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A Cooled 1–2 GHz Balanced HEMT Amplifier

Stephen Padin and Gerardo G. Ortiz

Abstract —**The design details and measurement results for a cooled L-band balanced HEMT amplifier are presented. The amplifier uses commercially available packaged HEMT devices (Fujitsu FHR02FH). At a physical temperature of 12 K the amplifier achieves noise temperatures between 3 and 6 K over the 1 to 2 GHz band. The associated gain is ~ 20 dB.**

I. Introduction

Cryogenically cooled high electron mobility transistor (HEMT) amplifiers have realized noise temperatures as low as the operating frequency of the amplifier expressed in GHz at frequencies up to 43 GHz [1], [2]. These amplifiers have now become the standard for radio astronomy applications. Cooled HEMT amplifiers are also used as the first IF stage in millimeter-wave superconductor–insulator–superconductor (SIS) and Schottky mixer systems. In these receivers the noise performance of the IF amplifier is very important because the mixer is usually lossy.

Most millimeter-wave receiver systems built for radio astronomy use an L-band IF amplifier with a bandwidth of ~ 500 MHz and a noise temperature of ~ 4 K [3], [4]. In a typical 115 GHz SIS receiver, the IF amplifier contributes ~ 30% of the total receiver noise. A bandwidth of 500 MHz is barely enough for observations of sources with high velocity dispersion and it limits the capability of systems which are able to observe several molecular transitions simultaneously. Increasing the receiver bandwidth by using a higher frequency IF amplifier is not viable because this would increase the IF noise contribution and degrade the sensitivity of the system. Our approach to this problem has been to develop a 1-GHz-bandwidth L-band cooled HEMT amplifier.

The main problem in the design of a wideband cooled amplifier is obtaining s parameters and noise parameters for transistors at low temperatures. For this work no facilities were available for measuring low-temperature s parameters, but the HEMT noise parameters were measured at a physical temperature of 12 K. The absence of s-parameter information precluded the design of a feedback amplifier, so a balanced configuration was adopted. This has the advantage of providing a good input match even though the amplifiers in the two arms of the balanced circuit are poorly matched. However, there are disadvantages. The loss of the input hybrid degrades the noise temperature and coupling errors in the hybrids, and differences between the amplifiers reduce the gain and result in a noise contribution from the input load. In the amplifier described here these effects degrade the noise temperature by less than 1 K.

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S. Padin is with the Owens Valley Radio Observatory, California Institute of Technology, Big Pine, CA 93513.

G. G. Ortiz is with the Jet Propulsion Laboratory, 4800 Oak Grove Drive, Pasadena, CA 91109.

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II. Noise in a Balanced Amplifier

The noise contributions in a balanced amplifier are explored in Fig. 1. Each hybrid directs a fraction, c, of the input power to the 0° port and the remaining power to the 90° port. The deviation from quadrature at the outputs is θ. To simplify the analysis the amplifiers are assumed to have similar gains but different transfer function phases. In practice this situation can be approached by selecting similar devices and by adjusting the bias.

The power gain of the balanced amplifier (with the input terminated in a matched source and the output terminated in a matched load) is

$$G = 2gc(1-c)(1+\cos\phi) \qquad (1)$$

where g is the power gain of each amplifier in the balanced structure and ϕ is the phase difference between the amplifier transfer functions. The output noise temperature with the input terminated in a matched source at 0 K is

$$T_{\text{out}} = gT_0\left[(2c^2-2c+1)-2c(1-c)\cos\phi\right] + gT_a \qquad (2)$$

where T_0 is the physical temperature of the input hybrid termination and T_a is the noise temperature of each amplifier. The first term is the contribution from the input hybrid termination and the second term represents the noise generated by the amplifiers. Note that the noise from the two amplifiers is uncorrelated. The noise temperature of the balanced amplifier is

$$T_n = \frac{T_{\text{out}}}{G} = \frac{T_0\left[(2c^2-2c+1)-2c(1-c)\cos\phi\right]+T_a}{2c(1-c)(1+\cos\phi)}. \qquad (3)$$

The loss of the input hybrid can be modeled as an attenuator at the amplifier input. Loss in the output hybrid affects only the overall gain (and hence the noise contribution of the next stage). With an input hybrid loss L, the noise temperature of the balanced amplifier is

$$T_n' = T_0(L-1)+LT_n \qquad (4)$$

where T_n is given by (3). As an example of what might be achieved, a balanced circuit containing amplifiers with $T_a = 4$ K and $\phi = 5°$ and hybrids with coupling errors of 1 dB ($c = 0.40$) and excess loss of 0.1 dB would have a noise temperature of 5.09 K at a physical temperature of 12 K. If the input load were at 4 K instead of 12 K the noise temperature would be 4.73 K.

