MOTOROLA
Semiconductor Products Inc.

AN-791
Application Note

A SIMPLIFIED APPROACH TO VHF POWER AMPLIFIER DESIGN

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This note discusses the design of 35-W and 75-W VHF linear amplifiers. The construction technique features printed inductors, the design theory of which is fully described. Complete constructional details, including a printed circuit layout, facilitate easy reproduction of the amplifiers.

Solid-state VHF amplifier design can be simplified by employing printed or etched lines for impedance matching. The lines, having a distant ground-plane reference and high Z_O, can be treated as lumped constant inductors, and make design and duplication easier than with wire-wound inductors.

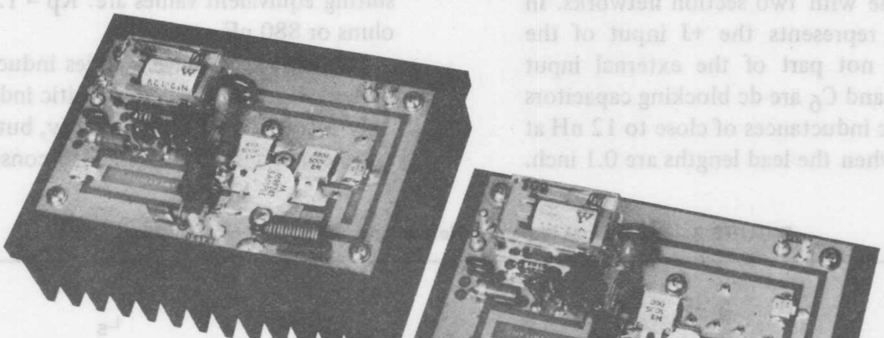
An example is an optimized 35-W amplifier which yields over 10 dB of power gain across the 2-meter amateur band. It employs an inexpensive, non-internally matched transistor, the MRF240, which has good linear characteristics for SSB operation.

A higher power version with the same board layout is concentrated around the MRF247, although this results in some compromise in the impedance matching.

A carrier operated T/R switch (COR) is incorporated, allowing applications such as a booster amplifier for hand-held and mobile radios.

Both designs are biased class AB for linear operation, but are suitable for FM operation as well. Figure 1 shows the two amplifiers.

FIGURE 1 — 35-watt and 75-watt Engineering Models



GENERAL

VHF solid state amplifier design is almost exclusively done with lumped constant LC matching networks. Broadband transformer matching is feasible when extremely wide bandwidths are required. Transmission lines for impedance transformation usually require quarter-wave electrical lengths and make designs bulky at VHF unless materials with high dielectric constant are used. Transmission lines can be realized with coaxial cable or printed lines (strip-lines) on a circuit board with a continuous ground plane, separated by a suitable dielectric material. The printed airlines discussed here are, in fact, high characteristic impedance transmission lines which, for the purposes of design calculation, are treated as inductors; therefore the quality of the board material is less critical. The printed airlines also have the advantage of repeatability and easy access for designing multi-element networks. The network calculations can be done in the same manner as if lumped-constant, round-wire inductors were used.

Input and output impedance matching in transistor amplifiers is required to transform the source impedance (usually 50 ohms) to the low complex input impedance of the device. The output load impedance, which is a function of the supply voltage and power level, must also be matched to a 50-ohm load except in multistage driver designs.

At VHF, the input and output impedances of a power transistor are both usually inductive in reactance (designated as +JX in data sheets), becoming capacitive (-JX) at lower frequencies. For transistors such as MRF240, 2N6084 and 2N5591, the crossover point is around 100 MHz. This is determined by the transistor die size, geometry and package type, and smaller devices can be capacitive up to UHF frequencies.

Since the bandwidth required here is only a fraction of an octave, (140-150 MHz) the impedance matching can be adequately done with two section networks. In Figure 2, X_1, which represents the +J input of the MRF240 transistor is not part of the external input matching network. C_1 and C_6 are dc blocking capacitors with measured parasitic inductances of close to 12 nH at the center frequency when the lead lengths are 0.1 inch. These inductances, as well as the relay inductance, are added to the values of L_1 and L_5.

If the relay were used in a 50-ohm system, it would result in 0.3 dB power loss due to impedance mismatch and losses. This can be minimized if the relay inductance is used as part of a resonant circuit, but the series inductance (37 nH per contact pair) obviously places an upper frequency limit.

The simplest approach to matching network design is with a purely resistive source and load. This can be accomplished by compensating the +J with an equal amount of capacitance (–J). C_3 and C_4 are used to accomplish the compensation in Figure 2. This is not always practical, however, especially when maximum bandwidths are required. In this case, only part of the inductive component may be cancelled, leaving the base and collector still inductively reactive. In either case, it may be considered that part of the impedance-matching occurs within the device package itself; this is more obvious with internally matched devices, which are discussed later.

