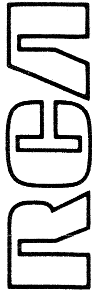
RCA

RF Power Transistor Manual



RF Power Transistor Manual

THIS Manual provides detailed information on the use of high-frequency power transistors in a variety of power-circuit applications at frequencies from the vhf range to well within the microwave region. It includes a general review of the basic requirements of all power transistors and explanations of the special features required of transistors that are especially designed for high-frequency power applications. The special requirements, device selection criteria, and design techniques are described for high-frequency transistor power-amplifier, oscillator, and frequency-multiplier circuits. Practical examples are shown and performance data are given for such circuits intended for use in single-sideband systems, military communications systems, commercial-aircraft radio, mobile and marine radio, community-antenna television, microwave relay links, and air-traffic-control systems.

THIS Manual explains the characteristics and capabilities of commercially available rf and microwave power transistors and describes current design practices employed in power-circuit applications of these devices. It will be found useful by circuit and system designers, educators, students, and others interested in rf and microwave power transistors and circuits.

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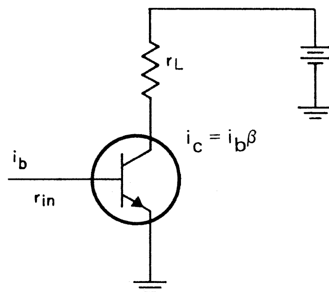
General Considerations for Power Transistors

THE performance of any power transistor in electronic equipment depends on many highly diverse design factors. In this section, the basic physical parameters that determine transistor power gain are described, the upper-frequency limitations of conventional power-transistor designs are defined, and the thermal considerations and maximum ratings that establish the power-output capabilities and electrical and environmental limitations of power transistors are defined. The basic considerations involved in the design of power-transistor structures and geometries to effect the optimum compromise among these factors are also discussed. This basic review of the general theory of power transistors should be helpful to an understanding of the special design features and tradeoffs involved in the development of transistors for high-frequency power applications.

BASIC GAIN FACTORS

Power gain in transistor circuits is usually obtained by use of a small control signal to produce larger signal variations in the output current. The gain parameter most often specified is the current gain (β) from the base to the collector. The power gain of a transistor operated in a common-emitter configuration is equal to the square of the current gain β times the ratio of the load resistance r_L to the input resistance r_{in}, as indicated in Fig. 1.

Although the input resistance r_{in} affects the power gain, as shown by the equations given in Fig. 1, this parameter is not usually specified directly in the published data on transistors because



INPUT CURRENT $= i_b$

INPUT VOLTAGE $= i_b r_{in}$

OUTPUT CURRENT $= i_c = i_b\beta$

OUTPUT VOLTAGE $= i_c r_L = i_b \beta r_L$

INPUT POWER $= i_b^2 r_{in}$

OUTPUT POWER $= i_c^2 r_L = i_b^2 \beta^2 r_L$

POWER GAIN $=$ *power output/power input*

$= i_b^2\beta^2 r_L / i_b^2 r_{in}$

$= \beta^2 r_L / r_{in}$

Figure 1. Test circuit and simplified power-gain calculation for a transistor operated in a common-emitter configuration.

of the large number of components of which it is comprised. In general, the input impedance is expressed as a maximum base-to-emitter voltage V_{be} under specified input-current conditions.

The performance of power transistors is significantly affected by the high current densities and large collector fields encountered in the operation of these devices. Operation under such conditions results in three basic effects, referred to as base conductivity modulation, current crowding, and base widening, which adversely affect the power gain and, consequently, the power-output capabilities of the transistors.

In general, high-frequency power transistors employ a diffused base, and base conductivity modulation is not significant in these devices. This factor, therefore, is not discussed further in this Manual.

Current-Gain Parameters

The current gain (or current transfer ratio) of a transistor is expressed by many symbols; the following are some of the most common, together with their particular shades of meaning:

1. beta (β)—general term for current gain from base to collector (i.e., common-emitter current gain)
2. alpha (α)—general term for current gain from emitter to collector (i.e., common-base current gain)
3. h_{fe}—ac gain from base to collector (i.e., ac beta)
4. h_{FE}—dc gain from base to collector. (i.e., dc beta)

Common-base current gain, a, is the ratio of collector current to emitter current (i.e., $a = I_C/I_E$). Although a is slightly less than unity, circuit gain is realized as a result of the large differences of input (emitter-base) and output (collector-base) impedances. The input impedance is small because the emitter-base junction is forward-biased, and the output impedance is large because the collector-base junction is reverse-biased.

Common-emitter current gain, β, is the ratio of collector current to base current (i.e., $\beta = I_C/I_B$). Useful values of β are normally greater than ten.

The current-gain parameters α and β are determined by three basic transistor parameters (emitter efficiency γ, collector efficiency α^*, and base transport factor β_O) that are directly dependent upon the physical properties of the device. The following equations express the dependence of the current-gain parameters α and β on the three physical parameters:

$$\alpha = \gamma\beta_o\alpha^* \quad (1)$$

$$\beta = \frac{\gamma\beta_o\alpha^*}{1 - (\gamma\beta_o\alpha^*)} \quad (2)$$

The collector efficiency α^* is generally very near unity and this value is usually assumed for this parameter in the calculations for α and β.

Emitter Efficiency—In a forward-biased n-p-n transistor, electrons from the emitter diffuse into the base, and holes from the base diffuse into the emitter. The total emitter current I_E is the sum of the hole-current component I_p and the electron-current component I_n and, therefore, may be expressed as follows:

$$I_E = I_p + I_n \qquad (3)$$

The potential collector current, I_C, is the difference between the drift currents and is given by

$$I_C = I_n - I_p \qquad (4)$$

The holes that diffuse from the base into the emitter originate in the base dc supply and add to the total base current. This hole current I_p, however, does not contribute to the collector current and, in effect, represents a loss in current gain that is directly attributable to poor emitter injection efficiency. The loss in current gain can be held to a minimum if the resistivity of the base is made much greater than that of the emitter so that the number of free holes in the base available to diffuse into the emitter is substantially smaller than the number of free electrons in the emitter available to diffuse into the base.

Base Transport Factor—If high emitter efficiency is assumed, the electrons injected from the emitter into the base diffuse to the collector junction. Some of these electrons, however, recombine with free holes in the base and, in effect, are annihilated. Base current must flow to replenish the free electrons used in the recombination process so that the emitter-to-base forward bias is maintained. In other words, charge neutrality must prevail.

The base transport factor β_0 is an indication of the extent of recombination that takes place in the base region of a transistor. For a high value of the base transport factor β_0, the lifetime of holes in the base (a function of the property of the material) must be long, or the time necessary for the electrons to reach the collector must be short. Any reduction in the time required for the electrons to reach the collector requires a decrease in base width or an increase in the accelerating field used to speed the holes through.

The value of the base transport factor should be in the order of 0.98 for the transistor to provide useful gain.

Current Crowding

The effect of current crowding causes a reduction in the usable gain of a power transistor because of the electrical reduction in the injecting emitter area. This reduction in area results because a diminished forward bias is induced in the central portions of the emitter by the transverse voltage drops in the base. The voltage drops are a result of the flow of base current (through the relatively high base resistance) required to replenish base recombination current and maintain charge neutrality in the base. The current-crowding effect causes the reduced-biased center of the emitter to stop injecting charge carriers, and the edge becomes the primary injecting area. Simply stated, the emitter edge is biased at a higher emitter-base forward voltage than the center of the emitter.

Physical factors which reduce current crowding are large emitter-periphery-to-area ratios, low base resistance, and wide base widths. (The latter two factors affect the magnitude of the transverse base voltage drop.) Fig. 2 shows a cross section of a typical power transistor under simulated current-crowding conditions.

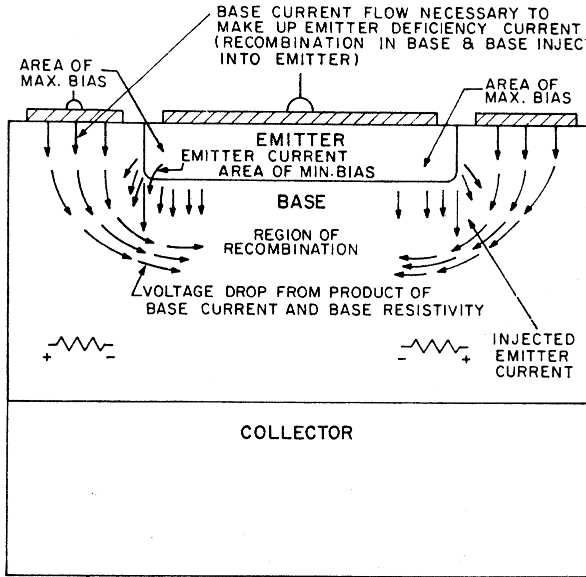


Figure 2. Cross section of a typical n-p-n power transistor under current-crowding conditions.

Base Widening

Base widening is the effect of the local widening of the electrical base in the areas directly underlying the high injecting regions of the emitter. This condition results when the concentration of injected charge from the emitter exceeds the background doping level of the collector. Fig. 3 shows a cross section of a

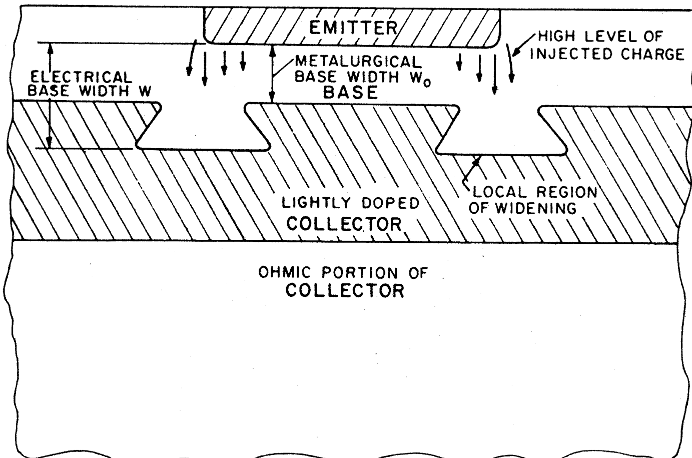


Figure 3. Cross section of a typical power transistor under base-widening conditions.

typical power transistor under simulated base-widening conditions.

As the level of the injected charge exceeds the doped collector impurity level during the transit of injected charges through the collector to the ohmic region, the injected charges dynamically compensate the uncompensated fixed charges (ions) in the collector side of the depletion layer. This compensation makes that local area electrically equivalent to a base region, as shown in Fig. 4. Complete base-widening analysis requires extensive computation, and the aid of a computer is mandatory for such analyses. Qualitatively, however, transistors with lightly doped collectors suffer more degradation in gain and saturation than transistors with heavily doped collectors, provided that the current density is the same for both types of transistors. The mechanism for reduced gain results from effects on the base transport factor caused by the widened or extended base width and charge pile-up in the vicinity of the collector-base junction.

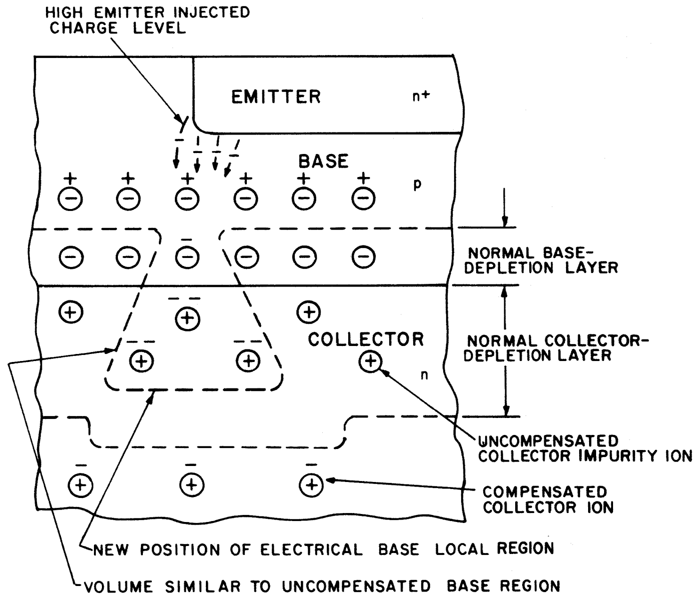


Figure 4. Effect of fixed-impurity compensation (ion compensation).

