SEPTEMBER 1959

APPLICATION NOTES

VHF TRANSISTOR POWER STAGES

Transistorized vhf power oscillators and amplifiers can be designed without difficulty to provide reasonable assurance of optimum performance. The essential requirements are proper load impedance, coils and capacitors having sufficiently high Q, a proper feedback path for the oscillator, and a physical layout compatible with the frequency.

The oscillators and amplifiers to be described are power stages. A large voltage swing at the collector is assumed.

Common-base operation is preferred for the following reasons. The transistor can generally be operated at a higher d-c collector voltage because the common-base breakdown voltage is higher. Experimental results show that both output and efficiency are somewhat higher than in a common-emitter stage. The gain of a common-base amplifier may be lower, but this is usually not significant in a power stage. In some instances, the common-base stage may have more gain. Finally, the feedback path for an oscillator is more easily established in a common-base connection.

DESIGN ANALYSIS

The load impedance, R_L, at the collector is dictated by the output power and the collector voltage. Its maximum value is given by:

$$R_L = \frac{E_{bb}^2}{2 P_O} \qquad (1)$$

where E_{bb} is the battery voltage and P_O is the output power. This assumes that the peak a-c voltage is equal to the battery voltage. It is often impossible to swing the collector voltage this much, particularly at higher frequencies. In this case, R_L will be less (often considerably less) than the value given above.

Since predicting large-signal behavior of a transistor at vhf is difficult, a load-matching network is required to obtain the desired performance. The network shown in Fig. 1 is frequently used for impedance matching. The range of load impedance may be determined with relative ease and a wide variety of load conditions may be accommodated.

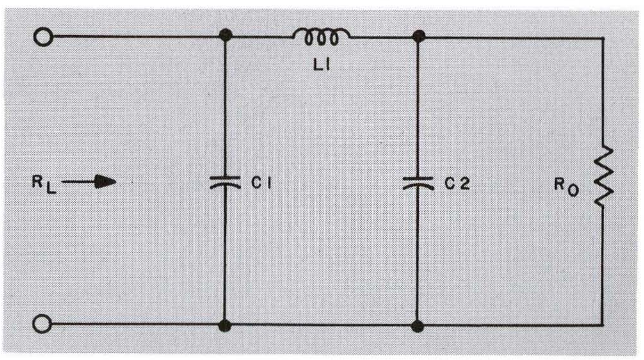

Fig. 1. Load-matching π network

Capacitors C1 and C2 can be 1.8-13 $\mu\mu$f (E. F. Johnson Type U). Losses in C3 are relatively unimportant since it is shunted by the 50-ohm load.

400—MC OSCILLATOR

While the π matching networks are still applicable at 400 mc, other arrangements also are attractive. For example, frequently we may use a small section of transmission line with a short-circuit termination (usually shaped like a hairpin) to provide an equivalent inductance with high Q, low loss, and convenient size.

The problem of a heat sink for a power stage is neatly solved by inserting the transistor into the end of the collector line. The loop reactance and size are set by the minimum Q desired at the operating power level and by the transistor and circuit capacitance. Loading adjustment is possible with a small variable capacitor.

Figure 5 shows the dimensions of a 400-mc oscillator. From 20 to 50 mw may be obtained from typical 2N1141 transistors with 20 volts from collector to base and 10-ma emitter current.

The capacitors can be 1.5-9 $\mu\mu$f (Johnson Type U). The emitter line is not strictly necessary for an oscillator.

This circuit may be used as an amplifier if an input is coupled to the emitter line in a manner similar to the load connection.

AVOIDING OSCILLATION IN TUNED AMPLIFIERS

The easiest way to avoid oscillation is to load the output and input sufficiently. At frequencies up to a few hundred megacycles, this is entirely satisfactory and is consistent with the foregoing designs. It begins to be difficult at higher frequencies because of the resultant loss in gain.

Neutralization is a possible alternative, but it is difficult at vhf. Also, neutralization may introduce more problems than it solves.

