

## LOW-DISTORTION 1.6 TO 30 MHz SSB DRIVER DESIGNS

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### GENERAL CONSIDERATION

Two of the most important factors to be considered in broadband linear amplifier design are the distortion and the output harmonic rejection.

The major cause for intermodulation distortion is amplitude nonlinearity in the active element. The non-linearity generates harmonics, and the fundamental odd-order products are defined as  $2f_1-f_2$ ,  $2f_2-f_1$ ,  $3f_2-2f_1$ , etc., when a two-tone test signal is used. These harmonics may not always appear in the amplifier output due to filtering and cancellation effects, but are generated within the active device. The amplitude and harmonic distortion cannot really be distinguished, except in a case of a cascaded system, where even-order products in each stage can produce odd-order products through mixing processes that fall in the fundamental region.<sup>2</sup> This, combined with phase distortion—which in practical circuits is more apparent at higher frequencies—can make the distortion analysis extremely difficult;<sup>5,2</sup> whereas, if only amplitude distortion was present, the effect of IMD in each stage could easily be calculated.

In order to expect a low harmonic output of the power amplifier, it is also important for the driving source to be harmonic-free. This is difficult in a four-octave bandwidth system, even at 10–20 watt power levels. Class A biasing helps the situation, and Class A push-pull yields even better results due to the automatic rejection of even harmonics.

Depending on the application, a full Class A system is not always feasible because of its low efficiency. The theoretical maximum is 50%, but practical figures are not higher than 25% to 35%. It is sometimes advantageous to select a bias point somewhere between Class AB and A which would give sufficiently good results, since filtering is required in the power amplifier output in most instances anyway.

In order to withstand the high level of steady dc bias current, Class A requires a much larger transistor die than Class B or AB for a specific power output. There are sophisticated methods such as generating the bias voltage from rectified RF input power, making the dc bias proportional to the drive level.<sup>1</sup> This also yields to a better efficiency.

### 20 W, 25 dB AMPLIFIER WITH LOW-COST PLASTIC DEVICES

The amplifier described here provides a total power gain of about 25 dB, and the construction technique allows the use of inexpensive components throughout. The plastic RF power transistors, MRF475 and MRF476, featured in this amplifier, were initially developed for the CB market. The high manufacturing volume of these

TO-220 packaged parts makes them ideal for applications up to 50 MHz, where low cost is an important factor.

The MRF476 is specified as a 3-watt device and the MRF475 has an output power of 12 watts. Both are extremely tolerant to overdrive and load mismatches, even under CW conditions. Typical IMD numbers are better than -35 dB, and power gains are 18 dB and 12 dB, respectively, at 30 MHz.

The collectors of the transistors are electrically connected to the TO-220 package mounting tab which must be isolated from the ground with proper mounting hardware (TO-220 AB) or by floating heat dissipators. The latter method, employing Thermalloy 6107 and 6106 heat dissipators, was adapted for this design. Without an airflow, the 6106 and 6107 provide sufficient heat sinking for about 30% duty cycle in the CW mode. Collector idle currents of 20 mA are recommended for both devices, but they were increased to 100 mA for the MRF475 and to 40 mA for the MRF476 to reduce the higher order IMD products and to achieve better harmonic suppression.

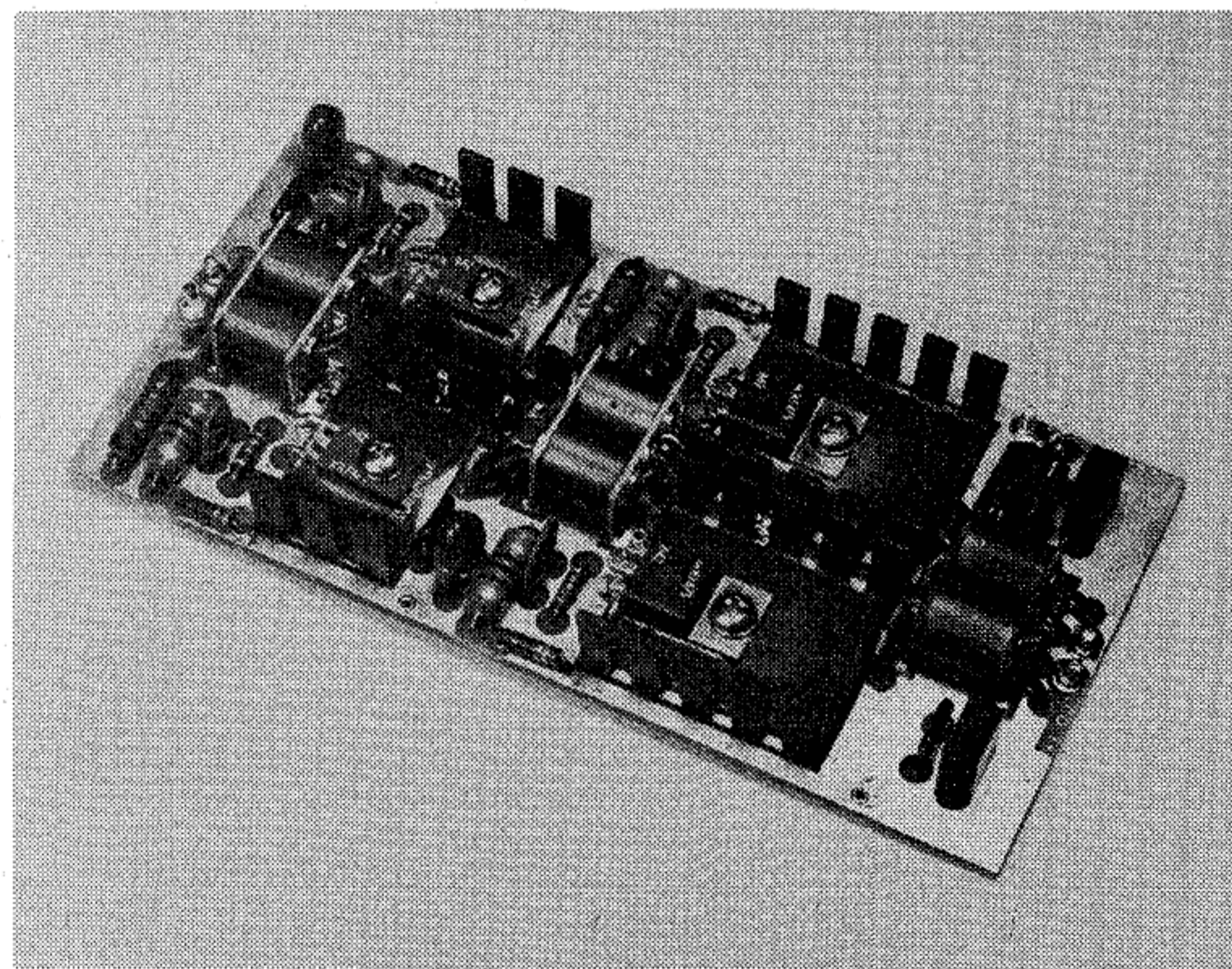


FIGURE 1

### Biasing and Feedback

The biasing is achieved with the well-known clamping diode arrangement (Figure 2). Each stage has its own diode, resistor, and bypass network, and the diodes are mounted between the heat dissipators, being in physical contact with them for temperature-tracking purposes. A better thermal contact is achieved through the use of silicone grease in these junctions.