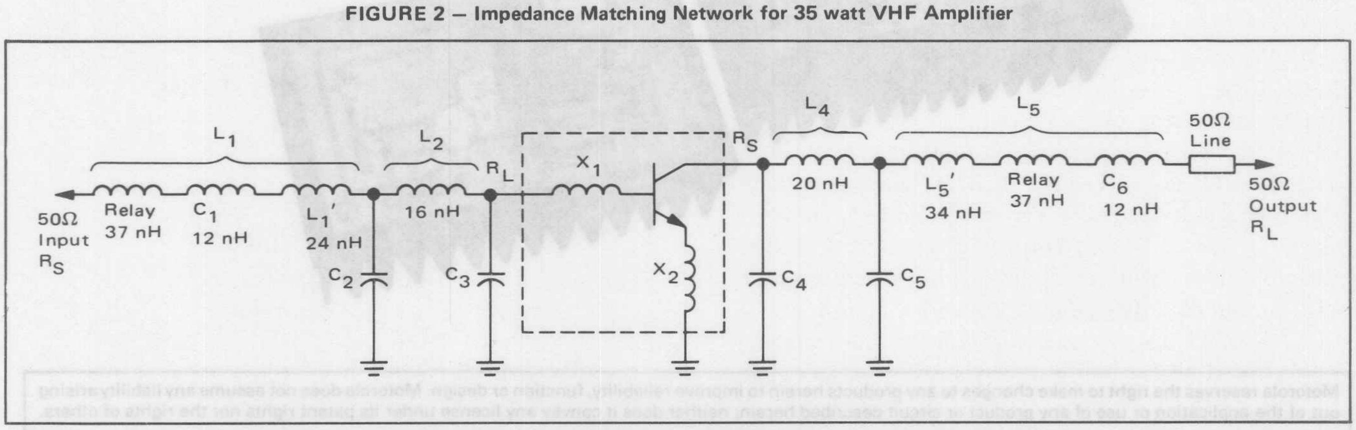
35-W LINEAR AMPLIFIER

The MRF240 was chosen for this application due to its ruggedness against load mismatch and inherently high power gain for a non-internally matched device. The transistor is rated for an output power of 40 W and a power gain of 8 dB at 175 MHz. A typical power gain at 145 MHz is 10 to 11 dB. At this frequency the input and output impedances of the MRF240 are 0.6 +J 0.8 ohms and 2.0 +J 0.1 ohms respectively (P_{out} = 35 W).

Before designing the matching networks, the values of C_3 and C_4 must be established to cancel the inductive reactance components at the base and the collector. For the input, the series numbers 0.6 +J 0.8 must be converted to parallel equivalent values, either by using a Smith chart or equations in references 3 and 4. The resulting equivalent values are: R_P = 1.67 ohms, X_P = 1.25 ohms or 880 pF.

All capacitors have a series inductive reactance component, normally called parasitic inductance. It could be only a fraction of a microhenry, but at VHF its effect is large enough to be taken into consideration. The para-

FIGURE 2 — Impedance Matching Network for 35 watt VHF Amplifier



sitic inductance results in an increased effective value of capacitance, and is frequency and impedance-level dependent.

The unencapsulated mica capacitors, widely used in VHF power applications, range from 1 to 2 nH in parasitic inductance for a single plate type, (up to 360 to 390 pF nominal values) depending on the mounting technique. Assuming a parasitic inductance of 1.5 nH, the equivalent low-frequency value can be calculated with Equation 1 as:

$$C_{Equiv} = \frac{C}{1 + [(2\pi f)^2 \ LC] \ 10^{-9}} \qquad (1)$$

where C = effective capacitance required in pF
L = parasitic inductance in nH
f = frequency in MHz

Substituting the values in equation (1):

$$C_{Equiv} = \frac{880}{1 + [(910)^2 \times 1.5 \times 880] \ 10^{-9}} = 420 \text{ pF}$$

Thus, for the required 880 pF, a capacitor of this type with equivalent low-frequency value of 420 pF, or the closest standard (390 pF), should be used.

Similarly, converting the output impedance (2.0 +J 0.1 ohms) to parallel form, R_P = 2.01 ohms and X_P = +J 26.8 ohms. The X_C represents a capacitance value of 47 pF for C_4 (from Equation 2), or a 43 pF nominal value.

$$C = \left(\frac{1}{\frac{X_C}{2\pi f}}\right) 10^6 \qquad (2)$$

where X_C = capacitive reactance in ohms
C = capacitance in pF
f = frequency in MHz

This high reactance in parallel with the low collector impedance had no noticeable effect and was completely omitted in later functional tests of the unit. It would be easy to see from a Smith chart that the resistive components of 1.67 ohms and 2.01 ohms remain unchanged, and can be treated as a purely resistive load and source for the matching network calculations.

At high frequencies the base-emitter impedance of the transistor die itself is always lower than the collector output impedance. With power devices, both can be only a fraction of an ohm. The input impedance is increased by the base and emitter bonding wire and package lead frame inductances, which are effectively in series with the transistor base (Figure 2, X_1 and X_2). The collector has normally much less series inductance since it is attached directly to the package bonding pad.

From this it can be seen that part of the matching network is actually built into the transistor package, and it is obvious that the amplifier bandwidth cannot be accurately determined by calculating the Q values of the external matching networks. (See the discussion of a 75-W linear amplifier.)

As an approximation, the 3 dB bandwidth can be used to obtain a starting point. Assuming a 15 MHz bandwidth at ±1.5 dB is desired at 145 MHz center frequency, a loaded Q of approximately 9 is required. For simplicity this number is applied to both input and output network design.