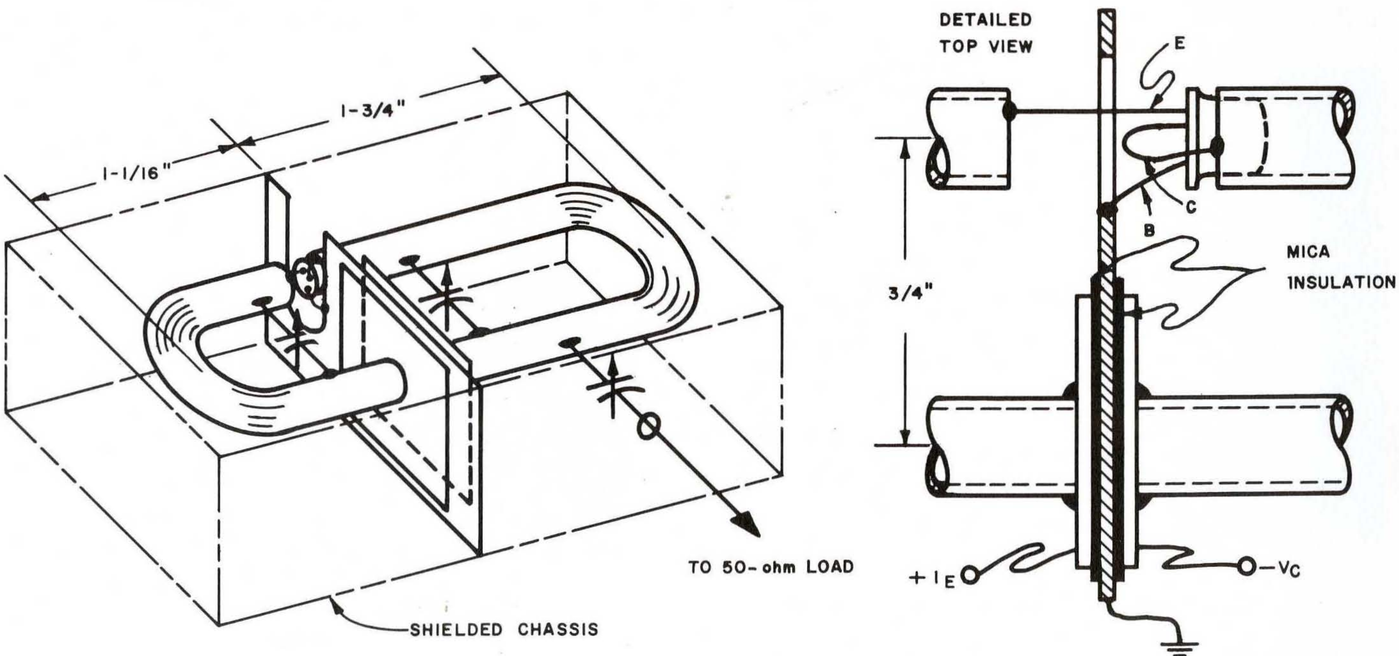

Fig. 5. 400-mc Oscillator Layout

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SEMICONDUCTOR-COMPONENTS DIVISION
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The maximum value of C2 is determined from Eq.(5).

$$X_{C2(max)} = \frac{141}{1 + \sqrt{\frac{1600}{50} - \left(\frac{141}{50}\right)^2}} = \frac{141}{1 + 4.9}$$

$$= 23.9 \text{ ohms}$$

$$C_{2(max)} = 33 \ \mu\mu f$$

C1 is selected to tune the network to resonance at 200 mc. In this example, the value of C1 is about 5 $\mu\mu$f.

The complete network is shown in Fig. 2.

The minimum Q (to a first approximation) is:

$$Q_{(min)} \cong \frac{141}{50} \cong 3$$

Although this circuit Q is too low to be generally suitable as an output matching network where harmonic rejection is required, it will be adequate for investigating transistor performance.

INPUT MATCHING SECTION

While a π network is possible, the Q is very low for a low-impedance source. Therefore a resonant matching system is best. A simple and effective scheme is shown in Fig. 3.

