A Compact 1-kW 2-50 MHz Solid-State Linear Amplifier

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Solid-state high-power linear amplifiers are becoming more and more popular in the field of ham radio as the prices of HF power transistors continue to fall. 250-W devices are now available for almost half the price they were selling for a few years ago. RF power FETs are still more expensive, but eventually their prices will also fall, although not as fast since they are still novelty items and the manufacturing yields are low due to ESD problems and requirement for cleaner facilities for wafer processing.

General

It is much easier to design wideband power amplifiers with FETs than bipolar transistors mainly due to their higher input impedances at least up to VHF. Their input impedance also varies less with frequency than that of bipolar devices and changes in the output load line are reflected back to the input to a lesser degree because of the much lower value of feedback capacitance (collector to base vs drain to gate). Practically all RF power FETs on the market today are of the enhancement MOS type, meaning that positive voltage at the gate in respect to the source is required to turn the device on.

The 1-kW amplifier described here would be difficult, if not impossible, to design to cover four and a half octaves with comparable performance using inexpensive bipolar transistors. In addition, a series of power splitters and combiners would be required to reach high power levels. Biasing to class AB linear operation is also much simpler with FETs since the gate does not draw any dc current, whereas a current equal to $I_C(peak)/h_{FE}$ must be supplied to the base of a bipolar device. One example of this and the splitter-combiner complexity is presented in the Application Note AN-758 by Motorola, Inc.

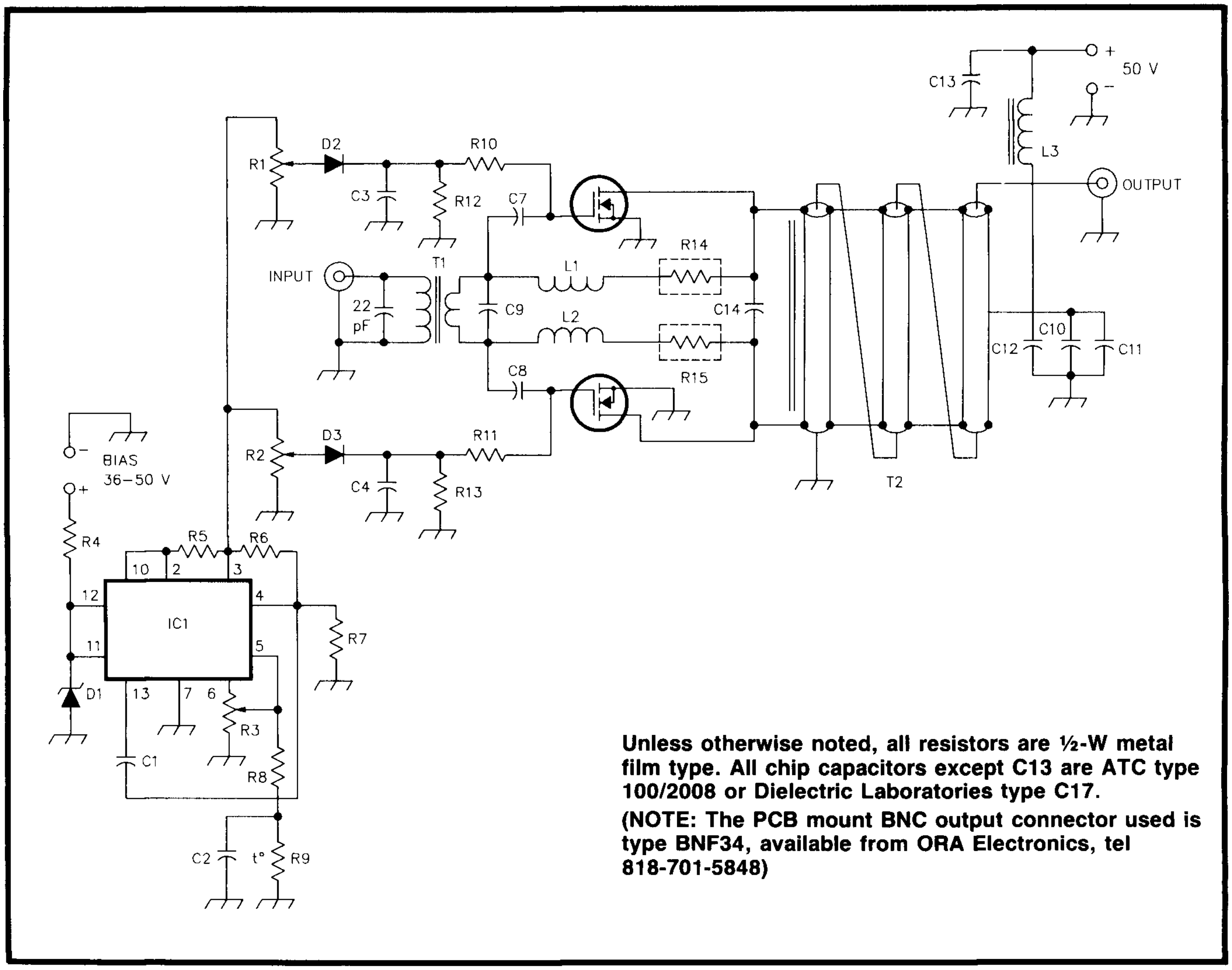
This article features a state of the art extremely compact design using a pair of FETs rated for 600 W of power output each. It would be capable of a power output of 1.2 kW as a push-pull circuit, but with the output matching employed, which is optimized at around 800 W, the unit starts saturating at around 1 kW at a 50-V dc supply, resulting in high IM distortion. Similarly at a 40-V supply, it would be usable up to 800 W. The type output matching transformer employed allows only integers as 1:4, 1:9. 1:16, etc. The 1:16 impedance ratio transformer would make the output matching optimized at 1400 W, which would result in a poor efficiency at 1200 W and lower power levels. The only way to compensate for this would be to adjust the supply voltage accordingly, in this case 45-46 V. However, the 1:16 ratio transformer of this type is physically much more difficult to fabricate than the lower ratio ones, and may not be available in the commercial market.

