A 100-W MOSFET HF Amplifier

We had the power supply in Mar/Apr[1] and the diplexer filters in Jul/Aug,[2] here's the main event: a reliable FET power amplifier that needs only 10 dBm of drive to produce a pristine 100 W output.

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The two-stage amplifier described in this article and shown in Fig 1 is intended for SSB/CW/Data operation on all nine HF amateur bands. An input ($R_{in} \approx 50\ \Omega$) of about 10 mW (+10 dBm) is amplified to 100 W (PEP or average), continuous duty, with a gain of 40 ± 1.0 dB from 1.8 to 29.7 MHz. Third-order, two-tone intermodulation distortion (IMD) products are 35 to 40 dB below 100 W, and higher-order products are also within the high-quality range for amateur SSB equipment, as shown in Fig 2. The main goal for this amplifier is to operate as a driver, with low adjacent-channel interference, for a legal-limit 1500-W linear amplifier. At this power level,

[1]Notes appear on page 40.

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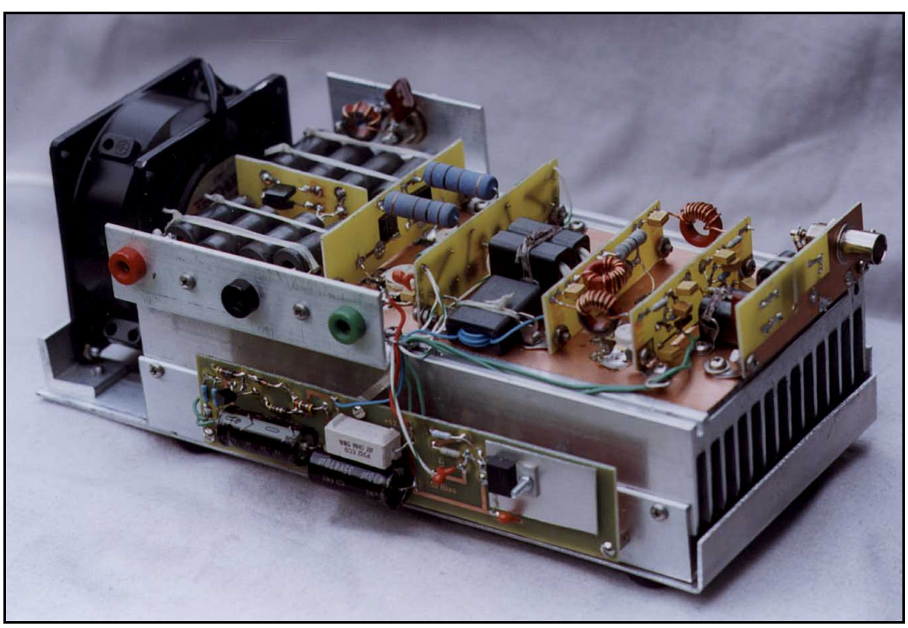


Fig 1—The 100-W broadband amplifier.

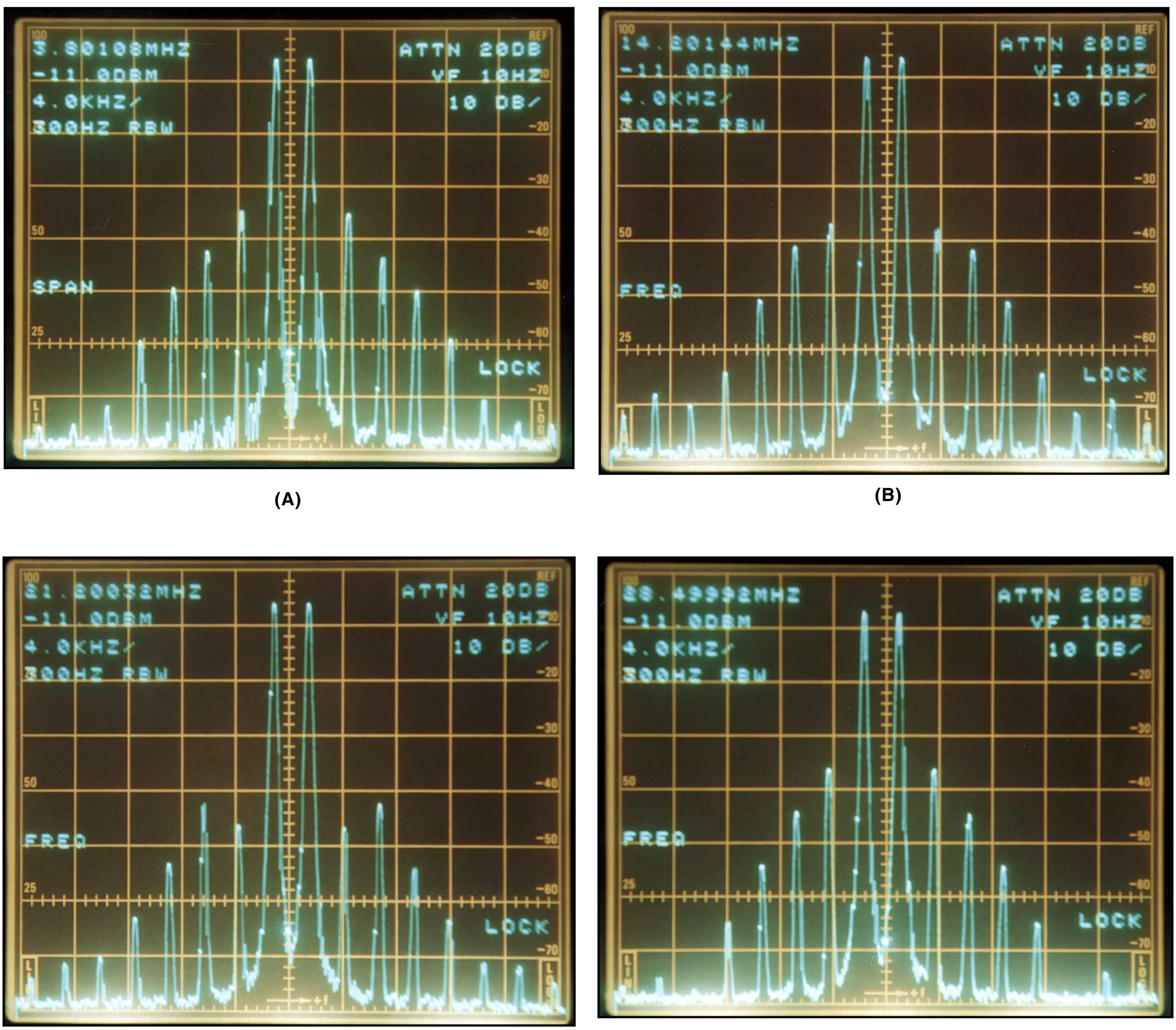
adjacent-channel reduction is especially important. And, of course, it is used in the "barefoot" mode as well.