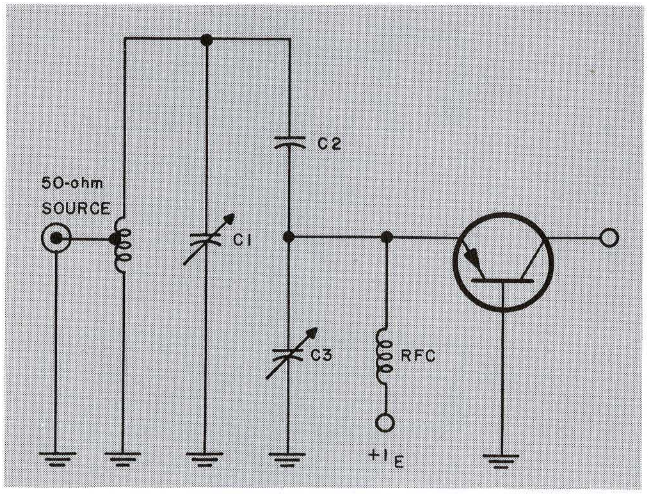

Fig. 3. Resonant input system for vhf amplifier

At high frequencies, the input resistance of the transistor is approximately equal to r_b'. C2 is selected for the minimum Q desired:

$$X_{C2} \cong 2Q_{(min)} \ r_b'$$

The factor 2 is included to account for the loading of the source at impedance match. L1 and C1 are selected to resonate with C2 at the operating frequency. C3 will serve as an impedance matching capacitor. Typical values for X_{C3} might range from 1/3 to 3 times r_b'. The input tap on L1 should be located experimentally to provide the maximum available power at the transistor input.

OSCILLATOR DESIGN

Adding a feedback circuit to the collector matching section will complete the oscillator design.

The common-base circuit is well suited for oscillation since the emitter current is nearly in phase with the collector current. Direct collector-emitter feedback through the internal transistor capacitance (at high frequencies) or with an externally connected collector-emitter capacitor (at low frequencies) will always produce oscillation.

Oscillation occurs at that point on the resonant-circuit impedance curve where the phase of the feedback is correct. An adjustment of the magnitude of the feedback voltage is needed. This may be a small variable capacitor from emitter to ground (at high frequencies) or from collector to emitter (at low frequencies).

The circuit of a typical 200-mc oscillator is shown in Fig. 4. The 2N1141 will give 60 to 100 mw in this circuit with 20 volts from collector to base and 10-ma emitter current.

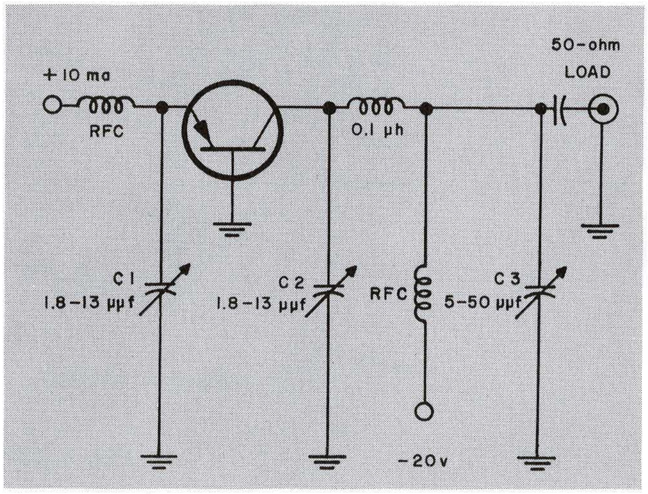

Fig. 4. 200-mc Oscillator

The maximum value for R_L is determined by Eq.(1). R_O, the output load resistance, is known (usually 50 ohms). The loaded Q of the entire circuit is subject to certain limitations. Since a low value of loaded Q permits excessive transmission of harmonic frequencies, it should not be less than 5, and preferably not less than 7. Maximum Q is limited by the realizable Q of the circuit elements; therefore, C1 must be a high-Q capacitor. The Q of L1 is usually the limiting factor and the loaded circuit Q should not exceed 10 percent of the inductor Q so that not more than about 10 percent of the available power will be lost in L1. If inductor size is not too small, Q's of 200 to 300 are possible.