The Bias Regulator

The gate bias regulator (IC1 in Fig 1) allows the main supply voltage to be varied or the use of an unregulated supply while keeping the gate bias voltages and the FET idle currents constant. Since the maximum operating voltage of the regulator IC is only 40 V, a Zener diode (D1) is employed to keep it at a safe level. The regulator supply terminals are separated from the main power supply permitting the use of a separate bias supply if desired. There is also an option for a thermistor connection to stabilize the idle currents against temperature changes. The thermistor should be in a physical contact preferably with a mounting flange of one of the FETs. The gate voltages are individually adjustable (R1, R2) making gate threshold voltage matching of the devices unnecessary. In case of a device failure, such as a drain-gate short, D2 and D3 block the full supply voltage from being fed back to destroy the regulator. R10, R11 and C3, C4 are merely RC filters to protect the regulator from possibly strong RF fields. To set the idle currents, R1 and R2 must be adjusted to minimum. R3 is then adjusted for a regulator output voltage of about double the FET gate threshold voltages (IC1, pin 3). The current is monitored at the main supply voltage point while adjusting R1 for a desired idle current, typically 800 mA-1.0 A. R2 is then advanced until the current is doubled, resulting in equal idle currents for both devices. After this procedure, the settings of R1 through R3 should remain until one or both FETs must be replaced.

The RF Path

The amplifier is designed to operate into the industry standard 50-ohm input and output interface. The impedance matching to the low impedance levels of the



Unless otherwise noted, all resistors are ½-W metal film type. All chip capacitors except C13 are ATC type 100/2008 or Dielectric Laboratories type C17.
(NOTE: The PCB mount BNC output connector used is type BNF34, available from ORA Electronics, tel 818-701-5848)