In Figure 2, X_1 and X_2 represent the inductive impedance component of the transistor and are shown only to give an idea of the transistor internal structure. The values of L_1, L_2 and C_2 can be obtained from the Appendix, or calculated by using Equation 3:

$$XL_1 = R_S B$$

$$XL_2 = R_L Q$$

$$XC_2 = \frac{A}{Q + B} \qquad (3)$$

$$A = R_L (1 + Q^2)$$

$$B = \sqrt{\frac{A}{R_S} - 1}$$

where R_S = source impedance
R_L = load impedance

For Q = 9:

$$XL_1 = R_S B = 50 \times 1.32 = 66 \text{ ohms}$$

$$XL_2 = R_L Q = 1.67 \times 9 = 15 \text{ ohms}$$

$$XC_2 = \frac{A}{Q + B} = \frac{137}{9 + 1.32} = 13.3 \text{ ohms}$$

$$A = 1.67 (1 + 9^2) = 137$$

$$B = \sqrt{\frac{A}{50 - 1}} = 1.32$$

where R_S = 50 ohms, R_L = 1.67 ohms

$$\text{Since} \quad L = \left(\frac{XL}{2\pi f}\right) 10^3 \qquad (4)$$

where XL = inductive reactance in ohms
L = inductance in nH
f = frequency in MHz

we have from Equations 3 and 4:

$$L_1 = 73 \text{ nH}$$

$$L_2 = 16 \text{ nH}$$

$$C_2 = 82 \text{ pF}$$

Subtracting the relay inductance (37 nH) and the parasitic inductance of the blocking capacitor C_1 (12 nH) from the total value of L_1, $L_1{}' = 24$ nH. This means the total printed line inductance must be $L_1{}' + L_2 = 24 + 16 = 40$ nH.

Calculating the values of the output network in a similar manner, the values for L_4, L_5 and C_5 are obtained as 20 nH, 83 nH and 70 pF, respectively, and $L_5{}'$ becomes 34 nH.

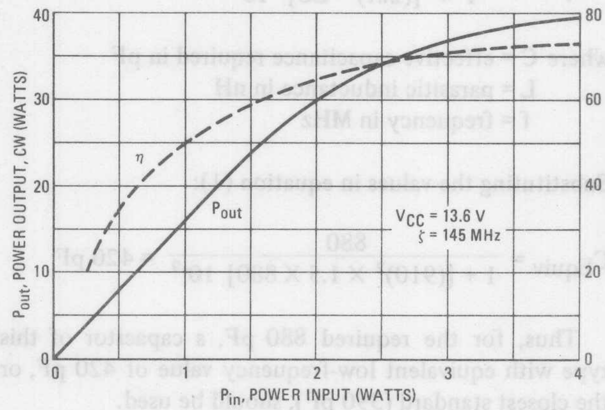
The capacitors employed for C_2 and C_5 are of the same unencapsulated mica type as C_3, but smaller in size, and their parasitic inductance is only about 1 nH. The equivalent values for C_2 and C_5 would then be 77 pF and 66 pF according to Equation 1. These are nonstandard values, and considering a 5% tolerance, a 68 pF marked value can be used for both.

Inductors L_1, L_2, L_4 and L_5 are comprised of etched lines on the circuit board. To determine their widths and lengths, the inductance of each line per unit length must be established. From the tables in the Reference section, it can be extrapolated that the inductance of #25 round wire is 24 nH per inch and #26 wire nearly 26 nH per inch. When a ground plane is 0.15 inch below, which in this case is the heat sink, and the side grounds are off an equal distance, the inductance is about one-half of this, which has been verified by measurement.

If the circuit board is made of 1-ounce, copper-clad material, (one ounce of copper per one square foot) the copper thickness is 1.4 mils. With a one mil solder plating, the total thickness is 2.4 mils, and a 100-mil-wide strip would be equivalent to a #26 round wire having a 240 square mil cross sectional area. Similarly, a 130-mil-wide strip would be equivalent to a #25 round wire with 312 square mil area. A wider line would have lower losses but would also be physically longer for a given inductance. As a compromise, a narrow line was used for the input in this design, and a wider line for the output, where the losses due to the high RF currents are more evident. Bends in the line have a minimal effect to the inductance compared to the presence of the ground plane.

From the above, the resulting inductances for the 100 mil and 130 mil lines are 13 nH per inch and 12 nH per inch, respectively. This means that for $L_1 + L_2$ a total length of 3.1 inches is required, and 4.4 inches for $L_4 + L_5{}'$. Then, for $L_2 = 16$ nH, C_2 should be located 1.3 inches from the transistor base along the input line. For $L_4 = 20$ nH, C_5 should be 1.6 inches from the collector along the output line. The Power Output and Efficiency vs. Power Input of the 35-W amplifier is shown in Figure 3.

FIGURE 3 — Power Input vs. Power Output and Collector Efficiency of 35-W Amplifier

