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DESIGN OF A LOW NOISE, BALANCED, 2-4 GHz GaAsFET AMPLIFIER

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I. Introduction

The advance of radio astronomy into the submillimeter region requires wider instantaneous bandwidth than for millimeter wave observations. For low-noise receivers utilizing SIS or Schottky-diode mixers, bandwidth will be limited by the low-noise IF amplifier. At present, the amplifier utilized in most is the 1-5 GHz FET amplifier with 0.5 GHz bandwidth described by S. Weinreb, et al. [1]. The critical element is a feedback source inductance to get both input power matching and noise matching.

This report describes the design of a 3 GHz FET amplifier with 2 GHz bandwidth. As it is difficult to get an appropriate feedback inductance for such a wide bandwidth, a balanced amplifier [2] with a Lange coupler [3], [4] as 3dB hybrid is utilized. This choice is superior to a single-ended amplifier with an isolator because the Lange coupler is easy and has wider bandwidth than the isolator, especially when cryogenic cooling is considered. For the input circuit design, only noise matching need be considered. The input network is a three-section transmission line which has output impedance relatively independent of frequency as described by N. Takahashi [5]. The noise temperature is within 4K of the minimum noise temperature of the FET over the 2- to 4- GHz range. FET parameters are for the MGF1412A at room temperature, but it is intended to cool the amplifier to 15K.

In this design, the HP9816 was used with the software of FARANT [6] established at NRAO. In Section II, an effect of the deviation from the ideal 3dB coupler is discussed. In Section III, the design of the matching networks for the input and the output are described.

II. Balanced Amplifier with Lange Coupler

The microwave balanced amplifier that is formed by two identical, single-ended amplifiers in cascade with a 3dB hybrid was studied in detail by K. Kurokawa [2]. This configuration is shown in Figure 1 and is made very suitable for wide bandwidth by using the Lange coupler. Even if the two amplifiers (a) and (b) have poor match at the input and output, the balanced amplifier can maintain good power match over the wide frequency range of 3dB Lange couplers [3]. The ideal 3dB coupler has $t^2 = 0.5$ and $\emptyset = \pi/2$, where t is the amplitude transmission factor and $\emptyset$ is the transmission phase.

Even though a Lange coupler is used, the factor t^2 varies from 0.50 to 0.55 over 2 to 4 GHz. This deviation causes a deterioration of the power matching that is induced from [2], Equation 11. The resulting standing wave ratio, ρ_{SWR} at the input or the output is

$$\rho_{SWR} = \frac{1 + (2t^2 - 1)\ |\Gamma a|}{1 - (2t^2 - 1)\ |\Gamma a|}\ , \tag{1}$$

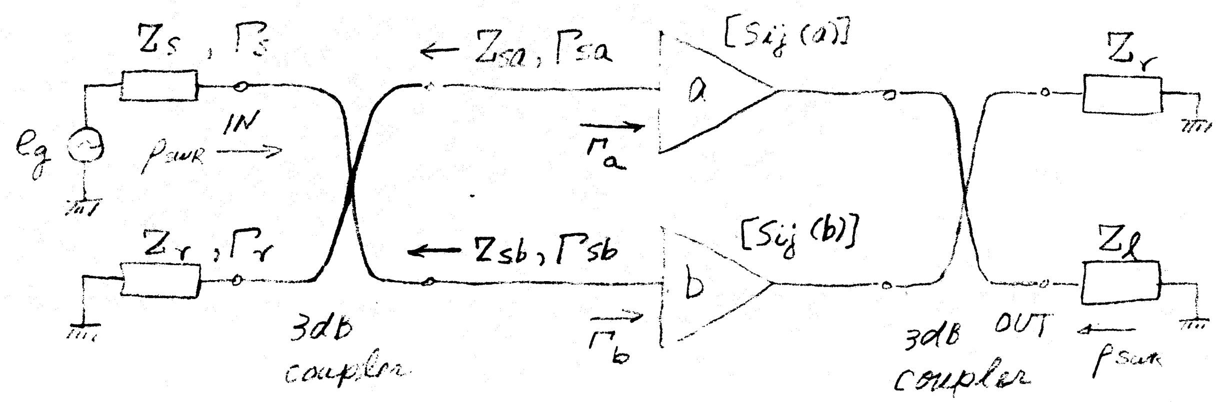

Fig. 1. The configuration of the balanced amplifier.

where Γa is the reflection coefficient of the single-ended amplifier at the input or the output, respectively. Figure 2 shows ρ_{SWR} vs t^2. Even with $|\Gamma a|$ equal to 1, we can expect $\rho_{SWR} \leq 1.2$ as a worst case. The deviation of t from ideal also gives an excess noise T_{e1} from the termination R as is shown in [2], Equations 15-21. Assuming $\phi = \pi/2$, T_{e1} is represented by

$$T_{e1} = \frac{2t^2 - 1}{4t^2(1 - t^2)} T_r \quad , \tag{2}$$

where T_r is the noise temperature of the input termination R. T_{e1}/T_r is plotted in Figure 3.