Fig 1—Circuit Diagram—2 to 50 MHz Amplifier

2-50 MHz Amplifier Components List

R1,R2—1 kΩ single-turn Trimpots
R3—10 kΩ single-turn Trimpot
R4—470 Ω, 2 watts
R5—10 Ω
R6,R12,R13—2 kΩ
R7—10 kΩ
R8—Exact value depends on thermistor R9 used (typically 5-10 kΩ)
R9—Thermistor, Keystone RL1009-5820-97-D1 or equivalent
R10,R11—100 Ω, 1 W carbon
R14,R15—EMC Technology model 5308 or KDI Pyrofilm PPR 870-150-3 power resistors, 25 Ω
D1—1N5357A or equivalent
D2,D3—1N4148 or equivalent
IC1—MC1723 (723) voltage regulator
C1—1000-pF ceramic disc capacitor
C2,C3,C4—0.1-μF ceramic disc capacitor
C5—0.01-μF ceramic chip capacitor
C6,C12—0.1-μF ceramic chip capacitor
C7,C8—Two 2200-pF ceramic chip capacitors in parallel each
C9—820-pF ceramic chip capacitor
C10,C11—1000-pF ceramic chip capacitor
C13—0.47-μF ceramic chip capacitor or two smaller values in parallel
C14—Unencapsulated mica, 500 V. Two 1000-pF units is series, mounted under T2.
L1,L2—15 ηH, connecting wires to R14 and R15, 1.5 cm each #20 AWG
L3—10 μH, 10 turns #12 AWG enameled wire on Fair-Rite Products Corp ferrite toroid #5961000401 or equivalent
T1,T2—9:1 and 1:9 impedance ration RF transformers, types RF800-3 and RF2067-3 R, respectively (RF Power Systems, 3038 E Corrine Dr, Phoenix, AZ 85032)

FETs is accomplished with broadband RF transformers. Both the input transformer (T1) and the output transformer (T2) are of the so-called conventional type in contrast to transmission line transformers.[1,3,4] Both employ only one turn in the low impedance winding. T2 is far more critical than T1 because it determines the efficiency and the high frequency end gain characteristics, plus it must be able to handle a large amount of RF power. For increased bandwidth characteristics, its low impedance, one turn winding consists of three

[1]Notes appear on page 8.