CUTOFF FREQUENCIES

For all transistors, there is a frequency f at which the output signal cannot properly follow the input signal because of time delays in the transport of the charge carriers. The three principal cut-off frequencies, shown in Fig. 5, may be defined as follows:

1. The **base cut-off frequency** $f_{\alpha b}$ is that frequency at which

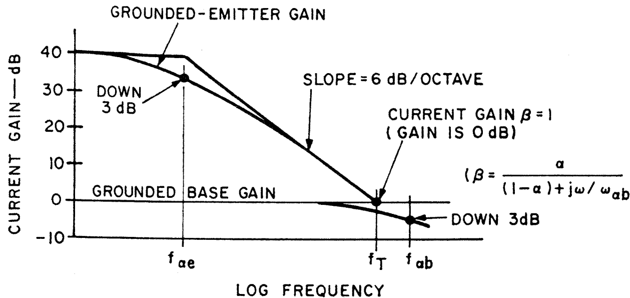


Figure 5. Cut-off frequencies.

alpha (α) is down 3 dB from the low-frequency value.

2. The **emitter cut-off frequency** $f_{\alpha e}$ is that frequency at which beta (β) is down 3 dB from the low-frequency value.

3. the frequency f_T is that frequency at which beta theoretically decreases to unity (i.e., 0-dB gain) with a theoretical 6-dB-per-octave fall off. This term, which is a useful figure of merit for transistors, is referred to as the **gain-bandwidth product.**

The frequency f_T is related to the time delays in a transistor by the following expression:

$$f_T \approx \frac{1}{2\pi \Sigma t_d} \quad (5)$$

where Σt_d is the sum of the emitter-delay time constant t_e, the base transit time t_b, the collector depletion-layer transit time t_{xm}, and the collector-delay time constant t_c.

The gain-bandwidth product f_T is the term that is generally used to indicate the high-frequency capability of a transistor. Other parameters that critically affect high-frequency performance are the capacitance or resistance which shunts the load and the input impedance, the effect of which is shown by the equations given in Fig. 1.

The specification of all the characteristics which affect high-frequency performance is so complex that often a manufacturer does not specify all the parameters, but instead specifies transistor performance in an rf-amplifier circuit. This information is very useful when the transistor is operated under conditions very similar to those of the test circuit, but is difficult to apply when the transistor is used in a widely different application. Some manufacturers also specify transistor performance characteristics as a function of frequency, which alleviates these problems.

THERMAL CONSIDERATIONS

The maximum allowable power dissipation in a solid-state device is limited by the temperature of the semiconductor pellet (i.e., the junction temperature). An important factor that assures that the junction temperature remains below the specified maximum value is the ability of the associated thermal circuit to conduct heat away from the device. For this reason, solid-state power devices should be mounted on a good

thermal base (usually copper), and means should be provided for the efficient transfer of heat from this base to the surrounding environment.

When current flows through a solid-state device, power is dissipated in the semiconductor pellet that is equal to the product of the voltage across the junction and the current through it. As a result, the temperature of the pellet increases. The amount of the increase in temperature depends on the power level and how fast the heat can flow away from the junction through the device structure to the case and the ambient atmosphere. The rate of heat removal depends primarily upon the thermal resistance and capacitance of the materials involved. The temperature of the pellet rises until the rate of heat generated by the power dissipation is equal to the rate of heat flow away from the junction; i.e., until thermal equilibrium has been established.

Thermal resistance can be compared to electrical resistance. Just as electrical resistance is the extent to which a material resists the flow of electricity, thermal resistance is the extent to which a material resists the flow of heat. A material that has a low thermal resistance is said to be a good thermal conductor. In general, materials which are good electrical conductors are good thermal conductors, and vice versa.

The methods of rating solid-state power devices under steady-state conditions are indicated by the following definition of thermal resistance: The thermal resistance of a solid-state device is the ratio of the temperature drop to the heat generated through internal power dissipation under steady-state conditions; the temperature drop is measured between the region of heat generation and some reference point.

The over-all thermal resistance of an assembled device is usually expressed as the rise in junction temperature above the case temperature per unit of power dissipated in the device. This information, together with the maximum junction-temperature rating, enables the user to determine the maximum power level at which the device can be safely operated for a given case temperature. Subtraction of the case temperature from the maximum junction temperature indicates the allowable internal temperature rise. If this value is divided by the specified thermal resistance of the device, the maximum allowable power dissipation is determined.

It should be noted that thermal resistance is defined for steady-state conditions. If a uniform temperature over the entire semiconductor junction is assumed, the power dissipation required to raise the junction temperature to a predetermined value, consistent with reliable operation, can be determined. Under conditions of intermittent or switching loads, however, such a design is unnecessarily conservative and expensive. For such conditions, the effect of **thermal capacitance** should also be considered.

When a solid-state device is mounted in free air, without a heat sink, the steady-state thermal circuit is defined by the junction-to-free-air thermal resistance given in the published data on the device. Thermal considerations require that there be a free

flow of air around the device and that the power dissipation be maintained below that which would cause the junction temperature to rise above the maximum rating. When the device is mounted on a heat sink, however, care must be taken to assure that all portions of the thermal circuit are considered.

MAXIMUM RATINGS

All solid-state devices undergo irreversible changes if their temperature is increased beyond some critical limit. A number of ratings are given for power transistors, therefore, to assure that this critical temperature limit will not be exceeded on even a very small part of the silicon chip. The ratings for power transistors normally specify the maximum voltages, maximum current, maximum and minimum operating and storage temperatures, and maximum power dissipation that the transistor can safely withstand.

Voltage Ratings

Maximum voltage ratings are normally given for both the collector and the emitter junctions of a transistor. A V_{BEO} rating, which indicates the maximum base-to-emitter voltage with the collector open, is usually specified. The collector-junction voltage capability is usually given with respect to the emitter, which is used as the common terminal in most transistor circuits. This capability may be expressed in several ways. A V_{CEO} rating specifies the maximum collector-to-emitter voltage with the base open; a V_{CER} rating for this voltage implies that the base is returned to the emitter through a specified resistor; a V_{CES} rating gives the maximum voltage when the base is shorted to the emitter; and a V_{CEV} rating indicates the maximum voltage when the base is reverse-biased with respect to the emitter by a specified voltage. A V_{CEX} rating may also be given to indicate the maximum collector-to-emitter voltage when a resistor and voltage are both connected between base and emitter.

If a maximum voltage rating is exceeded, the transistor may "break down" and pass current in the reverse direction. The breakdown across the junction is usually not uniform, and the current may be localized in one or more small areas. The small area becomes overheated unless the current is limited to a low value, and the transistor may then be destroyed.

The collector-to-base or emitter-to-base breakdown (avalanche) voltage is a function of the resistivity or impurity doping concentration at the junction of the transistor and of the characteristics of the circuit in which the transistor is used. When there is a breakdown at the junction, a sudden rise in current (an "avalanche") occurs. In an abruptly changing junction, called a step junction, the avalanche voltage is inversely proportional to the impurity concentration. In a slowly changing junction, called a graded junction, the avalanche voltage is dependent upon the rate of change of the impurity concentration (grade constant) at the physical junction. Fig. 6 shows the two types of junction breakdowns. The basic transistor voltage-breakdown mechanisms and their relationship to external circuits are the basis for the various types of voltage ratings

used by transistor manufacturers.

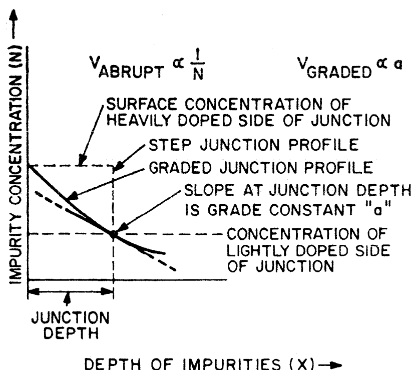


Figure 6. Step-junction and graded-junction breakdown.

The collector voltage can be limited below its avalanche breakdown value if the depletion layer (space-charge region) associated with the applied collector voltage expands through the thin base width and contacts the emitter junction. The doping in the base (under the emitter) and the base width in relation to the magnitude of applied voltage govern whether punch-through occurs before avalanche. Higher doping concentrations and wider bases increase punch-through voltage V_{PT} in accordance with the following relationship:

$$V_{PT} = \frac{qNW^2}{2kE_o} \qquad (6)$$

where q is the electronic charge, N is the doping-level concentration in the base, W is the base width, k is the dielectric constant, and E_o is the permittivity of free space (kE_o is approximately 1×10^{-12} farad per centimeter for silicon).

Current and Temperature Ratings

The physical mechanisms related to basic transistor action are temperature-sensitive. If the bias is not temperature-compensated, the transistor may develop a regenerative condition, known as **thermal runaway,** in which the thermally generated carrier concentration approaches the impurity carrier concentration. [Experimental data for silicon show that, at temperatures up to 700°K, the thermally generated carrier concentration n_i is determined as follows: $n_i = 3.87 \times 10^{16} \times T \times (3/2) \exp(-1.21/2kT)$.] When this condition becomes extreme, transistor action ceases, the collector-to-emitter voltage V_{CE} collapses to a low value, and the current increases and is limited only by the external circuit.

If there is no current limiting, the increased current can melt the silicon and produce a collector-to-emitter short. This condition can occur as a result of a large-area average temperature effect, or in a small area that produces hot spots or localized thermal runaway. In either case, if the intrinsic temperature of a semiconductor is defined as the temperature at which the thermally generated carrier concentration is equal to the doped impurity concentration, the absolute maximum temperature for transistor action can be established.

The intrinsic temperature of a semiconductor is a function of the impurity concentration, and the limiting intrinsic temperature for a transistor is determined by the most lightly doped region. It must be emphasized, however, that the intrinsic temperature acts only as an upper limit for transistor

action. The maximum operating junction temperature and the maximum current rating are established by additional factors such as the efficiency of heat removal, the yield point and melting point of the solder used in fabrication, and the temperature at which permanent changes in the junction properties occur.

The **maximum current rating** of a transistor indicates the highest current at which, in the manufacturer's judgment, the device is useful. This current limit may be established by setting an arbitrary minimum current gain or may be determined by the fusing current of an internal connecting wire. A current that exceeds the rating, therefore, may result in a low current gain or in the destruction of the transistor.

The basic materials in a silicon transistor allow transistor action at temperatures greater than 300°C. Practical transistors, however, are limited to lower temperatures by mounting systems and surface contamination. If the **maximum rated storage** or **operating temperature** is exceeded, irreversible changes in leakage current and in current-gain characteristics of the transistor result.

Power-Dissipation Ratings

A transistor is heated by the electrical power dissipated in it. A maximum power rating is given, therefore, to assure that the temperature in all parts of a transistor is maintained below a value that will result in detrimental changes in the device. This rating may be given with respect to case temperature (for transistors mounted on heat sinks) or with respect to "free-air ambient" temperature. Case temperature is measured with a small thermocouple or other low-heat-conducting thermometer attached to the outside of the case or preferably inserted in a very small blind hole in the base so that the measurement is taken as close to the transistor chip as possible. Very short pulses of power do not heat the transistor to the temperature which it would attain if the power level was continued indefinitely. Ratings of maximum power consider this factor and allow higher power dissipation for very short pulses.

The dissipation in a transistor is not uniformly distributed across the semiconductor wafer. At higher voltages, the current concentrations become more severe, and hot spots may be developed within the transistor pellet. As a result, the power-handling capability of a transistor is reduced at high voltages. The power rating of a transistor may be presented most easily by a limiting curve that indicates a peak-power safe operating region. This curve shows power-handling capability as a function of voltage for various time durations.