A straightforward analysis of Fig. 1 gives the following expression for the ratio of input resistance, R_L, to output resistance, R_O:

$$\frac{R_L}{R_O} = \left(1 - \frac{X_L}{X_{C2}}\right)^2 + \left(\frac{X_L}{R_O}\right)^2 \quad . \; . \; . \; . \; . \quad (2)$$

$R_{L(min)}$, the minimum value of R_L, and the minimum ratio of $\frac{R_L}{R_O}$ occurs when $X_L = X_{C2}$. At this point:

$$X_L = \sqrt{R_O R_{L(min)}} \quad . \; . \; . \; . \; . \; . \; . \quad (3)$$

$$X_{C2(min)} = X_L \quad . \; . \; . \; . \; . \; . \; . \; . \; . \; . \; . \quad (4)$$

This is the minimum value of C2. R_L, the input resistance to the network, will rise as C2 is increased.

The maximum value of C2 is determined by the maximum value of R_L, designated $R_{L(max)}$.

$$X_{C2(max)} = \frac{X_L}{1 + \sqrt{\frac{R_{L(max)}}{R_O} - \left(\frac{X_L}{R_O}\right)^2}} \quad . \; . \; . \; . \quad (5)$$

When a minimum loaded Q is specified, the minimum values of X_L and R_L are also specified. This sets a limit on the maximum output power using a single π network with a given R_O. If greater power is needed, other matching schemes may be used or a second π network added. The second network is calculated in the same manner as the first but with the ends reversed to transform R_O to a lower impedance.

The approximate minimum values for X_L and R_L are given by:

$$X_{L(min)} \cong Q_{(min)} R_O \quad . \; . \; . \; . \; . \; . \; . \; . \; . \quad (6)$$

$$R_{L(min)} \cong [Q_{(min)}]^2 R_O \quad . \; . \; . \; . \; . \; . \; . \; . \; . \quad (7)$$

200 — MC NETWORK

A typical example may clarify the design procedure. Given a transistor output stage that is operating as follows, design a π matching network. Let $V_{CB} = 20$ volts, $R_O = 50$ ohms, f = 200 mc, and $P_O = 0.2$ watt.

From Eq.(1),

$$R_L = \frac{20^2}{2 \times 0.2} = 1000 \text{ ohms}$$

Since a full 20-volt swing is unlikely, design the π matching section to provide loads from 400 to 1600 ohms. Loads higher than 1000 ohms will not permit 0.2 watt to be developed, but may help realize maximum output from marginal transistors. From Eq.(3),

$$X_L = \sqrt{R_O R_{L(min)}} = \sqrt{50 \times 400} = 141 \text{ ohms}$$

$$L \cong 0.125 \; \mu h$$

$$X_{C2(min)} = X_L = 141 \text{ ohms}$$

$$C2_{(min)} \cong 5 \; \mu\mu f$$

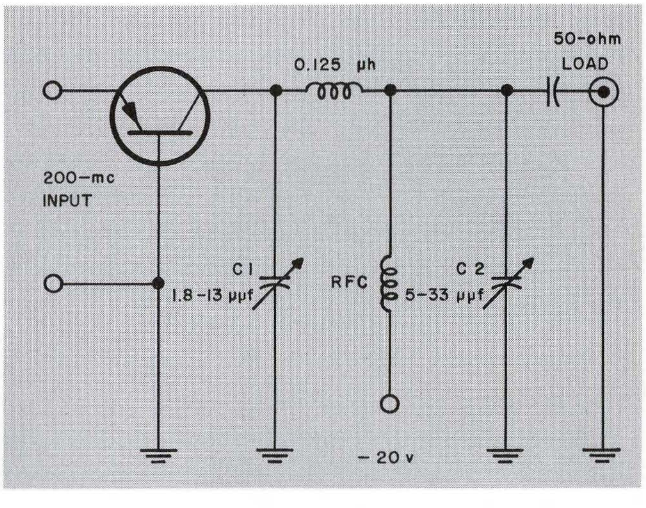

Fig. 2. Output π network for 200 mc