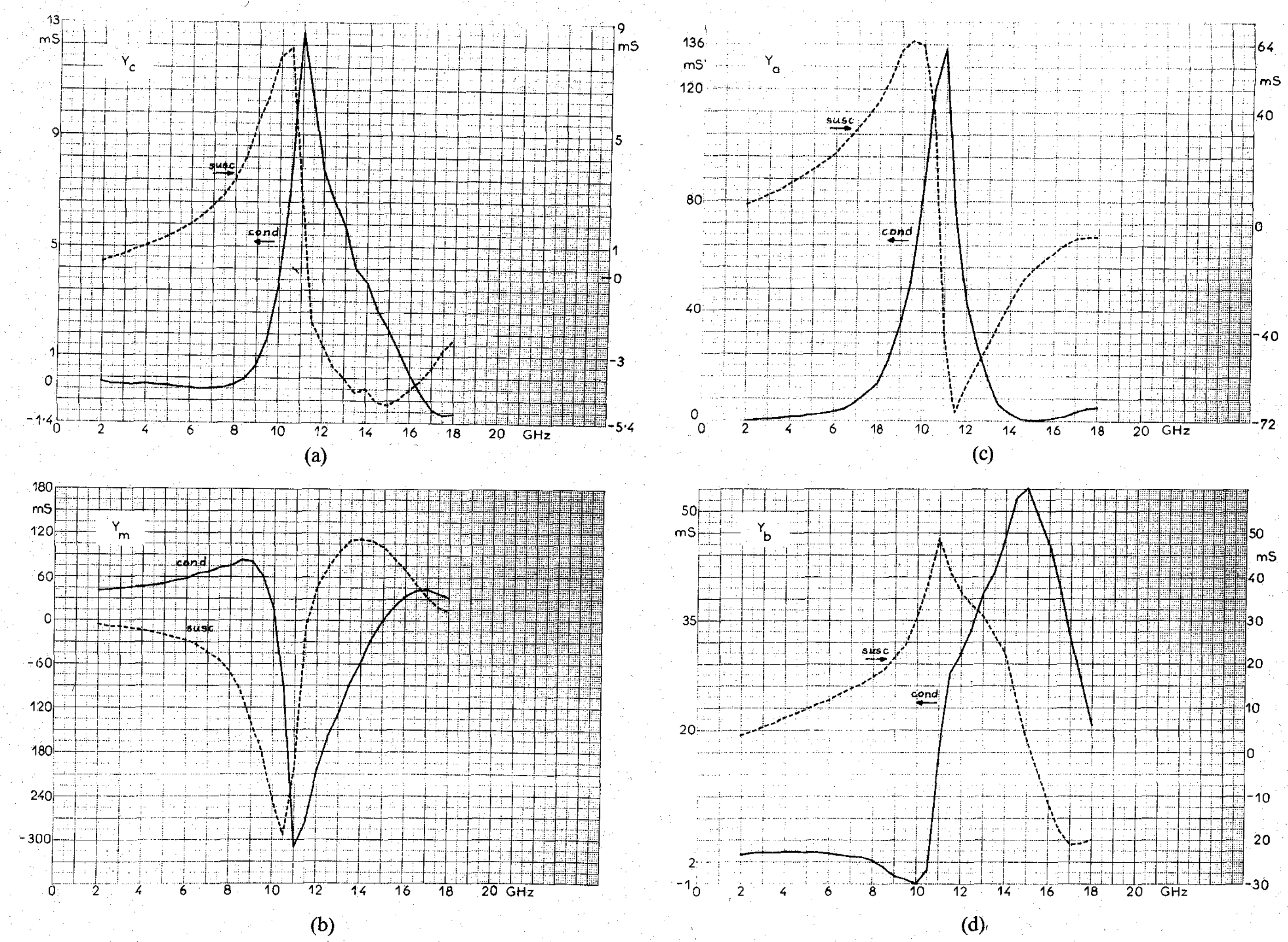

Fig. 2. Broad-band transistor representation admittances computed from measured S-parameters. (a) y_c, (b) y_m, (c) y_a, and (d) y_b. NB. negative conductance of y_b near 10 GHz.

ponents have been calculated then plotted in Fig. 2(a)–(d). These are, in the author's experience, the smoothest curves of branch admittances yet seen, and tend to substantiate the claims [5] made for the new transistor test fixture. The other striking features are the rapid variations in the neighborhood of 11 GHz in all graphs. This has been traced to the frequency dependence of the common denominator of all the expressions for the admittances as functions of the S-parameters. As could be expected, the smallest computed value of the denominator occurs at 11 GHz, where it is some 200 times less than at 2 GHz. Since the denominator is given by

$$(1+S_{11})(1+S_{22})-S_{12}S_{21}$$

it is important to use a set of S-parameters measured on one device, but not values obtained by averaging S-parameters measured on many devices, as primary data.