Second Breakdown

Second breakdown is a potentially destructive phenomenon that can occur in all power transistors within the maximum current and voltage ratings of the device. A simplified explanation is that localized thermal regeneration occurs, and the transistor exhibits a lower value of breakdown voltage, referred to as the "second breakdown." The lower value of voltage results from thermal generation of charge-carrier pairs (holes and electrons) at high lo-

calized temperatures which alter the conductivity of the semiconductor in that vicinity. This localized effect reduces the ability of the transistor to support the applied voltage. Fig. 7 shows qualitatively what happens under primary or secondary breakdown.

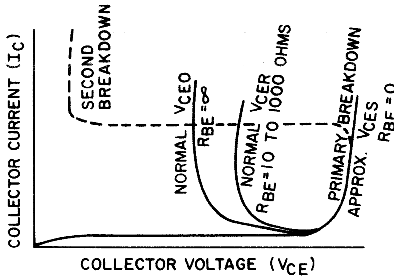


Fig. 7. Primary and secondary breakdown voltages.

Reverse-Bias Second Breakdown—Reverse-bias second breakdown is a phenomenon that may occur when the collector current continues to flow under reverse-bias conditions and causes the injected current to be concentrated in the central portions of the emitter, in contrast to the normal edge injection of the current. If the injected current is severely restricted to a very small central area by a large reverse emitter-base bias, the current density can rise to very large levels—in the order of thousands of amperes per square centimeter. If the collector of the transistor is of high-resistivity silicon, the high current density may inject a density of charge carriers that is equal to or greater than the collector impurity density. In this local region, the base widens and the collector depletion layer expands until the injected current density is smaller than the collector impurity density. If the current density is sufficiently high, the collector depletion layer expands to a more heavily doped collector region, such as an epitaxial substrate or a collector diffused region of a triple-diffused device. When the collector depletion layer expands, the collector breakdown voltage is governed by the impurity gradient related to the base doping and the heavily doped collector. The collector breakdown voltage normally supports only a fraction of the original voltage, and the second breakdown voltage results. The thermal effects from the large current densities also contribute to the regeneration process. Fig. 8 shows the process of reverse-bias second breakdown.

In an inductive circuit, a situation exists such that collector current flows in the forward direction while the transistor is being turned off, and a high voltage is induced across the device. As a result, the transistor enters the sustaining region. The hot spot that

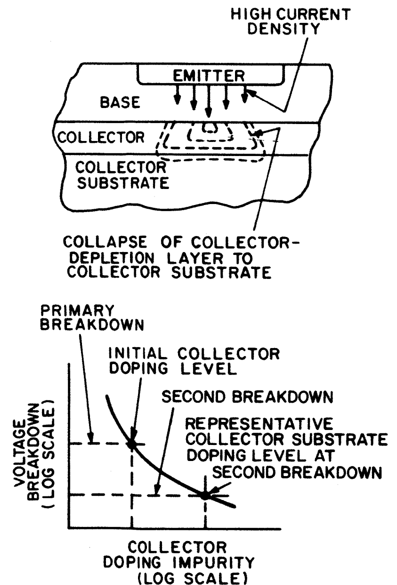


Fig. 8. Reverse-bias second breakdown.

forms during reverse-bias second breakdown may then be generated by current crowding in the depletion region, as shown in Fig. 9

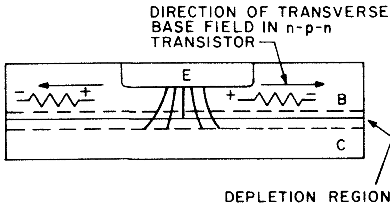


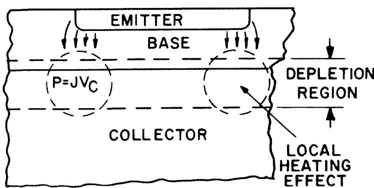
Fig. 9. Cross section showing current crowding that occurs during reverse-bias second breakdown.

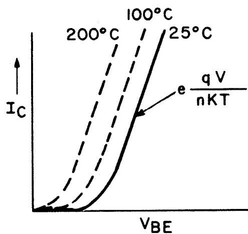
Forward-Bias Second Breakdown—Forward-bias second breakdown is somewhat different from reverse-bias second breakdown. As shown in Fig. 10, the localized heating results because the current density J crosses the depletion region (collector field) V_c to yield a power density P. As P increases, more current is injected into the localized area. The increase in current is caused by a decrease in the localized V_{BE}, at an approximate rate of 2 millivolts per ^{o}C. The local system becomes regenerative as more heat from the increased power density reduces V_{BE} and thereby increases the current injection.

The forward-bias second-breakdown current, $I_{S/b}$, is defined as the current at the onset of second breakdown, and is closely related to the collector field V_c, the current density J, and other properties of the transistor. Forward-bias second breakdown is also related to charge-carrier transit time across the base region, and is controlled by base width and any accelerating fields that exist in the base. The longer the transit time required for the charge carrier to cross the base, the more lateral diffusion of the charge and thus the greater the reduction in the current density at the edge of the collector depletion layer. This diffusion effect, referred to as "fan-out," is enhanced by wide base widths and homogeneously doped bases. Because the forward-bias second breakdown is related to the base width, it is also related to frequency response. For a given structure, this frequency relationship is expressed by the following empirical equation:

$$I_{S/b} \approx \left(\frac{1}{\sqrt{f_T}}\right) K \qquad (7)$$

Operation in the forward-bias region subjects the transistor to simultaneous current and voltage. This condition causes current concentrations as previously discussed. This type of rating must





$I_{S/b} \sim \frac{1}{\sqrt{f_T}}$ — FACTORS:
BASE WIDTH
DRIFT FIELD
CURRENT DENSITY
VOLTAGE

Fig. 10. Forward-bias second breakdown.

be considered for all linear applications of transistors. High-frequency transistors, particularly those intended for use at uhf and microwave frequencies, require very narrow base widths and graded bases to achieve the required f_T. For this reason, protective features must be built into these devices to avoid second breakdown. One such feature is emitter ballasting. The effect of and the techniques used to achieve this ballasting are discussed in the section on **Special Features of High-Frequency Transistors.**

Safe-Operating-Area Ratings

During normal circuit operation, power transistors are often required to sustain high current and high voltage simultaneously. The capability of a transistor to withstand such conditions is normally shown by use of a safe-operating-area rating curve. This type of rating curve defines, for both steady-state and pulsed operation, the voltage-current boundaries that result from the combined limitations imposed by voltage and current ratings, the maximum allowable dissipation, and the second-breakdown ($I_{s/b}$) capabilities of the transistor.

If the safe operating area of a power transistor is limited within any portion of the voltage-current characteristics by thermal factors (thermal impedance, maximum junction temperatures, or operating case temperature), this limiting is defined by a constant-power hyperbola ($I = KV^{-1}$) which can be represented on the log-log voltage-current curve by a straight line that has a slope of —1.

The energy level at which second breakdown occurs in a power transistor increases as the time duration of the applied voltage and current decreases. The power-handling capability of the transistor also increases with a decrease in pulse duration because thermal mass of the power-transistor chip and associated mounting hardware imparts an inherent thermal delay to a rise in junction temperature.

Fig. 11 shows a forward-bias safe-area rating chart for a typical silicon power transistor, the RCA-2N3585. The boundaries defined by the curves in the safe-area chart indicate, for both continuous-wave and nonrepetitive-pulse operation, the maximum current ratings, the maximum collector-to-emitter forward-bias avalanche breakdown-voltage rating [$V\alpha M = 1$, which is usually approximated by $V_{CEO}(sus)$], and the thermal and second-breakdown ratings of the transistors.

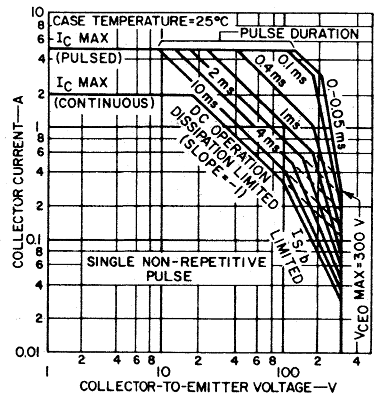


Fig. 11. Safe-area rating chart for the 2N3585 silicon power transistor.

As shown in Fig. 11, the thermal (dissipation) limiting of the 2N3585 ceases when the collector-to-emitter voltage rises above 100 volts during dc operation. Beyond this point, the safe operating area of the transistor is limited by the second-breakdown ratings. During

pulsed operation, the thermal limiting extends to higher values of collector-to-emitter voltage before the second-breakdown region is reached, and as the pulse duration decreases, the thermal-limited region increases.

If a transistor is to be operated at a pulse duration that differs from those shown on the safe-area chart, the boundaries provided by the safe-area curve for the next higher pulse duration must be used, or the transistor manufacturer should be consulted. Moreover, as indicated in Fig. 11, safe-area ratings are normally given for single nonrepetitive pulse operation at a case temperature of 25°C and must be derated for operation at higher case temperatures and under repetitive-pulse or continuous-wave conditions.

Fig. 12 shows temperature derating curves for the 2N3585 safe-area chart of Fig. 11. These curves show that thermal ratings are affected far more by increases in case temperature than are second-breakdown ratings. The thermal (dissipation - limited) derating curve decreases linearly to zero at the maximum junction temperature of the transistor [T_J(max) = 200°C]. The second-breakdown ($I_{S/b}$-limited) temperature derating curve, however, is less severe because the increase in the formation of the high current concentrations that cause second breakdown is less than the increase in dissipation factors as the temperature increases.

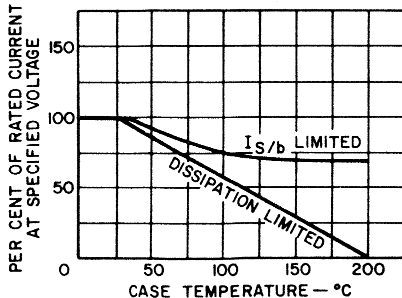


Fig. 12. Safe-area temperature-derating curves for the 2N3585 silicon power transistor.

STRUCTURE AND GEOMETRIES

The ultimate aim of all transistor fabrication techniques is the construction of two parallel p-n junctions with controlled spacing between the junctions and controlled impurity levels on both sides of each junction. A variety of structures and geometries have been developed in the course of transistor evolution.

In power transistors, structure refers to the junction depth, the concentration and profile of the impurities (doping), and the spacings of the various layers of the device. Geometry refers to the topography of the transistor. These factors and the method of assembly of the semiconductor pellet into the over-all transistor package have an important bearing on the types of applications in which a power transistor can be used to optimum advantage. The proper choices of trade-offs among these factors determine the gain, frequency, voltage, current, and dissipation capabilities of power transistors.

Various structures have been developed to provide different electrical, thermal, or cost properties, with each having certain advantages or compromises to offer. Table I lists the principal structures available for silicon power transistors, together with some of the advantages and disadvantages of each type.

All RCA high-frequency power transistors employ a double-

Table I—Types of Structures for Silicon Power Transistors

Structure	Advantages	Disadvantages
Alloy	Generally rugged	Low speed, high cost
Hometaxial-base	Rugged, low cost	Low speed
Double-diffused mesa	High speed	Poor saturation resistance
Double-diffused planar	High speed, low leakage	Poor saturation resistance
Triple-diffused mesa	High speed, low-saturation resistance	Moderate cost, moderate leakage
Triple-diffused planar	High speed, low leakage, low saturation resistance	Moderate to high cost
Double-diffused epitaxial mesa	High speed, low-saturation resistance	Moderate cost, moderate leakage, less rugged
Double-diffused epitaxial planar	High speed, low leakage, low saturation resistance	Higher cost, less rugged
Epitaxial-base mesa	Moderate speed, low saturation resistance	Low voltage, moderate leakage
Multiple epitaxial-base mesa	Moderate speed, low saturation resistance, rugged, high voltage	Moderate cost
Double-diffused multiple-epitaxial mesa	High speed, rugged, low saturation resistance	Moderate cost, moderate leakage

diffused epitaxial planar structure which permits passivation of the emitter-base and collector-base junctions to achieve low leakage and a high-frequency collector design.