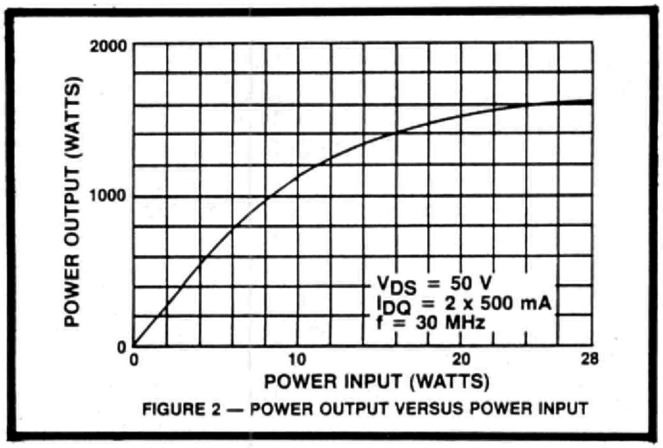


FIGURE 2 — POWER OUTPUT VERSUS POWER INPUT

Fig 2

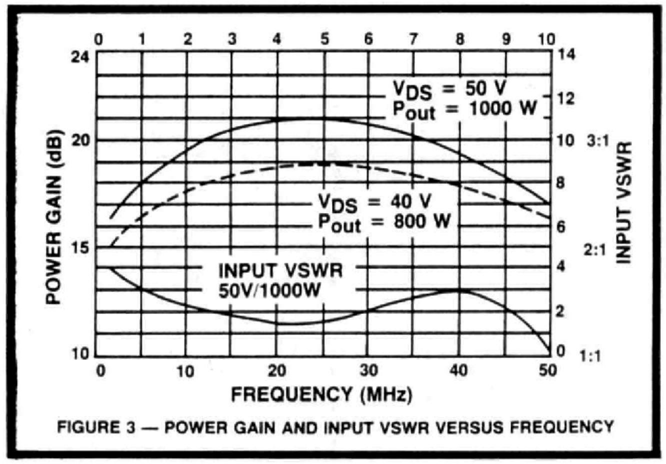


FIGURE 3 — POWER GAIN AND INPUT VSWR VERSUS FREQUENCY

Fig 3

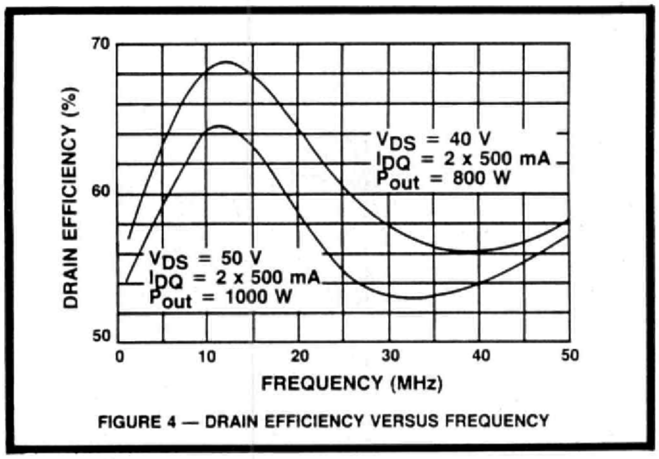


FIGURE 4 — DRAIN EFFICIENCY VERSUS FREQUENCY

Fig 4

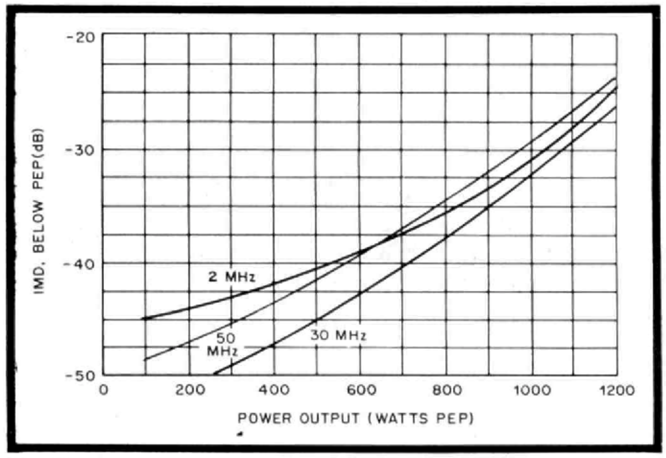
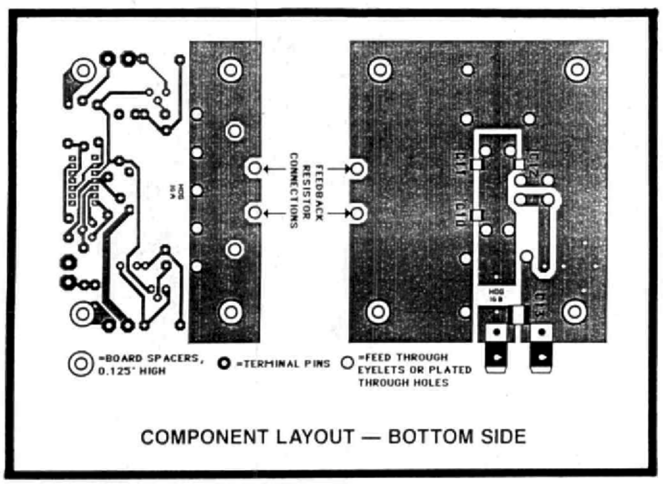
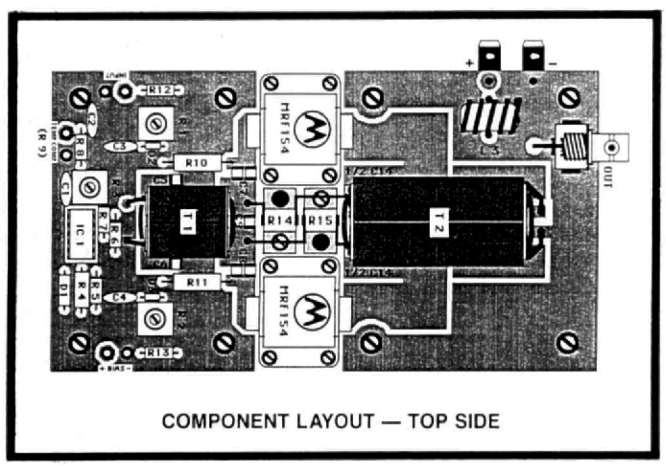


