transfer function can be synthesized into a low-pass prototype network containing integral cross coupling in the form of a "tri-section" of impedance inverters. Each tri-section realizes one integrated pole.

Comparisons made between conventional symmetrical techniques and filters built using the methods described here show that a significant reduction in degree can be obtained for a given selectivity by utilizing an asymmetric response.

An ordinary Chebyshev response can be recovered from the integrated pole transfer function by letting all the ω_i tend to infinity since

$$\lim_{\omega_i \to \infty} \frac{1 - \omega\omega_i}{\omega - \omega_i} = \omega.$$

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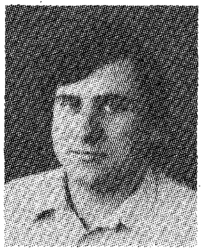

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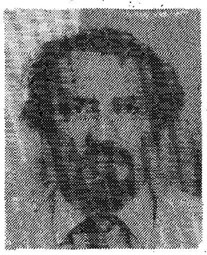

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Electronically Cold Microwave Artificial Resistors

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Abstract —A large percentage of microwave field-effect transistors (FET's) are shown to act as a broad-band artificial resistor with a resistance of about 25 Ω when their drain is connected to their gate. The resistance appears between the gate–drain lead and the source lead. This resistance can be raised to 50 Ω with its reactive components eliminated over a reasonable bandwidth by using a matching transmission line of the proper impedance and a length near a quarter-wave at midband. An HFET-1000 constructed in this configuration showed an impedance of 18 ± 3 Ω over an octave bandwidth, and when transformed with a 30-Ω quarter-wave transmission line produced a resistance of 51 ± 1 Ω from 8 to 13 GHz. A noise analysis shows that, at some frequencies, some FET's in this configuration will produce artificial resistors with an effective noise temperature as low as 67 K. The addition of a transmission line in the gate lead and in the feedback line allows the effective temperature of any FET to be made substantially below room temperature. No cryogenic cooling is required. The addition of a half-wave to either transmission line will produce a low-noise negative resistor which can be used in a negative resistance amplifier circuit to produce an amplifier with the gain of a tunnel diode and the noise figure of a FET. These "electronically cold" artificial resistors could improve the overall noise figure of a microwave receiver if they replaced the first-stage amplifiers and the standard room-temperature loads in critical front-end circuits using loads in conjunction with circula-

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tors or balanced amplifiers and switches. They can also be used to terminate long-signal lines, where they will not cause the usual 3-dB loss in signal-to-noise ratio that occurs when a standard noisy resistor is used. Finally, they may also find use as an artificial "cold load" in radio astronomy and microwave calibration test setups without requiring a cryogenic capability.

I. Introduction

It has long been known that room-temperature active artificial resistors can be made with effective noise temperatures significantly below room temperature. The first "electronically cold" artificial resistors were made using vacuum tubes [1], [2] and reached effective temperatures of 77 K. With the advent of semiconductors, it was found that the effective noise temperature of a semiconductor device was typically $T_{\mathrm{eff}} \sim 0.5\text{–}0.6T_0$ [3]. For example, a forward-biased Schottky diode will act as a 50-Ω microwave load with an effective temperature of 165 K.

Later, with the advent of research on detectors for gravitational waves and Newtonian gravity gradient fields, a number of additional designs for artificial resistors were generated [4]–[9] that demonstrated effective noise temperatures well below 10 K at audio frequencies.

More recent work has concentrated on achieving at microwave frequencies effective noise temperatures below the 165 K obtainable with forward-biased Schottky diodes. A recent paper [10] used inductive feedback in the source lead of a FET to achieve an effective temperature of 50 K at L-band. Our paper discusses the use of drain-to-gate feedback in a FET to produce a low-noise temperature artificial resistor, as well as the use of transmission lines from an input port to the gate and the input port to the drain to minimize the effective noise temperature.