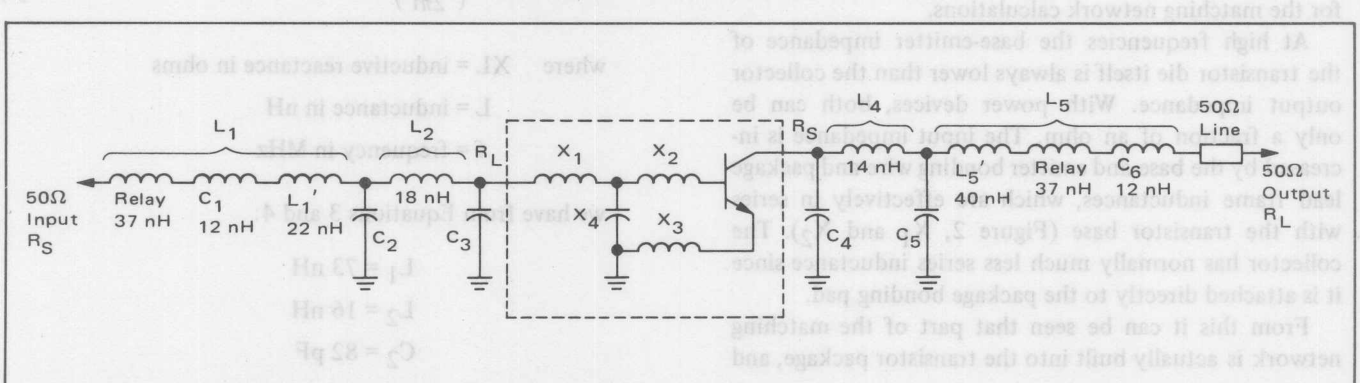


75-W LINEAR AMPLIFIER

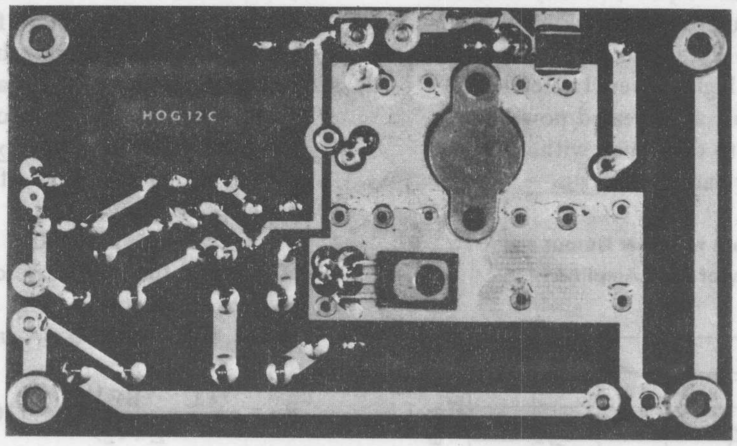
The MRF247 employed in this design is a version of the well-known MRF245, which has been reprocessed to improve the linear characteristics. It is a much larger device than the MRF240, resulting in lower input and output impedances. However, it employs internal base matching with a built-in MOS capacitor to bring the base impedance up to a level where external low loss matching networks can be realized.

In Figure 4 the dashed line encircles the specially designed T matching network, including the metal oxide capacitor X_4. X_1, X_2, and X_3 represent the bonding wires whose inductances can be varied by controlling the loop heights. This network will be part of the total matching network designed to match the transistor to function in a practical circuit.

FIGURE 4 — 75-Watt Amplifier Impedance Matching Network



Underside View of 75-W Amplifier



The internal matching still leaves the input impedance inductively reactive.

The MRF247 input impedance under forward biased conditions (100 mA) is 0.45 +J 0.85 ohms at 145 MHz, which translates to 2.06 +J 1.08 ohms in parallel form. A capacitive reactance of -J 1.08 ohms, converting to 1018 pF is required for C_3. The nominal value equivalent value, using Equation 1, is obtained as 450 pF.

Since the remaining resistive component of the base impedance (2.06 ohms) is only slightly higher than that of the MRF240, only minor changes in the input matching network are necessary. When $L_1 + L_2$ is fixed, and only their ratio can be varied, the resulting Q will be lower for the increased R_L. If only $L_1 + L_2$ is known, the Q can be calculated with Equation 5 as:

$$Q = \frac{[4X_T^2 + (R_S^2/R_L + X_T^2/R_L - R_S)\,4(R_S - R_L)]^{1/2} - 2X_T}{2(R_S - R_L)} \qquad (5)$$

where

$X_T = XL_1 + XL_2$ or $XL_4 + XL_5$

R_S = source impedance
R_L = load impedance
Reverse for output network calculations

Therefore,

$$Q = \frac{\sqrt{[26244 + (1214 + 3185 - 50)(192)]} - 162}{95.88}$$

$$Q = \frac{928 - 162}{95.88} = 7.99$$

$$Q = 8$$

where

$X_T = XL_1 + XL_2 = 81$ ohms

RS = 50 ohms

RL = 2.06 ohms

Then, with Equations 1, 2, 3 and 4, the values for L_1, L_2 and C_2 can be calculated as: L_1 = 71 nH, L_2 = 18 nH, C_2 = 63 pF (56 pF nearest standard). The position of C_2 will be approximately 1.6 inches from the transistor base. (See line inductance calculations in the discussion for the 35-watt amplifier.)

The measured output impedance of MRF247 is 0.65 +J 0.45 ohms, which is much lower and more reactive than the values shown for MRF240. The output matching must also be done with the existing total line inductance, ($L_4 + L_5$) and it can be expected that a higher factor of compromise in the output matching is evident regarding the network bandwidth.

The above impedance numbers convert to 0.96 ohms resistive and -J 1.39 ohms reactive in parallel form. Since -J 1.39 ohms = 790 pF, a nominal value of 400 pF (C_4) is required at the collector. To find the Q:

X_T = 94 ohms ($XL_4 + XL_5$)
R_S = 0.96 ohms
R_L = 50 ohms

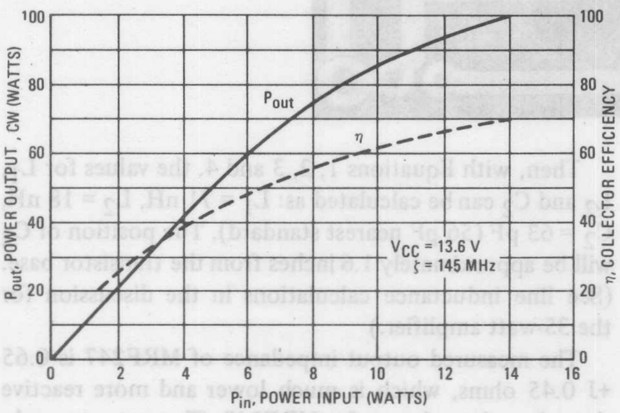
Then:

Q = 13.7 (Eq. 5), and:
L_4 = 13 ohms = 14 nH
L_5 = 81 ohms = 89 nH
C_5 = 11.8 ohms = 93 pF

A practical value of 82-91 pF can be used for C_5, and it should be located 1.1 inches along the output line from the collector, to give the above inductance values for L_4 and L_5.