Turning now to the embedding network, the topology and initial element values were taken from a published circuit [7]. The final values obtained by optimizing the input and output matching networks to the measured NE 70083 are given in Fig. 1. Reasonably flat gain has been achieved from 2 to 11 GHz as can be seen from Fig. 3, which also shows the input and output return losses.

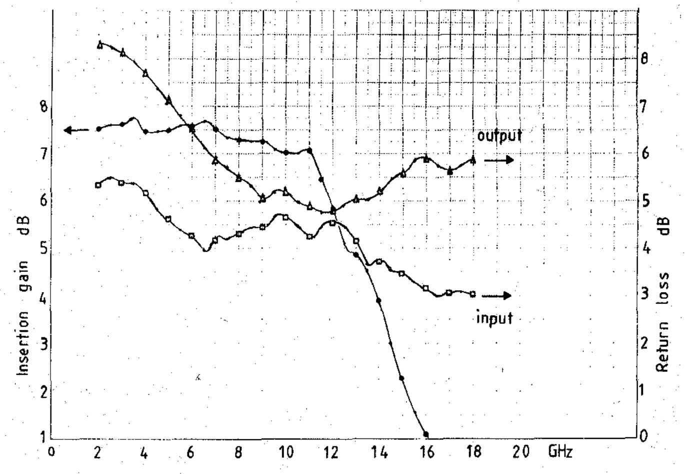

Fig. 3. Computed insertion gain and return losses of amplifier.

It is probable that better VSWR's could be achieved using different matching networks and recent work [8] by H. J. Carlin and others on broad-band matching.

The main point of the present paper is to show, for the first time, the results of calculating the return ratio for a measured transistor in a simple circuit. This has been done

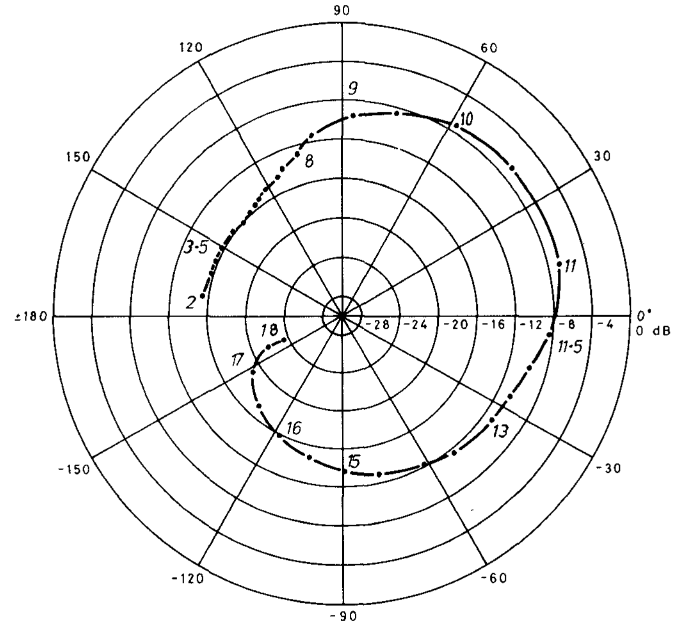

Fig. 4. Polar plot of computed return ratio for y_m of Fig. 1.

using the method outlined in Section II, modified by the use of a circuit diagram of the embedding network rather than measurements on the physical amplifier. The 2×2 matrices $[RTR]$ and $[DS]$ for the NE 70083 are combined with a data file of conventional R, L, C elements in a nodal analysis program, and return voltage ratio computed. The results are illustrated in the Nyquist plot of Fig. 4 covering the band from 2 to 18 GHz at half-gigahertz intervals. In this diagram, pure negative feedback occurs when the phase angle is $0°$, whereas the critical point is at 0 dB, $180°$ at the left-hand side of the circle. The maximum computed value of return ratio is only -6.90 dB, which corresponds to 2.77 dB of feedback (return difference, $1 + RR$)—a modest amount, but very close to the result (2.86 dB) for a similar amplifier [1].