Fig 5



COMPONENT LAYOUT — BOTTOM SIDE

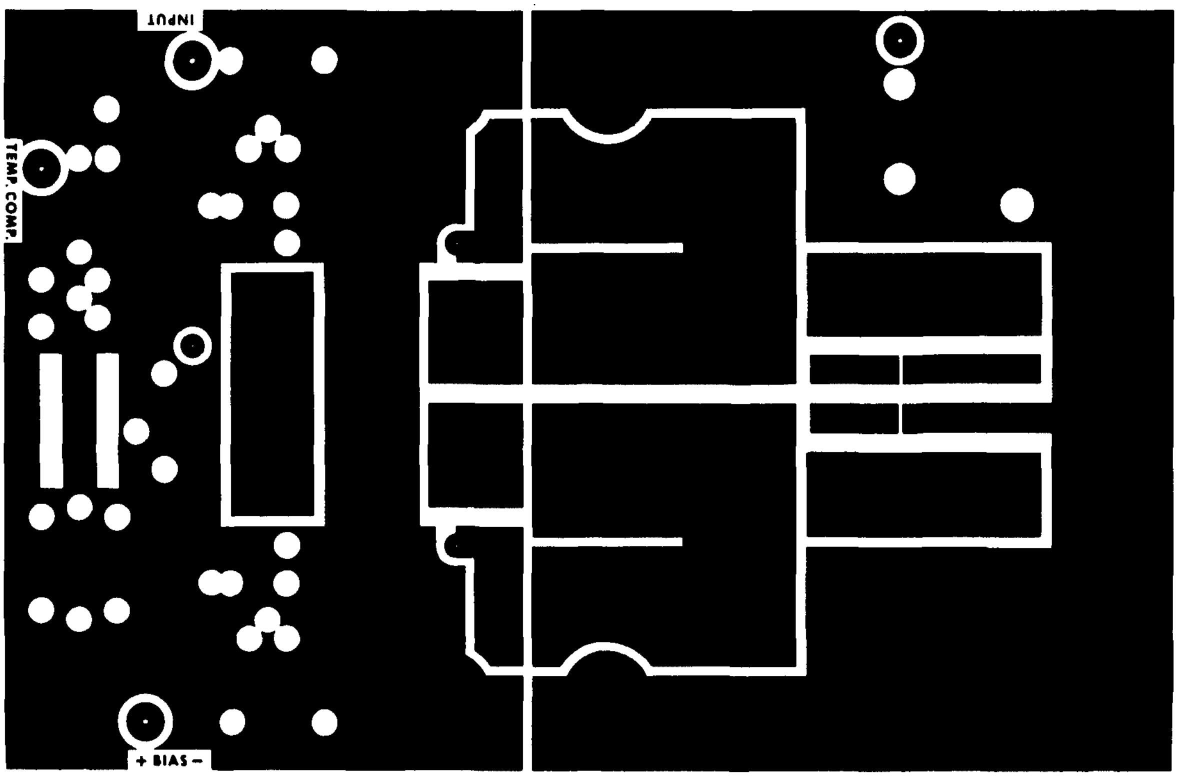
Fig 6



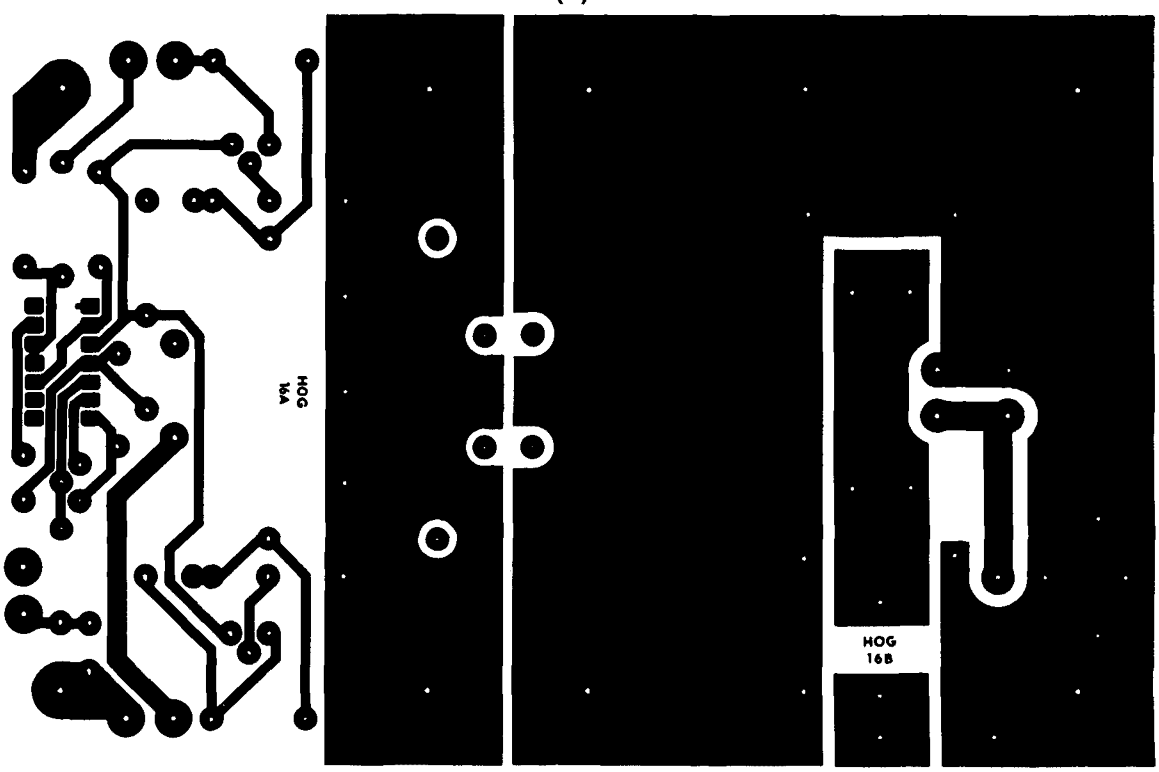
COMPONENT LAYOUT — TOP SIDE

Fig 7

paralleled 10-ohm coaxial cables, resulting in a tight and controllable coupling between the primary and secondary. According to formulas given in Reference 2, approximately twice the present 4.7 cm² ferrite cross sectional area would be required in order for the core not to saturate with the calculated 127 gauss flux density. The saturation mainly occurs at the lowest frequencies, in this case at 2-3 MHz. Unfortunately most ferrite manufacturers do not give information on saturation flux densities that applies to applications such as this. However, it is known that high permeability ferrites, in general, saturate easier than low permeability materials. Thus, the lowest permeability material should be selected that will satisfy the minimum inductive



(A)



(B)

1 INCH

Figs 8A and 8B—Circuit Board—Top Side and Bottom Side

reactance requirement at the lowest frequency of operation. The formula to calculate this is $NX_L = 2R_{S(L)}$, where: X_L = inductive reactance for one turn, N = number of turns, $R_{S(L)}$ = source or load impedance. Low permeability material is also less lossy at high frequencies, resulting in less heat generated in the transformer. T1, which must handle only 8-12 W of power, is made of higher permeability ferrite. This makes it possible to make the unit physically small as well. In T1, the secondary consists of metal tubes (see Ref 1), where three turns of the primary wire is threaded through. Metal tubes are also used in T2, but only to hold the structure mechanically together.