II. Artificial Resistors from FET's

A typical microwave FET has an equivalent circuit given by Fig. 1, where typically R_g is in tens of ohms, C_g is in tenths of picofarads, and g_m is in tens of millisiemens. If we connect the drain to the gate, the resulting circuit is given by Fig. 2. We now have a single-port device. When a voltage V_{in} is applied across the two terminals of the port, a current I_{in} flows into the port. The input impedance of the device is then just

$$Z_{\mathrm{in}} = \frac{V_{\mathrm{in}}}{I_{\mathrm{in}}}$$

where

$$I_{\mathrm{in}} = I_g + I_d = \frac{V_{\mathrm{in}}}{R_g + jX_g} + g_m V_g$$

and

$$V_g = \frac{jX_g}{R_g + jX_g} V_{\mathrm{in}}.$$

Defining

$$r_m \equiv \frac{1}{g_m}$$

and substituting into the previous equation yields the following expression for the input impedance:

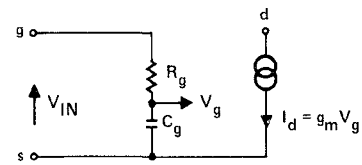

Fig. 1. Equivalent circuit of a microwave FET.

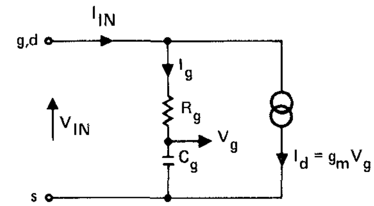

Fig. 2 Equivalent circuit of a microwave FET with the drain connected to the gate

$$Z_{\mathrm{in}} = r_m \frac{\left(R_g r_m + X_g^2\right) + jX_g\left(r_m - R_g\right)}{X_g^2 + r_m^2}.$$

If the reactance

$$X_g = -\frac{1}{\omega C_g}$$

is significantly greater than the resistances r_m and R_g, which is increasingly true at lower frequencies, then the input impedance is dominated by the transconductance of the FET

$$Z_{\mathrm{in}} \to r_m = \frac{1}{g_m}.$$

Since the transconductance of a commercially available microwave FET is typically 40 mS, the typical equivalent resistance of a fed-back FET is 25 Ω.

The lumped circuit model used for a FET in Fig. 1 is a simplified model; it served to show how, in a fed-back FET circuit, the input admittance is dominated by the transconductance. To calculate the actual input impedance of a real FET, one uses the manufacturer's scattering coefficient (S-parameter) data. For circuit analysis, we convert these data to an equivalent admittance matrix $[Y_{mn}]$ using standard handbook formulas or a microwave analysis program such as COMPACT.

III. Calculation of Input Impedance

Fig. 3 shows a circuit schematic for calculating the input impedance of a fed-back FET in terms of its admittance matrix $[Y_{12}]$. For clarity in relating the circuit equations to the figure, we specifically call out the two port voltages separately even though they are equal to each other.

The input impedance of the circuit is obtained by calculating the ratio of the input voltage to the input current

$$Z_{\mathrm{in}} = \frac{V_{\mathrm{in}}}{I_{\mathrm{in}}}.$$

Using Fig. 3 and defining

$$Y_{\mathrm{in}} = \left(Y_{11} + Y_{12} + Y_{21} + Y_{22}\right)$$

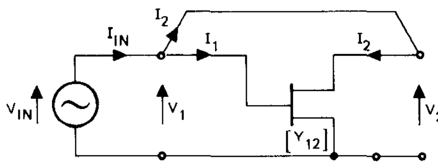

Fig. 3. Schematic for calculating input impedance.

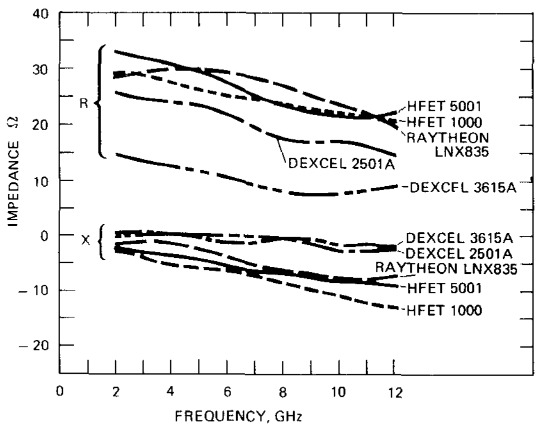

Fig. 4. Resistive and reactive components of impedance looking into typical microwave FET's with drain connected to gate.

we find that the input impedance of the fed-back FET is just

$$Z_{in} = \frac{1}{Y_{in}} = \frac{Y_{in}^*}{|Y_{in}|^2}.$$

Using the scattering parameter data available from various manufacturers, we calculated the real and imaginary parts of the impedance seen in five selected microwave FET's at frequencies ranging from 2 to 12 GHz. These results are shown in Fig. 4. As can be seen, the input impedance is predominantly resistive, becoming more so at lower frequencies. The equivalent resistances range from 7 to 34 Ω with most values being around 25 Ω.