Another noise temperature increase, T_{e2}, is caused when the scattering parameters $S_{21}(a)$ and $S_{21}(b)$ of each single-ended amplifier respectively are not equal. The excess noise T_{e2} has the value

$$T_{e2} = 4\left|\frac{r_s - 1}{r_s + 1}\right|^2 T_r \quad , \tag{3}$$

where r_s is defined as $|S_{21}(a)/S_{21}(b)|$ and $t^2 = 1/2$, $\phi = \pi/2$ are assumed. Figure 4 shows T_{e2}/T_r vs r_s. The total excess noise from the input termination R is then given by $T_{e1} + T_{e2}$. If we assume that $t^2 = 0.55$ and $r_s = 1.4$ are practical limits, the excess noise is over 20% of T_r. Therefore, the termination R should be cooled to cryogenic temperatures.

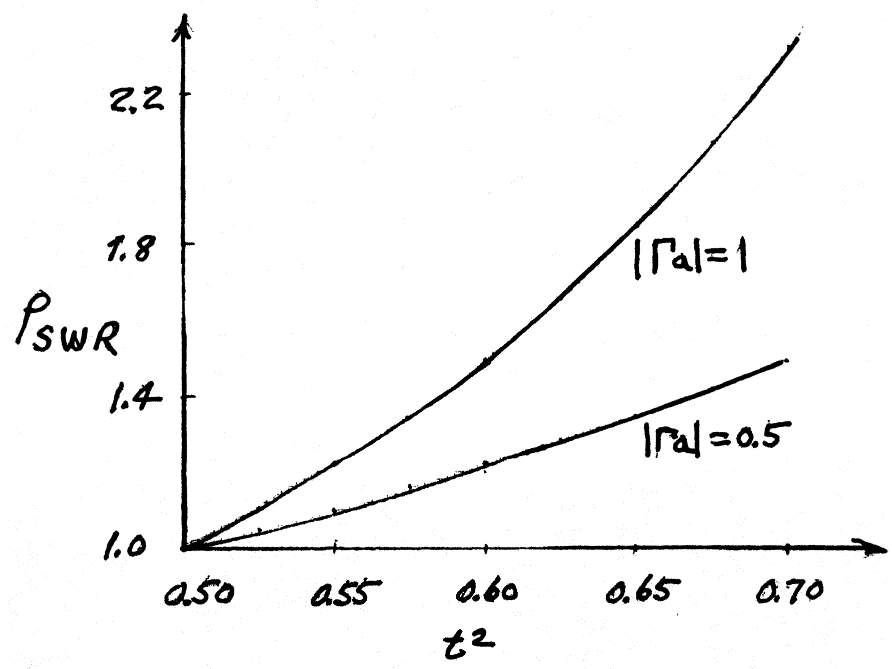

Fig. 2. Standing wave ratio ρ_{SWR} vs power transmission factor t^2.

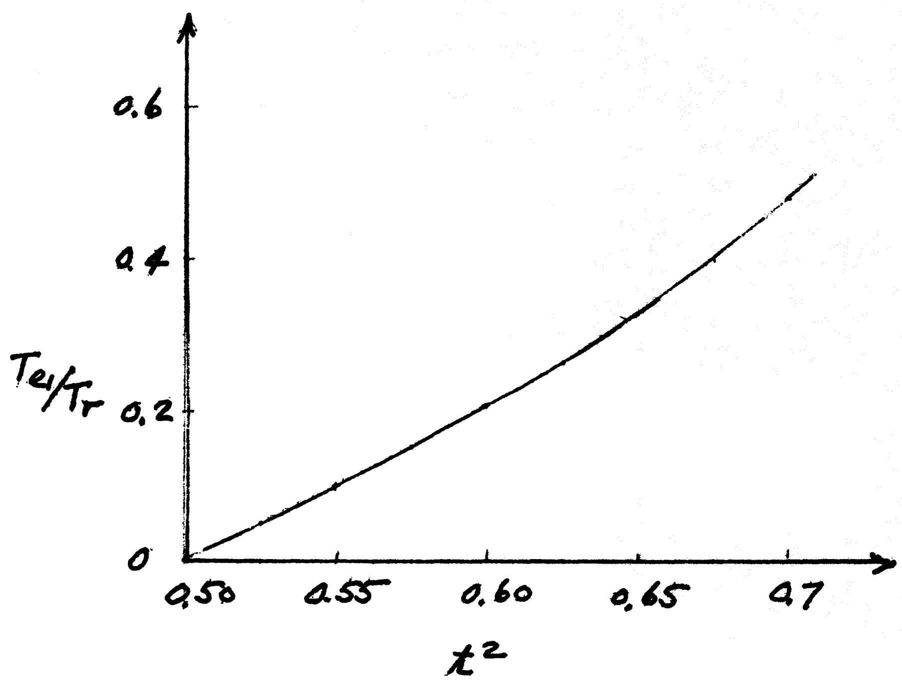

Fig. 3. Excess noise T_{el}/T_r vs power transmission factor t^2.