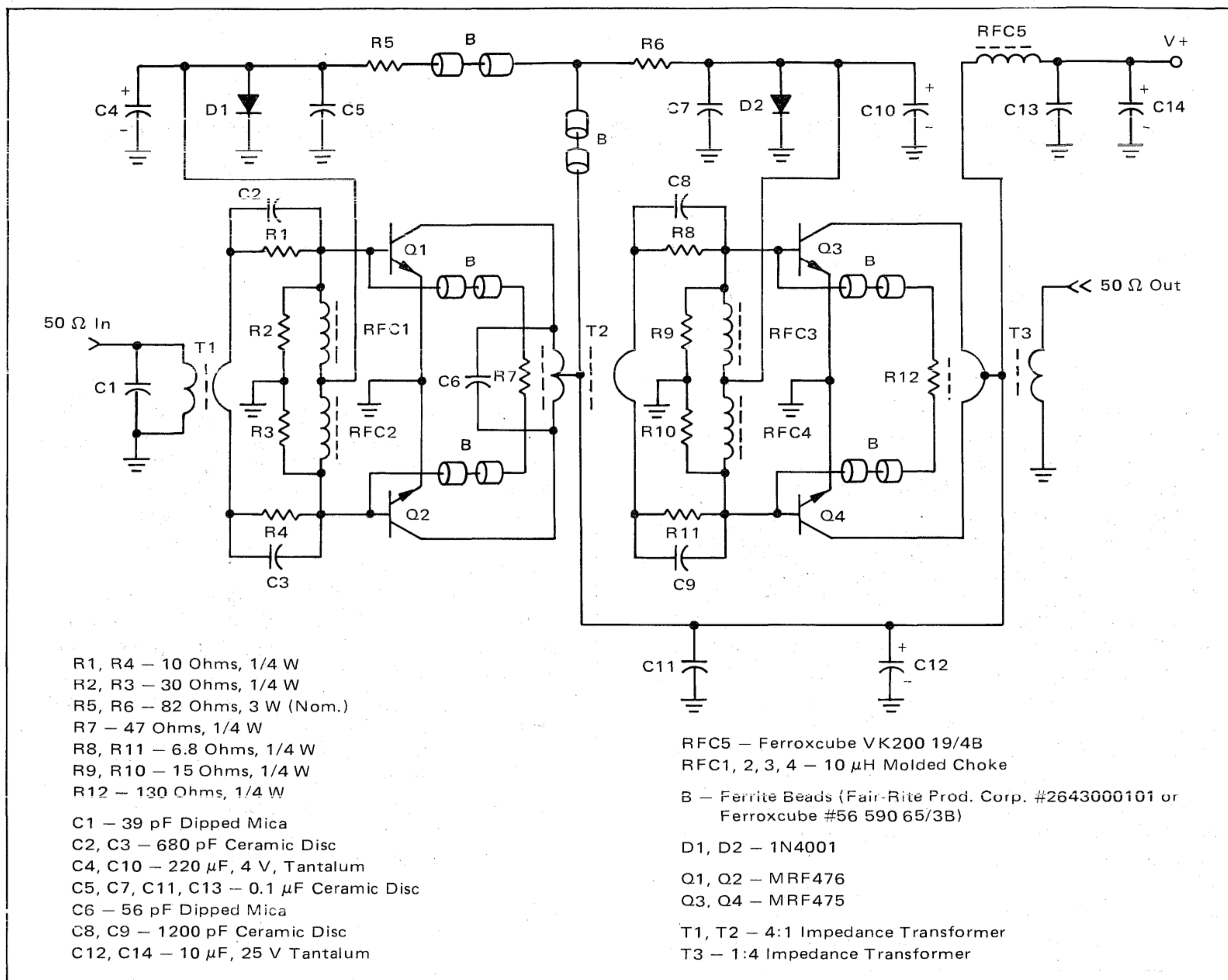


FIGURE 2

The bias currents of each stage are individually adjustable with R5 and R6. Capacitors C4 and C10 function as audio-frequency bypasses to further reduce the source impedance at the frequencies of modulation.

This biasing arrangement is only practical in low and medium power amplifiers, since the minimum current required through the diode must exceed  $I_C/h_{fe}$ .

Gain leveling across the band is achieved with simple RC networks in series with the bases, in conjunction with negative feedback. The amplitude of the out-of-phase voltages at the bases is inversely proportional to the frequency as a result of the series inductance in the feedback loop and the increasing input impedance of the transistors at low frequencies. Conversely, the negative feedback lowers the effective input impedance presented to the source (not the input impedance of the device itself) and with proper voltage slope would equalize it. With this technique, it is possible to maintain an input VSWR of 1.5:1 or less from 1.6 to 30 MHz.

#### Impedance Matching and Transformers

Matching of the input and output impedances to 50 ohms, as well as the interstage matching, is accomplished with broadband transformers (Figures 3 and 4).

Normally only impedance ratios such as 1:1, 4:1, 9:1, etc., are possible with this technique, where the low-impedance winding consists of metal tubes, through which an appropriate number of turns of wire is threaded to form the high-impedance winding. To improve the broadband characteristics, the winding inductance is increased with magnetic material. An advantage of this design is its suitability for large-quantity manufacturing, but it is difficult to find low-loss ferrites with sufficiently high permeabilities for applications where the physical size must be kept small and impedance levels are relatively high. Problems were encountered especially with the output transformer design, where an inductance of 4  $\mu$ H minimum is required in the one-turn winding across the collectors, when the load impedance is

$$\frac{2(V_{CE} - V_{CEsat})^2}{P_{out}} = \frac{2(13.6 - 2.5)^2}{20} = 12.3 \text{ ohms.}^{4,8}$$

Ferrites having sufficiently low-loss factors at 30 MHz range only up to 800-1000 in permeability and the inductance is limited to 2.5-3.0  $\mu$ H in the physical size required. This would also limit the operation to approximately 4 MHz, below which excessive harmonics are

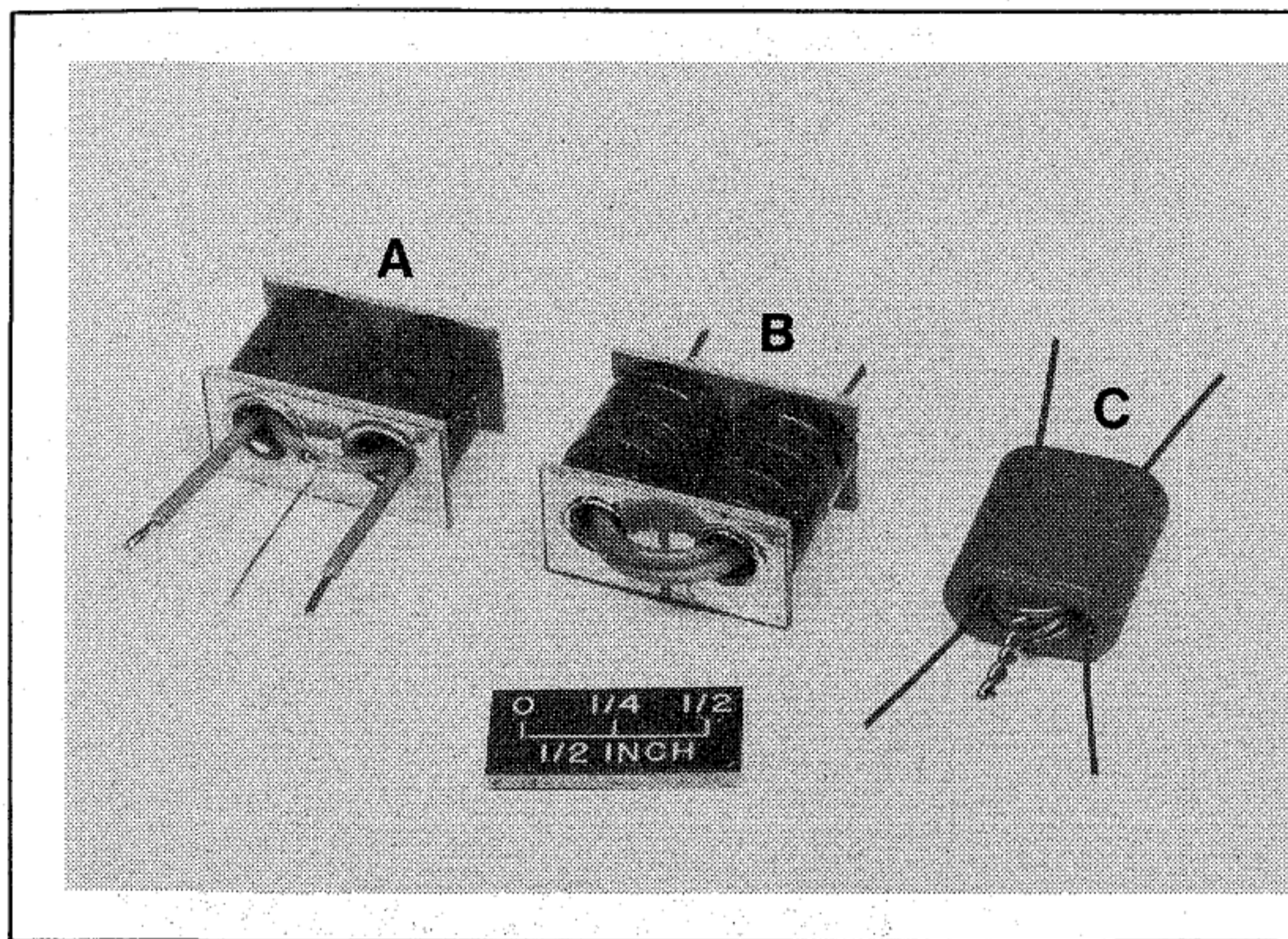


FIGURE 3

Examples of broadband transformers. Variations of these are used in all designs of this article (see text). All ferrites in transformers are Fair-Rite Products Corp. #2643006301 ferrite beads.\* The turns ratios shown in Figure 4 are imaginary and do not necessarily lead to correct design practices.

generated and the efficiency will degrade. One possible solution is to increase the number of turns, either by using the metal tubes for only part of the windings as in Figure 4B, or simply by winding the two sets of windings randomly through ferrite sleeves or a series of beads (Figures 3C and 4C). In the latter, the metal tubes can be disregarded or can be used only for mounting purposes. T3 was eventually replaced with a transformer of this type, although not shown in Figure 1.