The topography of a transistor is referred to as its geometry. This transistor geometry, in conjunction with its structure, establishes most of the fundamental transistor electrical, thermal, and economic properties. Proper geometric design of a transistor allows for many compromises, which may result in a variety of advantages and disadvantages from different structures.

The basic premise for most geometry designs for power transistors is to increase current handling per unit area of device. This condition results in greater power gain and collector efficiency at high operating frequencies because of the smaller device areas.

Power-transistor geometries have evolved from the very early inefficient "ring-dot" configurations to the present-day sophisticated "overlay" concepts. Fig. 13 shows some typical geometry milestones in this evolutionary cycle.

The early geometries were characterized by simple shapes, large dimensional tolerances, and poor utilization of active regions. As the state of the art in fine-line mask making and wafer printing improved, the geometries became more involved, with much finer dimensions.

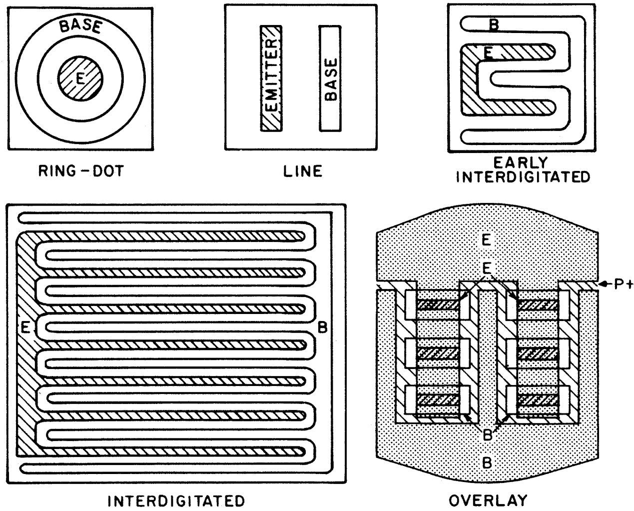


Figure 13. Typical geometries in the development of transistors.

Special Features of High-Frequency Power Transistors

POWER transistors are used in high-frequency amplifiers for military, industrial, and consumer applications. They are operated class A, B, or C, with frequency- or amplitude-modulation, single sideband or double sideband, in environments ranging from airborne to marine.

The increasing number of rf power transistors available today offers the circuit designer a wide selection from which to determine the optimum type for a particular application. The choice is based on factors such as maximum power output, maximum operating frequency, operating efficiency, power gain, reliability, and cost per watt of power generated. The ultimate choice of the transistors produced by any manufacturer, therefore, is dependent upon how well the devices perform in relation to these critical factors. RCA "overlay" silicon power transistors offer significant advantages for rf power applications at frequencies that extend well into the microwave region.

PHYSICAL DESIGN

During the past several years, considerable effort has been expended to improve the quality and reliability of high-frequency power transistors and simultaneously to advance the power-frequency capability of these devices. The technological developments that resulted from this effort have made possible transistor stuctures that can provide substantial power output and gain and high operating efficiency at frequencies that extend well into the microwave region. Such devices can now be produced with confidence in high-volume productions for use in applications in which high-quality performance and high reliability are primary requirements.

Basic Design Considerations

At high current levels, the emitter current of a transistor is concentrated at the emitter-base edge. As explained in the section on **General Considerations for Power Transistors**, this current-

crowding effect results because the current flow through the base region, between the emitter and the collector, causes a voltage drop that produces the maximum forward bias at the edge of the emitter closest to the base contact. The center of the emitter, therefore, injects very little current. Because of this edge-injection phenomenon, a high emitter periphery-to-area ratio is essential to the achievement of high current-handling capability. This requirement has been a major factor in the evolution of transistor emitters from the circle type, to the line type, to the comb type, and finally to the overlay type of structure.

In addition to the requirement for a high emitter periphery-to-area ratio, power transistors intended for use in high-freauency applications must also exhibit a low capacitance and a short carrier transmit time between emitter and collector. These latter factors critically affect the frequency capabilities of the device and, therefore, must be considered in the design of a high-frequency transistor structure.

The maximum frequency of oscillation f_{max} of a junction-transistor structure is determined as follows:

$$f_{max} = (PG)^{1/2} f = \frac{1}{4\pi}\left[\frac{1}{r_{bb}' C_c T_{ec}}\right] \quad (8)$$

where PG is the power gain, f is the frequency of operation, r_{bb}' is the base spreading resistance, C_c is the collector capacitance, and T_{ec} is the emitter-to-collector transit time (i.e., the signal delay time). The latter two terms are very small.

Eq. (10) was derived from analyses of the transistor as a low-level class A amplifier, but it can also serve as a guide to performance in the class C circuit more commonly encountered in high-frequency power applications. To a first approximation, the frequency f_{max} at which the power gain is unity is independent of collector area. Although the collector capacitance C_c is directly dependent on the collector area, the resistance r_{bb}' varies inversely with collector area. As a result, the length of a transistor can be extended and the power dissipation and current-handling capability improved without any increase of the $r_{bb}'C_c$ product.

The collector-to-emitter transit time T_{ec} has four components: the charging time of the emitter capacitance, the transit time through the collector depletion region, and the charging time of the collector capacitance and the collector series resistance. The last term is usually negligible in devices made with triple-diffused or epitaxial construction. Of the remainder, only the emitter transit time T_e is current-dependent, as shown by:

$$T_e = r_e C_{ea} = \frac{kT}{qI_e} C_{ea} \quad (9)$$

where r_e is the emitter resistance, C_{ea} is the emitter capacitance, k is Boltzmann's constant, T is the temperature in degrees Kelvin, q is the electron charge, and I_e is the emitter current. However, if the emitter edge is increased proportionally as the area of the emitter is increased, the fraction C_{ea}/I_e remains a constant. With suitable scaling, transistors can be enlarged to increase power-handling capability without deterioration of frequency response.

Overlay Transistor Structure

The exceptional high-frequency-power capabilities of the overlay power transistors result from the unique emitter construction used in these devices. In overlay transistors, the size of the emitters is substantially reduced, and a large number (from 16 to several hundred) of separate emitter sites are connected in parallel. This method of construction results in the high emitter periphery-to-area ratios, and makes possible the high current-handling capabilities, low capacitances, and short transit times between emitter and collector, that are required for rf power transistors.

The overlay transistor takes its name from the emitter metallization, shown in Fig. 14, that lies over the base instead of adjacent to it as in the interdigitated structure. The actual base and emitter areas beneath the metal pattern are insulated from one another by a silicon dioxide layer. The overlay arrangement provides a substantial increase in over-all emitter periphery without increasing the physical area of the device, and thus improves the power-frequency capability of the device.

In addition to the standard base and emitter diffusions, an added diffused region in the base serves as a conductor grid. This p+ region offers three advantages: (1) it distributes base current uniformly over all the separate emitter sites, (2) it reduces the base-contact resistances between the aluminum metallization and the silicon material, and (3) it permits the use of larger emitter sites which makes possible higher design ratios and wider emitter metallizing fingers that result in lower current density.

For lower-power hf/vhf and small-signal uhf RCA transistors, an interdigitated structure is used. In this structure, as shown in Fig. 15, the emitters and bases are built like a set of

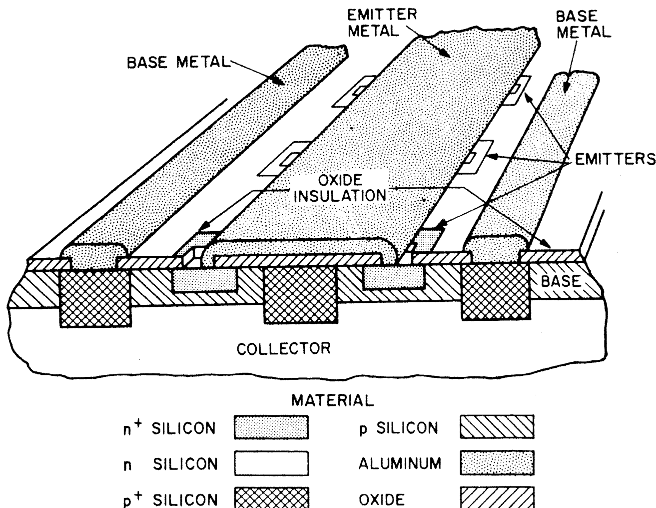


Figure 14. Top and cross-sectional view of a typical overlay transistor.

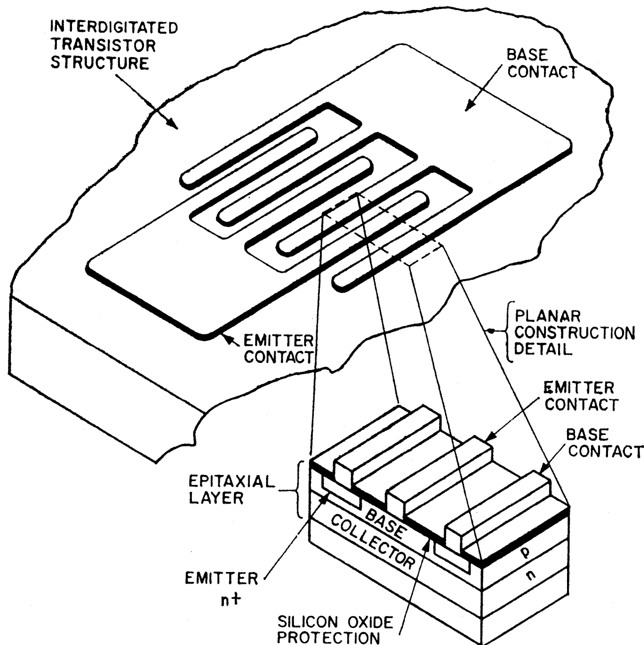


Figure 15. Top and cross-sectional view of a typical interdigitated transistor.

interlocking combs. The sizes of the emitter and base areas are controlled by masking and diffusion. The oxide deposit, formed of silicon heated to a high temperature, masks the transistor against either an n- or p-type impurity. This oxide is removed by the usual photoetching techniques in areas where diffusion is required.

Polycrystalline Silicon Layer

The broad emitter fingers and the noncritical metal definition of the overlay transistor structure makes possible the introduction of additional conducting and insulating layers between the aluminum metallization and the shallow diffused emitter sites that are required for good high-frequency performance. RCA has developed a technique in which a polycrystalline silicon layer (PSL) is used to separate these regions. Fig. 16 shows a cross-sectional diagram of an overlay transistor structure in which this interlayer is used.

Use of the polycrystalline silicon layer between the aluminum metallization and the shallow diffused emitter region forms an insulating barrier that substantially reduces the possibility of "alloy spikes" that result from intermetallic formations of silicon and aluminum under severe hot-spot conditions. Such intermetallic formations can cause transistor failures because of emitter-to-base shorts.

As shown in Fig. 16, the polycrystalline silicon layer also forms a barrier between the aluminum emitter fingers and the

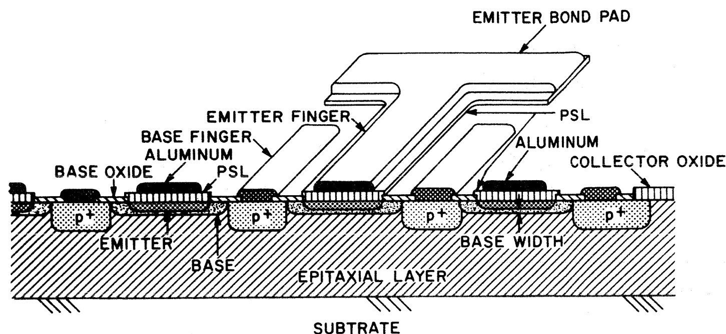


Figure 16—Overlay transistor structure that contains a polycrystalline silicon layer (PSL).

oxide insulation layer over the base. This barrier minimizes the possibility of dielectric failure, which can also lead to emitter-to-base shorts because of an interaction between the aluminum and the silicon diozide.