Although the output Q is higher than the value calculated earlier for the 40 W unit, the total bandwidth of this version is increased as shown in Figure 7. The input matching network is usually dominant in determining the total bandwidth since the impedance transformation required is greater than the output requires, although the output circuit also has secondary effect. The internal matching elements of the device further make the total effective Q even lower than the calculated value, which in this case was 8. The higher output Q usually results in higher collector efficiency and better harmonic suppression, but at the same time the circulating RF currents will increase, resulting in higher overall circuit losses which is especially noticeable at increased power levels. These factors are difficult to determine without knowing all the internal transistor parameters.

FIGURE 5 – Power Input vs. Power Output and Collector Efficiency of 75-W Amplifier



CLASS AB BIASING AND OTHER CONSIDERATIONS

The biasing system, as seen in Figure 6, uses a forward biased transistor, Q_2, to provide a voltage source of 0.6 to 0.7 volts. When the collector is connected to the base, a second current path is formed, decreasing the base current according to the h_{FE}, and thus lowering the voltage drop across the base-emitter junction. In this manner the voltage drop can be adjusted by selecting the appropriate h_{FE} for Q_2. For the 2N5190 series h_{FE} is typically in the range of 80-100, although the minimum spec is 20-25.

Typical h_{FE}'s for the MRF240 and MRF247 are 50-60, and the worst case collector currents around 4A and 9A respectively. The minimum base currents required, $I_E(Q_2)$ are 80 mA and 180 mA (I_C/h_{FE}).

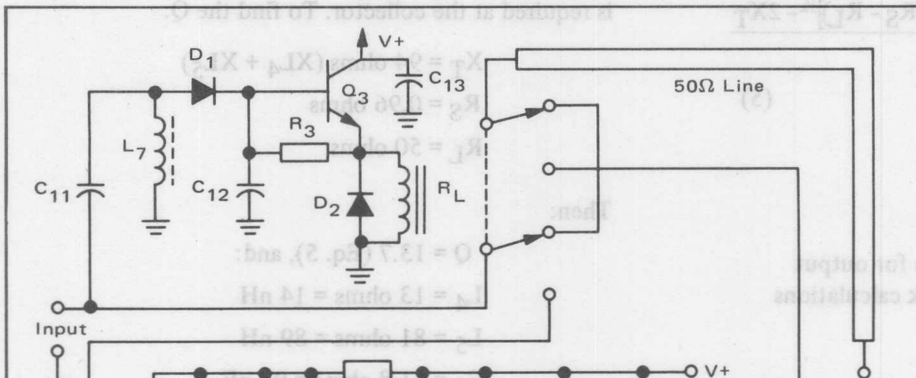
$$R_2 = \frac{V_{CC} - V_{BE}\,(Q_2)}{I_E\,(Q_2)} = 160 \text{ ohms and 75 ohms.}$$

The bias, which should not exceed 50 mA for MRF240 and 150 mA for MRF247, can be further adjusted by varying the value of R_2, but the minimum $I_E(Q_2)$ should be maintained.

It should be noted that since Q_2 is attached to the heat sink for temperature tracking purposes, its collector must be electrically isolated from the ground. The anodized surface of the heat sink is normally sufficient, or a separate insulating washer can be employed.

The 0.3 dB relay insertion loss mentioned earlier amounts to a VSWR of 1.7:1. However, the reflected power is only 0.2% (VSWR = 1.1:1) in a straight-through mode (receive), indicating that most of the relay losses are due to contact resistance and the dielectric insula-

FIGURE 6



PARTS LIST

- C_1, C_6, C_{13} – 1000 pF, 300 V dipped mica
- C_2 – 56 pF or 68 pF *, see text
- C_3, C_4 – 390 pF or 470 pF **, see text
- C_5 – 68 pF or 91 pF *, see text
- C_7 – 1000 pF ceramic chip or small disc
- C_8 – .01 µf ceramic disc
- C_9 – .1 µf ceramic disc
- C_{10} – 220 µf/6 V tantalum
- C_{11} – 3 pF dipped mica
- C_{12} – 10 µf/25 V tantalum
- R_1 – 10 ohms/¼ W carbon
- R_2 – 120 to 200 ohms/3 W
- R_3 – 3.3K/½ W carbon
- Q_1 – see text
- Q_2 – 2N5190
- Q_3 – 2N4401
- D_1 – 1N914 or 1N4148
- D_2 – 1N4001
- L_1, L_2, L_4, L_5 – see text
- L_3 – 12T AWG #16 enameled wire, ¼" I.D.
- L_6 – 10 µh, molded choke
- L_7 – .33–1.0 µh, molded choke
- RL – Arrow M, NF2-12 or equivalent– 'OMRON, LZN2-UA-DC12 or equivalent

Elecromotive Corp. (ELMENCO) *Type MCM01/010
Lauter Avenue **Type EMJ10
Florence, S.C. 29501

Standex Electronics *Type 3HS0006
P.O. Box 67 **Type J-101
Casey, Ill. 62420

tion resistance, rather than impedance mismatch.