Although not all FET's display this behavior, a significant fraction do. These artificial resistors can then be used (with suitable impedance-matching transmission lines) as microwave loads.

IV. Noise Analysis

We want to carry out a noise analysis of these artificial resistors to determine the voltage noise appearing across the port and to compare that with the thermal Johnson voltage noise appearing across the terminal of a comparable resistor. For the noise analysis, we will use the conventions of the IRE Standard for Noise in Linear Twoports, 1959 [11] and its companion theoretical paper [12]. In these papers, it is shown that a noisy amplifier can be represented by a noiseless amplifier with the same admittance parameters, plus three noise sources arranged at the input. These are the equivalent input noise voltage e, the equivalent input uncorrelated noise current i_u, and the equivalent input correlated noise current $i_c(e)$.

The noise analysis of a FET with its drain connected to its gate by a simple short uses the schematic shown in Fig. 5.

Solving for V_{out}, letting $Y_{in} = Y_{11} + Y_{12} + Y_{21} + Y_{22}$ as be-

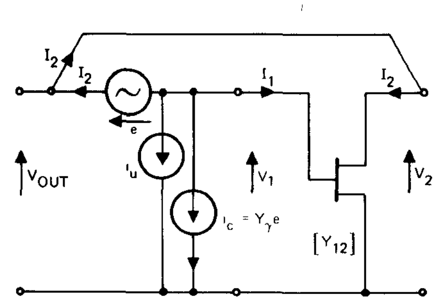

Fig. 5. Schematic for calculation of Thevenin open-circuit equivalent voltage (i.e., V_{out} with $I_{out} = 0$).

fore, and defining $Y_e = Y_\gamma - Y_{11} - Y_{21}$, we get

$$V_{out} = -\frac{Y_e}{Y_{in}}e - \frac{i_u}{Y_{in}}.$$

Since e and i_u are by definition uncorrelated, the mean square noise voltage put out by the artificial resistor is calculated by squaring each of the noise terms separately and adding

$$\overline{V_{out}^2} = \frac{|Y_e|^2}{|Y_{in}|^2}\overline{e^2} + \frac{\overline{i_u^2}}{|Y_{in}|^2}.$$

The "available" noise power from an impedance at a temperature T in the bandwidth B, regardless of the value of the impedance, is obtained when the impedance is connected to a matching load and is given by

$$P_{avail} = kTB$$

where $k = 1.38 \times 10^{-23}$ J/K is Boltzmann's constant.

The maximum noise power obtainable from the artificial resistor is also obtained when it sees a matching load and is given in terms of the output noise voltage as

$$P_{max} = \frac{\overline{V_{out}^2}}{4\mathrm{Real}\left(Y_{in}^{-1}\right)}$$

where $\mathrm{Re}(Y_{in}^{-1}) = R_{in}$.

We can now define an equivalent noise temperature T_{eq} for the artificial resistor as that temperature where the "available" power of the equivalent impedance equals the maximum noise power obtainable from the circuit

$$kT_{eq}B = \frac{\overline{V_{out}^2}}{4R_{in}}$$

or

$$\overline{V_{out}^2} = 4kT_{eq}BR_{in}.$$

Substituting this equation and the equations for $\overline{e^2}$ and $\overline{i_u^2}$ into the mean-square output noise voltage equation, and dividing through by $4kT_0BR_{in}$, we obtain an equation for the equivalent noise temperature of the fed-back FET

$$\frac{T_{eq}}{T_0} = \frac{|Y_e|^2R_n + G_u}{|Y_{in}|^2R_{in}}.$$

V. Example Noise Calculation

To demonstrate that the equivalent noise temperature of the artificial resistor can be less than room temperature, let us take the noise and admittance parameters for a

Hewlett–Packard Microwave FET type HFET-1000 (chip) at 11 GHz, when operated at its low-noise point ($V_{DS} = 3.5$ V, $I_{DS} = 15$ percent of I_{DSS}).

Measured Device Noise Model Parameters:

Noise resistance	$R_n = 0.53$ (26.5 Ω).
Optimum reflection coefficient	$\rho_o = 0.632@115°$.
Optimum noise figure	$F_o = 3.27$ dB (2.12 ratio).

