A 0.15–12-GHz Matched Feedback Amplifier Using Commercially Available FET's

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Abstract —**Using a graphical approach followed by a simple computer optimization, an ultra-broad-band feedback amplifier covering the frequency range from 150 MHz to 12 GHz has been designed. Experimental results show that using commercially available FET's and parallel configuration, 5-dB flat gain and small VSWR's can be obtained.**

I. INTRODUCTION

Feedback techniques have been used in the design of ultra-broad-band amplifiers. Ulrich [1] achieved noteworthy results for ultra-broad-band microwave amplifiers using negative feedback with GaAs FET's; recently, Niclas *et al.* [2] have described design methods and results for GaAs MESFET feedback amplifiers up to 14 GHz, obtaining five or more octave bandwidth. However, to obtain reasonable gain in the whole band, FET's with very high transconductance must be employed because of the decrease of gain due to the feedback circuit. This does not hold when common commercially available transistors are used. However, the required high transconductance can be obtained using two transistors in a parallel configuration. Using this configuration, ultra-broad-band matched amplifiers with reasonable gain can be made.

II. AMPLIFIER DESIGN

The design procedure in the mentioned papers relies upon a known model of transistor that works up to relatively low frequencies and then on doing a lot of computer optimization. Our graphical method relies on the knowledge of the measured S-parameters and leads to design values very close to the final ones, so that very little computer optimization is needed. A complete description of this graphical method can be seen in [3].

Let us consider the parallel-feedback circuit shown in Fig. 1. When the feedback network is series type RL, RC, $\cdots (Y'_{11} = Y'_{22} = -Y'_{12})$, the following expressions may be easily obtained:

$$\frac{2S_{11}^T}{S_{21}^T} \simeq \frac{2S_{11}}{S_{21}} - Y'_{12}\left[1 - \frac{Y_{11}+Y_{22}}{Y_{21}}\right] \simeq \frac{2S_{11}}{S_{21}} - Y'_{12} \tag{1}$$

$$\frac{2}{S_{21}^T} \simeq \frac{2}{S_{21}} + Y'_{12}\left[1 + \frac{2+Y_{11}+Y_{22}}{Y_{21}}\right] \simeq \frac{2}{S_{21}} + Y'_{12} \tag{2}$$

$$\frac{2S_{22}^T}{S_{21}^T} \simeq \frac{2S_{22}}{S_{21}} - Y'_{12}\left[1 - \frac{Y_{11}+Y_{22}}{Y_{21}}\right] \simeq \frac{2S_{22}}{S_{21}} - Y'_{12} \tag{3}$$

where S_{ij} and S_{ij}^T are the S-parameters of the transistor alone and that of the transistor plus feedback network, respectively.

These expressions indicate that the feedback network produces a translation in the complex plane of the functions $2/S_{21}$, $2S_{11}/S_{21}$, and $2S_{22}/S_{21}$ depending on Y'_{12}. Plotting the values of $2/S_{21}$, $2S_{11}/S_{21}$, and $2S_{22}/S_{21}$ for different frequencies, the problem is to find out the feedback circuit that transforms the $2/S_{21}$ curve into a constant $|S_{21}^I|$ curve, placing $2S_{11}^T/S_{21}^T$ and $2S_{22}^T/S_{21}^T$ near the origin. It is easy to prove that constant $|S_{21}^T|$ loci are a family of circles centered in the origin whose radii are $2/|S_{21}^T|$.

Manuscript received February 17, 1982; revised April 20, 1982.

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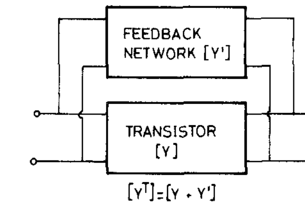

Fig. 1. Parallel-feedback network

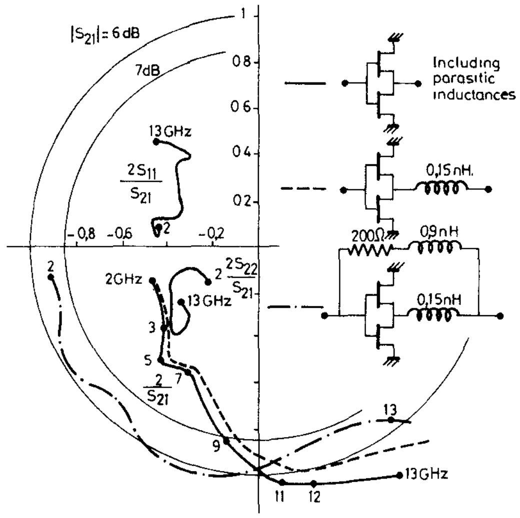

Fig. 2. Graphical construction for parallel transistor cell, cell plus series inductance, and complete feedback configuration.

III. SIX-OCTAVE-BANDWIDTH FET AMPLIFIER

We use a commercially available transistor NE-38800 biased with $V_{DS} = 3$ V, $I_D = 30$ mA. The low-frequency analysis (using an elemental FET model) or the former graphic one indicated that only a feedback resistance tends to equalize and match the transistor in the low-frequency band. However, a strong feedback is needed, making the gain very low. The problem is associated with the low transconductance of the device ($g_m \simeq 25$ m℧). Using two devices in a parallel configuration a greater g_m is obtained.

Using the S-parameters given in data sheets, we have drawn in Fig. 2 the function $2/S_{21}$, $2S_{11}/S_{21}$, and $2S_{22}/S_{21}$ for the parallel configuration. The parasitic inductances associated with the bondwires used to connect the transistor have been included. It is seen that a parallel-feedback resistance ($Y'_{12}/Y_0 = -1/RY_0 = -0.3$) tends to match and equalize the device but the gain is still low for frequencies higher than 7 GHz.