Below approximately 100 MHz, the input impedances of devices of the size of MRF475 and smaller are usually capacitive in reactance, and the  $X_S$  is much smaller than the  $R_S$ , (Low Q). For practical purposes, we can then use the formula  $\sqrt{R_S^2 + X_S^2}$  to find the actual input impedance of the device. The data-sheet numbers for 30 MHz are 4.5,  $-j2.4$  ohms, and we get  $\sqrt{4.5^2 + 2.4^2} = 5.1$  ohms. The base-to-base impedance in a push-pull circuit would be four times the base-to-emitter impedance of one transistor. However, in Class AB, where the base-emitter junction is forward biased and the conduction angle is increased, the impedance becomes closer to twice that of one device. The rounded number of 11 ohms must then be matched to the driver output. The drive power required with the 10 dB specified minimum gain is

$$P_{out}/\text{Log}^{-1}(G_{PE}/10) = 2.0 \text{ W}$$

and the driver output impedance using the previous formula is  $2(11.1^2)/2 = 123$  ohms. The 11 ohms in series with the gain-leveling networks (C8, R8 and C9, R11) is 17 ohms. The closest practical transformer for this interface would be one with 9:1 impedance ratio. This would present a higher-than-calculated load impedance to the driver collectors, and for the best linearity the output load

\*Wallkill, N.Y. 12589

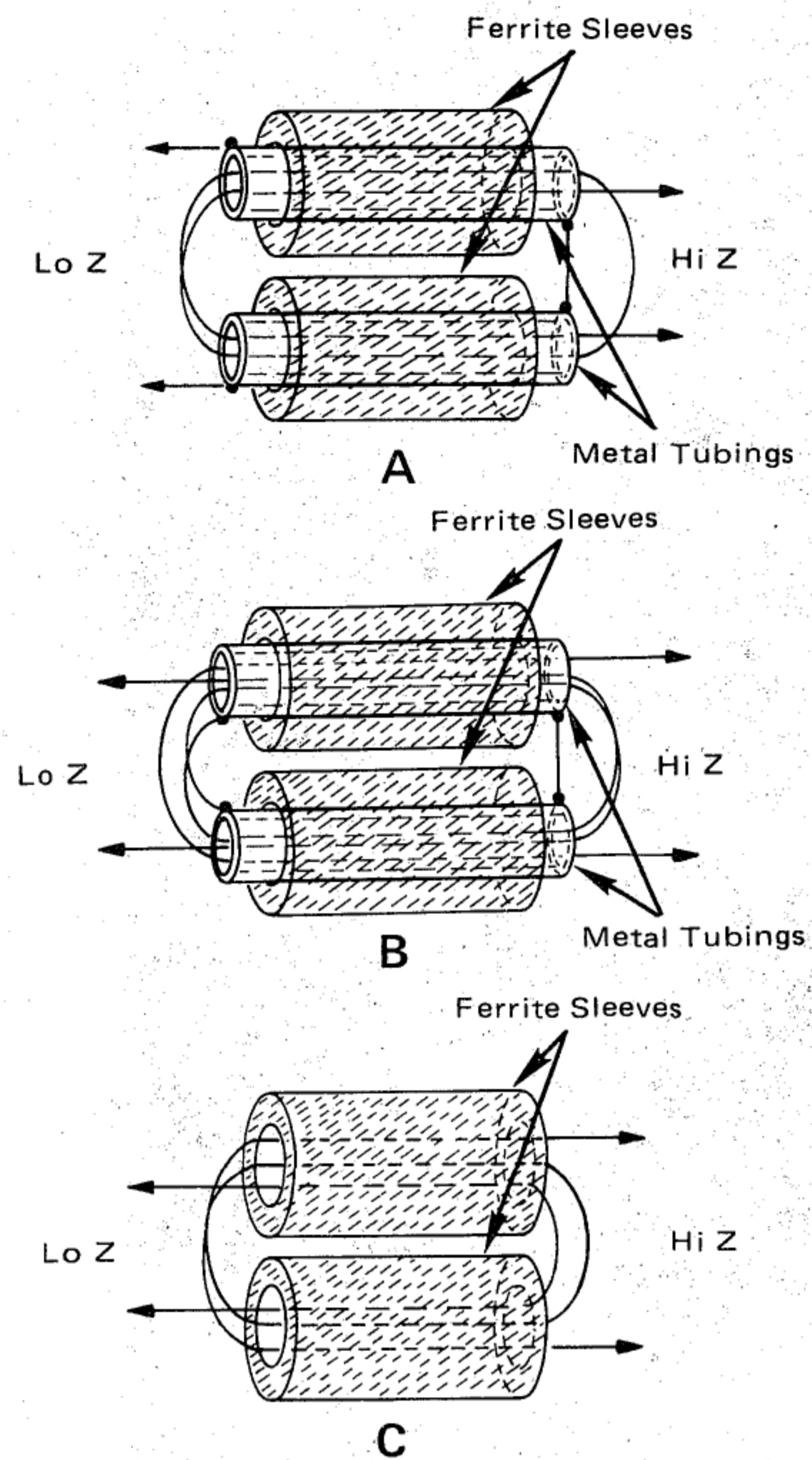


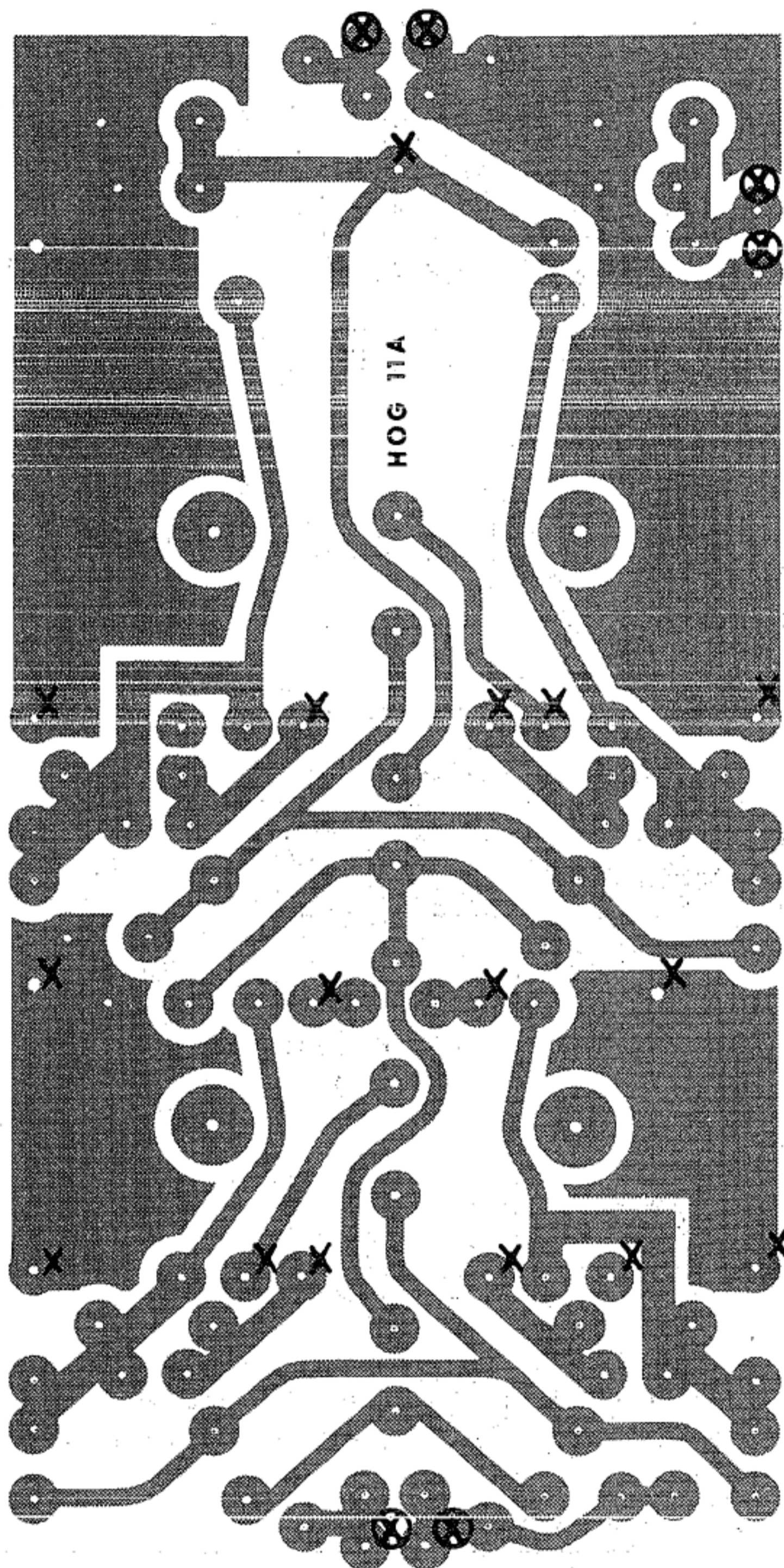
FIGURE 4

should be lower than required for the optimum gain and efficiency. Considering that the device input impedance increases at lower frequencies, a better overall match is possible with a 4:1, especially since the negative feedback is limited to only 4 dB at 2 MHz due to its effect on the efficiency and linearity.