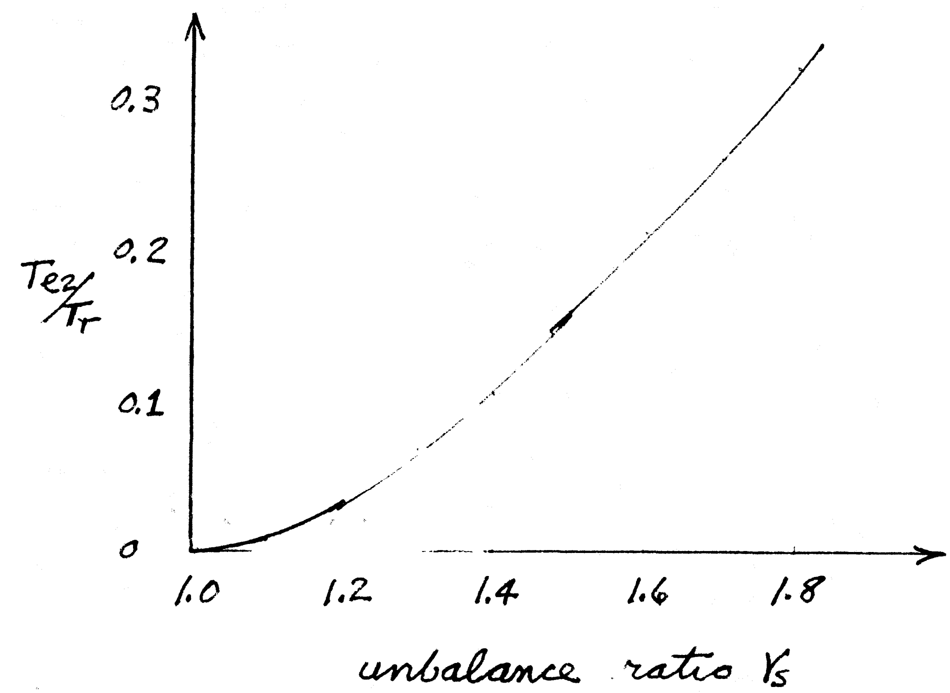

Fig. 4. Excess noise T_{e2}/T_r vs unbalance ratio $r_s = |S_{21}(a)/S_{21}(b)|$.

The signal source impedances, Z_{sa}, Z_{sb}, presented to each amplifier (see Figure 1) have a value different from Z_s when $t^2 \neq 0.5$ and $\emptyset \neq \pi/2$. Let the reflection coefficients Γsa and Γsb, for Z_{sa} and Z_{sb} respectively, be

$$\Gamma_{sa} = \Gamma_s \alpha^2 + \Gamma_r \beta^2 + (\Gamma_s + \Gamma_r)\,\Gamma_b\, \alpha^2 \beta^2$$

$$\Gamma_{sb} = \Gamma_r \alpha^2 + \Gamma_s \beta^2 + (\Gamma_s + \Gamma_r)\,\Gamma_a\, \alpha^2 \beta^2 \quad ,$$

where Γ_s and Γ_r are the reflection coefficients of Z_s and Z_r respectively and

$$\alpha = j \sqrt{1 - t^2}\; e^{-j\emptyset} \; , \qquad \beta = t\; e^{-j\emptyset} \quad .$$

Of course, when Z_s and Z_r are equal to the line characteristic impedance Z_o, Z_{sa} and Z_{sb} are equal to Z_s. If Z_s is different from Z_o, amplifiers (a) and (b) have different noise temperatures depending on Z_{sa} and Z_{sb}. Setting $|\Gamma_a|$ and $|\Gamma_b|$ equal to 1 as a worst case, the maximum of $|\Gamma_{sa}|$ or $|\Gamma_{sb}|$ is then

$$|\Gamma_{sa}| \text{ or } |\Gamma_{sb}| = \frac{1}{2}|\Gamma_s| \left(1 + \frac{1}{2}|\Gamma_s|\right) .$$

Thus the source impedance driving each amplifier is a function of the source impedance driving the balanced amplifier. This is not the case for an amplifier driven with an isolator, but in that case, the mismatch between source and isolator must also be considered.

III. Matching Networks

In order to minimize the noise temperature, T_n, of the amplifier, the signal source impedance Z_s presented at the input of the FET should be equal to the optimum noise impedance Z_{opt} of the FET. When Z_s is equal to Z_{opt}, T_n is equal to the minimum noise temperature T_{min} of the FET. In general, it is difficult to obtain this condition over all frequencies. Then T_n is given as follows:

$$T_n = T_{min} + 290 \frac{G_n}{R_s} \{(R_s - R_{opt})^2 + (X_s - X_{opt})^2\} ,$$

where $Z_s = R_s + jX_s$ and $Z_{opt} = R_{opt} + jX_{opt}$ are defined and G_n is the noise conductance of the FET.

Figure 5 shows Z_{opt} of the MGF1412A without package capacitance on the Smith chart from 2 GHz to 4 GHz. The starting point "0" is indicated for 2 GHz. Note that Z_{opt} rotates counterclockwise as the frequency increases. On the other hand, the signal source impedance Z_{sn} of a transmission line rotates clockwise when Z_s is not equal to the characteristic impedance Z_o. To make Z_{sn} close to Z_{opt} over a wide bandwidth, it is necessary to give a counterclockwise rotation to Z_{sn}.

Lumped circuit elements may be used but are rather lossy in this frequency range. In this design, the networks are all composed of microstrip transmission lines except for some parts of the DC bias circuit.