Both amplifier designs may be employed in FM applications without modification. The bias networks may be omitted and L_6 connected to ground, which modifies the operation to Class C. The increased input impedance of the device operating class C results in increased input VSWR, but it will still remain less than 1.5:1 for the 145-150 MHz band.

FIGURE 7 — IMD vs. Power Output

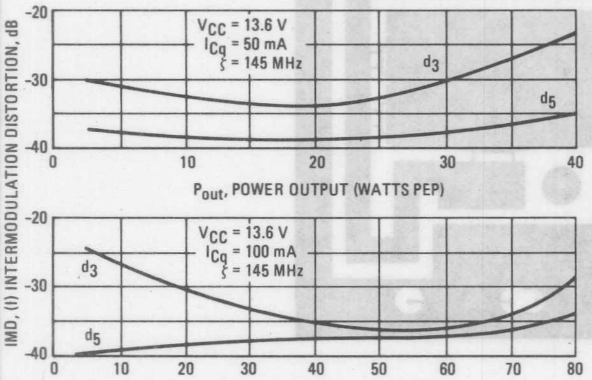
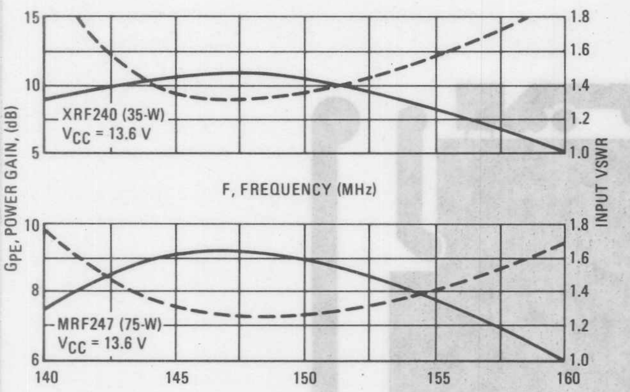


FIGURE 8 — Power Gain and VSWR vs. Frequency



The two amplifiers may be connected in cascade to provide a total power gain of around 20 dB; however, an attenuator of 4 to 6 dB is required between the two units to prevent overdrive of the MRF247. Since 10 to 20 watts will be dissipated in the attenuator, it cannot be built from discrete resistors. Most convenient, size and costwise, are the thin film attenuators such as those manufactured by Pyrofilm.

The COR circuit requires 400 to 500 mW for the relay to switch. At this drive level, without the attenuator, the second amplifier would already produce full power output.

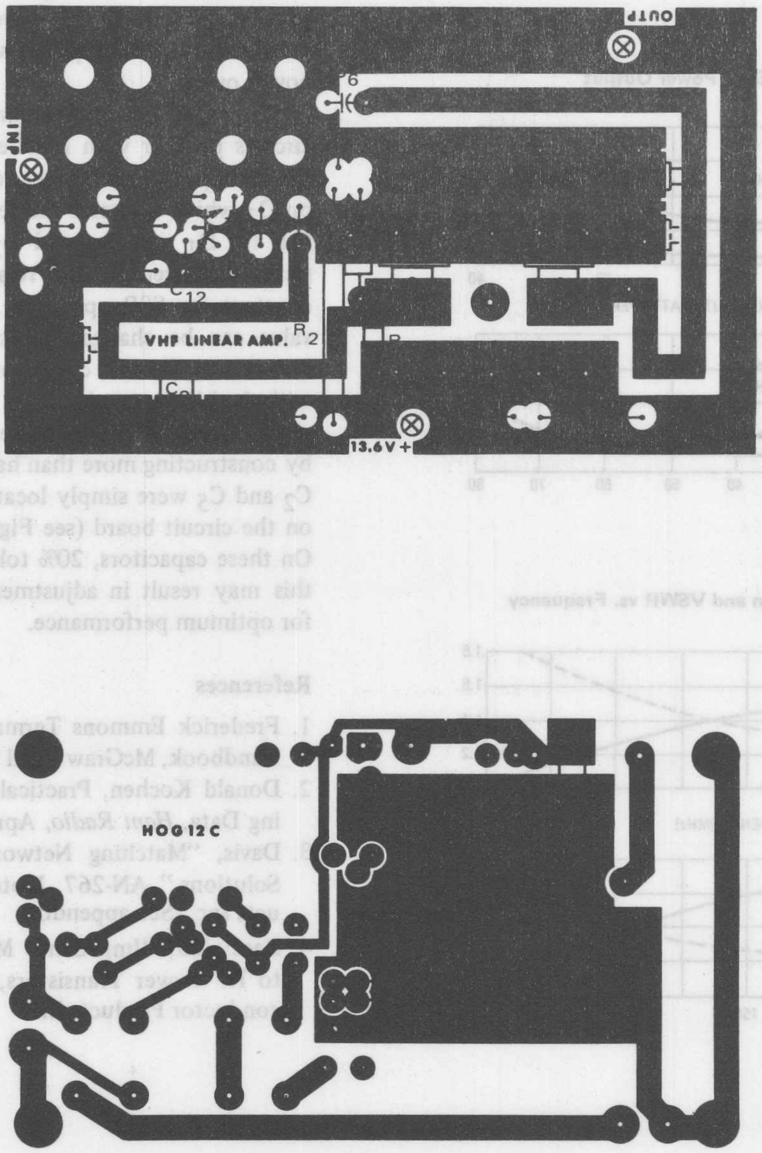
The COR (Figure 5) incorporates one of the standard circuits popular with mobile add-on amplifiers. Part of the RF input signal is being rectified by D_1. The dc turns on Q_3 which activates the relay. L_7 and R_3 provide the bias for D_1 and Q_3, and D_2 suppresses inductive transients produced by the relay coil inductance. A time constant for SSB operation is provided by C_{12}, whose value can be changed according to individual requirements. For FM this capacitor can also be omitted along with the bias network.

