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

Prepared By
Helge Granberg
Motorola Inc., Semiconductor Products Sector

Reprinted from QEX, July 1990 issue.
© 1990, ARRL Publishing.
All rights reserved.
A Compact 1-kW 2-50 MHz Solid-State Linear Amplifier

By H. O. Granberg, K7ES/OH2ZE
ARRL Technical Advisor
Member of the Technical Staff
Motorola, Inc
Phoenix, AZ 85062

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 water 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 \( I_C(\text{peak})/h_{FE} \) 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
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 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
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
\[ X_L = \frac{2\pi f L}{R_S(L)} \]
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. \( T_1 \), 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 \( T_1 \), 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 \( T_2 \), 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, \( C_{14} \) 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 shortened. The location of \( C_9 \) 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 \( C_{10} \) through \( C_{12} \) must also be of good quality. The center tap of \( T_2 \) 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. \( L_3 \) and \( C_{13} \) 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 \( L_1-R_{14} \) and \( L_2-R_{15} \). 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 \( R_{14} \) and \( R_{15} \). 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 x 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. 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 x 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.

Fig 9—Amplifier Mounted to the Heat Spreader

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.
1Motorola, Inc. Semiconductor Sector Application Notes AN-749 and AN-1035.
6Wakefield Engineering, 60 Audubon Road, Wakefield, MA 01880 and 7261 Mars Drive, Huntington Beach, CA 92647.
Motorola reserves the right to make changes without further notice to any products herein to improve reliability, function or design. Motorola does not assume any liability arising out of the application or use of any product or circuit described herein; neither does it convey any license under its patent rights nor the rights of others. Motorola products are not designed, intended, or authorized for use as components in systems intended for surgical implant into the body, or other applications intended to support or sustain life, or for any other application in which the failure of the Motorola product could create a situation where personal injury or death may occur. Should Buyer purchase or use Motorola products for any such unintended or unauthorized application, Buyer shall indemnify and hold Motorola and its officers, employees, subsidiaries, affiliates, and distributors harmless against all claims, costs, damages, and expenses, and reasonable attorney fees arising out of, directly or indirectly, any claim of personal injury or death associated with such unintended or unauthorized use, even if such claim alleges that Motorola was negligent regarding the design or manufacture of the part. Motorola and are registered trademarks of Motorola, Inc. Motorola, Inc. is an Equal Opportunity/Affirmative Action Employer.

**MOTOROLA**

Literature Distribution Centers:
USA: Motorola Literature Distribution; P.O. Box 20912; Phoenix, Arizona 85036.
EUROPE: Motorola Ltd.; European Literature Centre; 88 Tanners Drive, Blakelands, Milton Keynes, MK14 5BP, England.
JAPAN: Nippon Motorola Ltd.; 4-32-1, Nishi-Gotanda, Shinagawa-ku, Tokyo 141 Japan.
ASIA-PACIFIC: Motorola Semiconductors H.K. Ltd.; Silicon Harbour Center, No. 2 Dai King Street, Tai Po Industrial Estate, Tai Po, N.T., Hong Kong.
OUTPUT TRANSFORMER MOUNTING PROCEDURE FOR 2-50 MHz, 1kW AMPLIFIER USING 2xMRF154's