The method [5] used to give reverse rotation for the wide bandwidth $|\Delta f/f_o| \leq 2/3$ is realized by a short-circuited $\lambda/4$ line at a point where the reflection coefficient toward the signal source has a pure real value at the center frequency. When Z_{opt} is inductive, a shunt stub can be connected at the point where Γ is real and negative and when Z_{opt} is capacitive, a series stub can be inserted at the point where Γ is real and positive. As a simple example, reverse rotation is approximately produced by the circuit of Figure 6, which has three lines. The first line, Y_t, may be used to give real negative Γ_t toward the signal source at the center frequency f_o. Assuming

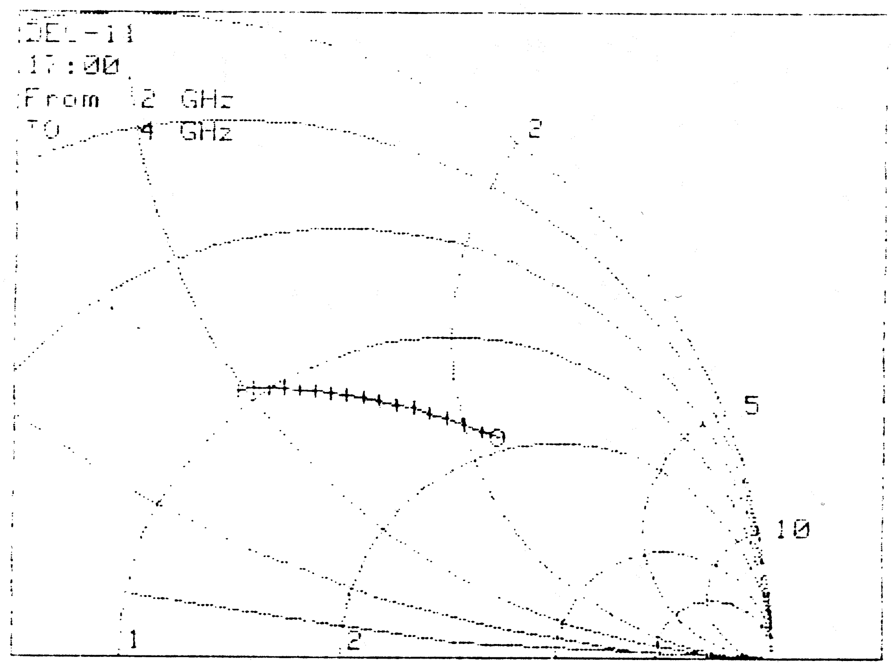

Fig. 5. Z_{opt} for MGF1412.

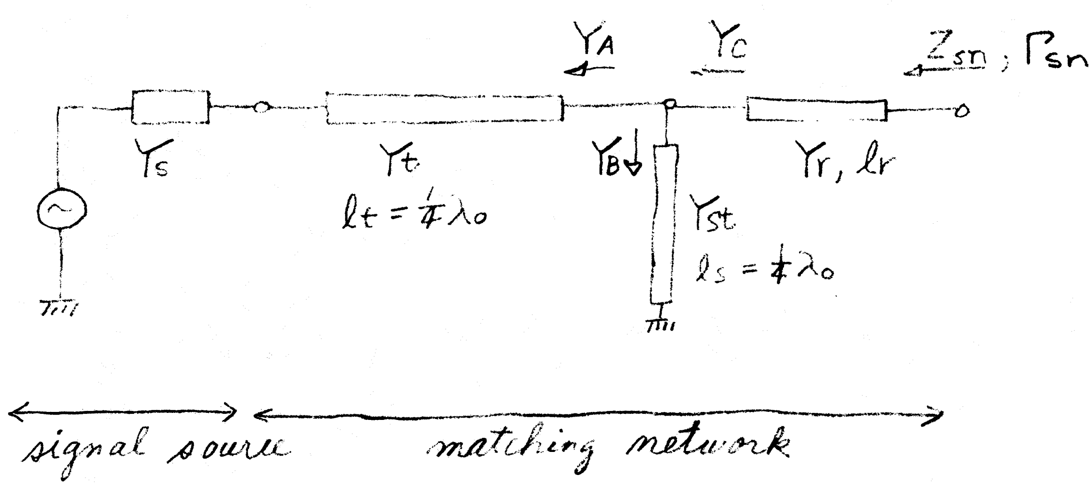

Fig. 6. The matching network with three lines.

$Y_s < Y_t$ and all of Y_s, Y_t, Y_{st} and Y_r are real, the admittance Y_A without Y_{st} and Y_r is

$$Y_A = \frac{Y_t^2}{Y_s} - jY_t\left\{\left(\frac{Y_t}{Y_s}\right)^2 - 1\right\}\frac{\pi}{2}\,\frac{\Delta f}{f_o} \quad ,$$

where the frequency $f = f_o + \Delta f$. The admittance Y_B for the stub Y_{st} is

$$Y_B = jY_{st}\,\frac{\pi}{2}\cdot\frac{\Delta f}{f_o} \quad .$$

Then Y_C for the sum $Y_A + Y_B$ is,

$$Y_C = \frac{Y_t^2}{Y_s} + j\left[Y_{st} - Y_t\left\{\left(\frac{Y_t}{Y_s}\right)^2 - 1\right\}\right]\frac{\pi}{2}\cdot\frac{\Delta f}{f_o} = \alpha + j\beta\frac{\Delta f}{f_o}$$

The real part α of Y_C represents a transformer. The imaginary part factor β of Y_C can be made positive with a large Y_{st}. Finally, Γ_{sn} through Y_r is

$$\Gamma_{sn} = |\Gamma_{sn}|\, e^{j\phi} \quad .$$

It follows that

$$\phi = -\frac{4\pi}{\lambda_o} \ell r + \pi + (\tan^{-1} \frac{\beta}{|Y_c|^2 - 1} \frac{\Delta f}{f_o}) - \frac{4\pi}{\lambda_o} \ell r \cdot \frac{\Delta f}{f_o}$$

Under some conditions the phase ϕ of Γ_{sn} can increase with frequency so that CCW rotation is realized.

