Ultra-Low-Noise 1.2- to 1.7-GHz Cooled GaAsFET Amplifiers

SANDER WEINREB, FELLOW, IEEE, DAN L. FENSTERMACHER, MEMBER, IEEE, AND RONALD W. HARRIS

Abstract — **A 3-stage GaAsFET amplifier operating at 13 K and utilizing source inductance feedback is described. The amplifier has a noise temperature of < 10 K and input return loss > 15 dB over the 1.2- to 1.7-GHz frequency range.**

I. Introduction

IN THE microwave frequency range between 1 and 3 GHz, the antenna noise temperature due to cosmic and atmospheric sources is quite low, ~5 K, and it is a challenge to receiver engineers to design low-noise amplifiers and low ground-radiation pickup antennas to take advantage of this low source temperature. The noise figure, in decibels, of a receiver is equal to the signal-to-noise (S/N) degradation caused by the receiver when the source temperature is 290 K. When the source temperature including antenna noise is 10 K, a reduction of receiver noise figure from 1.0 dB (or a noise temperature of 75 K) to 0.1 dB (6.8 K) causes a S/N improvement of $(75+10)/(6.8+10)=5.1$ or 7.0 dB.

The amplifier described in this article has a noise figure of approximately 0.1 dB at 1.3 GHz. This is achieved with wide bandwidth, impedance match, excellent stability, and relatively small cost and complexity compared with masers which have been required in the past for this noise performance. Cryogenic cooling to a temperature of ~15 K is required but this can now be obtained with closed cycle helium refrigerators [1] at a cost of under $5000 and a weight less than 42 kg. Some specific applications for the amplifier are: 1) as a radio astronomy front end for observations of the 1.42-GHz hydrogen line and OH lines at 1.61, 1.66, and 1.72 GHz; 2) as an IF amplifier for millimeter wave or infrared receivers utilizing cooled mixers; 3) as a front end for detection of extraterrestrial civilizations communicating in the "optimum" frequency range of 1.42 to 1.67 GHz [2].

In a previous article [3] a cryogenically cooled L-band amplifier utilizing source inductance feedback to achieve input match was described. This present paper reports on the further development of this amplifier utilizing computer-aided design techniques to increase bandwidth to 500 MHz and provide stability for any source and load impedance. Photographs and a schematic of the amplifier are shown in Figs. 1–3.

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S. Weinreb and R. Harris are with the National Radio Astronomy Observatory, Charlottesville, VA 22903.

D. L. Fenstermacher was with the National Radio Astronomy Observatory, Charlottesville, VA 22903. He is now with the Electrical Engineering Department, Cornell University, Ithaca, NY 14853.

II. Design

A first step in the amplifier design concerned the question of how to handle the mismatch which results when a FET device is driven by its optimum-noise generator impedance Z_{opt}. For a FET in the lower microwave range, this difference is quite large; typical values are $Z_{opt}=42+j175$ and the matched source impedance, $Z_{in}^*=10+j145$, giving an input voltage reflection coefficient magnitude of 0.73. It is, of course, possible to operate the amplifier with a large input mismatch, but small variations of the antenna feed impedance will then cause large ripples in the gain versus frequency response. Possible remedies to this problem are 1) an isolator, 2) a balanced amplifier, or 3) use of feedback. Feedback was chosen because it results in the most compact, fewest component amplifier, and because of previous experience with this technique [3], [4]. It should be mentioned that a cooled isolator in this frequency range has recently been developed [5].

The feedback to produce simultaneous noise and power match is realized in the form of the source inductor $L10$ of the first stage. The effect of this feedback upon noise performance is to reduce the optimum-noise generator reactance X_{opt} by the reactance of $L10$ [6]; this effect is fairly minor and is compensated by reducing the value of the gate circuit inductor $L1$. The noise figure of the amplifier is, to a large degree, unaffected by the feedback because the noise measure of the first stage is invariant to lossless feedback [7], the stage gain is fairly high, ~10 dB, and the second stage has similar noise figure as the first stage.