The power supply (40 V at 8 A) for the push-pull, class-AB MRF150MP (matched-pair) MOSFET output stage was described previously (see Note 1). Six diplexer filters (see Note 2)—also known as invulnerable filters—provide more than adequate harmonic attenuation for all nine HF bands. They present a broadband load impedance to the MOSFETs that helps to assure freedom from regeneration and oscillation, and good IMD performance. A resistive load impedance between 45 Ω and 55 Ω is recommended for best performance.

The MRF150 was chosen because it is designed for linear, class-AB SSB operation and because it has high gain (g_m) at the 30-MHz end of the HF spectrum. A desirable feature of the MOSFET power transistor is its ability to achieve low values of the higher-order IMD products.[3,4] As mentioned previously, these products contribute to adjacent channel SSB interference. The first stage uses high-gain, class-A push-pull MRF426 (matched-pair) BJTs that require 13.5 V at about 1.0 A from a separate supply. This supply is also the main supply for other system components. Matched pairs of both transistors are available for a small extra fee from at least two sources.[5,6] They are both listed in the current Motorola manual.[7]

One main idea for this amplifier is to operate it in a very low-stress manner that helps assure a low probability of failure for a very long time, which offsets the initial cost of the high-quality transistors. The MRF150 is a 50-V transistor operated at 40 V; the MRF426 is a 28-V transistor operated at 13.5 V. The required input level is low enough that most of the amplification at the signal frequency occurs in one "gain block." Because of the good layout, circuit design and decoupling, the 40-dB gain value does not result in any stability problems.

The balanced amplifier greatly



(A) (B) (C) (D)

Fig 2—Two-tone IMD products: (A) 3.8 MHz, (B) 14.2 MHz, (C) 21.2 MHz, (D) 28.5 MHz.

reduces even-order harmonics—especially the second—prior to any output filtering, as shown in Fig 3 for a 7.0-MHz signal. It should be 40 dB or more below a 100-W CW signal for each amateur band. This reduction has been found reliable, once achieved. The low-level signal source that drives this amplifier must have at least 50 dB of second-harmonic attenuation, since this amplifier will not suppress that harmonic. This is easy to accomplish, but must be considered during the equipment system design (see Fig 9) and while bench-testing the amplifier as shown in Fig 4.

Circuit General Discussion

Fig 5 is the schematic of the two-stage amplifier. It utilizes 1:1 choke baluns and 1:4 (impedance) step-up and step-down transmission line (Guanella) transformers. The choke baluns significantly improve the balance of the input and output stages. Notice also that T2, T3 and T4 have "floating" center taps rather than bypasses to ground. This is recommended to improve the even-harmonic balance[8] (verified). T4 and T5 run slightly warm as compared to conventional transformers that get quite hot. The 5-W feedback resistors get warm, but their large surface area limits their temperature rise. They also receive cooling air from the fan. All power resistors in the RF circuits are the excellent metal-oxide types (see Note 5) that are quite stable with age and temperature, and have very low reactance at 30 MHz (measured). Metal-film 1% resistors are used in several critical locations.

The first stage has resistance and inductance loading from collector to collector. A powerful free-running oscillation in the first stage at about 32.5 MHz was being triggered occasionally at the moment of dc supply turn-on while a fairly large 28.0 to 29.7 MHz input was present. The loading reduces stage gain and suppresses the parasitic collector circuit resonance that produced the oscillation. It also helps to flatten the amplifier frequency response.

The second stage has a very low value of RF resistance, 15 Ω, from each gate to ground that helps to assure stability. Because these resistors are in parallel with the large input capacitance of the MOSFETs, they also help to flatten the frequency response. Both stages have negative feedback networks that further assure stability and flatness of frequency response.

Two biasing networks are employed. The LM317 provides a highly regulated gate voltage for the second stage. The

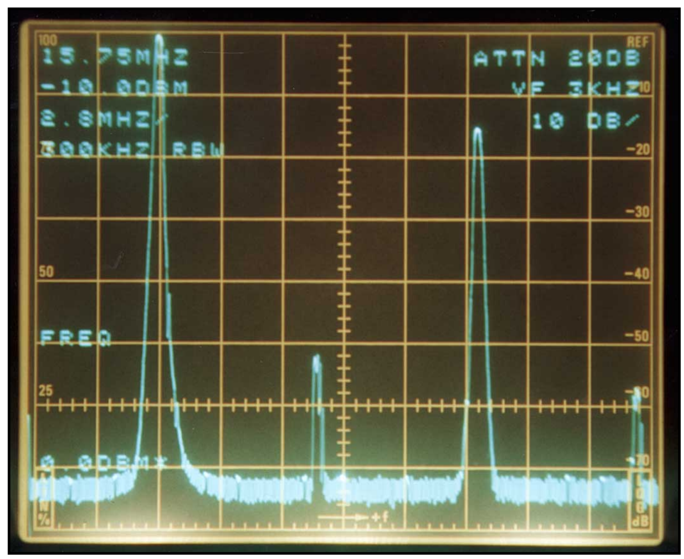


Fig 3—Wideband spectrum for a 7-MHz signal.

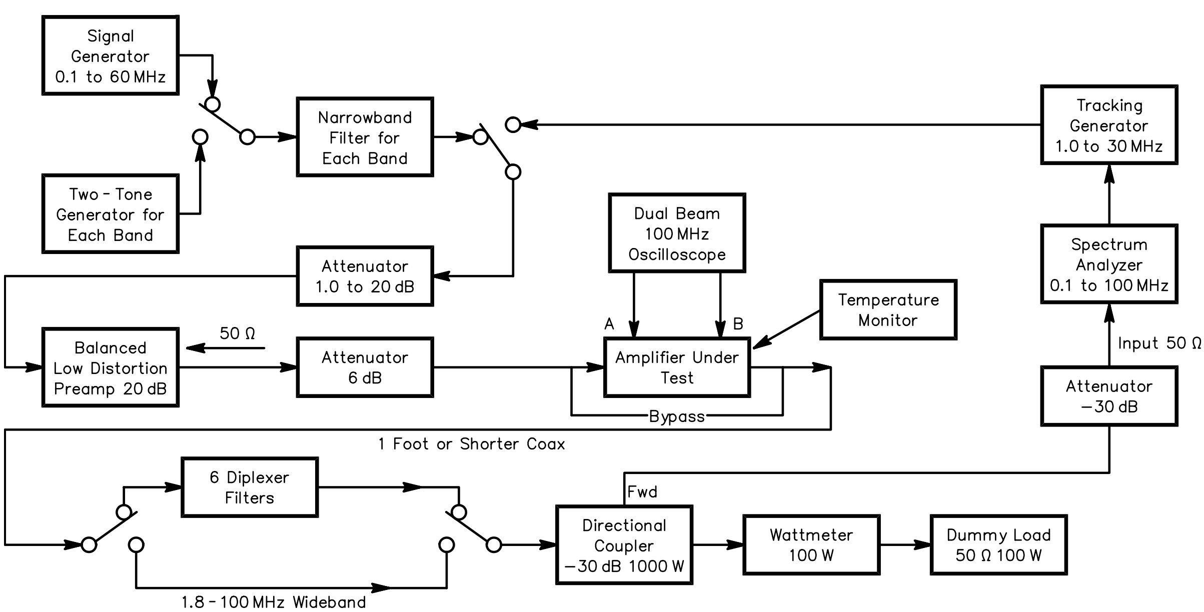


Fig 4—Block diagram for equipment setup, design and testing of the MOSFET amplifier.