At high-power levels generated with solid-state devices, which operate at relatively low voltages, the impedance levels automatically become low. This creates a problem for finding passive components, especially capacitors to handle the high RF currents involved. In vacuum tube circuits a similar problem exists, but in the form of high voltages. In this design, C14 gets the roughest treatment. It must be able to carry RF currents in excess of 10 amperes at the higher frequencies, although the voltage across it is only 75 V rms. At first, several good quality ceramic chip capacitors were tried in parallel, but temperature excursions caused them to crack resulting in RF arcs that burned the circuit board in the area as well. Finally, two unencapsulated mica capacitors (brand names such as Unelco, Underwood, Standex, Elmenco and Semco) were soldered in series by attaching the terminal tabs together, making it a symmetrical structure. Since each is double the total value required and with double the number of plates, this increases the RF current carrying capability and provides a larger area to be soldered to the board metal foil to make the cooling more efficient. The low impedance winding terminals are then soldered to the tops of the capacitor metal casings, leaving the effective capacitance across the winding. For further fine tuning, an Arco (Elmenco) #469 or Sprague #GM-40900 compression mica trimmer can be soldered to the front-top terminals of the transformer. Slot openings in the metal foil (Fig 7) located on each side of the output transformer, next to the drain terminals, were provided to increase the series inductance for certain high-frequency narrow-band applications. This tunes out some of the FET output capacitance, resulting in increased efficiency. At lower frequencies (below 80 MHz) however, they only add to the IR loss and should be shorted. The location of C9 is also critical and should be placed approximately as shown in Fig 7. This will affect the input VSWR at frequencies above 30 MHz.

Bypass capacitors C10 through C12 must also be of good quality. The center tap of T2 should be free of RF if the circuit is balanced. This may not always be the case, in which case these capacitors will aid this function. L3 and C13 form an additional filter, ensuring that no RF energy is being fed back to the power supply. Switch-mode power supplies especially are sensitive against RF and may actually get damaged from it.

Negative feedback is provided through the networks L1-R14 and L2-R15. Its purpose is to produce a relatively flat power gain versus frequency response. It also improves the input return loss and helps to stabilize the amplifier at low frequencies, where the power gain would be 25-30 dB without it. The feedback is at its minimum at the high frequency end and at maximum at low frequencies, where most power is dissipated in R14 and R15. This power is roughly the difference in power input without the feedback between 2 and 50 MHz assuming a constant power output (in this case 25-30 W). A simple formula for calculating the feedback resistor values as well as their dissipation ratings is given in Reference 5. Reference 5 also includes information on physical construction of RF transformers such as used here.

Thermal Aspects

Assuming a 50% worst case efficiency for the unit, each FET dissipates 500 W of heat in an area of 1 × 1.5 inch. It is imperative that the transistors are mounted on the surface of a material with low thermal resistance such as copper. This is called a heat spreader as it is then attached to a heat sink made of material with poorer thermal resistance. It should extend about one inch beyond the edges of the FET mounting flanges at least on three sides. It is even more practical to make the heat spreader as large or larger than the amplifier itself. This would allow all circuit-board spacers to be an equal height of 0.125 inch. The thickness of the heat spreader should be a minimum of 0.375 inch. The heat spreader is then separately attached to the actual heat sink, which can be a 12-inch length of Wakefield Engineering type 4559 extrusion or equivalent[6]. Heat sink compound must be applied to all thermal interfaces and the recommended transistor mounting procedure should be followed, including the screw torque. Fig 9 shows the amplifier mounted to the heat spreader. Although the heat sink is not shown, one must be used for continuous operation and for test periods longer than a couple of minutes. For continuous operation, two 5-inch muffin fans under the heat sink will suffice. They will keep the device case temperature at below 80°C, and the die temperature, which equals to device thermal resistance × power dissipation + case temperature = 0.13 × 500 + 80, at less than 145°C, which is well below the 200-degree maximum recommended value. We must realize that the 500-watt dissipation is only valid when the unit is operated into a 50-ohm load. Under mismatched conditions, depending on the phase angle, the dissipated power may be lower or higher than this value.