The addition of the polycrystalline silicon interlayer contributes substantially to the reliability of high-frequency power transistors. Reliability studies of uhf power transistors operated under overstress condition (i.e., at a junction temperature greater than 200°C) have demonstrated an order of magnitude improvement in the mean time between failures of devices that employed the PSL technique over that of devices in which the PSL technique was not used. The PSL technique, therefore, is being used increasingly in RCA high-frequency power transistors.

Emitter-Site Ballasting

In many RCA high-frequency power transistors, the PSL technology is used as the medium for the introduction of emitter-site ballasting. The resistivity and contacting geometry of the aluminum to the polycrystalline silicon interlayer are controlled to form a ballast resistor in series with each emitter site. These resistors function as negative-feedback elements to prevent excessive current in any portion of the transistor and, in this way, minimize the possibility of hot spots in the device. Because the emitter in an overlay transistor is segmented into many separate sites connected in parallel, each hot spot may be isolated and controlled, and the injection of charge carriers across the transistor chip is made more uniform.

Emitter-site ballasting makes possible a more effective use of emitter periphery and, therefore, results in an increased transistor power-output capability. In addition, this ballasting makes high-frequency power transistors more immune to failures caused by high VSWR conditions such as may be encountered in some broadband amplifiers. Transistor failures because of high output VSWR conditions are often related to forward-bias second breakdown. The feedback action of the emitter ballasting resistors tend to minimize the formation of localized concentration of current that is characteristic of forward-bias second breakdown. As a result,

transistors that employ emitter-site ballasting have a substantially higher capability to withstand high VSWR conditions.

OPERATING CHARACTERISTICS

At rf and microwave frequencies, the operation of a power transistor is critically dependent upon the high-frequency capability of the device. The ability of the transistor to provide significant power gain, develop useful power outputs, operate efficiently in circuit applications, and to operate over substantial bandwidths are direct functions of this capability.

Output Power

The power-output capability of a transistor is determined by current- and voltage-handling capabilities of the device at the frequency range of interest. The current-handling capability of the transistor is limited by its emitter periphery and epitaxial-layer resistivity. The voltage-handling capability of the device is limited by the breakdown voltages, which are, in turn, limited by the resistivity of the epitaxial layer and by the penetration of the junction.

The breakdown voltage at high frequencies is substantially higher than the dc or static value, as indicated by the following equation:

$$V_{CEO} = V_{CBO} \Big/ \left(\frac{f_T}{f} + 1\right)^{\frac{1}{n}} \quad (10)$$

where n is the avalanche breakdown factor. Eq. (10) shows that the breakdown characteristic increases from the V_{CEO} value under dc conditions to a value approaching V_{CBO} at a frequency f equal to or greater than f_T.

Another parameter that limits the power-handling capability of the transistor is the saturation voltage. The saturation voltage $V_{CE(sat)}$ at high frequencies is significantly greater than the dc value because the active area is smaller than at dc.

In general, the operating voltage restrictions are the same for all high-frequency power transistors; therefore, only current-handling capability differentiates high-power transistors from lower-power units.

At high current levels the emitter current of a transistor is concentrated at the emitter-base edge; therefore, transistor current-handling capability can be increased by the use of emitter geometries which have high emitter-periphery-to-emitter-area ratios and by the use of improved techniques in the growth of collector substrate material. Transistors for large-signal applications are designed so that the peak currents do not cause base widening which would limit the current-handling capability of the device. Base-width widening is severe in transistors in which the collector side of the collector-base junction has a lower carrier concentration and higher resistivity than the base side of the junction. However, the need for low-resistivity material in the collector to handle high currents without base widening severely limits the breakdown voltages, as discussed previously. As a result, the use of a different-resistivity epitaxial layer for different operating voltages is becoming common.

Power Gain

The power gain of a high-frequency transistor power amplifier is determined by the dynamic f_T, the dynamic input impedance, and the collector load impedance, which depends on the required power output and the collector voltage swing. The power gain, P.G., of a transistor power amplifier can be expressed as follows:

$$P.G. = \frac{(f_T/f)^2 R_L}{Re(Z_{in})} \qquad (11)$$

where f_T is the dynamic gain-bandwidth product, f is the frequency of operation, R_L is the real part of the collector parallel equivalent load impedance determined by the required power output, and $Re(Z_{in})$ is the real part of the dynamic input impedance when the collector is loaded with Z_L.

Eq. (11) shows that for high-gain operation of large-signal or power transistors, the device should have high current gain at the frequency of operation under large current swing conditions. This performance is achieved with shallow diffusion techniques.

R_L is defined approximately by Eq. (12) for class B or C operation:

$$R_L \cong K \frac{[V_{CC} - V_{CE}\,(sat)]^2}{2P_O} \qquad (12)$$

where K is unity or less, depending on the class of operation. The real part of the dynamic input impedance, $Re(Z_{in})$, varies considerably with signal level, and varies inversely with the power output of the device. The package parasitic inductance also are important in determining the value of $Re(Z_{in})$.

Efficiency

The collector efficiency of a transistor amplifier is defined as the ratio of signal power output at the frequency of interest to the dc input power. It can be calculated as:

$$\eta_c = \eta_v\, \eta_i\, \eta_{ckt} \qquad (13)$$

where η_v is the efficiency of conversion of dc collector voltage to signal-frequency collector voltage (determined primarily by the ratio of V_{CC} to V_{CE} and the class of operation), η_i is the efficiency of conversion of dc collector current to signal-frequency collector current (determined primarily by the class of operation and the transit time in the collector depletion region for high-frequency transistors), and η_{ckt} is the circuit efficiency, which is determined by the loaded and unloaded Q's of the collector circuit. (The amplifier collector efficiency η_c, which is a function of both circuit and transistor parameters, should not be confused with the collector efficiency α^*, which is a basic transistor parameter determined solely by the physical structure of the device, as explained in the discussion of **General Considerations for Power transistors.)**

Bandwidth

The bandwidth of a transistor power amplifier is determined by the intrinsic frequency capability of the transistor (directly related to f_T), package parasitic elements, and the input and output matching circuits.

SPECIAL RATING CONCEPTS

Unlike low-frequency high-power transistors, many rf devices can fail within the dissipation limits set by the classical junction-to-case thermal resistance during operation under conditions of high load VSWR, high collector supply voltage, or linear (Class A or AB) operation. Failure can be caused by hotspotting, which results from local current concentration in the active areas of the device, and may appear as a long-term parameter degradation. Localized hotspotting can also lead to catastrophic thermal runaway.

The presence of hotspots can make virtually useless the present method of calculating junction temperature by measurements of average thermal resistance, case temperature, and power dissipation. However, by use of an infrared microscope, the spot temperature of a small portion of an rf transistor pellet can be determined accurately under actual or simulated device operating conditions. The resultant peak-temperature information is used to characterize the device thermally in terms of junction-to-case hotspot thermal resistance, $\theta_{JS\text{-}C}$.

The use of hotspot thermal resistance improves the accuracy of junction temperature and related reliability predictions, particularly for devices involved in linear or mismatch service.

DC Safe Area

The safe area determined by infrared techniques represents the locus of all current and voltage combinations within the maximum ratings of a device that produce a specified spot temperature (usually 200°C) at a fixed case temperature. The shape of this safe area is very similar to the conventional safe area in that there are four regions, as shown in Fig. 17: constant current, constant power, derating power, and constant voltage.

Regions I and IV, the constant-current and constant-voltage regions, respectively, are determined by the maximum collector current and V_{CEO} ratings of the device. Region II is dissipation-

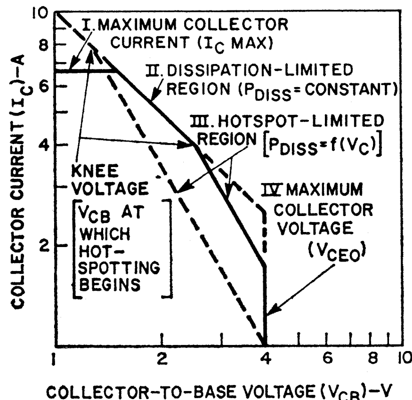


Figure 17. Safe-area curve for an rf power transistor determined by infrared techniques.

limited; in the classical safe area curve, this region is determined by the following relationship:

$$P_{max} = \frac{T_J(max) - T_C}{\Theta_{J-C}} \qquad (14)$$

where T_C is the case temperature.

This relationship holds true for the infrared safe area; P_{max} may be slightly lower because the reference temperature $T_{J(max)}$ is a peak value rather than an average value. The hotspot thermal resistance ($\theta_{JS\text{-}C}$) may be calculated from the infrared safe

area by use of the following definition:

$$\Theta_{JS-C} = \frac{T_{JS} - T_C}{P} \quad (15)$$

where T_{JS} is highest spot temperature [$T_{J(max)}$ for the safe area] and P is the dissipated power ($= I \times V$ product in Region II).

The collector voltage at which regions II and III intersect, called the knee voltage V_K, indicates the collector voltage at which power constriction and resulting hotspot formation begins. For voltage levels above V_K, the allowable power decreases. Region III is very similar to the second-breakdown region in the classical safe area curve except for magnitude. For many rf power transistors, the hotspot-limited region can be significantly lower than the second-breakdown locus. Generally V_K decreases as the size of the device is increased.

Fig. 18 shows the temperature profiles of two transistors with identical junction geometries that operate at the same dc power level. If devices are operated on the dissipation-limited line of their classical safe areas, the profiles show that the temperature of the unballasted device rises to values 130°C in excess of the 200°C rating. Temperatures of this magnitude, although not necessarily destructive, seriously reduce the lifetime of the device.

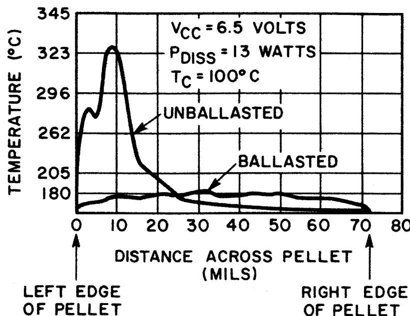


Figure 18. Thermal profiles of a ballasted and an unballasted power transistor during dc operation.

Effect of Emitter Ballasting

The profiles shown in Fig. 18 also demonstrate the effectiveness of emitter ballasting in the reduction of power (current) constriction. In the ballasted device, a biasing resistor is introduced in series with each emitter or small groups of emitters. If one region draws too much current, it will be biased towards cutoff, allowing a redistribution of current to other areas of the device.

The amount of ballasting affects the knee voltage, V_K, as shown in Fig. 19. A point of diminishing returns is reached as V_K approaches V_{CEO}.

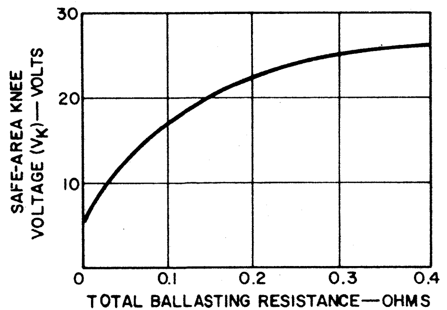


Figure 19. Safe-area voltage for an rf power transistor as a function of total ballasting resistance.

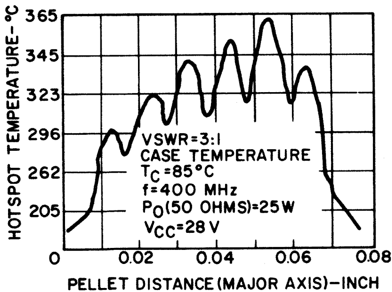
RF Operation

In normal class C rf operation, the hotspot thermal resistance is approximately equal to the classical average thermal resistance. If the proper collector loading (match) is maintained, $\theta_{JS\text{-}C}$ is in-

dependent of output power at values below the saturated- or slumping-power level, and is independent of collector supply voltage at values within + 30 per cent of the recommended operating level.

Power constriction in rf service normally occurs only for collector load VSWR's greater than 1.0. A transistor that has a mismatched load experiences temperatures far in excess of device ratings, as shown in Fig. 20(a) for VSWR = 3.0. For comparison, the temperature profile for the matched condition is shown in Fig. 20(b).