In the case of the 2-4 GHz FET amplifier, the three-line network can be used but the length of ℓr must be large to give a positive value for Γ_{sn}. For DC bias, a $\lambda_o/4$ transmission line is suitable. The connection point where the effect on matching is smallest is between Y_s and Y_t. Figure 7 shows a matching network which has been designed by such methods. The source impedance presented by the input network is shown in Figure 8. The noise temperature shown in Figure 9 is increased by less than 4K above T_{min}.

The network for the output of the FET is designed for power match utilizing the same technique. Figures 10 and 11 show the input and output impedances of the amplifier, respectively. In order to improve power matching for the input and output, a resistor was added to the output circuit, as shown in Figure 12, with results given in Figures 13 and 14. In this case, if the temperature of the resistor is assumed to be 15K, the noise temperature of the amplifier stage is only increased by 0.8K with a gain decrease of less than 3 dB.

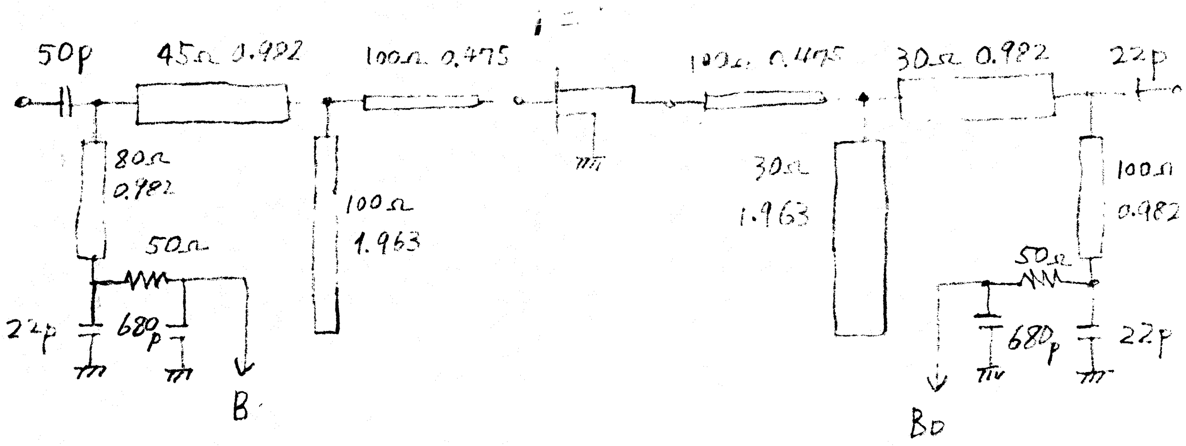

Fig. 7. The network designed for MGF1412 in chip form but including bond wire inductances. The lengths of lines shown are in inches for $\varepsilon_r = 1$.

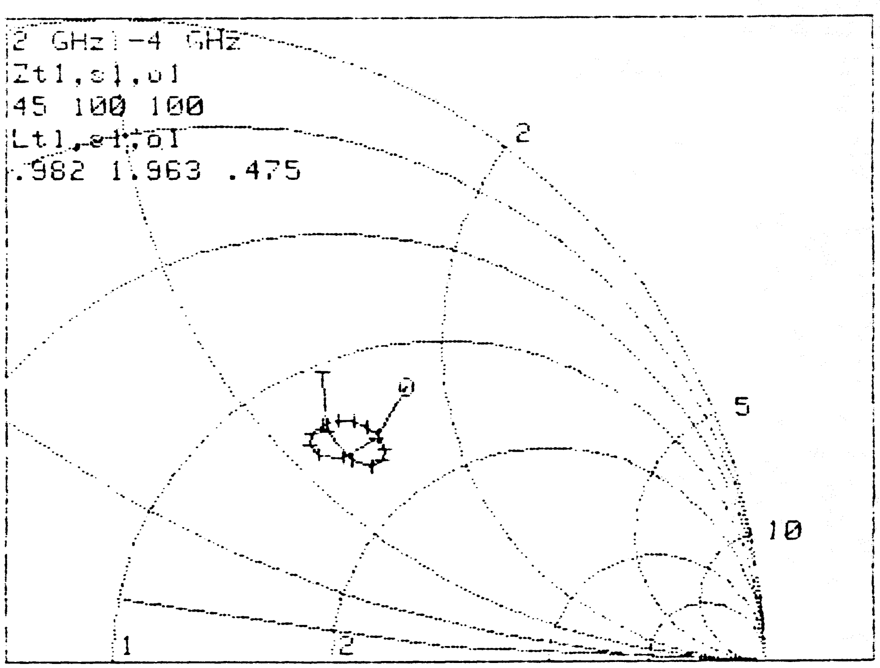

Fig. 8. The signal source impedance presented by the input network of Figure 7.