On the other hand, the effect of the source inductor upon input match can be represented by an impedance Z_{FB} added in series with the gate. This impedance can be calculated by a straightforward circuit analysis, is somewhat dependent upon first-stage load impedance, and, to a first approximation, is resistive and given by $g_m \cdot L10/C$ where g_m and C are the transistor transconductance and gate capacitance.

Optimization of the amplifier was performed using the FARANT program developed at NRAO [8]. The optimization tradeoff's are illustrated by the impedance plot of Fig.

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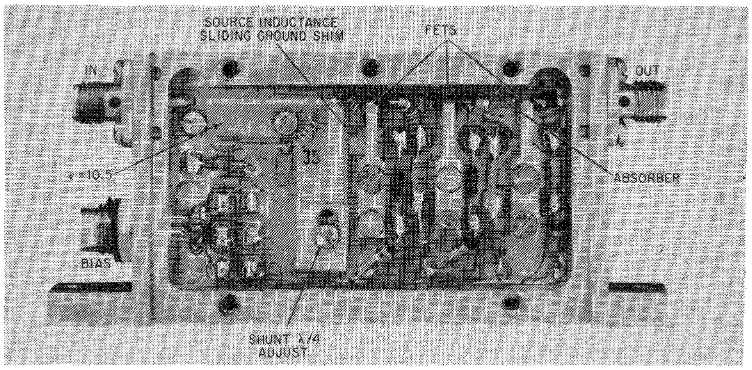

Fig. 1. Photograph of 3-stage amplifier with cover removed. Overall size is 9.15 cm×4.07 cm×1.27 cm. The FET's are soldered by their source leads to FET holders which are described in Fig. 3.

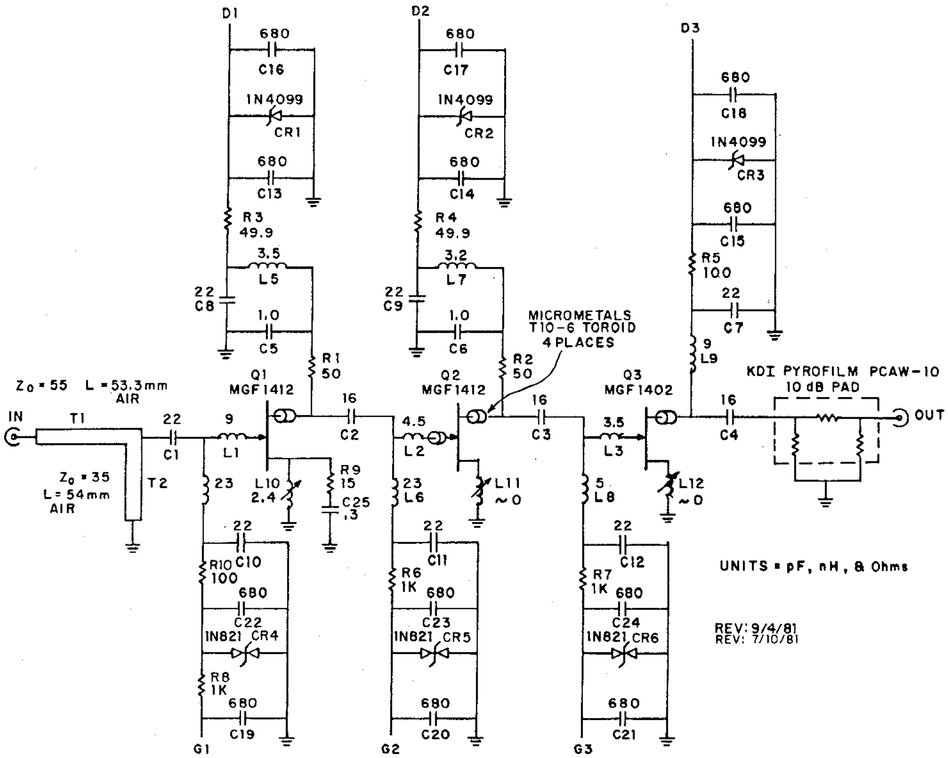

Fig. 2. Amplifier schematic. Bias voltages are supplied from a separate regulator which adjusts gate voltage for constant drain current.