III. HEMT Device Noise Parameters

Device noise parameters for this work were obtained from measurements of the noise temperatures of several single-ended amplifiers, each with a different input matching network. The same HEMT device was used for the entire set of measurements. Previous work with GaAs FET's at L-band [5] provided

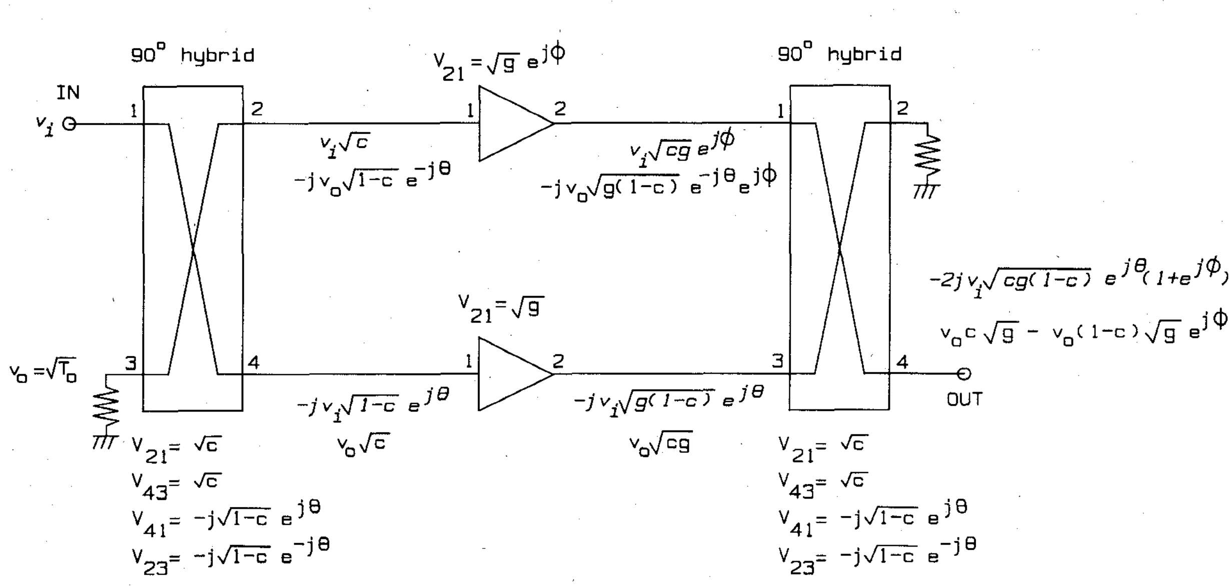

Fig. 1. Voltages in a balanced amplifier. V_{mn} is the voltage transfer function from port n to port m of a component in the circuit. The italicized terms indicate voltages caused by input v_i, the other terms indicate noise voltages caused by the input hybrid termination. The time-dependent and delay terms have been omitted. g is the power gain of each single-ended amplifier and ϕ is the difference between the amplifier transfer function phases. c is the fraction of the hybrid input power which appears at the 0° port and θ is the deviation from quadrature at the hybrid outputs.

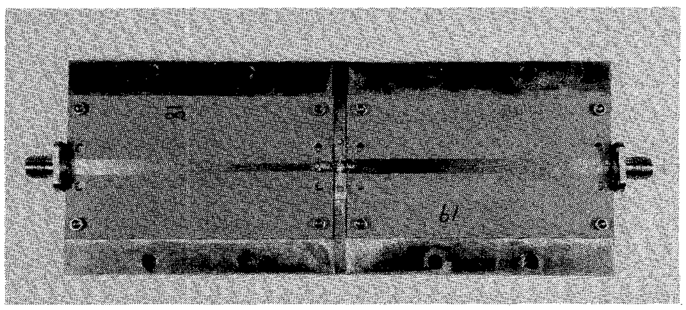

Fig. 2. Single-ended amplifier test fixture. The input is on the left. Bias voltages are supplied via external bias Tees.