Fig. 21 is a typical family of thermal-resistance curves that indicate the response of a device to various levels of VSWR and collector supply voltage. $\theta_{JS\text{-}C}$ responds to even slight increases in VSWR above 1.0 and saturates at a VSWR in the range of 3 to 6. The saturated level increases with increasing supply voltage.



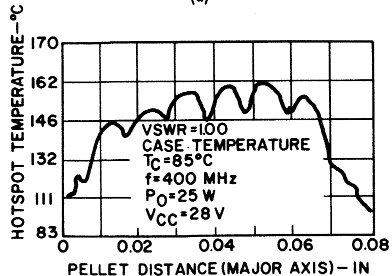


Figure 20. Thermal profile of a power transistor during rf operation: (a) under mismatched conditions; (b) under matched conditions.

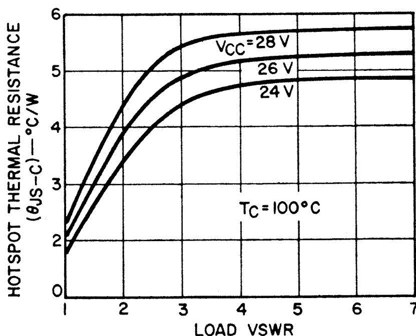


Figure 21. Mismatch-stress thermal characteristics for the 2N5071.

Devices with high knee voltages tend to show smaller changes of $\theta_{JS\text{-}C}$ with VSWR and supply voltage. $\theta_{JS\text{-}C}$ under mismatch is independent of frequency and power level, and reaches its highest values at load angles that produce maximum collector current. Power level does, however, influence the temperature rise and probability of failure.

Device failure can also occur at a load angle that produces minimum collector current. Under this condition, collector voltage swing is near its maximum, and an avalanche breakdown can result. This mechanism is sensitive to frequency and power level, and becomes predominant at lower frequencies because of the decreasing rf-breakdown capability of the device.

Collector mismatch can be caused by the following conditions:

1. Antenna loading changes in mobile applications when the vehicle passes near a metallic structure.
2. Antenna damage.

3. Transmission-line failure because of line, connector, or switch defects.
4. Variable loading caused by nonlinear input characteristics of a following transistor (particularly broadband) or varactor stage.
5. Supply-voltage changes that reflect different load-line requirements in class C.
6. Tolerance variations on fixed-tuned or stripline circuits.
7. Matching network variations in broadband service.

Case-Temperature Effects

The thermal resistance of both silicon and beryllium oxide, two materials that are commonly used in rf power transistors, increases about 70 per cent as the temperature increases from 25 to 200°C. Other package materials such as steel, kovar, copper, or silver, exhibit only minor increases in thermal resistance (about 5 per cent). The over-all increase in $\theta_{JS\text{-}C}$ of a device depends on the relative amounts of these materials used in the thermal path of the device; typically the increase of $\theta_{JS\text{-}C}$ ranges from 5 per cent to 70 per cent. Fig. 22

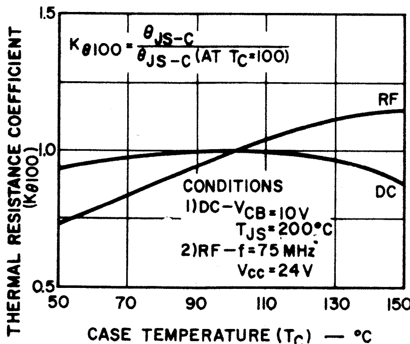


Figure 22. Thermal-resistance coefficient for the 2N5071.

shows the rf and dc thermal resistance coefficients for a typical rf transistor. For both cases, the coefficient is referenced to a 100°C case and is defined as follows:

$$K_{\theta 100} = \frac{\theta_{JS-C}}{\theta_{JS-C} \text{ at } T_C = 100°C} \quad (16)$$

The rf coefficient changes more than the dc coefficient, because of the power constriction that occurs in rf operation at elevated case temperature.

RELIABILITY CONSIDERATIONS

When the rf and thermal capabilities of a transistor have been established, the next step is to establish the reliability of the device for its actual application. The typical acceptable failure rate for transistors used in commercial equipment is 1 per cent per 1000 hours (100,000 hours MTBF); for transistors used in military and high-reliability equipment, it is 0.01 to 0.1 per cent per 1000 hours. Because it is not practical to test transistors under actual use conditions, dc or other stress tests are normally used to simulate rf stresses encountered in class B or class C circuits at the operating frequencies. Information derived from these tests is then used to predict the failure rate for the end-use equipment. The tests used to assure reliability include high-temperature storage tests, dc and rf operating life tests, dc stress step tests, burn-in, temperature cycling, relative humidity, and high-humidity reverse bias. The end-point measurement for these tests should include collector-to-emitter voltage V_{CEO} and emitter-

to-base voltage V_{EBO} in addition to the common end-point collector-to-emitter current I_{CEO}, collector-to-base voltage V_{CBO}, collector-to-emitter saturation voltage $V_{CE}(sat)$, power output, and power gain.

One of the common failure modes in rf power transistors is degradation of the emitter-to-base junction. The high-temperature storage life test and the dc and rf operating life tests can accelerate this failure mode, and it can be detected by measurement of V_{EBO}.

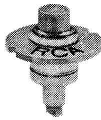
Plastic uhf power transistors are more sensitive to emitter-to-base-junction degradation than similar hermetic devices. The enhancement of this failure mode in plastic devices can be caused by moisture penetration into the very close geometries used in uhf power transistors. Thermal fatigue is also a problem that affects the reliability of uhf plastic power transistors because large thermal-expansion differences exist between the plastic encapsulant and the fine bonding wires (usually 1 mil) used in the devices.

PACKAGES

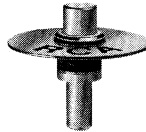
The package is an integral part of an rf power transistor. A suitable package for rf applications should have good thermal properties and low parasitic reactances. Package parasitic inductances and resistive losses have significant effects on such circuit performance characteristics as power gain, bandwidth, and stability. The most critical parastics are the emitter and base lead inductances. Table II gives the inductances of some of the more important commercially available rf power-transistor packages. Photographs of the packages are shown in Fig. 23. The TO-60 and TO-39 packages were first used in devices such as the 2N3375 and the 2N3866. The base and emitter parastic inductance for both TO-60 and TO-39 packages is in the order of 3 nanohenries; this inductance represents a reactance of 7.5 ohms at 400 MHz. If the emitter is grounded internally in a TO-60 package (as in the RCA-2N5016), the emitter

Table II—Summary of RCA Packages for RF Power Transistors

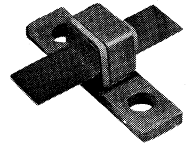
PACKAGE	APPROXIMATE INDUCTANCE (nH)	UPPER FREQUENCY OF OPERATION (MHz)
TO-39	3	500
TO-60 (isolated emitter)	3	400
TO-60 (internally grounded emitter)	0.6	500
UHF HERMETIC STRIPLINE		
HF-19 (STUD) = JEDEC TO-216AA	0.5	1000
HF-31 = STUDLESS JEDEC TO-216AA	0.5	1000
HF-32 (FLANGED)	0.5	1000
MICROWAVE HERMETIC STRIPLINE		
HF-28 (FLANGED)	0.2	2500
COAXIAL HERMETIC		
HF-11 = JEDEC TO-215AA	0.1	3000
HF-21 = JEDEC TO-201AA	0.2	2500



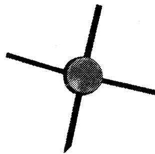
HF-21
Hermetic
Ceramic-Metal
Coaxial Package,
Large
(JEDEC TO-201AA)



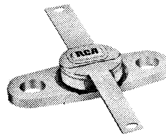
HF-11
Coaxial Package, Small
(JEDEC TO-215AA)



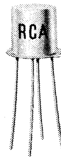
HF-28
Hermetic
Ceramic-Metal
Stripline Package
Grounded emitter
or base



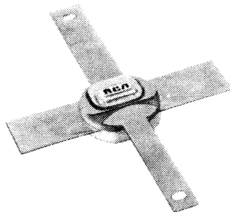
HF-33
Isolated
Electrodes



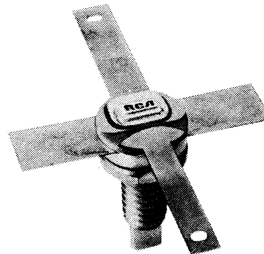
HF-32
Hermetic
Stripline Package



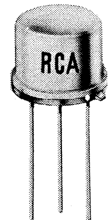
JEDEC TO-72



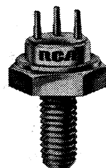
HF-31
Hermetic
Ceramic-Metal
Stripline Package
(Studless JEDEC TO-216AA)



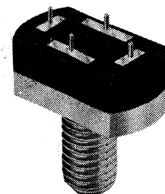
HF-19
Hermetic Strip-Line Type
Ceramic-to-Metal Package
with Isolated Electrodes
(JEDEC TO-216AA)



JEDEC TO-39



JEDEC TO-60



HF-12
Molded Silicone-Plastic Case
(JEDEC TO-217AA)

Figure 23. Commercially available high-frequency power transistor packages.

lead inductance is reduced to 0.6 nanohenry.

Hermetic low-inductance radial-lead packages are also available. The HF-19 package introduced by RCA for the 2N5919 utilizes ceramic-to-metal hermetic seals, has isolated electrodes, and has rf performance comparable to an rf plastic package. This package is also available in a studless version (HF-31) for miniaturized or low-power applications and in a grounded-emitter, flanged version (HF-32) for compact applications.

Low-parastic, hermetic packages are available for microwave applications. The HF-11, a medium-power, hermetic coaxial package first used for the RCA-2N5470, employs ceramic-to-metal construction and has parasitic inductances in the order of 0.1 nanohenry. A larger, higher-power version, the HF-21, uses the same constructional techniques and has parasitic inductances in the range of 0.2 nanohenry. The stripline equivalent of this package is the HF-28 and has approximately the same parasitic reactances as the HF-21.

Table III compares the performance of the TO-39 package, the HF-19 hermetric stripline package, and the HF-11 coaxial package with the same transistor chip. At a frequency of 1 GHz and an input power of 0.3 watt, the coaxial package performs significantly better than either the stripline or the TO-39 package. The coaxial package results in twice as much output power as the TO-39 package. In addition, the coaxial-package transistor is capable of delivering an output of more than 1 watt with a gain of 5 dB at 2 GHz.

Table III—Package Performances with Same Transistor Chip

	f-GHz	P_{in}-W	P_O-W	P.G.-dB	η_C(28V)-%
TO-39	1	0.3	1	5	35
HF-19	1	0.3	1.5	7	45
HF-11	1	0.3	2.2	8.6	50
HF-11	2	0.3	1	5	35

Design Considerations for High-Frequency Power Circuits

IN the design of silicon-transistor rf power amplifiers for use in transmitting systems, several fundamental factors must be considered. As with any rf power amplifier, the class of operation has an important bearing on the power output, linearity, and operating efficiency. The matching characteristics of input and output terminations significantly affect power output and frequency stability and, therefore, are particularly important considerations in the design of transistor power amplifiers. The selection of the proper transistor for a given circuit application is also a major consideration, and the circuit designer must realize the significance of the various transistor parameters to make a valid evaluation of different types.

CLASS OF OPERATION

The class of operation of an rf amplifier is determined by the circuit performance required in the given applications. Class A power amplifiers are used when extremely good linearity is required. Although power gain in this class of service is considerably higher than that in class B or class C service, the operating efficiency of a class A power amplifier is usually only about 25 per cent. Moreover, the standby drain and thermal dissipation of a class A stage are high, and care must be exercised to assure thermal stability.

In applications that require good linearity, such as single-sideband transmitters, class B push-pull operation is usually employed because the transistor dissipation and standby drain are usually much smaller and operating efficiency is higher. Class B operation is characterized by a collector conduction angle of 180 degrees. This conduction is obtained by use of only a slight amount of forward bias in the transistor stage. In this class of service, care must be taken to avoid thermal runaway.