Measured Device Y Parameters:

$$Y_{11} = 12.3 + 35.0j \text{ mS}$$
$$Y_{12} = 0.6 - 2.5j \text{ mS}$$
$$Y_{21} = 20.8 - 23.4j \text{ mS}$$
$$Y_{22} = 2.6 + 11.1j \text{ mS}.$$

From these we can calculate the input admittance

$$Y_{\text{in}} = Y_{11} + Y_{12} + Y_{21} + Y_{22} = 36.3 + 20.2j \text{ mS}$$

and the input impedance

$$Z_{\text{in}} = \frac{1}{Y_{\text{in}}} = \frac{Y_{\text{in}}^*}{|Y_{\text{in}}|^2} = 21.0 - 11.7j\ \Omega.$$

The normalized optimum noise source admittance (Y_o) is then obtained from the optimum noise source reflection coefficient (ρ_o) by

$$Y_o = \frac{1-\rho_o}{1+\rho_o} = 0.695 - 1.324j = G_0 + jB_0.$$

From these we can obtain the fundamental noise parameters B_γ, G_γ, and G_u (the unnormalized values are in parentheses)

$$B_\gamma = -B_0 = 1.324 \text{ (26.5 mS)}$$
$$G_\gamma = \frac{(F_0 - 1)}{2R_n} - G_0 = 0.362 \text{ (7.2 mS)}$$
$$G_u = R_n\left(G_0^2 - G_\gamma^2\right) = 0.187 \text{ (3.7 mS)}.$$

The correlation admittance is then

$$Y_\gamma = G_\gamma + jB_\gamma = 7.2 + 26.5j \text{ mS}$$

and the admittance coefficient of the voltage noise term is easily calculated to be

$$Y_e = Y_\gamma - Y_{11} - Y_{21} = -25.9 + 14.9j \text{ mS}.$$

We now gather all the factors needed to calculate the equivalent noise temperature

$$R_{\text{in}} = 21.0\ \Omega$$
$$R_n = 26.5\ \Omega$$
$$G_u = 3.7 \times 10^{-3} \text{ S}$$
$$|Y_{\text{in}}|^2 = 1.726 \times 10^{-3} \text{ S}^2$$
$$|Y_e|^2 = 0.893 \times 10^{-3} \text{ S}^2$$

and find that the normalized equivalent noise temperature is

$$\frac{T_{\text{eq}}}{T_0} = \frac{|Y_e|^2 R_n + G_u}{|Y_{\text{in}}|^2 R_{\text{in}}} = 0.653 + 0.102 = 0.755$$

which is three-fourths of room temperature. Assuming $T_0 = 290$ K, we find

$$T_{\text{eq}} = 219 \text{ K}.$$

Note that it is the first term that dominates the noise. If we neglect the second term and expand Y_e and Y_{in}, we get

$$\frac{T_{\text{eq}}}{T_0} \simeq \frac{|Y_\gamma - Y_{11} - Y_{21}|^2}{|Y_{11} + Y_{12} + Y_{21} + Y_{22}|^2} \frac{R_n}{R_{\text{in}}}.$$

In most microwave FET's, the admittance term Y_{21} is significantly larger than any of the other admittance terms, so to first order, $|Y_e|^2 \simeq |Y_{\text{in}}|^2$ and the equivalent temperature equation reduces to

$$\frac{T_{\text{eq}}}{T_0} \propto \frac{R_n}{R_{\text{in}}}.$$

VI. "Electronically Cold" Microwave Loads

Using the manufacturer's data for the HFET-1000, we calculate that it should have a low equivalent noise temperature at frequencies above 10 GHz, reaching a low of 80 K at 12 GHz. This is plotted in Fig. 6 along with the noise resistance R_n, the optimum noise figure F_o, and the unmatched input impedance Z_{in}. As can be seen, there is some correlation between the equivalent noise temperature T_{eq} and the noise resistance R_n.

At lower frequencies, we have calculated that the packaged FET, the 2N6680 (HFET-1101), has a low equivalent noise temperature in the region below 6 GHz, reaching temperatures of 67 K. The package causes an electrical resonance in the behavior of the fed-back FET artificial resistor configuration which makes it unsuitable for operation as an artificial resistor in the region from 8 to 10 GHz.