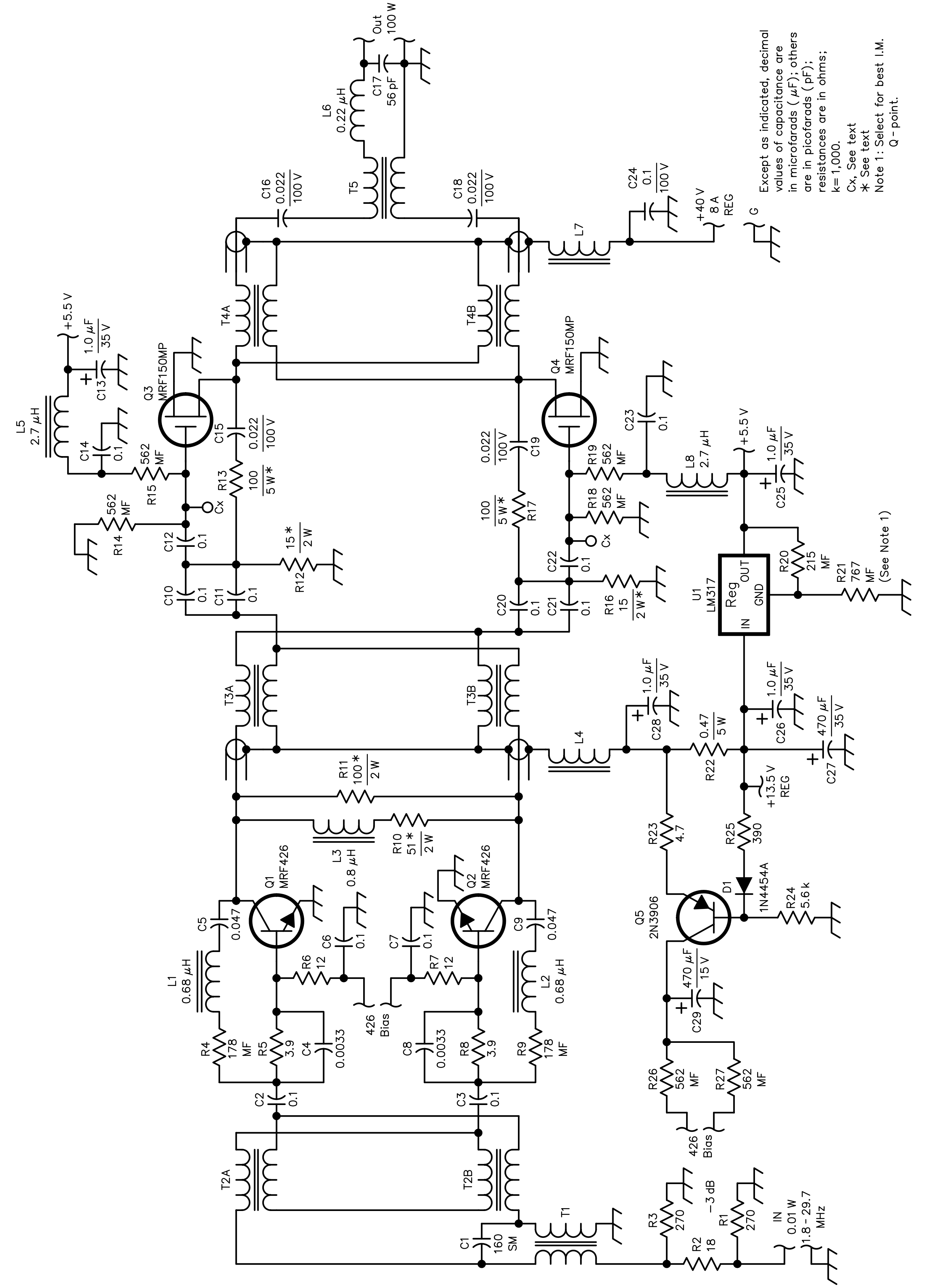

Out
100 W
L6
0.22 µH
C17
56 pF
C16
0.022
100 V
T5
C18
0.022
100 V
C24
0.1
100 V
+40 V
8 A
REG
G
L7
T4A
T4B
+5.5 V
1.0 µF
35 V
C13
L5
2.7 µH
C14
0.1
Q3
MRF150MP
Q4
MRF150MP
C15
0.022
100 V
C19
0.022
100 V
C23
0.1
R15
562
MF
R19
562
MF
L8
2.7 µH
+5.5 V
1.0 µF
35 V
C25
R13
100
5 W*
R17
100
5 W*
Cx
R18
562
MF
Cx
R14
562
MF
C12
0.1
R12
15*
2 W
C22
0.1
R16
15
2 W*
C10
0.1
C11
0.1
C20
0.1
C21
0.1
U1
LM317
Reg
IN
OUT
GND
R20
215
MF
R21
767
MF
(See Note 1)
T3A
T3B
1.0 µF
35 V
C28
1.0 µF
35 V
C26
L4
R22
0.47
5 W
470 µF
35 V
C27
R11
100*
2 W
+13.5 V
REG
L3
0.8 µH
R10
51*
2 W
R23
4.7
R25
390
Q1
MRF426
Q2
MRF426
D1
1N4454A
Q5
2N3906
R24
5.6 k
C5
0.047
C9
0.047
C6
0.1
C7
0.1
L1
0.68 µH
R6
12
R7
12
L2
0.68 µH
470 µF
15 V
C29
426
Bias
R4
178
MF
R5
3.9
C4
0.0033
C8
0.0033
R8
3.9
R9
178
MF
R26
562
MF
R27
562
MF
C2
0.1
C3
0.1
426
Bias
T2A
T2B
T1
C1
160
SM
R3
270
–3 dB
R2
18
R1
270
IN
0.01 W
1.8 - 29.7
MHz
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.
Cx, See text
∗ See text
Note 1: Select for best I.M. Q-point.

Fig 5

FET dissipation and linearity are very sensitive to this voltage; the bias voltage for SSB is fine-tuned for best signal purity using two-tone tests and a spectrum analyzer. Note the four 562-Ω resistors. If the LM317 fails short circuit, the voltage on the FET gates does not exceed 6.8 V, which will not damage the gates. The large drain current that results from this failure also does no damage because the 40-V power supply has current limiting and voltage fold-back that prevent harmful FET dissipation. The FETs are thus kept well within the safe-operating-area (SOAR) as defined in data sheets[9] (verified). It is important that each 562-Ω resistor from gate to ground be permanently attached directly between the gate and source tabs of the FETs themselves so that the gates are never floating. Wrap the resistor leads around the tabs so that they cannot come loose. This avoids accidental static charges that might ruin them. On the other hand, it is better to "tack" and not wrap the base-collector and gate-drain signal leads so that they can be easily disconnected. We want to be able to test individual segments of the circuitry easily. In my experience, the MRF150 has proven to be a rugged transistor, much more so when operated conservatively and with the power-supply safeguards that I mentioned.