In a class C transistor stage, the collector conduction angle is less than 180 degrees. The gain of the class C stage is less than that of a class A or class B stage, but is entirely usable. In addition, in the class C stage, standby drain is virtually zero, and circuit efficiency is the highest of the three classes. Because of the high efficiency, low collector dissipation, and negligible standby

drain, class C operation is the most commonly used mode in rf power-transistor applications.

For class C operation, the base-to-emitter junction of the transistor must be reverse-biased so that the collector quiescent current is zero during zero-signal conditions. Fig. 24 shows four methods that may be used to reverse-bias a transistor stage.

Figs. 24(b) and 24(c) show methods in which reverse bias is developed by the flow of dc base current through a resistance. In the case shown in Fig. 24(b), bias is developed across the base spreading resistance. The magnitude of this bias is small and uncontrollable because of the variation in r_{bb}' among different transistors. A better approach, shown

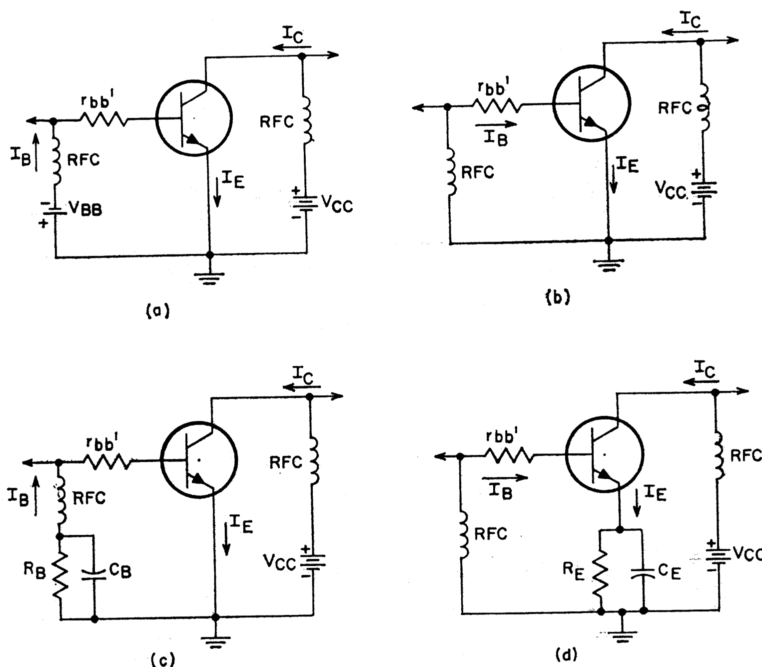


Figure 24. Methods for obtaining class C reverse bias: (a) by use of fixed dc supply V_{BB}; (b) by use of dc base current through the base spreading resistance r_{bb}'; (c) by use of dc base current through an external base resistance R_B; (d) by use of self bias developed across an emitter resistor R_E.

Fig. 24(a) shows the use of a dc supply to establish the reverse bias. This method, although effective, requires a separate supply, which may not be available or may be difficult to obtain in many applications. In addition, the bypass elements required for the separate supply increase the circuit complexity.

in Fig. 24(c), is to develop the bias across an external resistor R_B. Although the bias level is predictable and repeatable, the size of R_B must be carefully chosen to avoid reduction of the collector-to-emitter breakdown voltage.

The best reverse-bias method is illustrated in Fig. 24(d). In this

method, self-bias is developed across an emitter resistor R_E. Because no external base resistance is added, the collector-to-emitter breakdown voltage is not affected. An additional advantage of this approach is that stage current may be monitored by measurement of the voltage drop across R_E. This technique is very helpful in balancing the shared power in paralleled stages. The bias resistor R_E must be bypassed to provide a very-low-impedance rf path to ground at the operating frequency to prevent degeneration of stage gain. In practice, emitter bypassing is difficult and frequently requires the use of a few capacitors in parallel to reduce the series inductance in the capacitor leads and body. Alternatively, the lead-inductance problem may be solved by formation of a self-resonant series circuit between the capacitor and its leads at the operating frequency. This method is extremely effective, but may restrict stage bandwidth.

MODULATION (AM, FM, SSB)

Amplitude modulation of the collector supply of a transistor output stage does not result in full modulation. During downmodulation, a portion of the rf drive feeds through the transistor. Better modulation characteristics can be obtained by modulation of the supply to at least the last two stages in the transmitter chain. On the downward modulation swing, drive from the preceding modulated stages is reduced, and less feed-through power in the output results. Flattening of the rf output during up-modulation is reduced because of the increased drive from the modulated lower-level stages.

The modulated stages must be operated at half their normal voltage levels to avoid high collector-voltage swings that may exceed transistor collector-to-emitter breakdown ratings. RF stability of the modulated stages should be checked for the entire excursion of the modulating signal.

Amplitude modulation of transistor transmitters may also be obtained by modulation of the lower-level stages and operation of the higher-level stages in a linear mode. The lower efficiencies and higher heat dissipation of the linear stages override any advantages that are derived from the reduced audio-drive requirements; as a result, this approach is not economically practical.

Frequency modulation involves a shift of carrier frequency only. Carrier deviations are usually very small and present no problems in amplifier bandwidth. For example, maximum carrier deviations in the 50-MHz and 150-MHz mobile bands are only 5 kHz. Because there is no amplitude variation, class C rf transistor stages have no problems handling frequency modulation.

Single-sideband (SSB) modulation requires that all stages after the modulator operate in a linear mode to avoid intermodulation-distortion products near the carrier frequency. In many SSB applications, channel spacing is close, and excessive distortion results in adjacent-channel interference. Distortion is effectively reduced by class B operation of the rf stages, with close attention to biasing the transistor base-to-emitter junction in a near-linear region.

CHARACTERIZATION OF LARGE-SIGNAL RF POWER TRANSISTORS

The values of large-signal transistor parameters, such as the S and Y parameters, are different from those of small-signal transistors because (1) the values of transistor parameters change with power levels, and (2) the harmonic-frequency components that exist in a large-signal rf power amplifier must be considered in addition to the fundamental-frequency sinusoidal component in a small-signal amplifier. RF power-transistor characteristics are normally specified for a given circuit in a specific application.

The design of rf power-amplifier circuits involves the determination of dynamic input and load impedances. Before the input circuit is designed, the input impedance at the emitter-to-base terminals of the packaged transistor must be known at the drive-power frequency. Before the output circuit is designed, the load impedance presented to the collector terminal must be known at the fundamental frequency. These dynamic impedances are difficult to calculate at microwave frequencies because transistor parameters such as S_{11} and S_{22} vary considerably under large-signal operation and also change with the power level. Small-signal equations that might serve as useful guides for transistor design cannot be applied rigorously to large-signal circuits. Because large-signal representation of rf power transistors has not yet been developed, transistor dynamic impedances are best determined experimentally with slotted-line or vector voltmeter measurement techniques.

The system used for determination of transistor impedances under operating conditions is shown in Fig. 25. This system consists of a well-padded power signal generator, a directional coupler (or reflectometer) for monitoring the input reflected power, an input triple-stub tuner, an input low-impedance line section, the transistor holder (or test jig), an output line section, a bias tee, an

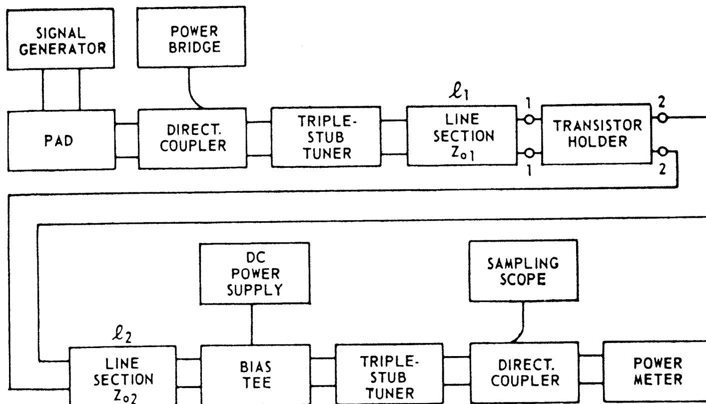


Figure 25. Set-up for measurement of rf transistor dynamic impedances.

output triple-stub tuner, another directional coupler for monitoring the output waveform or frequency, and an output power meter. For a given frequency and input power level, the input and output tuners are adjusted for maximum power output and minimum input reflected power. Once the system has been properly tuned, the impedance across terminals 1-1 (with the transistor disconnected) is measured at the same frequency in a slotted-line set-up or with the vector voltmeter. The conjugate of this impedance is the dynamic input impedance of the transistor. Similarly, the impedance across terminals 2-2 (with the transistor disconnected) is the collector-load impedance presented to the transistor collector. Such measurements are performed at each frequency and power level. It should be noted that the circuit arrangement of Fig. 25 is also useful for testing the performance of the transistor. Thus, power output, power gain, and efficiency are readily determined.

MATCHING REQUIREMENTS

A simplified high-frequency equivalent circuit of an "overlay" type of transistor is shown in Fig. 26. This circuit is similar to the hybrid-pi equivalent circuit of a transistor except for the addition of the capacitance C_{bc}. This capacitance represents the high collector-to-base capacitance in the overlay transistor which is created by the large area of the collector-to-base junction together with the active area under the emitter. This capacitance and the capacitance $C_{b'c}$ vary nonlinearly with the collector-to-emitter voltage.

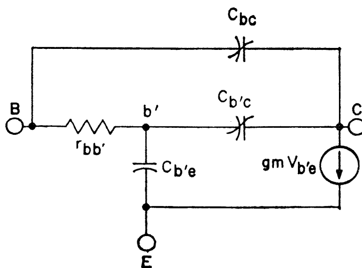


Figure 26. Simplified high-frequency equivalent circuit for an "overlay" transistor.

Maximum performance in a transistor rf amplifier can be obtained only if the base and collector terminals are properly terminated. The input network generally is required to match a 50-ohm source to the relatively low base-to-emitter impedance, which includes approximately 1 to 10 ohms of resistance and some series reactance. The output network must match a resistive component and the transistor output capacitance to a load impedance, which is generally about 50 ohms. In most applications, the output network also acts as a band-rejection filter to eliminate unwanted frequency components that may be included in the collector waveform. The filter presents a high impedance to these unwanted frequencies and also increases collector efficiency. The power output and collector-voltage swing determine the resistive component to be presented to the collector. The design and form of the output networks (resonant circuits for narrow-band operation or transmission lines for broad-band operation) are discussed in a later section.

MULTIPLE CONNECTION OF POWER TRANSISTORS

Many applications require more rf power than a single transistor

can supply. The parallel approach is the most widely used method for multiple connection of power transistors.

In parallel operation of transistors, steps must be taken to assure equal rf and thermal load sharing. In one approach, the transistors are connected directly in parallel. This approach, however, is not very practical from an economic standpoint because it requires the use of transistors that are exactly matched in efficiencies, power gains, terminal impedances, and thermal resistances. A more practical approach is to employ signal splitting in the input matching network. By use of adjustable components in each leg, adequate compensation can be made for variations in power gains and input impedances to assure equal load sharing between the transistors. For applications in which low supply voltages are used and high power outputs are desired, the output impedance of the rf amplifier is very low. For this reason, it is beneficial, in the interest of paralleling efficiency, to split the collector loads. By use of separate collector coils, the power outputs may be combined at higher impedance levels at which the effect of any asymmetry introduced by lead inductances is insignificant and resistive losses are less. The use of separate collector coils also permits individual collector currents to be monitored.

TRANSISTOR SELECTION

In selection of a transistor and circuit configuration for an rf power amplifier, the designer should be familiar with the following transistor and circuit characteristics:

(1) maximum transistor dissipation and derating,

(2) maximum collector current,

(3) maximum collector voltage,

(4) input and output impedance characteristics,

(5) high-frequency current-gain figure of merit (f_T),

(6) operational parameters such as efficiency, usable power output, power gain, and load-pulling capability.

Proper cooling must be provided to prevent destruction of the transistor because of overheating. Transistor dissipation and derating information reflect how well the heat generated within the transistor can be removed. This factor is determined by the junction-to-case thermal resistance of the transistor. A good rf power transistor is characterized by a low junction-to-case thermal resistance.