The maximum amount of feedback a circuit can tolerate depends much on the physical layout, the parasitic inductances, and impedance levels, since they determine the phase errors in the loop. Thus, in general, the high-level stages should operate with lower feedback than the low-level stages.

The maximum amount of feedback the low-level driver can tolerate without noticeable deterioration in IMD is about 12 dB. This makes the total 16 dB, but from the data sheets we find that the combined gain variation for both devices from 2 to 30 MHz is around 29 dB. The difference, or 13 dB, should be handled by the gain-leveling networks.

The input impedance of the MRF476 is 7.55,  $-j0.65$  ohms at 30 MHz resulting in the base-to-base impedance of  $2 \times \sqrt{7.55^2 + 0.65^2} = 15.2$  ohms. This, in series with networks R1, C1 and R4, C3 ( $2 \times 4.4$  ohms), gives 24 ohms, and would require a 2:1 impedance ratio transformer for a 50-ohm interface. However, due to the influence of strong negative feedback in this stage, a better overall matching is possible with 4:1 ratio. The input networks were designed in a manner similar to that described in Reference 8.



- = Feedthrough Eyelets
- ⊗ = Terminal Pins

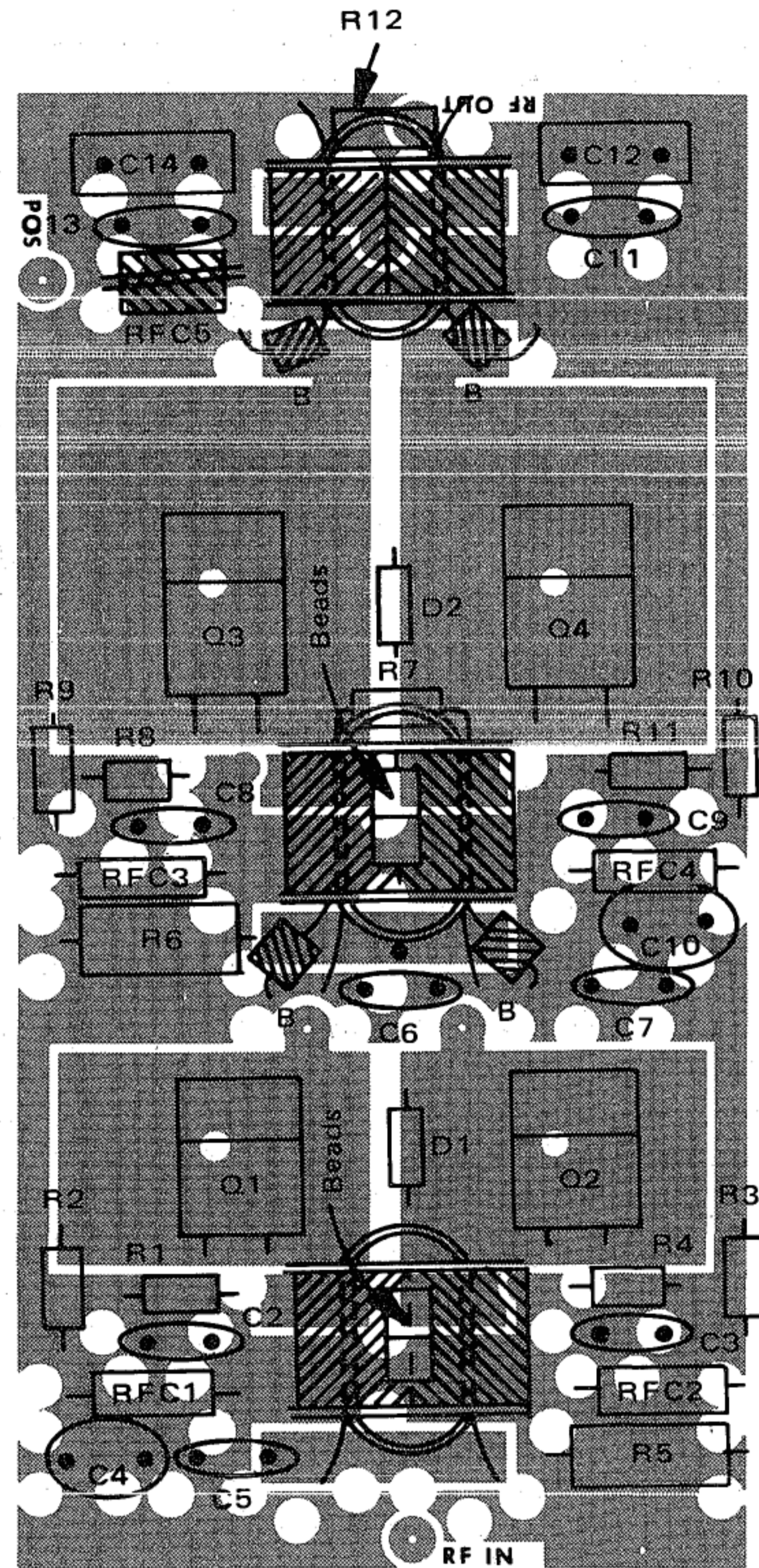


FIGURE 5

Component Layout Diagram of Low-Cost 20 W Amplifier

The leads of R7 and R12 form the one-turn feedback windings in T2 and T3. Ferrite beads in dc line can be seen located under T1 and T2.

Measurements and Performance Data

At a power output of 20 W CW, all output harmonics were measured about 30 dB or more below the fundamental, except for the third harmonic which was only attenuated 17 dB to 18 dB at frequencies below 5 MHz. Typical numbers for the higher order distortion products ( $d_9$  and  $d_{11}$ ) are in the order of  $-60$  dB above 7 MHz and  $-50$  dB to  $-55$  dB at the lower frequencies. These both can be substantially reduced by increasing the idle currents, but larger heat sinks would be necessary to accommodate the increased dissipation.

The efficiency shown in Figure 6 represents the overall figure for both stages. Currents through the bias networks, which are  $82/(13.6 - 0.7) = 0.16$  A each, are excluded. Modified values for R5 and R6 may have to be selected, depending on the forward voltage characteristics of D1 and D2.

Although this amplifier was designed to serve as a 1.6 to 30 MHz broadband driver, it is suitable for the citizens band use as well. With some modifications and design shortcuts, the optimization can be concentrated to one frequency.