TABLE I
NOISE PARAMETERS FOR THE GE HEMT

Frequency GHz	R_{opt} Ω	X_{opt} Ω	g_n ms	T_{min} K
1.0	100 ± 20	160 ± 20	0.07 ± 0.03	4.5 ± 1
1.5	80 ± 10	160 ± 20	0.07 ± 0.03	3.5 ± 1
2.0	80 ± 20	120 ± 20	0.13 ± 0.09	4.5 ± 1

an estimate of the optimum source impedance so it was necessary to explore only a small part of the source impedance plane to determine the HEMT noise parameters.

Different source impedances were provided using the fixture shown in Fig. 2. The input section consists of a microstrip transformer realized on 60-mil-thick RT/Duroid 6002[1] ($\epsilon_r = 2.94$) and a ~ 10 nH series chip inductor[2] close to the HEMT package. The output section is a 50 Ω microstrip line. Connec-

[1]RT/Duroid 6002, Rogers Corp., Microwave Materials Division, 100 S. Roosevelt Ave., Chandler, AZ 85226.
[2]Surface mount inductor series 1008CS, Coilcraft, 1102 Silver Lake Rd., Cary, IL 60013.

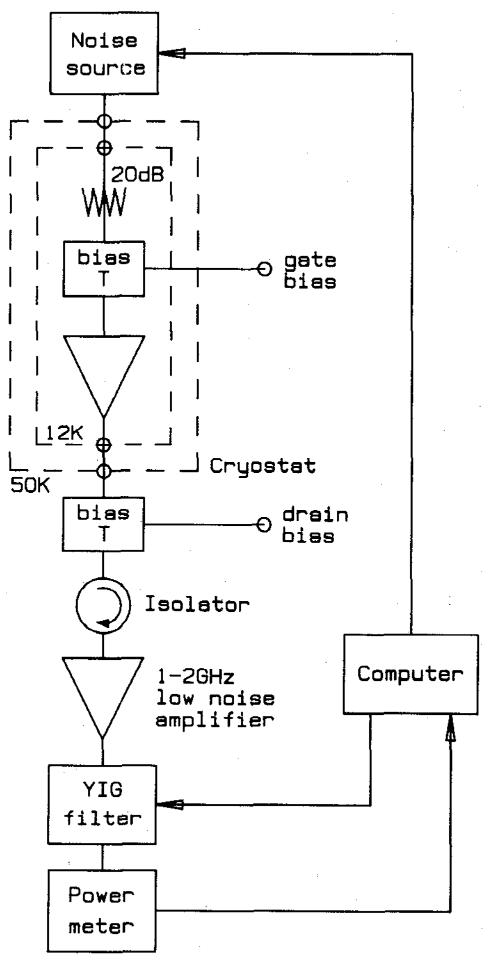

Fig. 3. Noise temperature measurement system.