The LM317 provides about 5.8 V for the particular pair of FETs that I used. Individual pairs of FETs will probably require an adjustment of this voltage for idling current and for IMD products that resemble Fig 2. The adjustment procedure is described later. This bias value is set to emphasize the reduction of the higher-order products rather than third- and fifth-order products. These lower-order products do not contribute as much to adjacent channel interference; in fact, an SSB speech processor will make them worse anyway. The higher-order products need to be reduced and the MRF150 has this capability.

The 2N3906 PNP transistor is a bias-current source. The value of this current is determined by the 4.7-Ω resistor and the base-to-ground voltage of the 2N3906. The two 562-Ω resistors force equal base currents into the MRF426s. This helps to assure equal performance. The 0.47-Ω resistor is part of a negative-feedback bias loop. If collector current increases, the voltage across this resistor increases and this reduces the base current. Thermal runaway is avoided by this strategy because the reduced base bias restricts the current increase to a small value. Because of the low dissipation of the MRF426s and the good heat sink, this method is very effective. These transistors are also well inside their SOAR requirements. This stage is very linear and contributes almost nothing to the overall IMD products. To verify this, it is necessary to turn off the drain voltage to the MRF150s and connect a spectrum analyzer across one of the 15-Ω resistors.

The diplexer filter method was decided upon—after a lot of experimentation with other low-pass filter methods—as a solution that is free of problems caused by complex interactions between the filters and the output transistors. A special peculiar problem is discussed later. The MRF150 has high gain into the VHF region; the diplexers eliminated all problems associated with this fact. This approach is recommended as a simple way to assure correct operation for HF amateur-band operation of these high-frequency MOSFETs. I was able to get "good-enough" operation with the more conventional low-pass filters, but this approach was by far the most satisfactory for an amateur-band amplifier, as confirmed by swept-frequency tests at all power levels into a 50-Ω load. More about complex loads later.

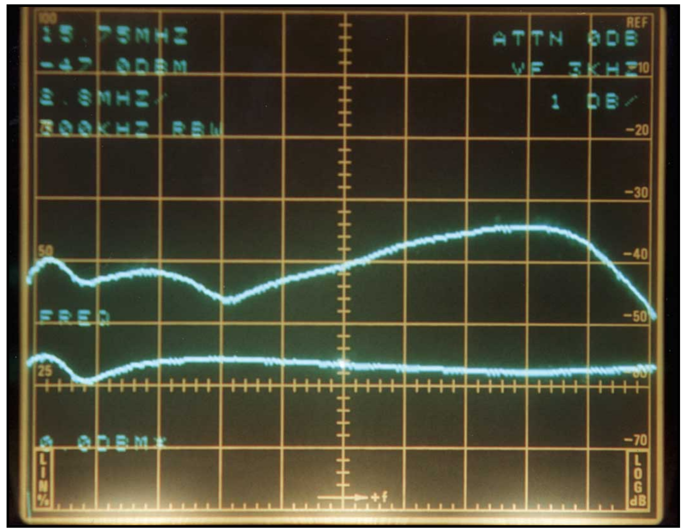


Fig 6—Gain variation, 1.8 MHz to 29.7 MHz, 1.0 dB per division vertical scale—The lower trace is a reference sweep that bypasses the 100-W amplifier—The response bump at low frequency is an artifact of the tracking generator.

Frequency Compensation

The bipolar transistors have gain values that decrease as frequency increases. The MOSFETS have large capacitances that also affect frequency-response roll-off. It was a major exercise to design networks that flatten the response from 1.8 to 29.7 MHz. Fig 5 shows the approach. The idea is to compensate smoothly from input to output in such a way that neither of the two stages is over-driven at any frequency. The criterion for this is to check IMD products and harmonics during the design process, which involves approximate analysis (see the "MOSFET Stage Simulation" sidebar) and negative feedback.[10] Beyond about 32 MHz, the gain falls off fairly rapidly (but not too rapidly). This is also desirable.

When testing the frequency response using the test setup of Fig 4, it is necessary to measure the frequency response of the signal path from tracking generator to spectrum analyzer while bypassing the 100-W amplifier. This reference response is then compared with the response with the amplifier in the path, as shown in Fig 6. Note the vertical scale: 1.0 dB per division. For the most credible results and ease of measurement, it is very desirable that the impedance looking back from the input be 50 Ω.

Fig 5—(left) Schematic diagram of a two-stage, 100-W amplifier. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. Resistors marked with an asterisk are metal-oxide units. MF indicates metal-film resistors. Both the MRF426 and MRF150 devices are matched pairs. All capacitors are 50 V, unless otherwise indicated.

Signal Level Testing

We want to verify that the first stage is operating normally by measuring its RF voltages at 4.0 MHz. The class-A first-stage input impedance is close to 50 Ω, and the input level is +10 dBm. Temporarily disconnect the signal leads to the gates of the FETs. The voltage at each output of T3 is about 1.35 V. Reconnect the gate signal leads. The final output over the entire range into an unfiltered, wide-band 50-Ω load is then observed using the setup in Fig 4. A spectrum analyzer with tracking generator is very valuable for this. The accuracy of the wattmeter should be verified or calibrated by some means at both the 100-W and the 25-W levels (the power of each tone of a two-tone, 100-W PEP signal) in each amateur band.

If the second stage is working correctly, the output power is 100 W, or 71 V RMS across 50 Ω. These procedures assure that both stages are working properly and amplifying as intended. Because of the variations in the fabrication of the transistors, the total gain can vary a decibel up or down despite the use of negative feedback in each stage. I suggest the 3-dB attenuator at the input not be modified for simplicity reasons.

The drain-to-drain load impedance of the class-AB output stage is 12.5 Ω and the CW output power is 100 W, so the drain-to-drain ac voltage is 35.4 V. Fig 7 shows the dual-trace scope waveforms (chop mode) on each drain, superimposed on the 40-V supply, and also the drain-to-drain waveform that is confined to the linear region. The third-harmonic content is visible. It is interesting to note that although these waveforms show considerable non-linearity, the fundamental component is quite linear with respect to the gate signal level. I also found that a 50-V dc supply created more heat, but at 100 W, did not improve linearity enough to make it worthwhile. The 35.4 V ac also appears across the two series-connected 100-Ω feedback resistors; they dissipate about $(35.4^2)/200 = 6.3$ W, or 3.2 W per resistor (64% of the 5 W rating).