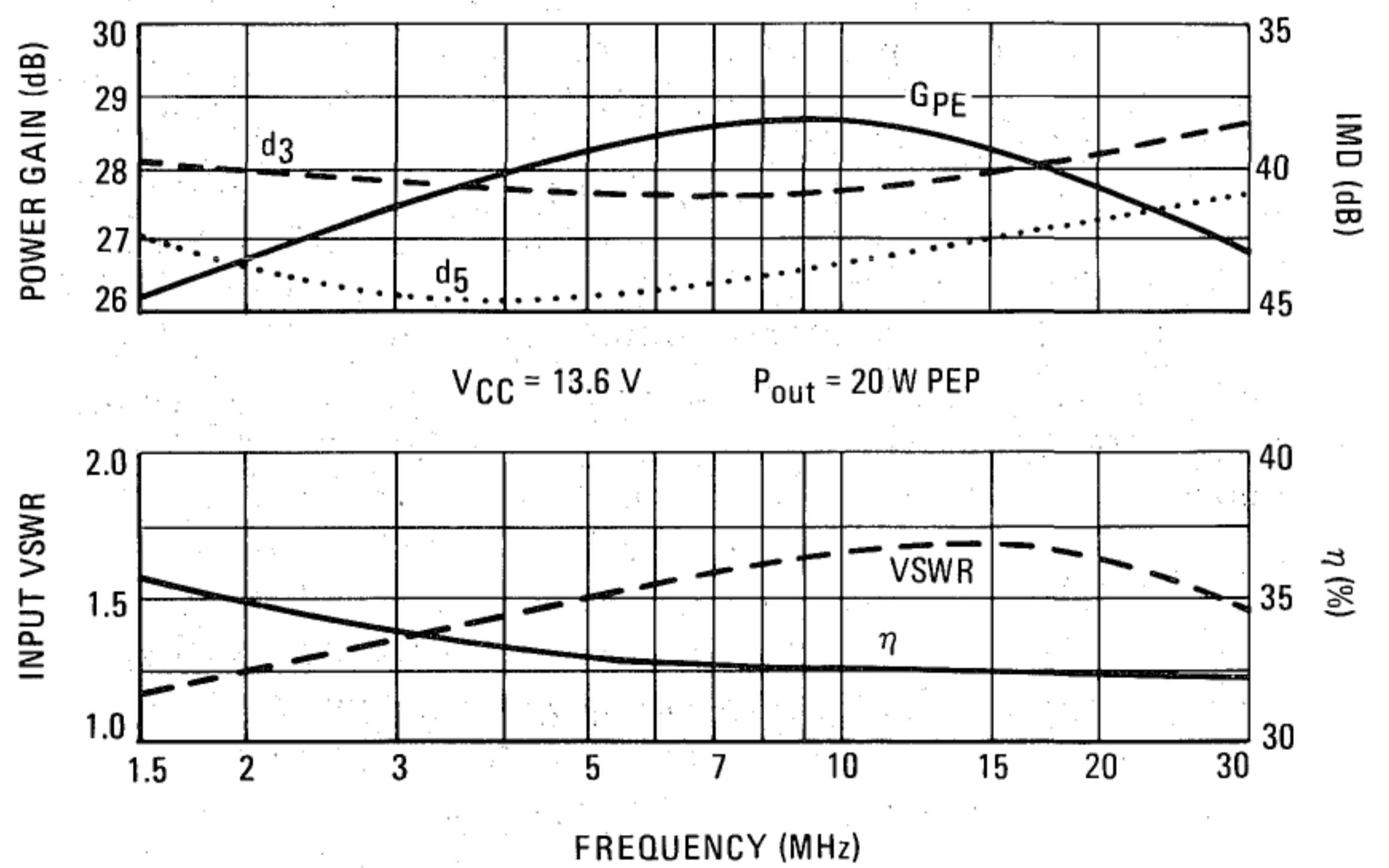


FIGURE 6

Intermodulation distortion and power gain versus frequency (upper curves).  
 Input VSWR and combined collector efficiency of both stages (lower curves).

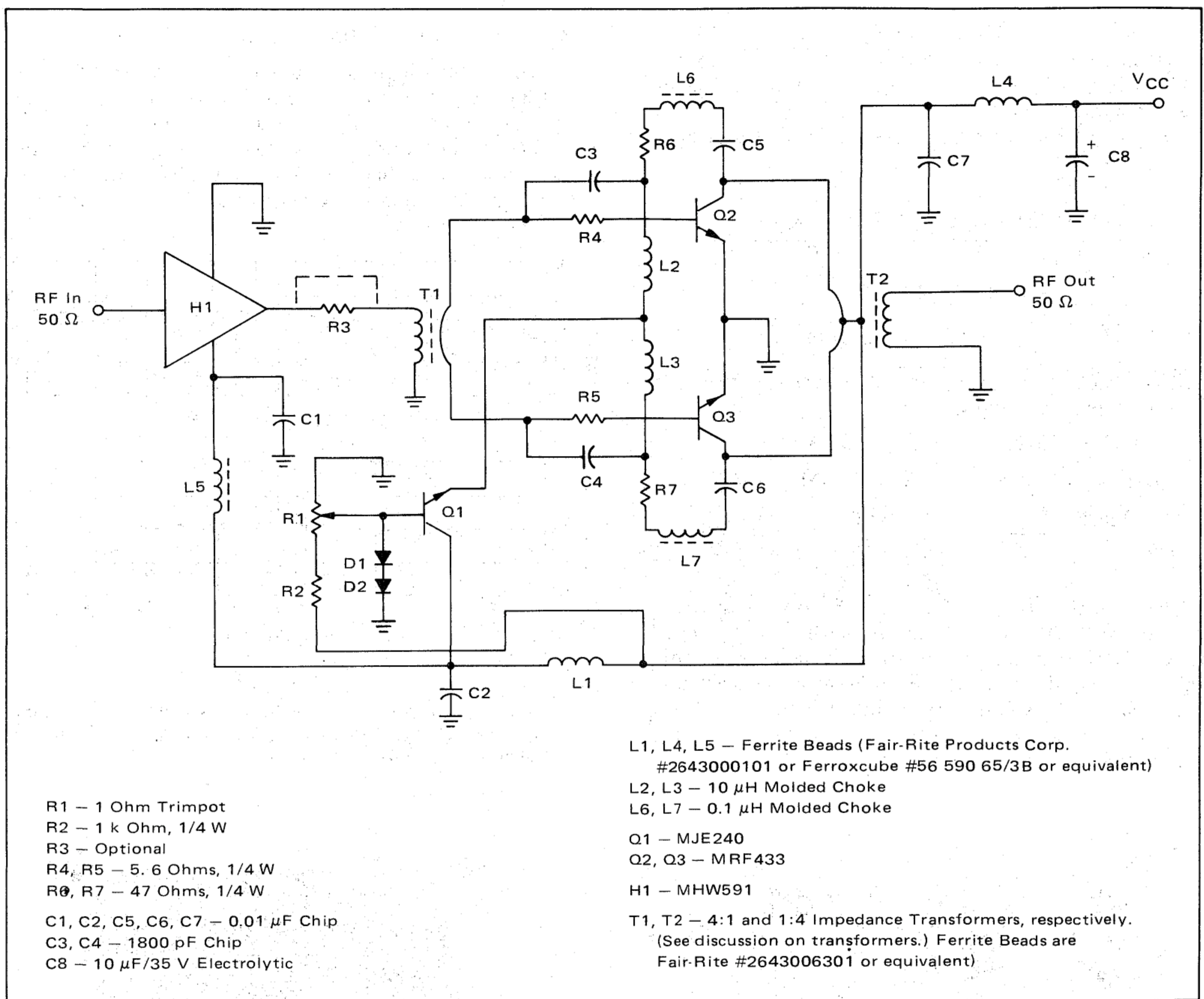


FIGURE 8

The output matching is done with a transformer similar to that described in the first part of this paper (Figures 4B, 4C). This transformer employs a multi-turn primary, which can be provided with a center tap for the collector dc feed. In addition to a higher primary inductance, more effective coupling between the two transformer halves is obtained, which is important regarding the even-order harmonic suppression.

#### 28-Volt Version

A 28-V version of this unit has also been designed with the MHW592 and a pair of MRF401s. The only major change required is the output transformer, which should have a 1:1 impedance ratio in this case. The transformer consists of six turns of RG-196 coaxial cable wound on an Indiana General F-627-8-Q1 toroid. Each end of the braid is connected to the collectors, and the inner conductor forms the secondary. A connection is made in the center of the braid (three turns from each end) to form the center tap and dc feed.

The MRF433 and MRF401 have almost similar input characteristics, and no changes are necessary in the input

circuit, except for the series feedback resistors, which should be 68–82 ohms and 1 W.

In designing the gain-leveling networks, another approach can be taken, which does not involve the computer program described in Reference 8. Although the input VSWR is not optimized, it has proved to give satisfactory results.

The amount of negative feedback is difficult to determine, as it depends on the device type and size and the physical circuit layout. The operating voltage has a minimal effect on the transistor input characteristics, which are more determined by the electrical size of the die. High-power transistors have lower input impedances and higher capacitances, and phase errors are more likely to occur due to circuit inductances.

Since the input capacitance is an indication of electrical size of the device, we can take the paralleled value ( $X_p$ ) at 2 MHz, which is  $X_s + (R_s^2/X_s)$  and for MRF433  $3.5 + (9.1^2/3.5) = 27$  ohms. The  $X_p$  of the largest devices available today is around 10 ohms at 30 MHz, and experience has shown that the maximum feedback should be limited to about 5 dB in such case. Using these figures

as constants, and assuming the  $G_{PE}$  is at least 10 dB, we can estimate the amount of feedback as:  $5/(10^2/27) + 5 = 6.35$  dB, although only 4 dB was necessary in this design due to the low  $\Delta G_{PE}$  of the devices.