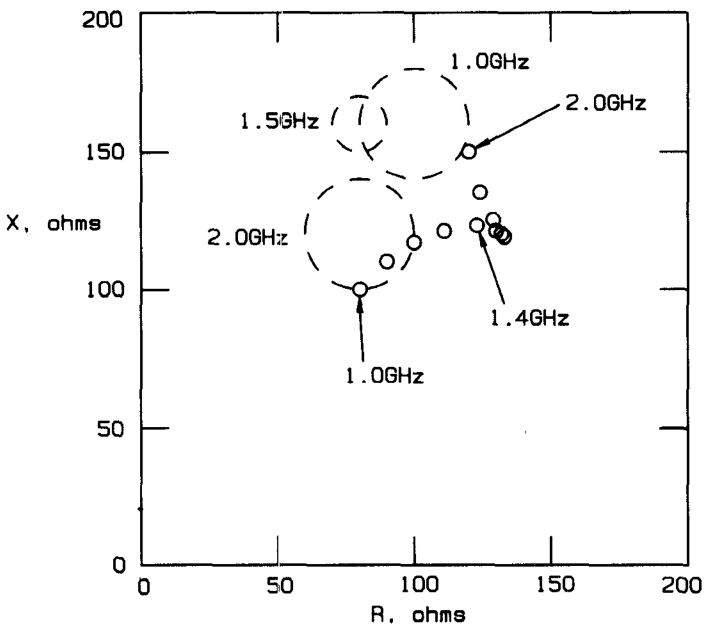

Fig. 4. Measured impedance of the amplifier input matching circuit at 100 MHz intervals (solid circles) and the optimum source impedance from Table I (dashed circles). The size of the dashed circles indicates the error in the optimum source impedance measurements.

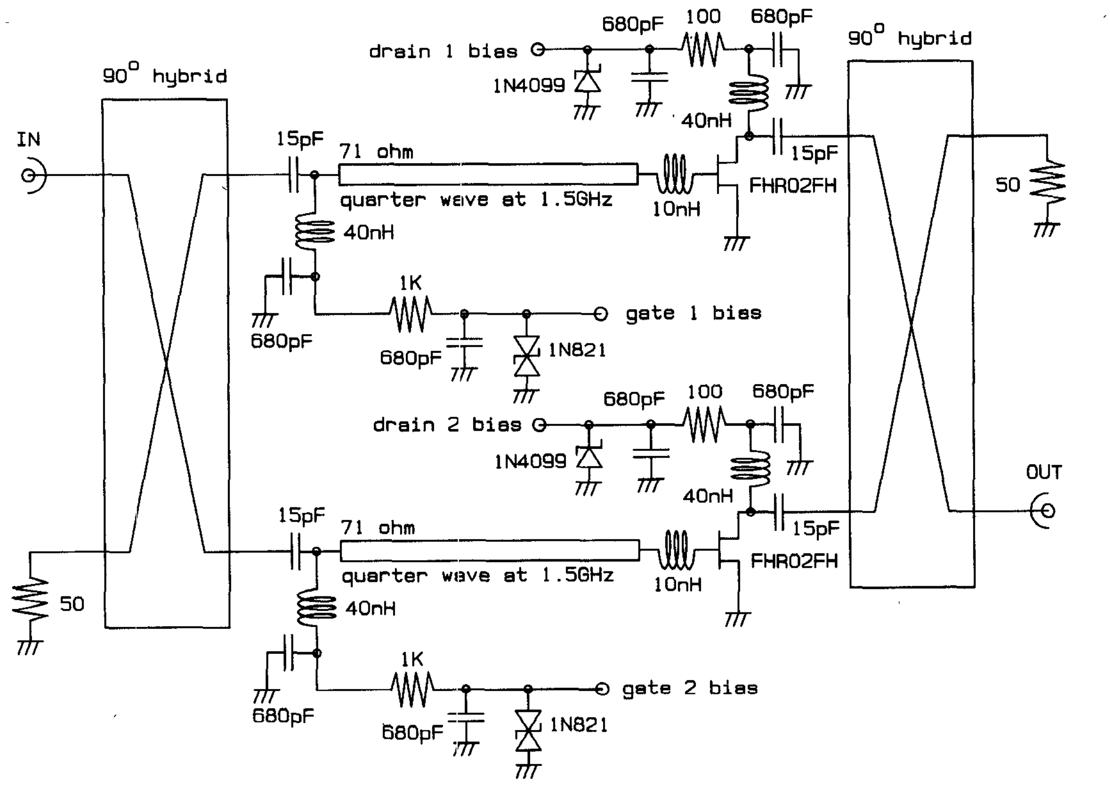

Fig. 5. Schematic of the 1–2 GHz balanced amplifier.

tions to the HEMT are simple pressure contacts that allow easy removal of the input matching network assembly. The assembly can be transferred to a test fixture fitted with SMA connectors to allow measurement of the source impedance presented to the HEMT.