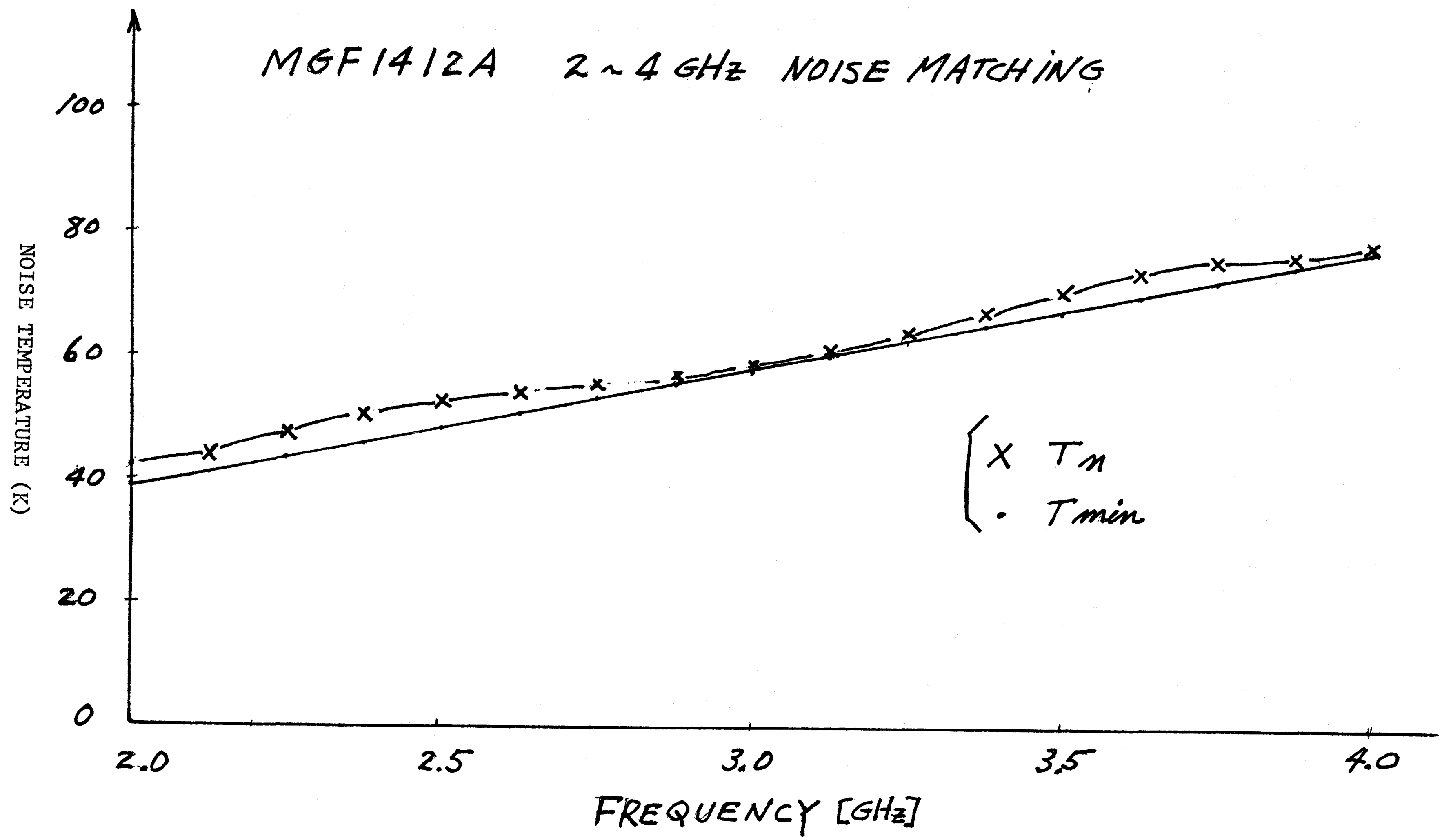

Fig. 9. Noise temperature of the amplifier.

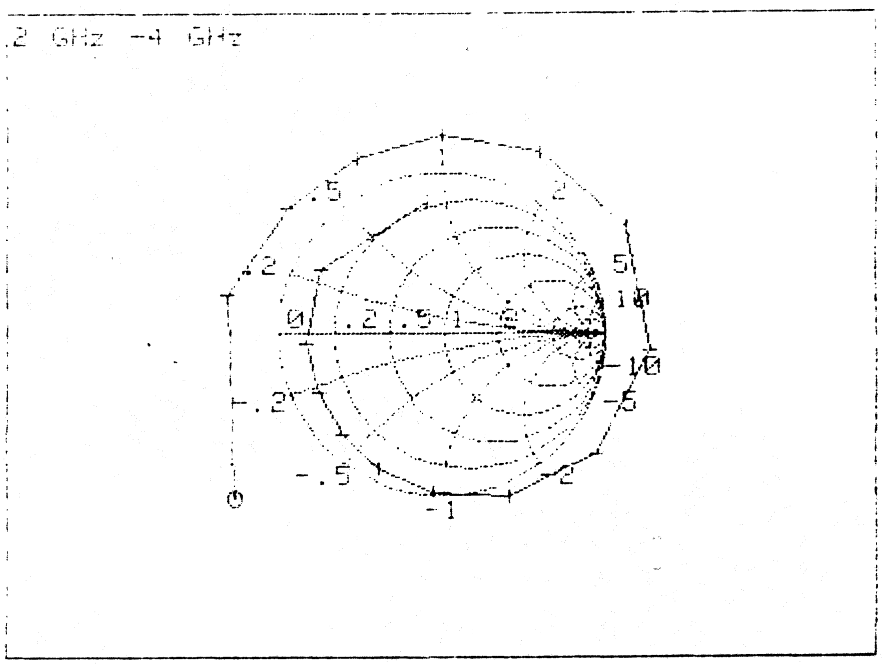

Fig. 10. The input impedance.