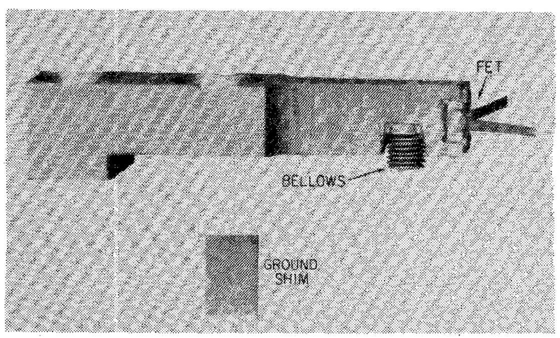

Fig. 3. Side view of FET holder and grounding shim. The shim allows adjustment of the path length from the FET source to ground, thus providing a variable source inductance. The bellows [19] contacting a 15-Ω chip resistor in series with a 0.3-pF chip capacitor soldered to the chassis is used only on the first stage; second and third stages require no additional source inductance and the grounding shim is positioned directly under the FET.

4 and consideration of the noise temperature dependence upon generator impedance $R_s + jX_s$, for any linear two-port

$$T_n = T_{\min} + 290 \times \frac{g_n}{R_s}\left[\left(R_s - R_{\mathrm{opt}}\right)^2 + \left(X_s - X_{\mathrm{opt}}\right)^2\right]$$

where $R_{\mathrm{opt}} + jX_{\mathrm{opt}}$ is the optimum-noise generator impedance and g_n is the amplifier noise conductance as given in Table I. The first-stage source inductance $L10$, and load reactances $C5$, $L5$, and $L2$, have been adjusted to make the FET input resistance R_{in}, approximately equal to R_{opt}. The series and shunt $\lambda/4$ line lengths and characteristic impedances and the input inductor $L1$, are then adjusted to make R_s as close as possible to R_{in} and R_{opt} and to bring the source reactance X_s to a value between $-X_{\mathrm{in}}$ and X_{opt}. If the shunt stub is not present, X_s is a linear increasing

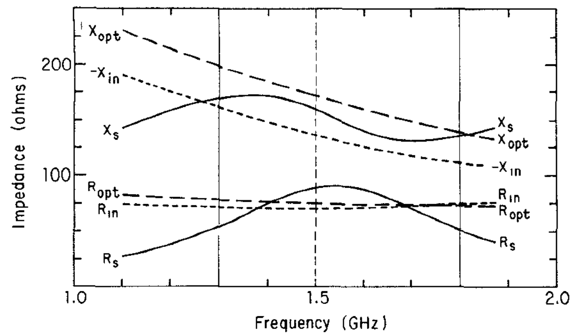

Fig. 4. Source impedance optimization example. The minimum noise impedance is $R_{opt} + jX_{opt}$, the FET input impedance is $R_{in} + jX_{in}$, and the generator impedance is $R_s + jX_s$. The negative slope in X_s is due to the shunt $\lambda/4$ transmission line. All quantities are computed from the circuit model and the S and noise parameters of the FET.

TABLE I
NOISE PARAMETERS OF MITSUBISHI MGF-1412 AT 1.6 GHz, TEMPERATURES OF 300 K AND 15 K, AND DRAIN BIAS OF 5 V, 10 mA

AMBIENT TEMP °K	T_{min}	$1/g_n$ ohms	R_{opt}	X_{opt}	COMMENT
300	63 ± 3	250 ± 100	42 ± 4	-	Measured
300	20.3	1785	60.7	221	Theory [10], [11]
15	8 ± 2	2200 ± 1500	29 ± 2	-	Measured
15	4.2	5882	39.3	196	Theory [10], [11]

function of frequency and R_s has little frequency variation. This results in narrower bandwidth in both noise temperature and match due to the difference between X_s and X_{opt} or $-X_{in}$.

At the time the input impedance optimization was performed, R_{opt} was thought to be 75 Ω; later measurements, described in Table I, showed $R_{opt} = 42$ Ω. However, a higher value of generator resistance gives less noise temperature variation with frequency and a center-frequency value of 60.5 Ω is used in the amplifier.