The series base resistors (R4 and R5) can be calculated for 4 dB loss as follows:

$$\frac{[(V_{in} \times \Delta 4dB) - V_{in}]}{I_{in}} = \frac{[(0.79 \times 1.58) - 0.79]}{0.04}$$

$$= 11.45 \text{ ohms, or}$$

$$11.45/2 = 5.72 \text{ ohms each.}$$

$$Z_{in}(2 \text{ MHz}) = \sqrt{(9.1^2 + 3.5^2)} = 9.75 \text{ ohms, in Class AB push-pull } 19.5 \text{ ohms.}$$

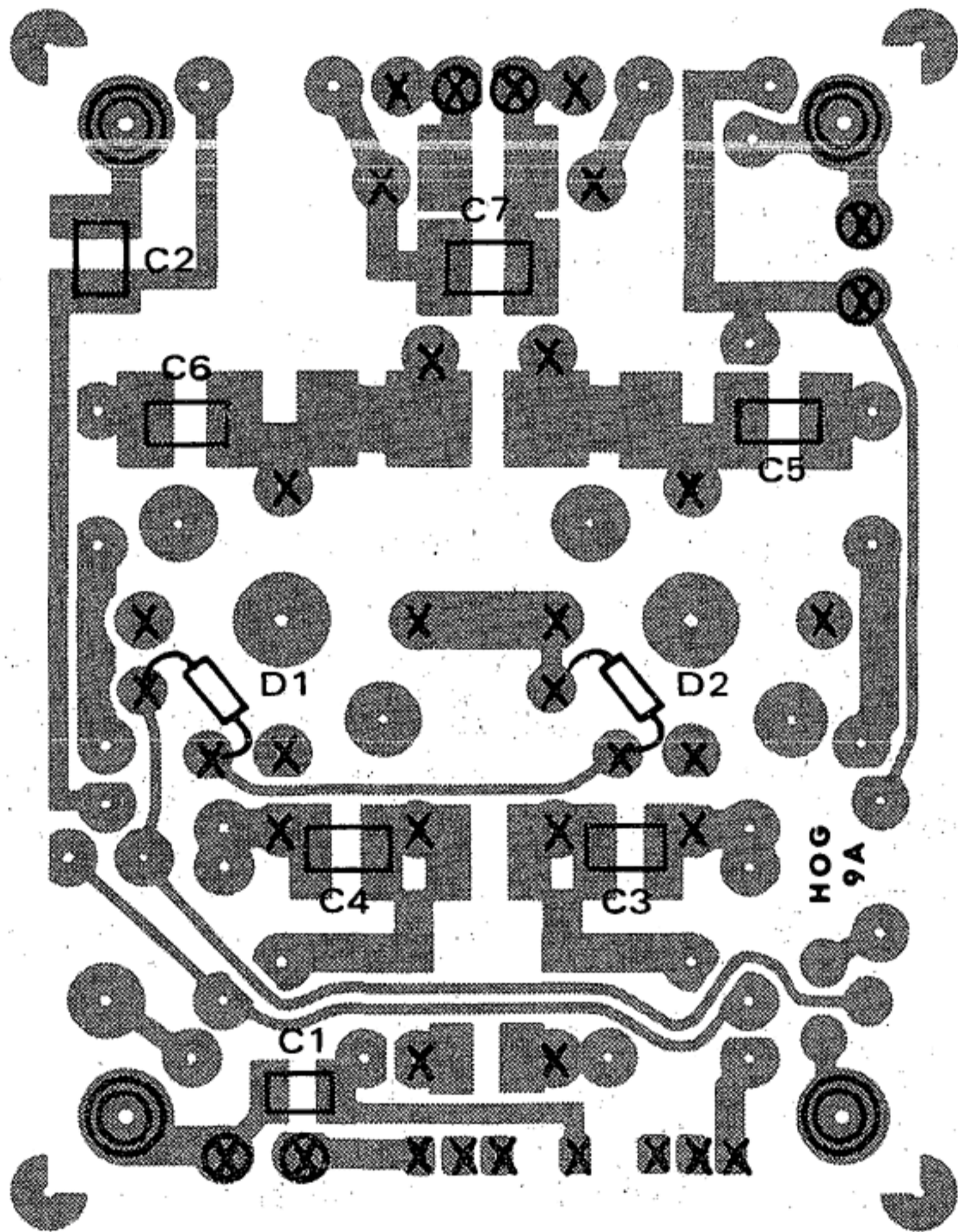
$$P_{in} = 20 \text{ W} - 28 \text{ dB} = 20/630 = 0.032 \text{ W}$$

$$V_{RMS} \text{ (base to base)} = \sqrt{(0.032 \times 19.5)} = 0.79 \text{ V}$$

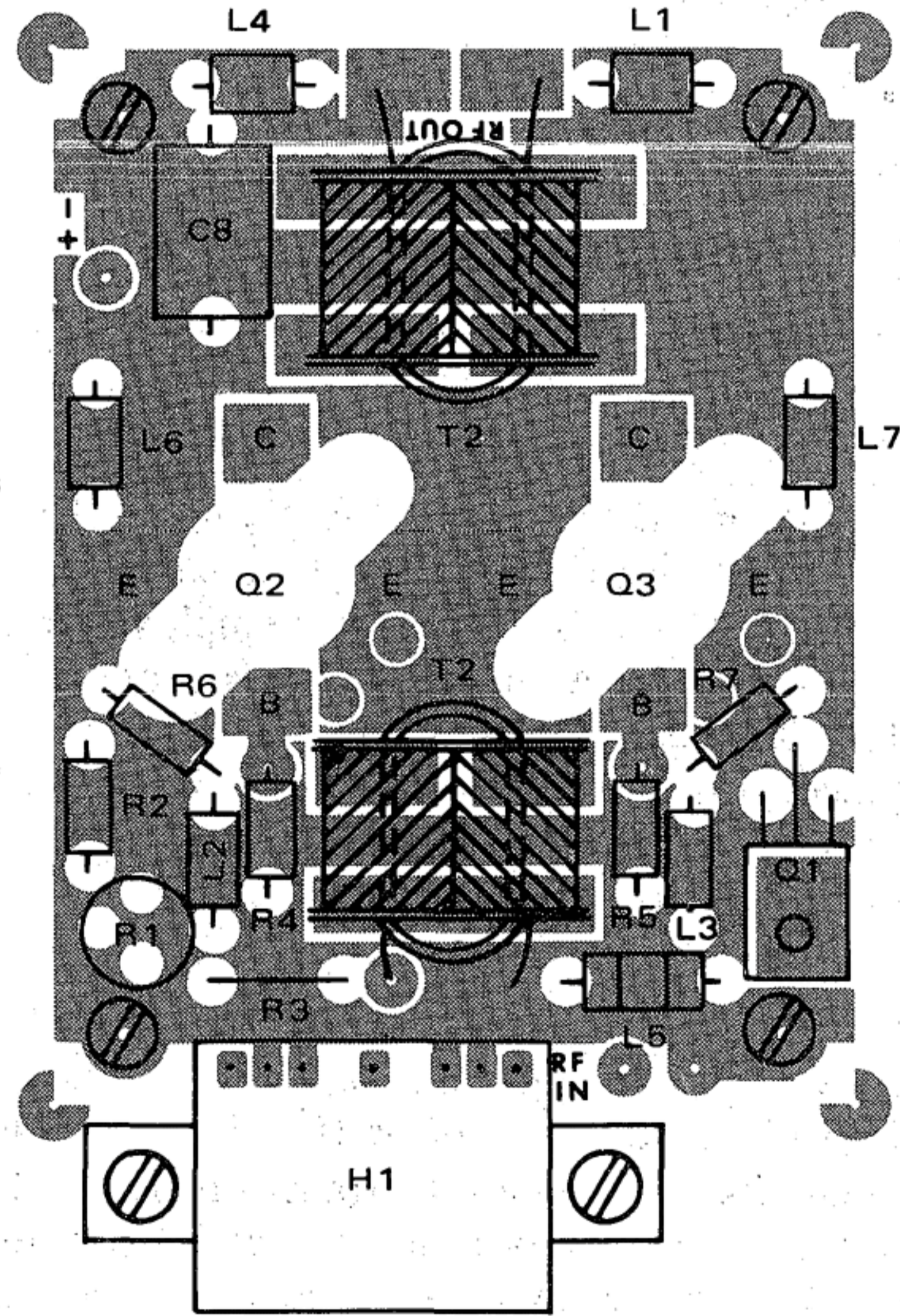
$$I_{in} = V_{in}/R_{in} = 0.79/19.5 = 0.04 \text{ A}$$

$$\Delta V 4 \text{ dB} = \sqrt{[\text{Log}^{-1}(4/10)]} = 1.58 \text{ V}$$

The parallel capacitors (C3 and C4) should be selected to resonate with R (5.7 ohms) somewhere in the mid-band. At 15 MHz, out of the standard values, 1800 pF appears to be the closest, having a negligible reactance at 2 MHz, and 2.8 ohms at 30 MHz, where most of the capacitive reactance is cancelled by the transformer winding inductance.