These low-noise microwave loads can be used to improve overall system performance in critical front-end radar and communication receivers by replacing loads used in conjunction with balanced amplifiers or switches. They can also be used to terminate long-signal lines, where they will not cause the usual 3-dB loss in signal-to-noise ratio that occurs when a standard noisy resistor is used. Finally, they may also find use as an artificial "cold load" in radio telescopes [3] and microwave calibration test setups without requiring a cryogenic capability.

VII. Experiment

To check out the analysis, we fabricated an artificial microwave resistor using an HFET-1000 with its drain fed back onto its gate through a large blocking capacitor, keeping all leads as short as possible. By designing our experimental configuration to place the blocking capacitor on top of one of the bias current fingers, we were able to obtain a very stable very flat artificial resistor. As is shown in the top curve in the Smith chart in Fig. 7, the effective resistance varied from 21 Ω at 6 GHz to 15 Ω at 12 GHz, an octave bandwidth, while the reactance was near zero over most of the band. There was no change in impedance with bias voltage or current around the nominal operating values.

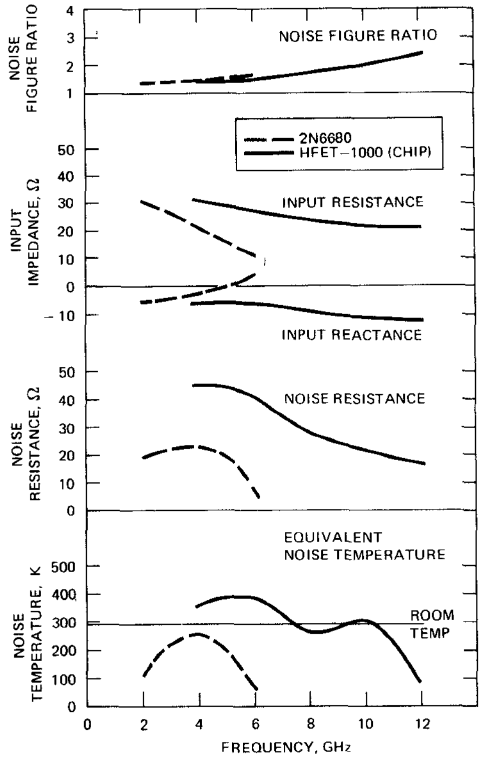

Fig. 6. Noise parameters for two FET's.

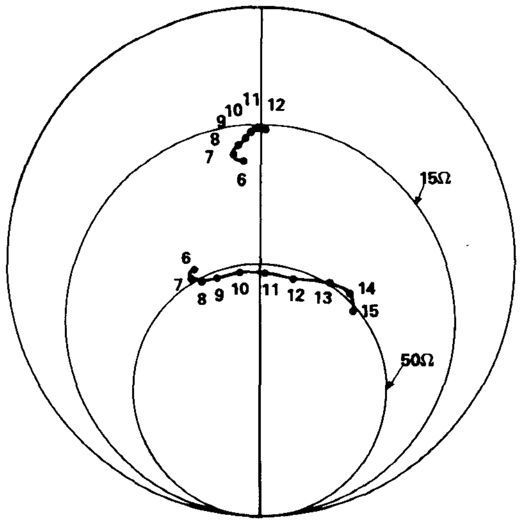

Fig. 7. Smith chart showing measured input impedances of artificial resistor before and after transformation to 50 Ω.

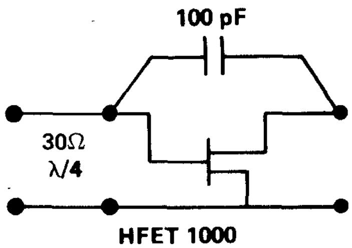

Fig. 8. Broad-band artificial 50-Ω load.

Measurements of the effective noise temperature of this resistor proved difficult. A slide tuner was used to convert the low effective resistance to the 50 Ω required by the noise measuring system, but the sharp response characteristics of the tuner made the bandwidth of the transformed resistor smaller than the bandwidth of the measuring system. However, by comparison with the noise of a room temperature resistor, it was estimated that at 9.5 GHz, the effective noise temperature was measurably below room temperature, while it was above room temperature at higher frequencies. Although we did see the below room temperatures predicted by the analyses, this trend with frequency was opposite to what was calculated using the noise data from the manufacturer.

We then added a 30-Ω quarter-wave transmission line, as shown in Fig. 8, and produced a broad-band match to 50 Ω. As can be seen in the lower curve on the Smith chart in Fig. 7, the measured impedance now goes smoothly from $50-25j$ Ω at 8 GHz, through $52+0j$ Ω at 11 GHz, to $50+25j$ Ω at 13 GHz, or a bandwidth of 5 GHz.