Noise temperature measurements were made using the arrangement of Fig. 3. A cooled 20 dB attenuator at the amplifier input provides a cold load, and signals enter and leave the cryostat via low-loss coaxial lines. The losses of the lines, connectors, input bias network, and 20 dB attenuator were measured at 12 K. These data along with measurements of the noise temperature of the receiver in Fig. 3 were used to calculate the noise temperature of the amplifier from measurements of the system noise temperature at the cryostat input. A solid-state noise source calibrated against liquid nitrogen and ambient loads was used for the measurements.

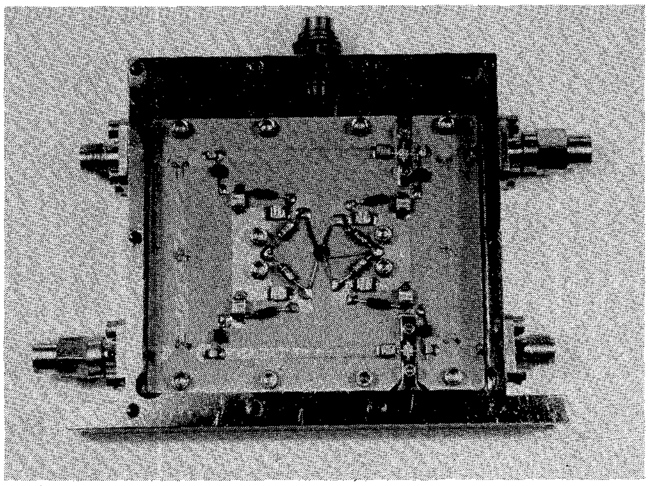

Fig. 6. Photograph of the 1–2 GHz balanced amplifier.

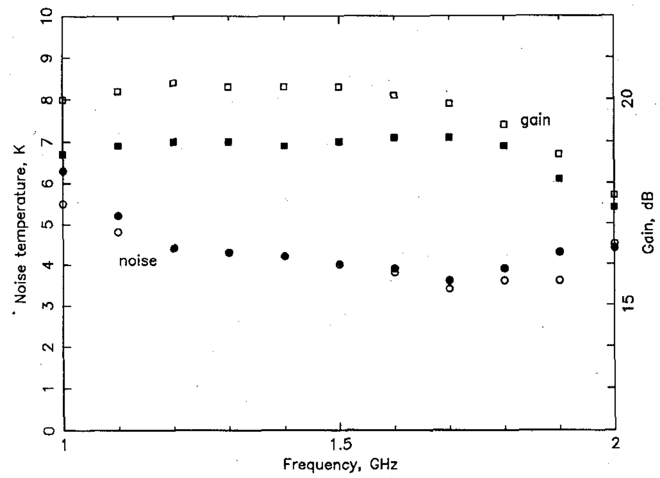

Fig. 7. Gain and noise temperature for two prototype balanced amplifiers at a physical temperature of 12 K. Filled symbols refer to amplifier s/n 00 (bias for minimum noise: $V_d = 3.4$ V, $V_g = -0.14$ V, and $I_d = 5.4$ mA). Open symbols refer to amplifier s/n 01 (bias for minimum noise: $V_d = 3.4$ V, $V_g = -0.12$ V, and $I_d = 6.5$ mA). The noise temperature measurement error is ± 1.2 K. s/n 01 noise temperature data for 1.2 to 1.5 GHz are coincident with those for s/n 00.

The amplifier noise temperature is related to the source impedance, $R + jX$, by

$$T = T_{\min} + \frac{290\, g_n\left[\left(R - R_{\mathrm{opt}}\right)^2 + \left(X - X_{\mathrm{opt}}\right)^2\right]}{R} \tag{5}$$

where $R_{\mathrm{opt}} + jX_{\mathrm{opt}}$ is the optimum source impedance, g_n is the noise conductance, and $T_{\min}$ is the minimum noise temperature [6]. The noise parameters $\{R_{\mathrm{opt}}, X_{\mathrm{opt}}, g_n, T_{\min}\}$ were obtained from the results of ten measurements of T with different source impedances using a least-squares fit to (5). Although the final amplifier was constructed using Fijitsu FHR02FH (200 μm$\times$ 0.25 μm) devices, the noise parameters were measured for a General Electric 300 μm$\times$0.25 μm AlGaAs/GaAs HEMT [7] in a standard 70 mil package. Recent work [8] with FHR02 chips indicates that the packaged GE and packaged Fujitsu devices have very similar R_{opt} and X_{opt} at L-band. The measured noise parameters for the GE HEMT are summarized in Table I. FHR02FH devices have $g_n \sim 0.03$ ms and are therefore less sensitive than the GE device to deviations from the optimum source impedance.