- = Board Standoffs
- = Terminal Pins
- = Feedthrough Eyelets



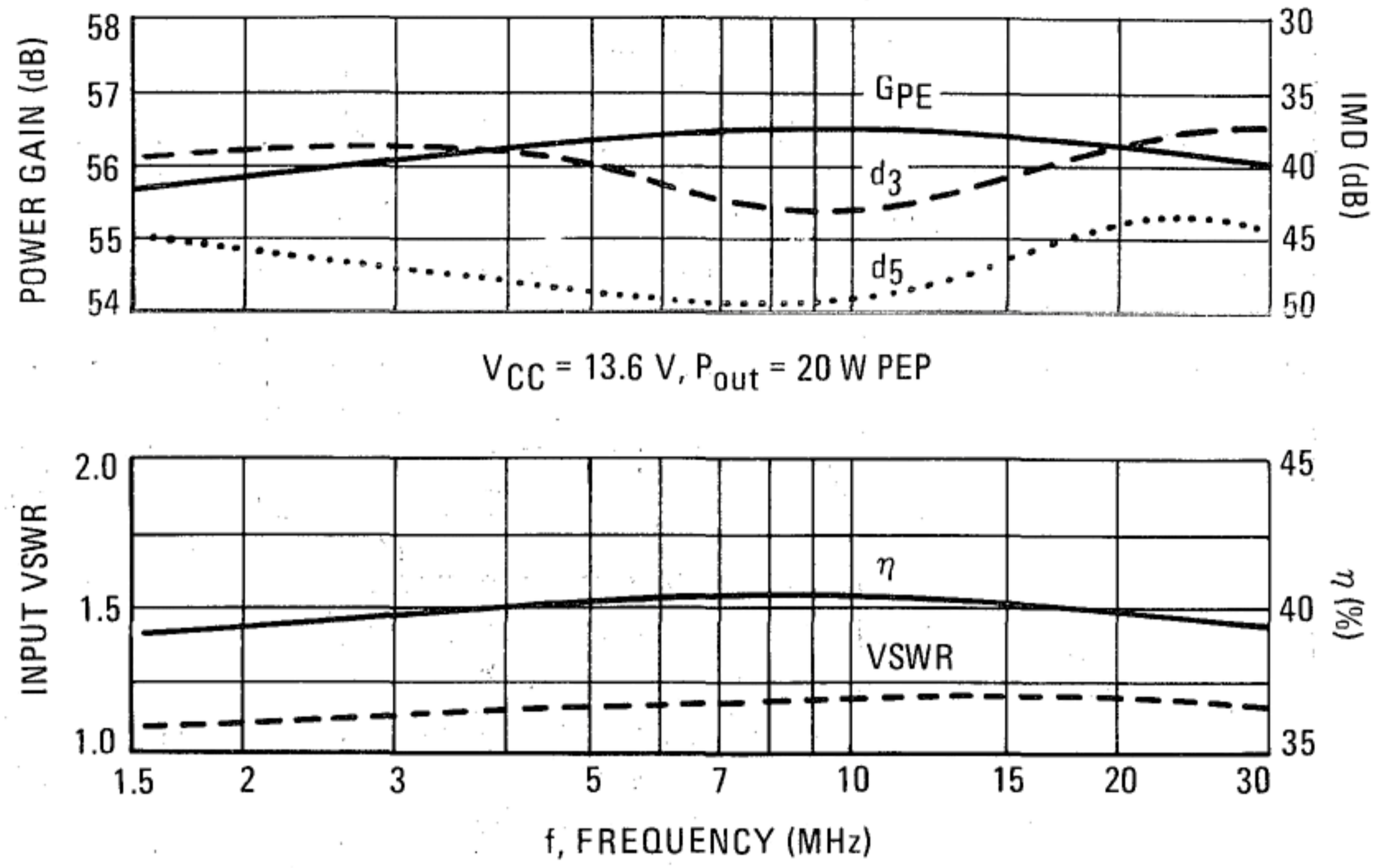
**FIGURE 9**  
Component Layout Diagram of  
20 W, 55 dB High-Performance Driver

The leads of D1 and D2 are bent to allow the diodes to contact the transistor mounting flanges.

Note that the mounting pad of Q1 must be connected to the lower side of the board through an eyelet or a plated through-hole.

**Measurements and Performance Data**

The output harmonic contents of this amplifier are substantially lower than normally seen in a Class AB system operating at this power level and having a 4.5-octave bandwidth. All harmonics except the third are attenuated more than 30 dB across the band. Between 20 and 30 MHz, -40 to -55 dB is typical. The third harmonic



**FIGURE 10**  
Intermodulation Distortion and Power Gain versus Frequency (Upper Curves). Input VSWR and Collector Efficiency (excluding MHW591) (Lower Curves).

has its highest amplitude (-20 to -22 dB), as can be expected, below 20 MHz. The measurements were done at an output level of 20 W CW and with 200 mA collector idle current per device. Increasing it to 400 mA improves these numbers by 3-4 dB, and also reduces the amplitudes of  $d_5$ ,  $d_7$ ,  $d_9$ , and  $d_{11}$  by an average of 10 dB, but at the cost of 2-3 dB higher  $d_3$ .

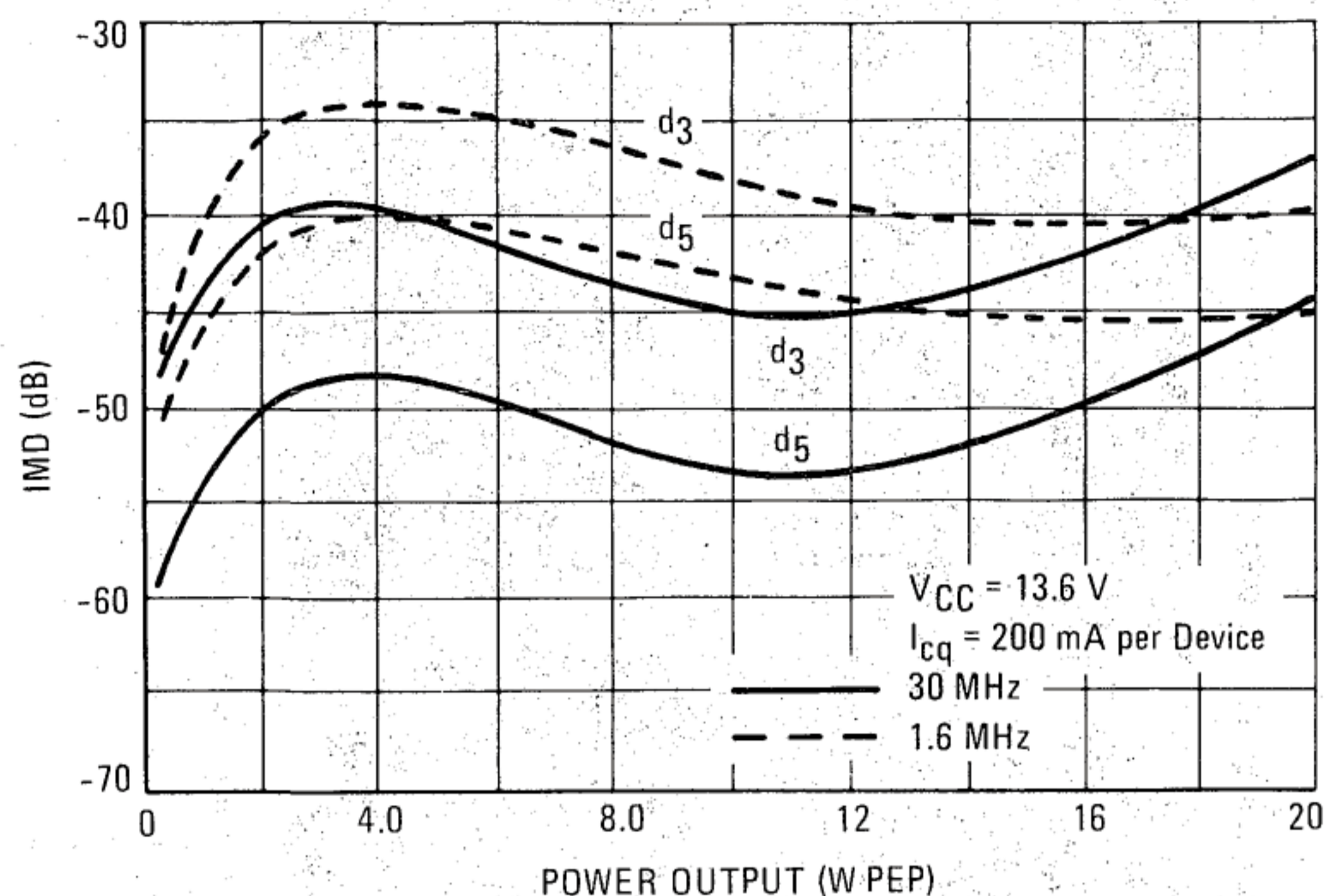


FIGURE 11 — IMD versus Power Output

#### CONCLUSION

The stability of both designs (excluding the 28 V unit) was tested into reactive loads using a setup described in Reference 8. Both were found to be stable into 5:1 load mismatch up to 7 MHz, 10:1 up to 30 MHz, except the latter design did not exhibit breakups even at 30:1 in the 20-30 MHz range. If the test is performed under two-tone conditions, where the power output varies from zero to maximum at the rate of the frequency difference, it is easy to see at once if instabilities occur at any power level.