VIII. Temperature Minimization Using Transmission Lines

We next examined the use of transmission lines with the fed-back FET to optimize the effective resistance and the equivalent temperature. The circuit we analyzed consists of two microwave transmission lines and a microwave FET. One transmission line is between the single input–output port of the device and the gate of the FET, while the other (acting as the feedback connection) is between the port and the drain of the FET (the source being common). In practice, there will also be a blocking capacitor in the feedback line and some inductance in the connections, but these reactances can be accommodated by slight changes in the lengths of the two transmission lines.

Since the available microwave analysis programs such as COMPACT are only suitable for two-port devices, we developed a detailed program that computes the effective input impedance, equivalent noise temperature, and stability for these one-port devices. Using this program, we found that by varying the lengths and impedances of the two transmission lines, we can not only change the effective resistance seen at the port, but we can also affect the noise temperature and stability.

One specific example we calculated is shown in Fig. 9. The circuit consists of an HFET-1000 with a 50-Ω transmission line of 180° at 11 GHz in the feedback loop and a 50-Ω 45° long line to the gate. The program indicates that with these parameters we should be able to produce an unconditionally stable resistor of $8.8 \pm 5.5j$ Ω with an effective noise temperature of 118 K. This resistor can be transformed to a standard 50-Ω artificial resistor with a near-quarter-wavelength 21-Ω transmission line. The temperature of this 50-Ω load will remain at 118 K.

We can turn this circuit into an amplifier by converting it into an electronically cold negative resistor. The negative resistor is obtained by leaving the feedback transmission line at 180° to insure circuit stability and increasing the gate transmission line by a half-wavelength to 235°. These

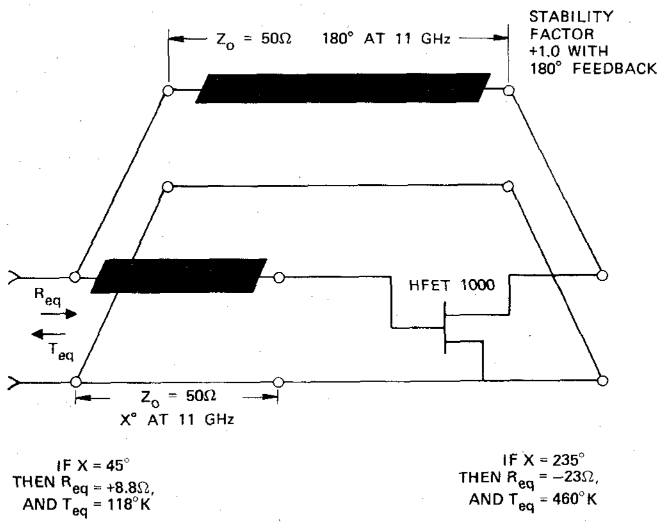

Fig. 9. Positive or negative electronically "cold" artificial resistor using transmission lines for tuning.

lengths for the transmission lines produce a negative resistor of $-23+32j\ \Omega$ with an effective noise temperature of 460 K (4.0 dB). Since the noise figure of the FET at 11 GHz is 3.27 dB or an effective noise temperature of 337 K, it is not surprising to see this noise temperature for the negative resistance. The negative resistor should be capable of producing an amplifier with the same noise temperature of 460 K or a 4.0-dB noise figure. Since it is a negative resistor, its gain in a negative resistance amplifier circuit (such as is used in tunnel diode amplifiers) is limited only by stability and bandwidth criteria. Using this approach, we should be able to get an amplifier with the gain of a tunnel diode and the noise figure of a FET.

Unfortunately, funding was not available to construct the electronically cold positive and negative resistors to check the accuracy of the analysis. In addition, recent advances in GaAs FET fabrication techniques have produced microwave FET's with significantly better noise figures than the HFET-1000. As soon as the appropriate microwave device parameters are available for these devices, they should be analyzed to determine their characteristics when used as an artificial resistor. With their better noise parameters, they should be significantly "colder" than the HFET-1000.

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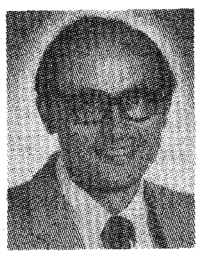

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