IV. Amplifier Design

The input matching circuit for each arm of the balanced amplifier consists of a 71 Ω line ($\lambda/4$ at 1.5 GHz) and a 10 nH series inductor. This gives a good noise match over the 1 to 2 GHz band and is reasonably compact and low loss. The impedance of the transformer and the value of the inductor were chosen to minimize $[(R - R_{\mathrm{opt}})^2 + (X - X_{\mathrm{opt}})^2]$ over the 1 to 2 GHz band. Fig. 4 shows measurements of the impedance presented by the matching circuit along with the optimum source impedance from Table I. The calculated increase in noise temperature caused by deviation of the source impedance from optimum is ~ 1 K.

The 90° hybrids in the balanced amplifier are seven-finger Lange couplers [9]. These are realized on the same 60 mil RT/Duroid 6002 substrate used for the matching networks. At room temperature the measured coupler insertion loss is ~ 0.2 dB, the coupling error is < 0.5 dB, and the deviation from phase quadrature is $< 2°$.

A schematic of the complete balanced amplifier is shown in Fig. 5, and the construction details are indicated in Fig. 6. To ensure good thermal coupling to the box, the HEMT devices are clamped to posts which protrude through the circuit board. The

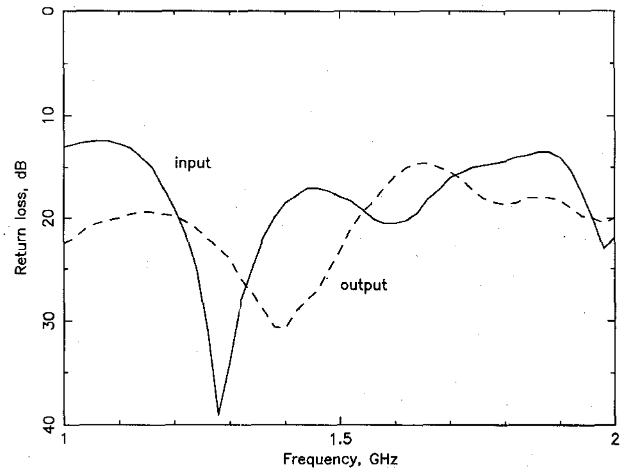

Fig. 8. Input and output return losses for amplifier s/n 00 at a physical temperature of 12 K.

bias circuits are configured to allow independent biasing of the two HEMT's, although in practice this has not been necessary.

V. Results

Two prototype balanced amplifiers were constructed with Fujitsu FRH02FH HEMT's, and the noise temperatures and gains at a physical temperature of 12 K are shown in Fig. 7. Both amplifiers have noise temperatures in the range 3 to 6 K. For these measurements the bias was adjusted to minimize the noise temperature. No other tuning was done, so the results indicate typical performance for a production amplifier. Some improvement in noise temperature might be obtained by adjusting the values of the inductors in the matching networks.

Input and output return losses were measured for one amplifier at 12 K and these data are shown in Fig. 8. The worst-case input return loss is 13 dB but over most of the band the return loss is better than 15 dB. All the results were obtained with no illumination of the HEMT's.

VI. Conclusions

We have described a 1 to 2 GHz cooled balanced HEMT amplifier. At a physical temperature of 12 K the amplifier has a noise temperature in the range 3 to 6 K and a gain of ~ 20 dB. The amplifier was designed primarily as a wide-band IF amplifier for millimeter-wave radio astronomy but it also has applications in wideband L-band receiver systems.