The repeatability of these amplifiers has been proven by constructing more than half a dozen units. Capacitors C_2 and C_5 were simply located within the marked areas on the circuit board (see Figure 9 and the photograph). On these capacitors, 20% tolerances can be allowed, but this may result in adjustments of each individual unit for optimum performance.

References

1. Frederick Emmons Terman, Sc. D. Radio Engineers Handbook, McGraw - Hill Co., Inc., 1943
2. Donald Kochen, Practical VHF and UHF Coil Winding Data, *Ham Radio,* April 1971
3. Davis, "Matching Network Design with Computer Solutions," AN-267, Motorola Semiconductor Products Inc. (See appendix).
4. Becciolini, "Impedance Matching Networks Applied to RF Power Transistors," AN-721, Motorola Semiconductor Products Inc.

FIGURE 9 – Component Layout Diagram of 35W and 75W Amplifiers

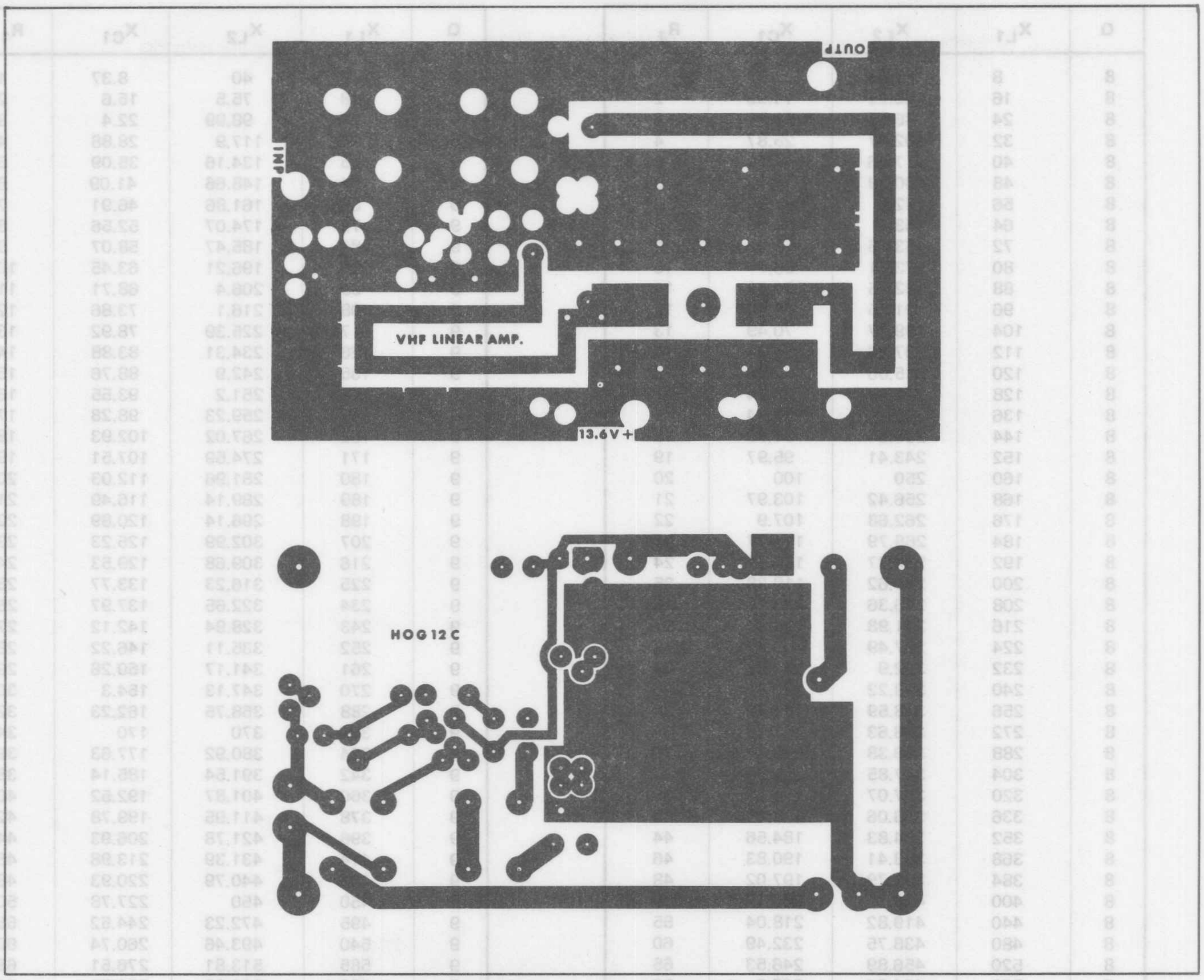


× DENOTES FEED THROUGH EYELETS OR PLATED THROUGH HOLES

⊗ DENOTES TERMINAL PINS

◎ DENOTES BOARD STANDOFFS

FIGURE 10 – Printed Circuit Board Layout



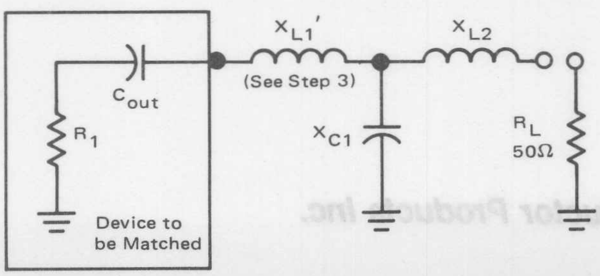
APPENDIX

This information was originally published in Motorola Application Note AN-267, "Matching Network Designs with Computer Solutions."