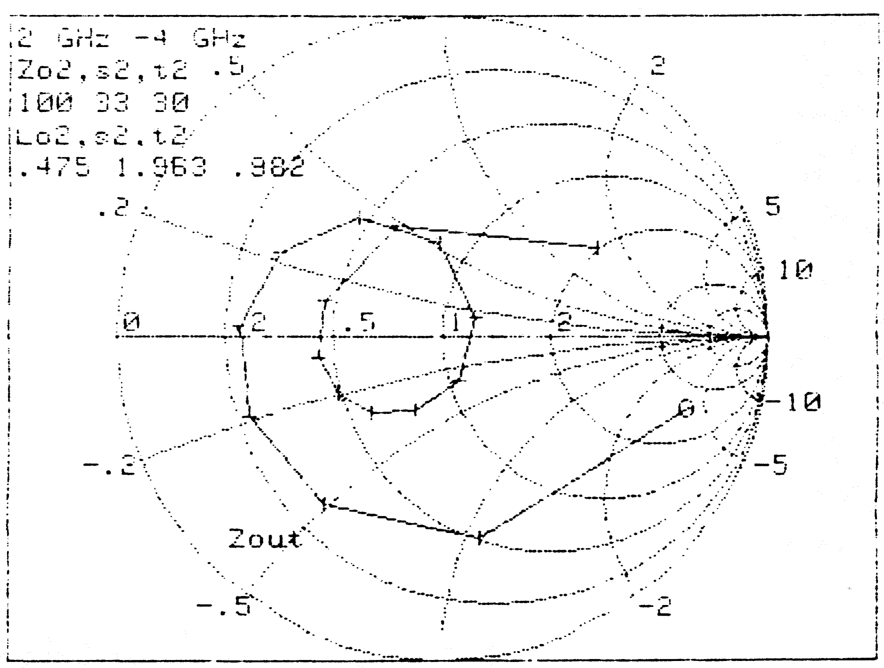

Fig. 11. The output impedance

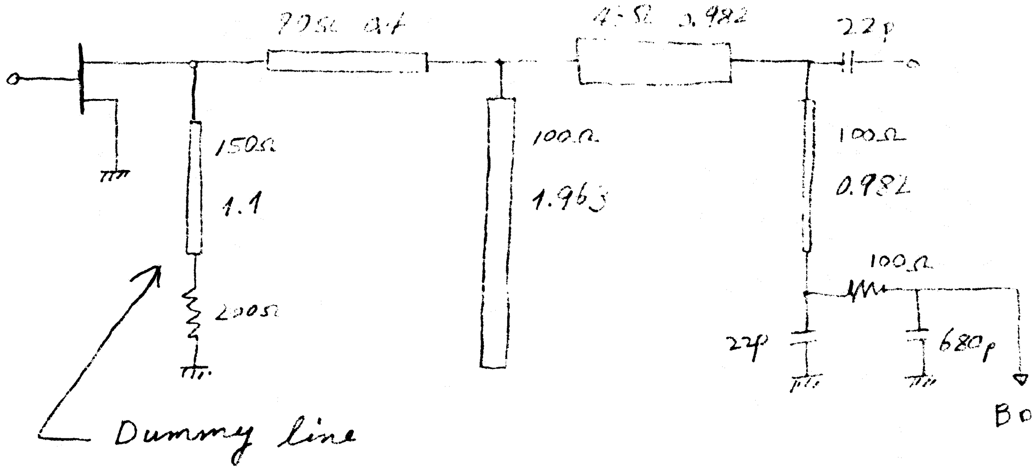

Fig. 12. The lossy output network to improve input and output match.

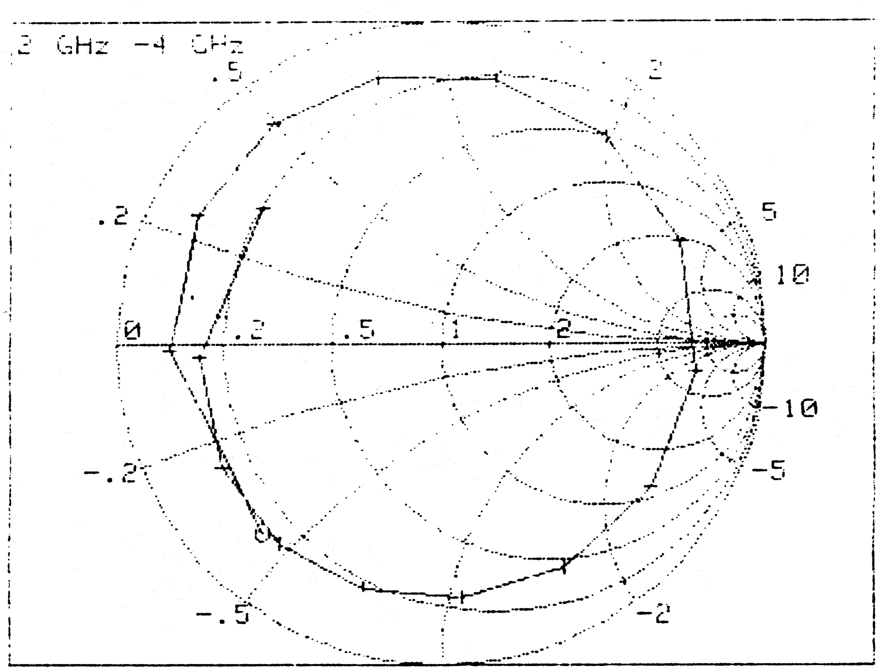

Fig. 13. The input impedance with resistor in output circuit of FET.