A second major decision was whether to use lumped or transmission-line matching elements. At 1.5 GHz this is a close decision as it is difficult to determine the parasitic reactances of lumped elements with sufficient accuracy, yet transmission lines are somewhat large. A compromise was made; the input network was constructed on high-dielectric constant (10.5) transmission-line circuit board [9] while the remainder of the amplifier was constructed with lumped elements. The input circuit board could be easily changed to vary the source resistance in a known manner.

A somewhat unconventional construction of the amplifier was based upon the following considerations.

1) The lowest-noise GaAsFET's for use at 1.5 GHz have considerable gain and a tendency to oscillate at 10 to 20 GHz, especially when source inductance feedback is utilized. The 1.5-GHz inductors have widely varying reactance in the 10- to 20-GHz frequency range and series RC bypass networks are necessary to stabilize the amplifier. For these bypass networks to be effective, their total length must be less than $\lambda/4$ at 20 GHz (~ 0.4 cm). Three of these bypass networks, $R9$-$C25$, $R1$-$C5$, and $R2$-$C6$ are constructed of very small chip resistors and capacitors and are located very close to the FET's; ideally they would be built into the FET package.

2) The first-stage source inductance and $\lambda/4$ shunt stub length are critical and should be easily adjustable. This is accomplished by small sliding shims as shown in Figs. 1 and 3.

3) The thermal resistance from the FET source leads to the amplifier case must be low at cryogenic temperatures to avoid self-heating. For this reason the FET source leads are soldered to a copper FET holder (see Fig. 3) with pure indium solder. The thermal resistance within the FET package is discussed in [10].

4) A "springy" mechanical design is needed to prevent fractures due to thermal stresses caused by the wide temperature range and materials with different thermal expansion coefficients. For this reason a dielectric-loaded Teflon substrate material [9] was used even though the temperature coefficient of its dielectric constant is much greater than that of alumina. (A 12-percent increase in dielectric constant was measured for cooling from 300 K to 15 K).

The Mitsubishi MGF-1412 transistors used in the design were selected on the basis of previous evaluations at 5 GHz [10] and also upon the results reported in [3]. Noise parameters of a MGF-1412 at approximately 1.6 GHz were measured at 300 K and 15 K by performing noise measurements with three different input circuit boards; these had no shunt $\lambda/4$ stub and had Z_0 values of 41, 51, and 69 Ω for the series $\lambda/4$ line. The results for a drain bias of 5 V and 10.8 mA are shown in Table I where they are compared with the theoretical values of Pucel *et al.* [11]. The agreement with theory is poor, in contrast to results reported at 5 GHz in [10], due to low frequency noise mechanisms not accounted for in the theory.

A MGF-1402 is used in the final stage because of its lower cost. The amplifier output is isolated from the load by an internal 10-dB chip attenuator [12]. Unlike an isolator, the attenuator terminates the amplifier over the entire microwave frequency range and thus insures that the amplifier will not oscillate for some particular out-of-band termination impedance.

III. CONSTRUCTION AND TUNING

The amplifier chassis is milled from copper which is then gold-plated for corrosion protection and to reduce absorption of thermal radiation. A rectangular bar, 1 mm×7 mm×27 mm, of iron-epoxy absorber material [13] is glued into the chassis as shown in Fig. 1. This material has been tested to retain its loss at cryogenic temperatures. Beryllium–copper finger stock material [14] is soldered to the top cover and contacts the FET holders near their mounting screws. Two of the coils, $L2$ and $L6$, are wound in reverse direction from the others to change the mutual coupling phase. All of these steps, as well as ferrite beads [15] on some of the leads, are for the purpose of suppressing high-frequency oscillations, particularly when the amplifier is cooled. Additional details concerning construc-