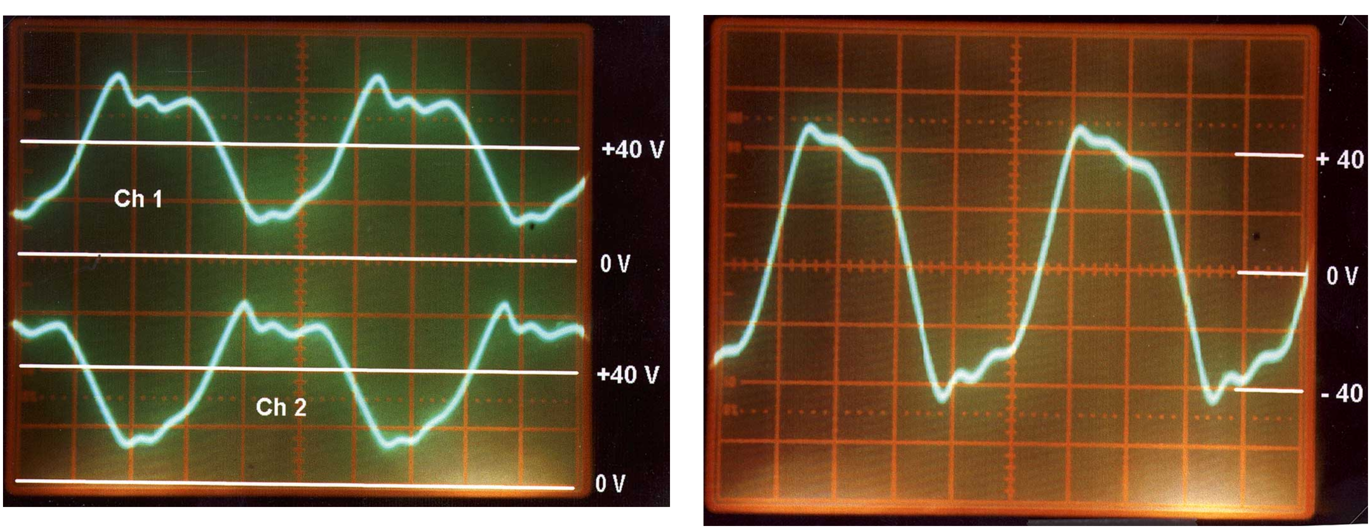
A broadband, untuned power amplifier with flat frequency response and low distortion is not, by necessity, especially energy-efficient (ratio of RF output power to dc power). The output stage is less than 40% efficient for this reason. For SSB use, where the average power is not more that 25 W, even with speech processing, this is no problem at all. For continuous key-down at 100 W, the cooling fan is more than adequate.

Here is an important caution about using oscilloscope probes at the FET drains and the output connector: A 10:1 probe could be damaged (it happened to me) if used directly at this RF voltage level, especially at the upper end of the HF range. Use instead the homemade probe described previously (see Note 2) with a 50-Ω terminating resistor.

The quality of balance is checked by looking at the second harmonic on a spectrum analyzer using the setup in Fig 4 with the diplexer filters out of the signal path. At 100 W output, check the harmonics in each of the nine amateur bands. Capacitor C_x in Fig 5 is used (if needed) to improve the second-harmonic phase balance on the higher amateur bands. If the value is not well below –40 dBc, use C_x at one gate or the other to achieve –45 dBc. Extra pads are provided on the PC board for this capacitor. A value of between 22 pF and 56 pF should be adequate. Overkill is neither necessary nor desirable: The output filters will do the rest. For each decibel of power output below 100 W, the second harmonic "normally" drops about two decibels.

Construction Notes

Fig 1 shows how my version of the amplifier is constructed. A 0.125-inch (or 0.062-inch) double-clad PC board, 3.25×8.0 inches, is firmly attached to the heat sink. I suggest using this compact size for best reproducibility. The heat sink shown may not be presently available from RF Parts (see Note 5), but I also purchased a Model 99 sink from CCI (see Note 6) that is 6.5×12 inches. This can be easily tailored to the appropriate size with a band saw that has a metal-cutting blade. Pieces of angle and sheet aluminum can then be creatively fashioned to accommodate the fan (RadioShack #273-242) and the PC board that contains the bias circuitry. The transistors are bolted directly to the heat sink (through cutouts in the PC board) using heat-sink compound and tapped (and carefully deburred) #4-40 holes. Careful mounting of the transistors is essential. The main PC board has small sections of copper removed underneath the base and collector tabs of the MRF426s and underneath the gate and drain tabs of the MRF150s, so that accidental grounding is avoided. Use a hobby



(A) (B)

Fig 7—(A) drain and (B) drain-to-drain oscilloscope waveforms.

knife to define the areas and a hot soldering iron to peel off the copper.

A bottom cover is advised, as shown, so that air is funneled through the heat-sink fins efficiently. For a 100-W, continuous, single-tone output, the exhaust air temperature reaches 50°C. I used a RadioShack #22-174 multimeter with its temperature probe mounted inside the fins to monitor temperature during the design and testing phase.

The components are mounted on a set of seven small PC boards, six of which are mounted vertically (as shown) and bolted to the drilled-and-tapped heat sink through the PC board, using #4-40 screws and small angle brackets. I used stiff, right angle #6 solder lugs that worked out very nicely. The seven PC boards are cut from a single 4×6-inch two-sided PC board, shown in Fig 8. The bias circuit board has a ground plane; the others do not. This set of boards is available from FAR Circuits.[11]

I chose this method because it is easy, makes the amplifier more compact, reduces stray L and C that can degrade the wide-band frequency response and reduces stray couplings that can impair stability and harmonic balance. It also allows the ground plane to be one continuous surface, which is a plus factor. This approach worked out very well and I recommend it as a simple approach.

Temperature Rise

The cooling fan is important. This amplifier, as designed, should have the fan running, and I have found that a *simple* and reliable way to keep everything safe. It has been tested at 100 W continuously, for several hours. The MOSFETs—well separated on a good heat sink and with efficient airflow—do not have a temperature-controlled gate-bias arrangement because I did not find it necessary at this power level. Because of the "spread" in threshold voltage of individual matched pairs, a one-time adjustment of gate bias is needed. The following simple procedure is used:

- Replace R21 with a resistor decade box set at 750 Ω.
- Place a 0-10 A meter in the +40-V line.
- With *no* signal input, switch on the amplifier and let the current reach its final value, which should reach about 1.5 A as the FETs warm up.
- Adjust the resistor value until I_{DD} is 1.5 A. After each adjustment, allow time for the FETs to reach a steady current value. This final value is not critical, but should be within 10%.
- Check the two-tone IMD at 100 W PEP (50 W average) and make further small changes, if needed, to resemble the IMD patterns in Fig 2.

In this amplifier, we are able to use the self-limiting feature of the MOSFET. This approach would not be appropriate in higher-power, higher-temperature amplifiers because of thermal runaway possibilities, where the decrease with temperature rise of the gate threshold voltage exceeds the decrease of the dc transconductance.[12, 13] If fan reliability is a concern, add a thermal cutout switch to the +40-V line.