NETWORK D

The following is a computer solution for an RF "Tee" matching network.

Tuning is accomplished by using a variable capacitor for C_1. Variable matching may also be accomplished by increasing X_{L2} and adding an equal amount of X_C in series in the form of a variable capacitor.



TO DESIGN A NETWORK USING THE TABLES

1. Define Q, in column one, as X_{L1}/R_1.
2. For an R_1 to be matched and a desired Q, read the reactances of the network components from the charts.
3. X_{L1}' is equal to the quantity X_{L1} obtained from the tables plus $|X_{C_{out}}|$.
4. This completes the network.

Q	X_{L1}	X_{L2}	X_{C1}	R_1
8	8	27.39	7.6	1
8	16	63.25	14.03	2
8	24	85.15	20.1	3
8	32	102.47	25.87	4
8	40	117.26	31.42	5
8	48	130.38	36.77	6
8	56	142.3	41.95	7
8	64	153.3	46.99	8
8	72	163.55	51.9	9
8	80	173.21	56.7	10
8	88	182.35	61.39	11
8	96	191.05	65.98	12
8	104	199.37	70.49	13
8	112	207.36	74.91	14
8	120	215.06	79.26	15
8	128	222.49	83.54	16
8	136	229.67	87.74	17
8	144	236.64	91.89	18
8	152	243.41	95.97	19
8	160	250	100	20
8	168	256.42	103.97	21
8	176	262.68	107.9	22
8	184	268.79	111.77	23
8	192	274.77	115.59	24
8	200	280.62	119.38	25
8	208	286.36	123.11	26
8	216	291.98	126.81	27
8	224	297.49	130.47	28
8	232	302.9	134.09	29
8	240	308.22	137.67	30
8	256	318.59	144.73	32
8	272	328.63	151.65	34
8	288	338.38	158.46	36
8	304	347.85	165.14	38
8	320	357.07	171.71	40
8	336	366.06	178.18	42
8	352	374.83	184.56	44
8	368	383.41	190.83	46
8	384	391.79	197.02	48
8	400	400	203.13	50
8	440	419.82	218.04	55
8	480	438.75	232.49	60
8	520	456.89	246.53	65
8	560	474.34	260.2	70
8	600	491.17	273.52	75
8	640	507.44	286.52	80
8	680	523.21	299.23	85
8	720	538.52	311.66	90
8	760	553.4	323.84	95
8	800	567.89	335.78	100
8	1000	635.41	392.36	125
8	1200	696.42	444.63	150
8	1400	752.5	493.49	175
8	1600	804.67	539.57	200
8	1800	853.67	583.29	225
8	2000	900	625	250
8	2200	944.06	664.96	275
8	2400	986.15	703.38	300

Q	X_{L1}	X_{L2}	X_{C1}	R_1
9	9	40	8.37	1
9	18	75.5	15.6	2
9	27	98.99	22.4	3
9	36	117.9	28.88	4
9	45	134.16	35.09	5
9	54	148.66	41.09	6
9	63	161.86	46.91	7
9	72	174.07	52.56	8
9	81	185.47	58.07	9
9	90	196.21	63.45	10
9	99	206.4	68.71	11
9	108	216.1	73.86	12
9	117	225.39	78.92	13
9	126	234.31	83.88	14
9	135	242.9	88.76	15
9	144	251.2	93.55	16
9	153	259.23	98.28	17
9	162	267.02	102.93	18
9	171	274.59	107.51	19
9	180	281.96	112.03	20
9	189	289.14	116.49	21
9	198	296.14	120.89	22
9	207	302.99	125.23	23
9	216	309.68	129.53	24
9	225	316.23	133.77	25
9	234	322.65	137.97	26
9	243	328.94	142.12	27
9	252	335.11	146.22	28
9	261	341.17	150.28	29
9	270	347.13	154.3	30
9	288	358.75	162.23	32
9	306	370	170	34
9	324	380.92	177.63	36
9	342	391.54	185.14	38
9	360	401.87	192.52	40
9	378	411.95	199.78	42
9	396	421.78	206.93	44
9	414	431.39	213.98	46
9	432	440.79	220.93	48
9	450	450	227.78	50
9	495	472.23	244.52	55
9	540	493.46	260.74	60
9	585	513.81	276.51	65
9	630	533.39	291.85	70
9	675	552.27	306.8	75
9	720	570.53	321.4	80
9	765	588.22	335.67	85
9	810	605.39	349.63	90
9	855	622.09	363.31	95
9	900	638.36	376.71	100
9	1125	714.14	440.24	125
9	1350	782.62	498.94	150
9	1575	845.58	553.81	175
9	1800	904.16	605.54	200
9	2025	959.17	654.64	225
9	2250	1011.19	701.48	250
9	2475	1060.66	746.36	275
9	2700	1107.93	789.51	300