Performance

Some of the amplifier performance characteristics are shown in Figs 2 through 5. Although at 30 MHz and above all harmonics are 25 dB or more below the fundamental, an output filter is required to comply with

FCC regulations. However, it can be a simpler one than required for the low frequencies, where the third harmonic may be only attenuated 12-15 dB. In push-pull amplifiers, the even harmonics are not usually a problem since they are attenuated by the balanced operation of the circuit. Information on high power low-pass filters for applications as this can be found in Reference 7. These filters are automatically relay switched with BCD code available in most modern transceivers.

References

The circuit boards and other components for this design are available from Communication Concepts, Inc, 508 Millstone Drive, Xenia, OH 45385, tel 513-429-3811/220-9677.

[1]Motorola, Inc, Semiconductor Sector Application Notes AN-749 and AN-1035.

[2]Hilbers, A.H., "Design of HF Wideband Power Transformers," Amperex (Philips) Application Laboratory Report ECO6907 and ECO7213.

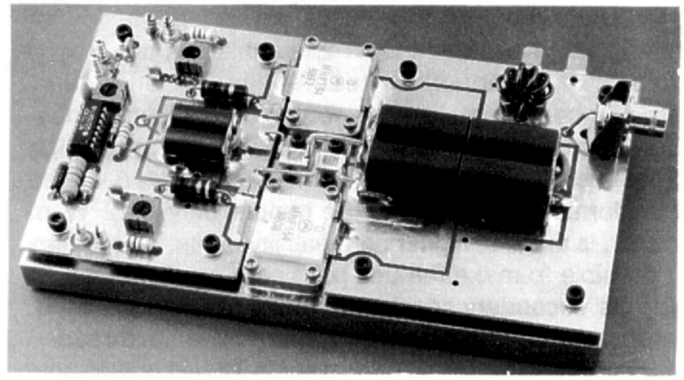


Fig 9—Amplifier Mounted to the Heat Spreader

[3]Blocksome, Roderick K., "Practical Wideband RF Power Transformers, Combiners and Splitters," *Proceedings of RF Expo*, February 1986.

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OSCAR Seminar

The Tandem Radio Amateur Club and Project OSCAR, Inc, invites everyone interested in Amateur Radio satellites (OSCAR) to attend the OSCAR Seminar which will be held on September 29 and 30, 1990. With so much information available and so many subjects to be covered, the entire weekend will be devoted to OSCAR! Speakers will cover all aspects of OSCAR from basic information necessary to get started, right up to advanced topics for the experienced OSCAR user. Computers will be available to demonstrate OSCAR software. Displays showing some of the equipment used by OSCAR users will also be available. Whether you are interested in RS satellites on 10 meters, worldwide DX available on OSCAR-13, telemetry sent from DOVE and other MICROSATs, or you just have many questions about OSCAR, this seminar is a must for you to attend.

Some of the subjects that will be covered include:

Understanding Keplarian Elements
How Operate MICROSATS and FO-20 Successfully
OSCAR—a Basic Tutorial about Using the Satellite
How to Use the S Satellites
Present Status and Future Plans for OSCAR
Preamplifers and How to use Them Properly
Successful Mode L Operating
An Open Forum Question and Answer Session
Successful Omni Antennas for MICROSATs
How Noise Figure Relates to Successful OSCAR Operation

The following topics may be added pending a speaker:

Understanding OSCAR Telemetry
Successful Mode S Operations
AMSAT-NA Forum

Preregistration is required so that adequate seating and supplies will be available. The funds raised will cover the minimal expenses to hold the seminar. All excess money raised will be allocated towards the building of future Amateur Radio satellites.

The seminar will be held right off highway 280 in Cupertino, CA. Motels, restaurants, and shopping are all close by. On Saturday evening, there will be time for informal discussions with those you want to learn more information from.

Registration fee for the seminar is $15.00 per person. This includes a copy of the papers published for the seminar. For complete registration forms and details of the seminar, send a business size, self-addressed, stamped envelope (SASE) to: **OSCAR Seminar, Project OSCAR, PO Box 1136, Los Altos, CA 94023-1136**. Here is the best opportunity to ask all the questions you ever had about any OSCAR!