Load Impedance

Swept- and fixed-frequency tests at many signal levels and voltage values—using various values of parallel conductance and capacitive/inductive susceptance—did not reveal instability problems at SWR values less than 2:1. A tendency to oscillate in a peculiar manner occurred in the 20.0 to 29.7 MHz range, but only with input signal applied (a "driven" oscillation)

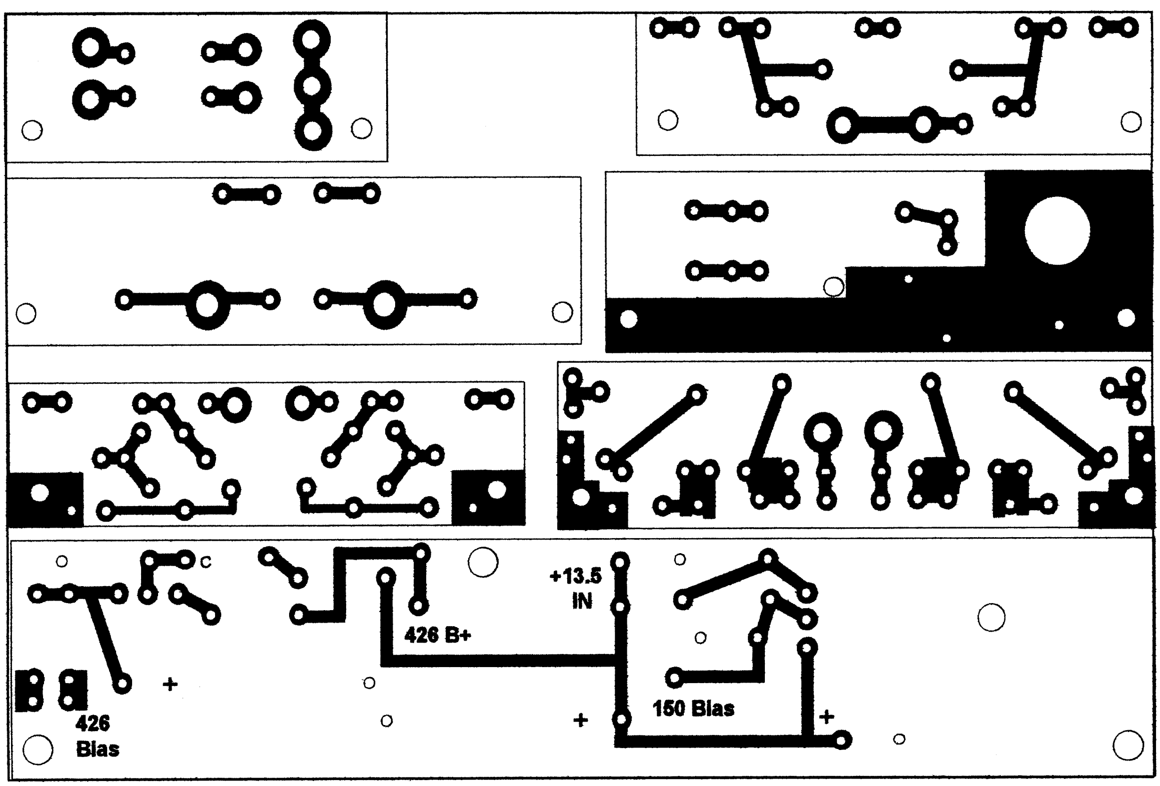


Fig 8—Etching pattern for a 4×6-inch (finished size) PC board that provides the seven individual boards.

and certain values of drive level, supply voltages less than 30 V, SWR of 3:1 or more, and a certain range of shunt capacitance at the output of the amplifier itself or at the output of the 12/10 meter diplexer. The instability shows up as a "Christmas tree" pattern in which discrete, uniformly-spaced 2- to 4-MHz sidebands appear on the main carrier. It is due to a parametric effect[14] involving the voltage-variable capacitances of the MOSFET transistors that is somehow enhanced by the shunt test capacitance. Shunt inductance did not produce the problem. The amplitude modulation of the carrier by the low-frequency oscillation is clearly visible on an oscilloscope. The low-frequency oscillation itself can be seen on a spectrum analyzer. At SWR values less than 2:1 and drain voltages greater than 35 V (40 V is recommended), the safety margin is large. The 0.22-µH, 56-pF network at the output greatly assisted with this problem. A coax cable from the PA to the diplexer assembly that is a foot or less duplicates my test setup. Under these conditions, there is no stability problem on any band caused by load-impedance values that my testing could identify.

Actually, a resistive load between 45 Ω and 55 Ω is recommended for best SSB linearity, as mentioned previously. While driving a high-power linear amplifier to its rated two-tone SSB output, adjust its input impedance to 50 Ω resistive as closely as possible on each band, using a 50-Ω directional wattmeter. A broadband (untuned) solid-state driver amplifier such as this one requires such attention to load value for best results as compared to pi-network vacuum-tube PAs that can transform a fairly wide range of complex load im-

MOSFET Stage Simulation

A simplified analysis of the second MOSFET stage is presented (Fig A) to illustrate the effect of the resistive negative feedback of the 100-Ω resistors and the 40-pF gate-to-drain capacitance. The simulation diagram shows the voltage-controlled current sources with a G_m of 6.0 S, which is assumed constant over frequency. The gate-to-source capacitance is 360 pF and the drain-to-source capacitance is 200 pF. These numbers are from the MRF150 data sheets.

The frequency plot (Fig B) obtained from the *ARRL Radio Designer* program shows the gain, MS_{21} (dB), with (lower trace) and without (middle trace) the 100-Ω feedback resistors. The gain variation is reduced from about 6 dB to about 1.5 dB. In this simplified model (good enough for this illustration), the gain drop is caused by the capacitors, especially the 40-pF capacitor whose influence is greatly magnified by the Miller effect. If this 40-pF capacitance is eliminated from the simulation, the frequency response variation—even without the feedback resistors—is less than 1.0 dB, as seen on the upper trace.

If the power output at 15 MHz is to be held constant, the turns ratio of the output transformer and the value of the feedback resistors can be manipulated to achieve that end. The simulation shows that a turns ratio of 1: $\sqrt{2}$ and a resistor value of 120 Ω does this. The FET load impedance is now 25 Ω instead of 12.5 Ω. The increased voltage gain makes the Miller effect greater and increases the gain variation (lower trace) to 2.3 dB. The drain-to-drain voltage is now 50 V instead of 35 V, and this results in higher drain efficiency for the MOSFETs. However, the push-pull 1: $\sqrt{2}$ -turns-ratio transmission-line transformer is not as simple as the 1:2 that uses 25-Ω coax, and for me, that was the determining factor.