What does Fig. 4 reveal about stability? For single-stage amplifiers having only one source of activity (the dependent generator of transadmittance y_m in Fig. 1), freedom from self-oscillation is assured if the locus of the return ratio for y_m does not encircle the critical point in the clockwise sense. In other words, a stable circuit satisfies the appropriate form of the Nyquist stability criterion [3]. Clearly, both at low frequencies (around 2 GHz) and at high frequencies (around 17 GHz), the locus lies well to the right of the critical point, denoting good margins. However, since a conventional return-ratio gain margin is the magnitude when the phase angle is $180°$, the conventional phase margin is the phase when the magnitude is 0 dB, neither margin can be defined in the frequency range shown since the locus never crosses the 0 dB circle—again, a feature seen before [1]. Instead, the points of nearest approach can be taken, and are seen to give margins of about 15.5 dB and 21.5 dB at low and high frequencies, respectively. These are substantial margins against self-oscillation for the circuit of Fig. 1. It is also gratifying that the high-frequency behavior of the return ratio appears to be tending towards the center (corresponding to a zero value) as would be expected on physical grounds. The behavior below 2 GHz could readily be found from a simple hybrid-pi model derived from the listed S-parameters between say 2 and 4 GHz. This would probably show the locus crossing the $180°$ line and continuing anticlockwise in an arc as frequency decreases.

TABLE I
EXTRACTS FROM COMPUTED VALUES OF ADMITTANCE BETWEEN GATE AND DRAIN PINS: (a) NORMAL CIRCUIT; (b) WITH ADD-ON COMPONENT

Freq.: Hz	Conduct.	Suscept.
1.6900E+10	1.03218E−06	−3.22025E−03
1.7000E+10	4.81788E−07	−3.19611E−03
1.7100E+10	−2.22256E−06	−3.17400E−03
1.7200E+10	−2.70704E−07	−3.15190E−03
1.7300E+10	8.17685E−07	−3.12690E−03

(a)

Admittance Between Nodes 12 *and* 22

Freq.: Hz	Conduct.	Suscept.
1.6900E+10	3.25474E−06	−8.33753E−05
1.7000E+10	2.70435E−06	−4.06777E−05
►1.7100E+10	−4.62100E−12	−2.28414E−09
1.7200E+10	1.95185E−06	4.06548E−05
1.7300E+10	3.04024E−06	8.42157E−05
CCTMODB1		

(b)

IV. COMMENTS

Experience suggests that the return ratio locus is likely to be radically different if the lumped elements in Fig. 1 are replaced by a mixed lumped-distributed circuit such as A. M. Pavio described [10]. His resulting circuit diagram is very much more complicated than its lumped element original, and almost certainly the return ratio would be as well. This supposition is supported by experience with apparently very simple circuits involving a single FET, and without any intentional feedback, which were very prone to oscillate, and whose return ratio computed from a distributed circuit diagram indicated instability at about 17 GHz. This return ratio plot was very complicated, despite which it gave a clear prediction of instability. In stark contrast, when a lumped-element 'approximation' was used, the circuit's return ratio plot was very much simpler and predicted a stability margin of some 9 dB. It would be interesting to see a return ratio locus derived from a measured embedding network and measured transistor between say 2 and 18 GHz.

Although the return ratio cannot be compared with measurements, it is sometimes possible to compare results from the same data base with a simulated experiment. One such case showed that the (internodal) admittances between gate and drain pins had negative real parts between 16.7 and 17.4 GHz. To simulate the effect of adding an admittance of the opposite sign across these terminals, thus nulling both real and imaginary components and causing self-oscillation, the frequency 17.1 GHz was chosen and the computed value of the added admittance was combined with the data file. The admittances between the gate and drain leads before and after the additions are shown in Table I(a) and (b), respectively. It can be seen that the

TABLE II
EXTRACT FROM CALCULATED RETURN RATIO FOR Y_m OF GAT 6 WITH ADD-ON COMPONENT BETWEEN GATE AND DRAIN PINS

Frequency: Hz	Return Ratio Voltage Gain-dB	Return Ratio Voltage Phase-Deg	Input Re-Ohms
1.690E+10	1.564	−166.420	1.000
1.700E+10	1.208	−175.500	1.000
▸1.710E+10	−0.000	179.997	1.000
1.720E+10	0.268	168.718	1.000
1.730E+10	−0.502	160.680	1.000

conductance and susceptance at 17.1 GHz are reduced by factors of about 10^6 to effectively zero. That this add-on component 'causes oscillation' was confirmed by calculating the return ratio for y_m in the modified circuit. The evidence is presented in Table II which shows a phase of 180° to six significant figures and a magnitude of 0 dB (probably to about the same accuracy) at 17.1 GHz, in accordance with theory [3].