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Equivalent-Circuit Parameter Extraction for Cold GaAs MESFET's

R. Anholt and S. Swirhun

***Abstract*—The physical basis of the cold-FET method for extracting parasitic resistances and inductances is examined. A method to obtain the source resistance from the gate-current dependence of the FET Z parameters is used to analyze FET's with different gate lengths. Inductance results for FET's with different gate widths suggest that inductance extrinsic to the gate fingers is dominant, and models of the gate inductance support this. The effects that possible dependences of the parasitic-FET equivalent-circuit parameters on the gate and drain bias can have on the extracted intrinsic-FET parameters are discussed.**

I. Introduction

The cold-FET method provides an elegant way of extracting FET equivalent-circuit parameters (ECP's) from S parameters at any bias [1], [2]. Parasitic source, drain, and gate resistances and inductances are first extracted from measured S parameters of FET's biased at a drain–source voltage $V_{ds} = 0$ and a gate voltage, V_g, greater than the barrier height. With those values fixed, the measured S-parameter matrix for any other bias can be converted to an intrinsic Y-parameter matrix that can be solved exactly for up to eight independent ECP's, depending on the intrinsic-FET circuit topology.

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R. Anholt is with Gateway Modeling, Inc., 1604 East River Terrace, Minneapolis, MN 55414.

S. Swirhun is with the Honeywell Systems and Research Center, Bloomington, MN 55420.

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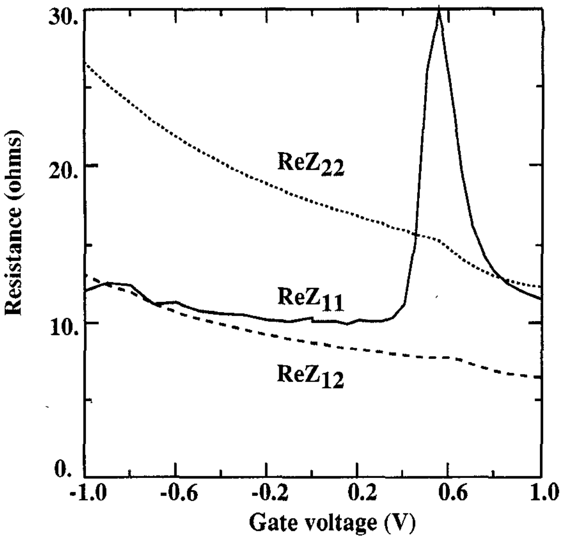

Fig. 1. Real parts of the Z matrices plotted against the gate voltage for an epitaxially doped $1.2 \times 200\ \mu m^2$ MESFET (biased at $V_{ds} = 0$).

There are two problems with this method. The first is in the cold-FET extraction technique. For the resistances, this can be reduced to a problem of determining two unknowns from only one equation, and we show that this can be solved at forward gate bias. This leads to the second question considered: Is it valid to assume that the cold-FET parameters are independent of bias? If they are not, how much are the extracted intrinsic-FET ECP's (g_m, C_{gs}, etc.) sensitive to possible errors?

In Sections II and III of this paper we examine the problem of extracting the resistances and inductances. Section IV examines the sensitivity of extracted intrinsic-FET ECP's to variations in cold-FET ones.

II. Source, Drain, and Gate Resistance Extraction

Fig. 1 shows extracted average real parts of Z parameters for FET's biased at $V_{ds} = 0$ plotted against gate potential. With the exception of $\mathrm{Re}\, Z_{11}$ near the peak, when measured S parameters from 1 to 26 GHz are converted to Z parameters, the values are nearly independent of frequency. Two regimes are evident in this figure. For $V_g < 0.6$ V, the Z matrix is given by (parts (c) and (d) of Fig. 2, $g_m = \tau = 0$, $R_{ds} = R_{ch}$, $C_{gs} = C_{gd} = C/2$)

$$\begin{aligned} Z_{11} &= R_s + R_{ch}/3 + R_g + j\omega(L_s + L_g) - \frac{1}{j\omega C} \\ Z_{12} &= Z_{21} = R_s + R_{ch}/2 + j\omega L_s \\ Z_{22} &= R_s + R_d + R_{ch} + j\omega(L_s + L_d) \end{aligned} \tag{1}$$

where R_s and R_d are the source and drain resistances; R_g is the distributed gate resistance; L_s, L_d, and L_g are the source, drain, and gate inductances; and C is total gate capacitance (including C_{pf} and C_{pg} in Fig. 2(c)). In this regime, the gradual