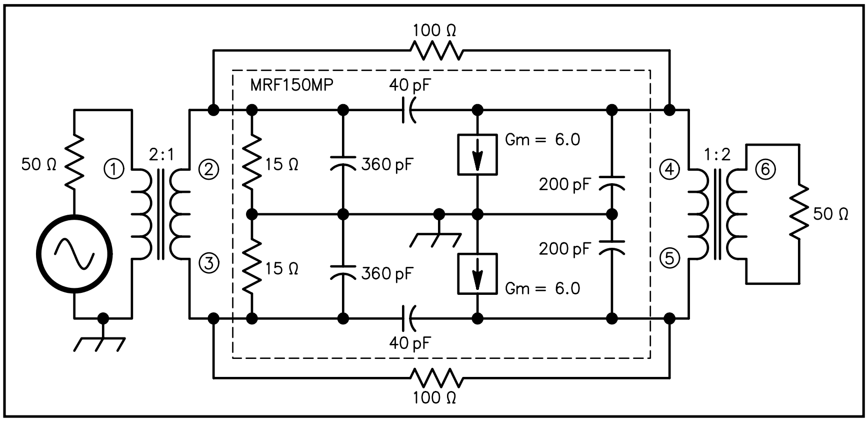


Fig A—A circuit used to analyze the PA output stage with ARD.

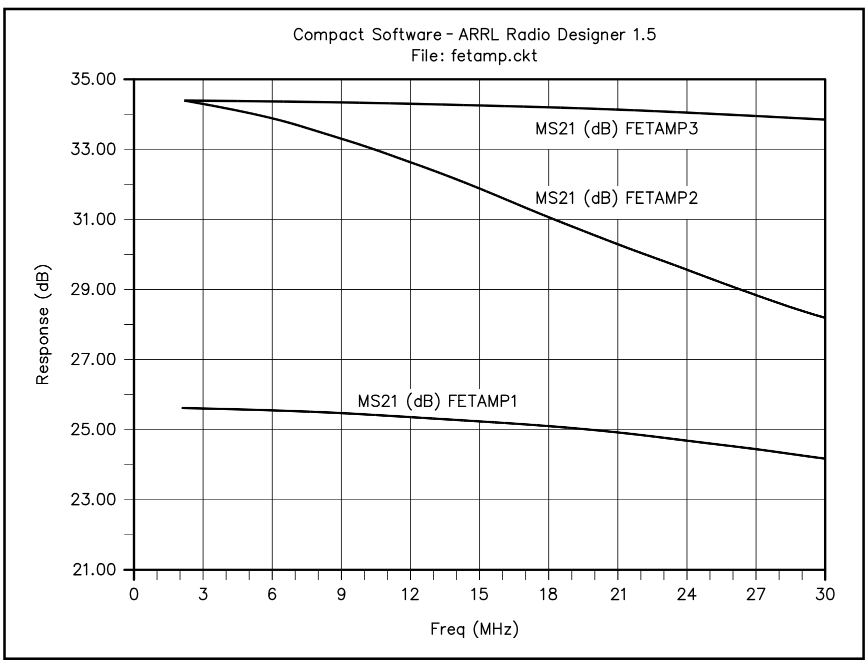


Fig B—The analysis results.

pedance values to the correct tube plate load resistance at resonance.

It is difficult to achieve stability in a broadband, transistor power amplifier. Having done so, it is still wise to operate into the correct 50-Ω load as closely as possible. The monitoring of forward and reflected power in Fig 9 is helpful for this and assures clean, stable and reliable operation. This circuitry will cut back the drive level to a value that protects the output stage from excessive drain current or gate drive. Incidentally, the first stage goes into saturation far below the level that would damage the MOSFETs.

System Design

Fig 9 suggests a system implementation of the amplifier. For the flattest frequency response of the complete system from 1.8 to 29.7 MHz, the circuitry that drives this amplifier should also have a flat response and a 50-Ω output resistance—both easy to achieve. If it is not perfectly flat or not exactly 50 Ω, it may be necessary to slightly adjust the drive level on each band, which is not difficult. As mentioned before, the second-harmonic production of the circuitry preceding this amplifier must be of sufficiently low level (–50 dBc) that balanced, push-pull operation is necessary. Narrow-band resonator filters are also very commonly used for amateur-band harmonic reduction in low-level exciter circuitry.[15, 16]

The directional coupler detects excessive forward and reflected power. Both of these are displayed on a panel meter. The directional coupler's forward port is also a source of ALC

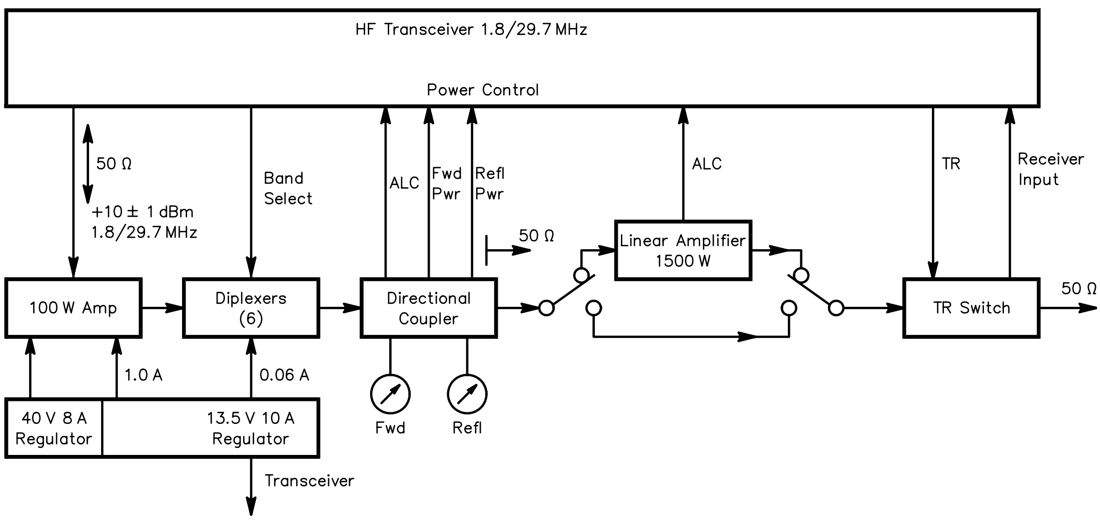


Fig 9—System implementation of the 100-W amplifier.