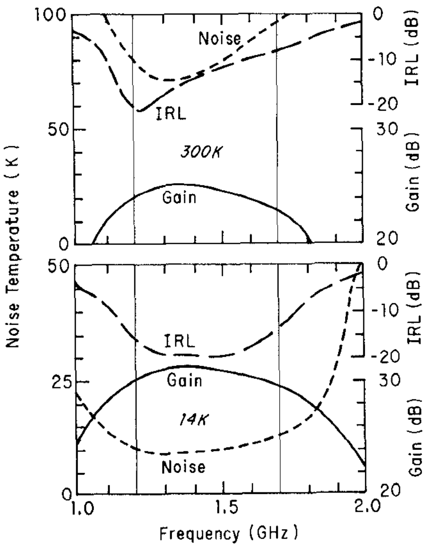

Fig 5. Input return loss (IRL), gain, and noise temperature of amplifier #74 at 300 K (top) and 14 K (bottom); all are referred to a room temperature dewar connector. Note scale change for noise temperature.

tion are presented in a report [16] available from the authors.

Tuning of the amplifier is necessary due to variability in the construction and installation of the inductors and to meet slightly different specifications for each application. Drain-bias voltage and current for stages 1, 2, and 3 are initially set at 5.5 V, 15 mA, 5.5 V, 12 mA, and 4 V, 10 mA, respectively. Inductor $L1$ and transformer $T2$ lengths are trimmed to minimize the noise at a desired frequency. The inductance of a coil may be raised or lowered by moving the turns closer or further apart; larger changes require more or less wire but this is usually not necessary. Inductors $L2$ and $L10$ are adjusted to achieve input match; it may be necessary to also change $L1$ and $T2$ for this requirement. Inductors $L5$, $L6$, $L8$, and $L9$ are adjusted to achieve the desired gain response.

The amplifier input match versus frequency changes appreciably as it is cooled as is shown in Fig. 5. This is due to the dielectric constant change in the input circuit board and to changes in gate-to-source capacitance caused by gate bias voltage changes necessary to keep drain current in an optimum noise range. To some extent the input match change can be compensated by pretuning at 300 K for the triangle shape shown in Fig. 5. This is often not satisfactory and iterations of tuning and cooling may be necessary.

IV. Measurement System and Results

Amplifiers are tested for input return loss and gain utilizing a scalar network analyzer [17]. For the return-loss measurement, test signal power level at the amplifier input is -25 dBm and is somewhat critical; a higher level causes an error due to second-stage overload and a lower level gives insufficient reflected power for the reflectometer bridge sensitivity. For gain measurement, the input signal level is -45 dBm.

Noise temperature and also gain are measured with the

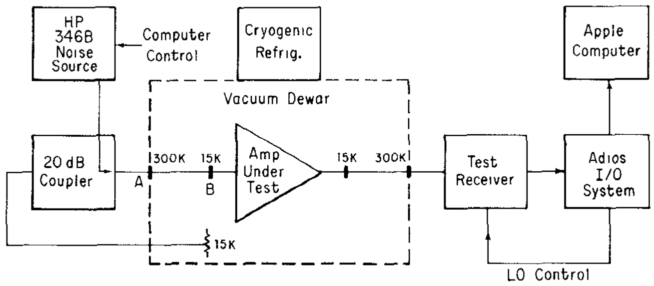

Fig. 6. Test setup for amplifier noise and gain measurements. Very low loss coaxial-line transitions were developed for connection, A to B, to the amplifier.

test setup shown in Fig. 6. The amplifier is mounted in a vacuum dewar and is thermally connected to a cryogenic refrigerator. The amplifier input is connected to the dewar exterior through a very low loss (0.08 dB, later reduced to 0.04 dB) coaxial line with APC 3.5 connectors and length 8.5 cm. This line has crystalline quartz inner conductor supports to form a vacuum seal and to insure that the amplifier end of the inner conductor is at the refrigerator temperature; the line will be described in detail in a future report.