The parallel resistance produces a negative feedback for points in the $2/S_{21}$ curves located in the third quadrant and a positive feedback for points in fourth quadrant. The parasitic inductances displace the high-frequency part of the $2/S_{21}$ curve into the fourth quadrant, an effect that can be reinforced with an inductance in the drain port, obtaining a gain peak at these frequencies. The inductance can be implemented by the connection bondwires. The inductance in series with the resistance produces the same effect and is implemented with other bondwire making

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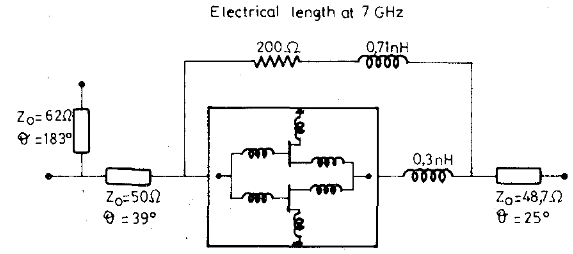

Fig. 3. 0.15–12-GHz amplifier module.

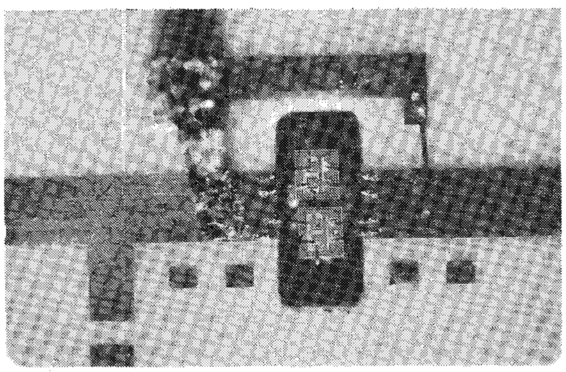

Fig. 4. Photograph of amplifier module.

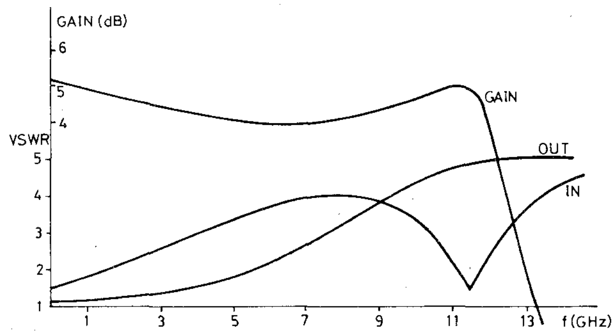

Fig. 5. Experimental results.

the practical realization of the feedback network easier. Definitely, the feedback diagrams allow us to determine the circuit that we present in Fig. 2.

Finally, two simple high-frequency matching networks were designed and the complete circuit was optimized by computer. The resulting final network is shown in Fig. 3. The stability factor was calculated obtaining $K > 1$ for all frequencies in the band.

IV. Experimental Results

A photograph of amplifier module is shown in Fig. 4. Resistors and microstrip lines are fabricated on 0.635-mm thin alumina ceramic substrates. The rated sheet resistance is 50 Ω/square and the coupling capacitor is a multilayer high-dielectric-constant ceramic capacitor. Experimental results are presented in Fig. 5, the power gain measured in a 50-Ω system is 5±0.5 dB in the 0.15–12-GHz band. The VSWR'S measured are higher than the calculated ones, this disagreement being attributed to the parasitics of the feedback network.

V. Conclusions

Using two NE-38800 chips, a feedback amplifier with 5-dB gain covering the frequency range from 0.15 to 12 GHz and low reflection coefficients has been designed. The experimental results obtained show that ultra-broad-band feedback amplifiers can be implemented with commercially available transistors.

These results are very important for the integration of this type of circuit in monolithic technology.

Acknowledgment

The authors wish to thank the members of the Laboratoire d'Etudes Microelectrique Hyperfrequence Thomson-CSF (Orsay, France), where the prototype was realized. The authors are indebted to Dr. J. Obregon who contributed to the success of this project.

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Comment on "A New Fin-Line Ferrite Isolator for Integrated Millimeter-Wave Circuits"

FRIEDRICH J. K. LANGE

In the above paper[1] the authors give a separation equation for "TE-eigenmodes" in a transversely magnetized ferrite as

$$k_{xn}^{(2)2} + \left(\frac{n\pi}{b}\right)^2 + \beta^2 = \omega^2 \epsilon^{(2)} \mu_{\text{eff}}. \quad (1)$$

For $n = 0$, this equation is correct when

$$\mu_{\text{eff}} = \mu_0 \left(\mu_1^2 - \mu_2^2\right) / \mu_1. \quad (2)$$

For $n \neq 0$, (1) is neither correct nor is it given in [1], as the authors try to make the readers believe. The correct treatment of the case $n \neq 0$ is mentioned on pp. 197, 198 in [1] and shown more detailed in [2], [3]. It leads to hybrid eigenmodes.

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Reply[2] by Adalbert Beyer and Klaus Solbach[3]

The comment is perfectly right in stating that the used separation equation is incorrect for higher order modes $n \neq 0$ because

Manuscript received February 10, 1982.

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[1]A. Beyer and K. Solbach, *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 1344–1348, Dec. 1981.

[2]Manuscript received April 2, 1982; revised April 4, 1982.

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