Parts List

C1—160 pF SM
C2, 3, 6, 7, 10-12, 14, 20-23—0.1 µF, 50-V CK05
C4, 8—0.0033 µF, 50-V CK05
C5, 9—0.047 µF, 50-V CK05
C13, 25, 26, 28—1.0 µF, 35-V tantalum
C15, 16, 18, 19—0.022 µF, 100-V CK05
C17—56 pF SM
C24—0.1 µF, 100-V CK05
C27—470 µF, 35-V aluminum
C29—470 µF, 15-V aluminum
D1—1N4454A or equivalent
L1, 2—0.68 µH, T50-2 core, 10 turns #26 AWG
L3—0.80 µH, T50-2 core, 12 turns #26 AWG
L4—BN-43-3312, 4 1/2 turns #22 hookup wire
L5, 8—2.7 µH molded
L6—0.22 µH, T50-2 core, 5 turns #26
L7—2 FB-43-5621 cores, 1 1/2 turns #12 stranded
Q1,2—MRF426 matched pair
Q3,4—MRF150 matched pair
Q5—2N3906 PNP
R1, 3—270 Ω, 1/4 W, 5% tolerance
R2—18 Ω, 1/4 W, 5% tolerance
R4, 9—178 Ω, 1/8 W metal film, 1% tolerance
R5, 8—3.9 Ω, 1/4 W, 5% tolerance
R6, 7—12 Ω, 1/4 W, 5% tolerance
R10—51Ω, 2 W metal oxide
R11—100 Ω, 2 W metal oxide
R12, 16—51 Ω, 2 W metal oxide
R13, 17—100 Ω, 5 W metal oxide
R14, 15, 18, 19, 26, 27—562 Ω, 1/8 W metal film, 1% tolerance
R20—215 Ω, 1/8 W metal film 1% tolerance
R21—767 Ω, 1/8 W metal film 1% tolerance, test selected
R22—0.47 Ω, 5W wire-wound
R23—4.7 Ω, 1/4 W 5% tolerance
R24—5.6 kΩ, 1/4 W, 5%
R25—390 Ω, 1/4 W, 5%
T1—BN-43-202, 2 1/2 turns #32 AWG, bifilar
T2A, B—BN-43-202, 2 1/2 turns #32 AWG, bifilar
T3A, B—BN-43-3312, 2 1/2 turns 25-Ω miniature coax*
T4A, B—(2) FB-43-5621, 1 1/2 turns 25-Ω miniature coax*
T5—(2) FB-43-5621, 2 1/2 turns 50-Ω miniature coax
U1—LM317 adjustable voltage regulator

Notes
*Microdot D260-4118-0000 available from Communication Concepts, Inc.
All cores are available from Amidon.
Closely matched transistor pairs are from RF Parts (see Note 5).

voltage. These control voltages are fed back to the appropriate gain-controlled stages in the exciter. Gain control of the transistors in the 100-W amplifier is not recommended because changes in base or gate bias will degrade amplifier linearity, IMD products and possibly the flat frequency response. There are better ways to accomplish the gain-control task, such as preferably in a low-level IF amplifier. With respect to IMD, we are trying to control the odd-order curvature of the MOSFET transfer characteristic—especially for the higher-order products—by setting the gate-bias point. The bias value is found by looking at IMD on each amateur band and selecting the best compromise value. This was discussed previously in this article.

The diplexer filters are linked to the other band-switch circuitry so that the correct filter is always switched in. The TR relay is at the output of the diplexers. Do not "hot-switch" the diplexer filters because this might damage the inexpensive relays.

Conclusion

In conjunction with the power supply (see Note 1) and the diplexer filters (see Note 2), the amplifier described here is a basic module for homebrew equipment that should satisfy the requirements for a 100-W power level, a 1.8 to 29.7 MHz bandwidth and a high-quality signal. It should run trouble-free for a very long time. The initial cost of the four transistors is about $180, but they will last indefinitely if cared for properly. The ones that I use were abused considerably during the experimentation and continue to work perfectly. The approach that I suggest is best implemented with diplexer filters and a preamplifier that has very low second-harmonic output and low IMD—both easy to get at +10 dBm. The payoff for me is excellent performance and reliability, once the design was completed.

This project is suggested for the homebrew enthusiast who has at least some part-time access to lab-quality test equipment. Others who do not care to build may find the article, together with numerous other sources, interesting background information regarding MOSFET power-amplifier design and test methods.

Any attempt at building and testing this amplifier should also use the 40-V power supply in the referent of Note 1 or something similar. The automatic current limiting at 8 A, the automatic reduction of drain voltage, the short-circuit limiting to 4 A and the manual control down to 24 V help to protect the MOSFETs from mishaps that are bound to occur.

Notes

1W. Sabin, W0IYH, "Power Supply for a MOSFET Power Amplifier," *QEX*, Mar/Apr 1999, pp 50-54.

2W. Sabin, W0IYH, "Diplexer Filters for an HF MOSFET Power Amplifier," *QEX*, Jul/Aug 1999, pp 20-26.

3N. Dye and H. Granberg, *Radio Frequency Transistors* (New York: Butterworth-Heinemann, 1993), pp 42-43.

4W. Sabin and E. Schoenike, *HF Radio Systems and Circuits* (Tucker, Georgia: Noble Publishing Co, 1998), Chapter 12, "Solid State Power Amplifiers," by R. Blocksome. This chapter is highly recommended regarding solid-state SSB power amplifiers. This book is also ARRL Order No. 7253. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line on the World Wide Web at **http://www.arrl.org/catalog**.

5RF Parts Co, 435 S Pacific St, San Marcos, CA 92069; tel 760-744-0700, 800-737-2787 (orders only), fax 760-744-1943; **rfp@rfparts.com**, **http://www.rfparts.com/**.

6Communication Concepts, Inc. (CCI), 508 Millstone Dr, Beavercreek, OH 45434; tel 937-426-8600, fax 937-429-3811; **cci.dayton@pobox.com**, **http://www.communication-concepts.com/**; ask for a catalog.

7*Motorola Wireless Semiconductor Solutions, Device Data*, Rev 9, Vol. II; Motorola #DL110/D. The data book is available from RF Parts (see Note 5). (A Motorola Selector Guide is available at **http://www.mot-sps.com/books/sguides/pdf/sg46rev18.pdf**. There is a MRF150 page at **http://www.mot-sps.com/products/rf_and_if/rf_transistors/high_power/power_mosfets/mrf150.html**. A MRF426 datasheet (MRF426/D)is available at **http://www.zettweb.com/CDROMs/cdrom013/pdf/mrf426rev0.pdf**, which is *not* a Motorola site.—*Ed.*)

8Dye and Granberg, pp 113-115.

9*Motorola Wireless Semiconductor Solutions*; see pp 4.2-133 for the MRF151, similar to the MRF150. (See Note 7 for an MRF150 URL—*Ed.*)

10Dye and Granberg, Chapter 12.

11The set of boards costs $12 plus shipping and handling. FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; tel 847-836-9148 (Voice mail), fax 847-836-9148 (same as voice mail); **farcir@ais.net**, **http://www.cl.ais.net/farcir/**.

12Dye and Granberg, Chapter 4.

13H. Granberg, "Wideband RF Power Amplifier," *RF design*, Feb 1988; also *Motorola RF Application Reports*, AR313, p 424.

14Dye and Granberg, similar to p 124, Fig 7.7.

15*The ARRL Handbook* (Newington, CT: 1995-2000 editions), ARRL Order No. 1832; pp 17.60-17.66 and Fig 17.69. See Note 4 for purchasing information.

16W. Sabin, W0IYH, "Designing Narrow Band-Pass Filters with a BASIC Program," *QST*, May 1983, pp 23-29.