The two-tone source employed in all tests consists of a pair of crystal oscillators, separated by 1 kHz, at each test frequency. The IMD ( $d_3$ ) is typically -60 dB and the harmonics -70 dB when one oscillator is disconnected for CW measurements.

HP435 power meters were used with Anzac CH-130-4 and CD-920-4 directional couplers and appropriate attenuators. Other instruments included HP141T analyzer system and Tektronix 7704A oscilloscope-spectrum analyzer combination.

#### REFERENCES

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The PCB layouts below are a supplement to Figures 5 and 9 and may be used for generating printed circuit artwork.

NOTE: The Printed Circuit Board shown is 75% of the original.

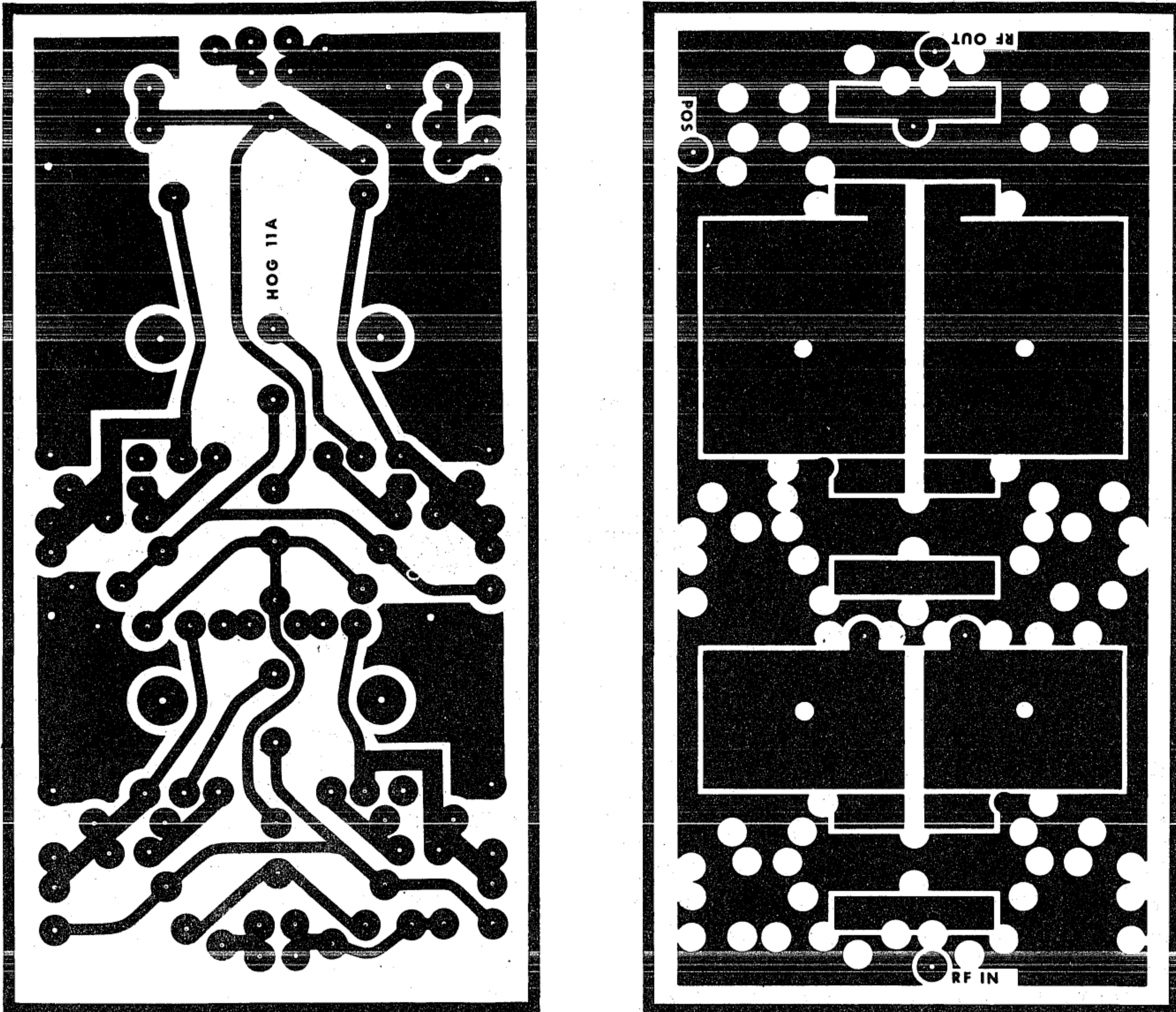


FIGURE 12 — PCB Layout of Low-Cost 20 W Amplifier

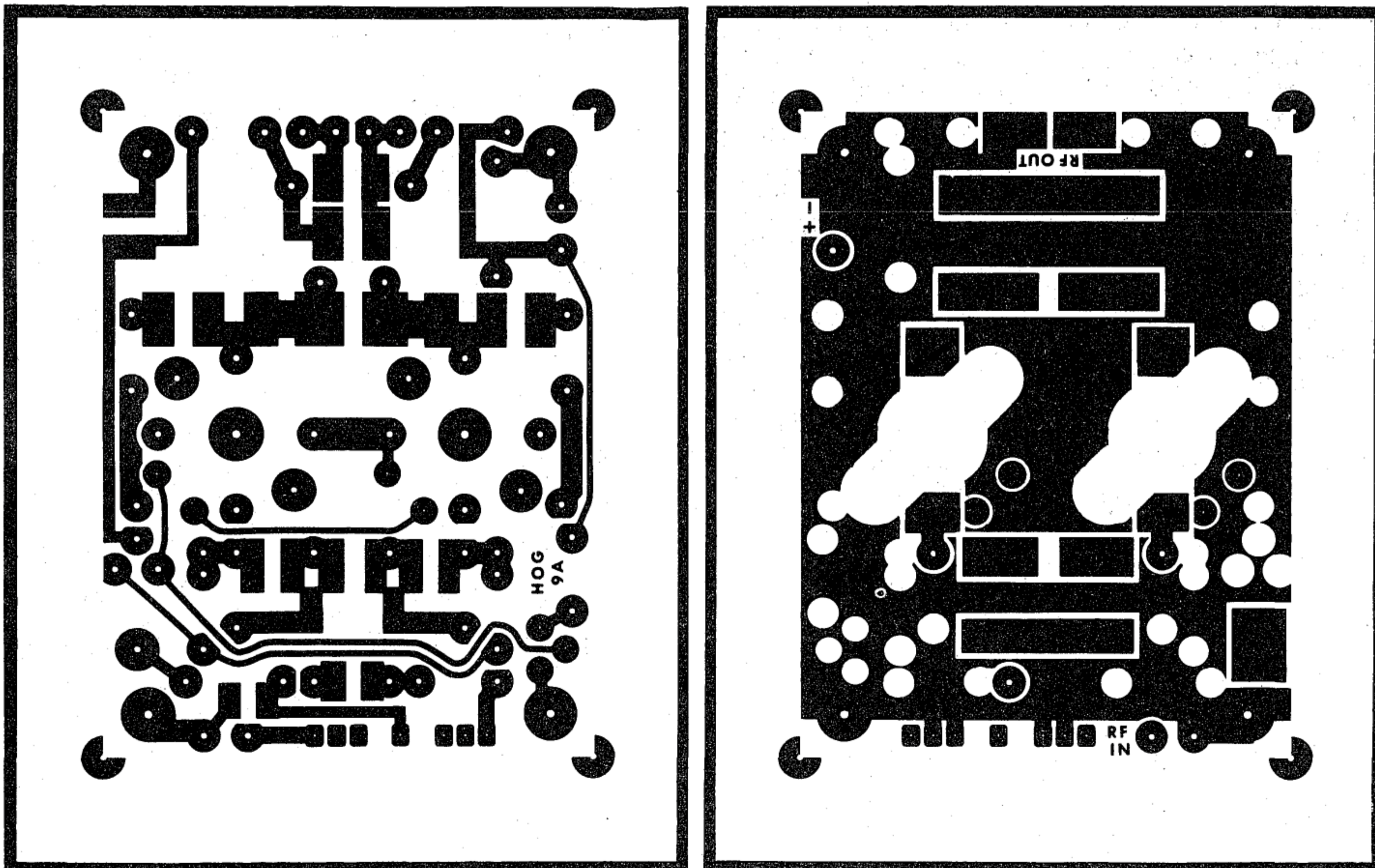


FIGURE 13 — PCB Layout of 20 W, 55 dB High-Performance Driver