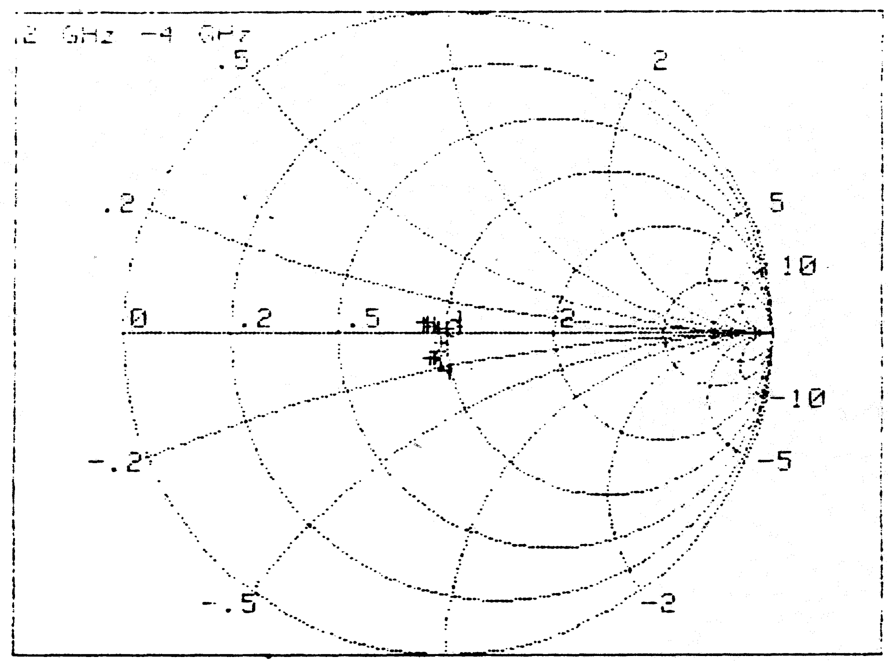

Fig. 14. The output impedance with resistor in output circuit of FET.

IV. Conclusion

A low-noise, balanced amplifier with a 2 GHz bandwidth from 2 to 4 GHz can be designed by using the Lange coupler and the MGF1412 FET. If the amplifier is cooled to 15K and the input termination is cooled to 4K, the increase in noise above the minimum noise temperature is estimated to be less than 5K over the frequency range.

The next steps to be performed in this design are the following:

1) Reoptimize the circuits of Figures 7 and 12 using FET noise parameters measured at 3 GHz and 15K temperature.

2) Design a microstrip layout of the circuit. A microstrip layout of a 2-4 GHz Lange coupler on .039" thick alumina is shown in Figure 15.

Dimension for Lange Coupler.

$\varepsilon_r = 10.1$ Alumina $1\,mm^t$

$W/H = 0.112$ $S/H = 0.08$ $\sqrt{\varepsilon_{eff}} = 2.593$

$\therefore W = 0.112$ mm $= 4.4$ mils (line width)

$\therefore S = 0.08$ mm $= 3.15$ mils (line space)

$f_0 = 3$ GHz $\frac{1}{4}\lambda_0 = 25$ mm

$\therefore l = 9.64$ mm $\doteqdot 0.38$ inches (line length)

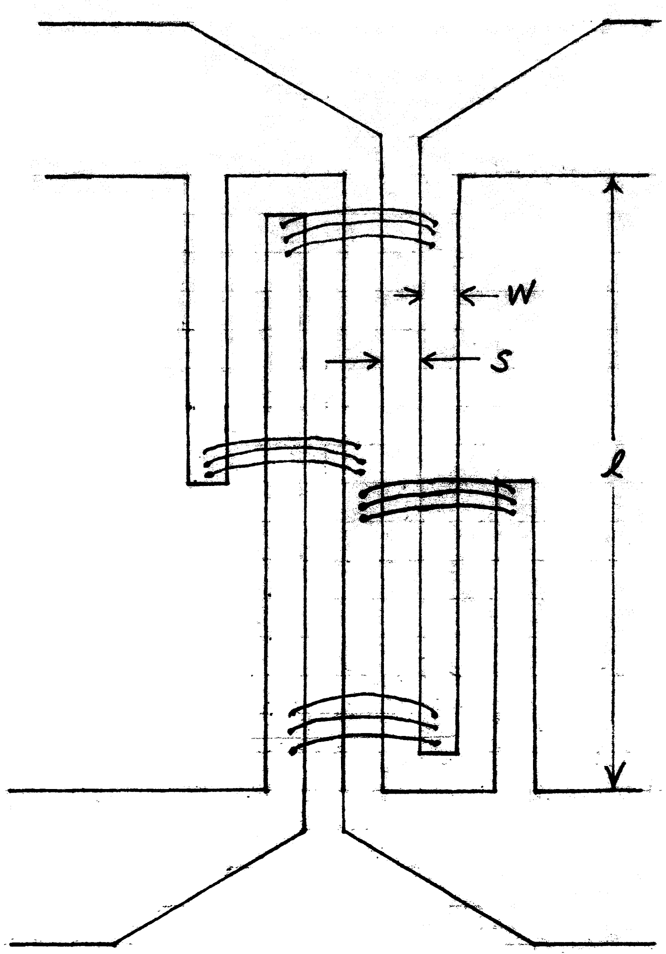

Fig. 15. Dimensions for Lange coupler.

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