Noise from a semiconductor-diode noise source [18] is coupled to the amplifier input through the side arm of a 20-dB directional coupler. The main arm of the coupler is connected to a cold termination in the dewar to increase accuracy by preventing addition of a large room temperature noise to the small amplifier noise. By replacing the coupler output, point A in Fig. 6, with hot and cold noise temperature standards, the noise temperature at point A with diode off, 28.8 K, and diode on, 94.1 K to 123.7 K dependent upon frequency, was calibrated. The noise diode was then turned on and off as receiver local-oscillator was scanned to give a swept-frequency measurement of noise temperature. A prior scan with the amplifier bypassed allows computation of the amplifier noise temperature, corrected for test receiver noise, and also the amplifier gain. This procedure is performed by an Apple II computer.

At the time of this writing, 30 amplifiers have been constructed. All amplifiers are stable for a sliding short of any phase connected to input or output. Oscillation of an amplifier can be detected either by observation of a bias value change or by observing the output of a broadband detector connected to the amplifier input or output port. Noise temperature, gain, and input return loss for one amplifier are shown in Fig. 5; this amplifier had the lowest noise temperature, 7.2 K, but is typical in gain and input match; typical noise temperatures are ~ 2 K higher. The noise temperature is referred to the cold amplifier input connector and is believed to be accurate within ± 0.5 K; a value of 8.3 K was measured at point A, outside of the dewar.

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[13] Eccosorb Type MF-124, Emerson and Cummings, Canton, MA.

[14] Type 97-500G Finger Contact Strip, Instrument Specialties Co., Delaware Water Gap, PA.

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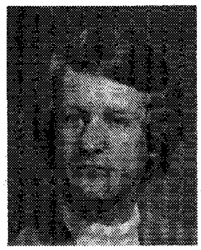
[18] Type 346B Noise Source, Hewlett-Packard Co., Palo Alto, CA.

[19] Type 2146 Bellows, Servometer Corp., Cedar Grove, NJ.

Sander Weinreb (SM'70-F'78) was born December 9, 1936, in New York, NY. He received the B.S. (1958) and Ph.D. (1963) degrees in Electrical Engineering from Massachusetts Institute of Technology, Cambridge, MA.

From 1960 to 1963 he was a Research Assistant at M.I.T. engaged in investigations of varactor frequency multipliers and digital autocorrelation techniques. In 1963 he joined Lincoln Laboratory where he was responsible for the radiometric equipment for the Haystack antenna. In 1965 he joined the National Radio Astronomy Observatory where he was Head of the Electronics Division and responsible for development of radio astronomy equipment for the Green Bank, WV, and Tucson, AZ observatories until 1977. He led the Electronics Design Group for the Very Large Array project from 1972 until 1975. In 1976 he took a two year leave at the Radio Astronomy Laboratory of the University of California, Berkeley and then returned to NRAO to specialize in the development of low noise devices.

Dr. Weinreb is a member of Sigma Xi, Eta Kappa Nu, Tau Beta Pi, and the International Scientific Radio Union. He is an advisor to the Netherlands Foundation for Radio Astronomy and the European Institute for Millimeter Wave Radio Astronomy.



Dan L. Fenstermacher (S'80–M'80) born on December 10, 1958 in Houston, TX. He received the B.S.E.E. degree from Cornell University, Ithaca, NY, in 1980.

Having previously been a co-op engineer for Corning Glass Works, he became involved in mircowave theory and computer aided design related to low-noise FET amplifiers at the National Radio Astronomy Observatory in Charlottesville, VA, during 1980 and 1981.

Mr. Fenstermacher is a member of Tau Beta Pi and is currently pusuring the M.S./Ph.D. degree in plasma physics at Cornell University.

Ronald W. Harris was born in Waynesboro, VA, on May 20, 1943. After graduating from Wilson Memorial High School, he attended classes at the University of Virginia School of General Studies, Charlottesville.

His first employment was with Sperry Marine Systems in the design and construction of electronic test equipment. Since that time, he has worked with ADF and LORAN receivers at Electronic Concepts, Inc., and in the bio-medical research department at the UVA Research Labs of Engineering Sciences. Since 1972, he has been employed by the National Radio Astronomy Observatory, Charlottesville, VA. His initial work was with the local oscillator group in the construction of the Very Large Array. Presently, he is engaged in the development of cooled low-noise GaAsFET amplifiers.