The important question of the sensitivity of the return ratio, for instance, to changes of transistor cannot be answered satisfactorily unless a statistically significant amount of measured S-parameter data is available to the designer, and this is seldom the case. One could imagine using the S-parameters of devices with highest S_{21}, lowest S_{21}, highest S_{12}, and lowest S_{12} to find the spreads in return ratio at some confidence level. Detailed published information reproduces scatter diagrams of S_{11} and S_{22} of a power FET at 6 and 11 GHz [11]. Other detailed results have been presented [6] for the S-parameter magnitudes in decibels of 17 ten-section amplifiers measured using a precision microwave wafer probe [6].

APPENDIX

The stability of microwave amplifiers is commonly assessed from Rollett's stability factor K computed from measured S-parameters or from a circuit diagram.

More recent works [11], [12] have shown that this factor is one of four criteria which, taken together, provide necessary and sufficient conditions for the bounded-input, bounded-output (BIBO) type of stability of linear n-ports with *any* passive terminations including open or short-circuits. In practice, an amplifier under investigation may have an active termination, for example the input impedance of a feedback amplifier. If so, the condition on the termination no longer applies, and doubt is cast on the validity of using these criteria.

On the other hand, the Nyquist criterion applied to a return ratio is concerned with zero-input stability with specific terminations, active or passive. For single-stage amplifiers, this one test will give an unambiguous prediction concerning self-oscillation, assuming only that there are no other sources of activity (e.g., Gunn domain effects) present apart from that represented by y_m.

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REFERENCES

[1] D. J. H. Maclean, "Comments on the matched feedback amplifier: Ultrawide-band microwave amplification with GaAs MESFET's," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 619–21, Jan. 1981.
[2] R. S. Pengelly, *Microwave Field-Effect Transistors-Theory Design and Applications.* Chichester, England: Research Studies Press (a division of John Wiley and Sons Ltd.), 1982.
[3] D. J. H. Maclean, *Broadband Feedback Amplifiers.* Chichester, England: Research Studies Press (a division of John Wiley and Sons Ltd.), 1982.
[4] Hewlett-Packard Application Note 95, "S-parameters...circuit analysis and design," Sept. 1968.
[5] R. Q. Lane, R. D. Pollard, M. A. Maury, and J. K. Fitzpatrick, "Broadband fixture characterizes any packaged microwave transistor," *Microwave J.*, pp. 95–97, 101–102, 104, 106, 108–109, Oct. 1982.
[6] E. W. Strid and K. R. Gleason, "A dc 12-GHz monolithic GaAs FET distributed amplifier," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, pp. 969–75, July 1982.
[7] R. S. Pengelly, "Application of feedback techniques to the realization of hybrid and monolithic broadband low-noise and power GaAs FET amplifiers," *Electron. Lett.*, vol. 17, no. 21, Oct. 15, 1981.
[8] H. J. Carlin and B. S. Yarman, "The double matching problem: Analytic and real frequency solutions," *IEEE Trans. Circuits Syst.*, vol. CAS-30, pp. 15–28, Jan. 1983.
[9] A. M. Pavio and S. D. McCarter, "Network theory and modelling method aids design of a 6–18 GHz monolithic multi-stage feedback amplifier," *Microwave Syst. News*, 8 pp., Dec. 1982.
[10] J. B. Klatskin, R. L. Camisa, and D. Haggis, "Fabrication of lumped-element broadband GaAs MESFET microwave power amplifiers," *RCA Rev.*, vol. 42, pp. 576–95, Dec. 1981.
[11] D. Woods, "Reappraisal of the unconditional stability criteria for active 2-port networks in terms of S-parameters," *IEEE Trans. Circuits Syst.*, vol. CAS-23, pp. 73–81, Feb. 1976.
[12] E. Zehib and E. Walach, "Necessary and sufficient conditions for absolute stability of linear n-ports," *Int. J. Circuit Theory Appl.*, vol. 9, pp. 113–30, 1981.

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