The current gain of an rf transistor varies approximately inversely with emitter current at high emitter-current levels. Peak collector current may be determined by the allowable amount of gain degradation at high frequencies. For applications in which amplitude modulation or low supply voltages are involved, peak current-handling capabilities are very important criteria to good performance.

The maximum collector voltage rating must be high enough so that junction breakdown does not occur under conditions of large collector voltage swing. The large voltage swing is produced under conditions of amplitude modulation or reactive loading because of load mismatch and circuit tuning operations.

Before the proper matching networks for an rf amplifier can be designed, transistor impedance

(or admittance) characteristics at the expected operating conditions of the circuit must be known. It is important that the value and dependence of transistor impedances on collector current, supply voltage, and operating frequency be defined.

The term f_T defines the frequency at which the current gain of a device is unity. This parameter is essential to the determination of the power-gain performance of an rf transistor at a particular frequency. Because f_T is current-dependent, it normally decreases at very high emitter currents. Therefore, it should be determined at the operating current levels of the circuit. A high f_T at high emitter or collector current levels characterizes a good rf transistor.

The operational parameters of an rf transistor can be considered to be those measured during the performance of a given circuit in which this type of transistor is used. The information displayed by these parameters is of a direct and practical interest. Operating efficiencies can normally be expected to vary between 30 and 80 per cent. Whenever possible, a circuit should employ transistors that have operational parameters specified at or near the operating conditions of the circuit so that comparisons can be made.

In some rf power applications, such as mobile radio, the transistors must withstand adverse conditions because high SWR's are produced by faulty transmission cables or antennas. The ability of a transistor to survive these faults is sometimes referred to as load-pulling or mismatch capability, and depends on transistor breakdown characteristics as well as circuit design. The load-pulling effects that the transistor may be subjected to can be determined by replacement of the rf load with a shorted stub and movement of the short through a half wavelength at the operating frequency. Dissipation capabilities of a transistor subjected to load pulling must be higher than normal to handle the additional device dissipation created by the mismatch.

CIRCUIT CONSIDERATIONS

Frequency stability is an important consideration in the design of high-frequency transistor circuits. Most instabilities occur at frequencies well below the frequency of operation because of the increased gain at lower frequencies. With the gain increasing at 6 dB per octave, any parasitic low-frequency resonant loop can set the circuit into oscillation. Such parasitic oscillations can result in possible destruction of the transistor. These low-frequency loops can usually be traced to inadequate bypassing of power-supply leads, circuit component self-resonances, or rf choke resonances with circuit or transistor capacitances. Supply bypassing can be effected by use of two capacitors, one for the operating frequency and another for the lower frequencies. For amplifiers operated in the 25-to-70-MHz range, sintered-electrode tantalum capacitors can provide excellent bypassing at all frequencies of concern. At uhf and higher frequencies, these capacitors may be lossy and therefore not effective for bypassing. High-Q ceramic bypass capacitors are better suited for uhf use. RF chokes, when used, should be low-Q types and should be kept

as small as possible to reduce circuit gain at lower frequencies. Chokes of the ferrite-bead variety have been used very successfully as base chokes. Collector rf chokes can be avoided by use of a coil in the matching network to apply dc to the collector.

Because of the variation of transistor parameters with changes in collector voltage and current, the stability of an rf transistor stage should be checked under all expected conditions of supply voltage, drive level, source mismatch, load mismatch, and, in the case of amplitude modulation, modulation swing.

Parametric oscillation is another form of instability that can occur in rf circuits that use power transistors. The transistor collector-to-base capacitance, as stated previously, is nonlinear and can cause oscillations that appear as low-level spurious frequencies not related to the carrier frequency.

Careful selection of components is necessary to obtain good performance in an rf transistor circuit. The components should be checked with an impedance bridge for parasitic impedances and self-resonances. When parasitic elements are encountered, their possible detrimental effects on circuit performance should be determined. This procedure helps the designer select coils and capacitances with low losses and high self-resonances (capacitors of the "bypass feed-through" or "mica postage stamp" variety can have very high self-resonances). Resistors used in rf current paths should have low series inductance and shunt capacitance (generally, low-wattage carbon resistors are quite acceptable).

Circuit layout and construction are also important for good performance. Chassis should be of a high-conductivity material such as copper or aluminum. Copper is sometimes preferable because of its higher conductivity and the fact that components can be soldered directly to the chassis. Another chassis approach now becoming popular is the use of double-side laminated printed-circuit boards. The circuit, in this approach, may be arranged so that all the conductors are on one side of the board. The opposite-side foil is then employed as an additional shield. Whenever possible, the chassis should be designed on a single plane to reduce chassis inductance and to minimize unwanted ground currents.

It must be remembered that, at rf frequencies, any conductor has an inductive and resistive impedance that can be significant when compared to other circuit impedances in a transistor amplifier. It follows, therefore, that wiring should be as direct and short as possible. It is also helpful to connect all grounds in a small area to prevent chassis inductance from causing common-impedance gain degeneration in the emitter circuit. Busses or straps may be used, but it should be remembered that these items have some inductance and that the point at which a component is connected to a buss can affect the circuit.

Coils used in input and output matching networks should be oriented to prevent unwanted coupling. In some applications, such as high-gain stages, coil orientation alone is not enough to prevent instability or strange tuning characteristics, and additional shielding between base and collector circuits must be used.

In common-emitter circuits, stage gain is very dependent on

the impedance in series with the emitter. Even very small amounts of inductive degeneration can drastically reduce circuit gain at high frequencies. Although emitter degeneration results in better stability, it should be kept as low as possible to provide good gain and to reduce tuning interaction and feedback between output and input circuits. The emitters of many rf power transistors are internally connected to the case so that the lowest possible emitter-lead inductance is achieved. This technique substantially reduces the problems encountered when the transistor is fastened directly to the chassis. If a transistor with a separate emitter lead is used, every attempt should be made to provide a low-inductance connection to the chassis, even to the point of connecting the chassis directly to the lead (or pin) as close to the transistor body as practicable. In extreme cases, emitter tuning by series resonating of the emitter-lead inductance is employed.

Another important area of concern involves the removal of heat generated by the transistor. Adequate thermal-dissipation capabilities must be provided; in the case of lower-power devices, the chassis itself may be used. Finned heat sinks and other means of increasing radiator area are used with higher-power devices. Consideration must also be given to ambient variations and mismatch conditions during tuning operations or load pulling, when transistor dissipation can increase. Under such conditions, the thermal resistance of the transistor may be the limiting factor, and may dictate either a change to another device of lower thermal resistance or a parallel mode of operation using the existing transistor.

Design Techniques for RF Power Amplifiers

THE design of an rf power amplifier consists primarily of selection of the optimum transistor type and determination of component values and desired circuit configuration for the input and output matching networks. The criteria for selection of the optimum type of transistor were discussed in the preceding section. Accordingly, this section is concerned primarily with the techniques and practices employed in the design of suitable matching networks.

DESIGN OBJECTIVES

Matching networks for rf amplifiers perform two important functions. First, they transform impedance levels as required by the active and fixed elements (e.g., transistor output to antenna impedance). Second, they provide frequency discrimination by virtue of the "quality factor" (Q) of the resonant circuit, transform harmonic energy into desired output-frequency energy, and prevent the presence of undesired frequency components in the output.

The design of matching circuits is based on the following requirements:

(1) desired or actual network output impedance specified by the series reactance X_s or shunt conductance G_p and shunt susceptance B_p;

(2) desired or actual network input impedance specified by R_s and X_s or G_p and B_p;

(3) loaded circuit Q calculated with input and output terminations connected.

The usual approach is to use L, T, or twin-T matching pads or tuned-transformer networks. More sophisticated systems may use exponential lines and balun transformers.

Input-Circuit Requirements

In practically all power-transistor stages, the input circuit must provide a match between a source impedance that is high compared to the transistor input impedance and the transistor input. When several stages are used, both the input and output impedance of a driver stage are usually higher than those of the following stage.

In most good rf transistors, the real part of the input impedance is usually low, in the order of a few tenths of an ohm to several

ohms. In a given transistor family, the resistive part of the common-emitter input impedance is always inversely proportional to the area of the transistor and. therefore, is inversely proportional to the power-output capability of the transistor, if equal emitter inductances are assumed.

The reactive part of the input impedance is a function of the transistor package inductance, as well as the input capacitance of the transistor itself. When the capacitive reactance is smaller than the inductive reactance, low-frequency feedback to the base may be excessive. It is not uncommon to use an inductive input for high-power large-area transistors because the input reactance is a series combination of the package lead inductance and the input capacitance of the transistor itself. Thus, at low frequencies, the input is capacitive, and at higher frequencies, it becomes inductive. At some single frequency, it is entirely resistive.

Output-Circuit Requirements

Although maximum power gain is obtained under matched conditions, a mismatch may be required to meet other requirements. Under some conditions, a mismatch may be necessary to obtain the required selectivity. In power amplifiers, the load impedance presented to the collector, R_L, is not made equal to the output resistance of the transistor. Instead, the value of R_L is dictated by the required power output and the peak dc collector voltage. The peak ac voltage is always less than the supply voltage because of the rf saturation voltage. The collector load resistance R_L may be expressed as follows:

$$R_L = (V_{CC})^2/2P_o \qquad (17)$$

Designs for tuned, untuned, narrow-band high-Q, and broad-band coupling networks are considered later under specific applications. In some cases, particularly mobile and aircraft transmitters, considerations for safe operation must include variations in the load, both in magnitude and phase. Safe-operation considerations may include protective circuits or actual test specifications imposed on the transistor to assure safe operation under the worst-load conditions.

NETWORK DESIGN

The basic components to be considered in the design of matching networks are shown in Fig. 27. For the input matching network, the source is assumed to be a generator that has a 50-ohm impedance. For the output matching

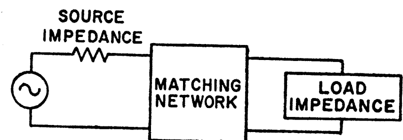


Figure 27. Basic components considered in the design of a matching network.

network, the source is the output of the transistor, which can be approximated as shown in Fig. 28

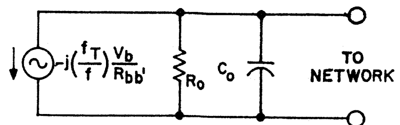


Figure 28. Equivalent circuit for the output of a transistor.

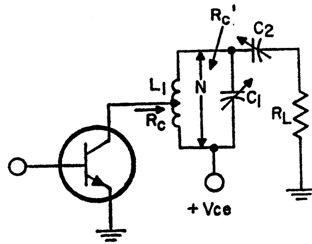
Output Circuit

When the dc supply voltage and power output are specified, the circuit designer must determine the load for the collector circuit [$R_L = (V_{CE})^2/2P_o$]. Because an rf power amplifier is usually designed to amplify a specific frequency or band of frequencies, tuned circuits are normally used as coupling networks. The choice of the output tuned circuit must be made with due regard to proper load matching and good tuned-circuit efficiency.

As a result of the large dynamic voltage and current swings in a class C rf power amplifier, the collector current contains a large amount of harmonics. This effect is caused primarily by the nonlinearity in the transfer characteristics of the transistor. The tuned coupling networks selected must offer a relatively high impedance to these harmonic currents and a low impedance to the fundamental current.

Class C rf power amplifiers are reverse-biased beyond collector-current cutoff; harmonic currents are generated in the collector which are comparable in amplitude to the fundamental component. However, if the impedance of the tuned circuit is sufficiently high at the harmonic frequencies, the amplitude of the harmonic currents is reduced and the contribution of these harmonic currents to the average current flowing in the collector is minimized. The collector power dissipation is therefore reduced, and the collector-circuit output efficiency is increased.

Figs. 29 and 30 illustrate the use of parallel tuned circuits to couple the load to the collector



FOR N:1 TURN RATIO

$$(1)\ R_C = \frac{V_{ce}^2}{2P_O}\ \text{(FOR CLASS C)}$$

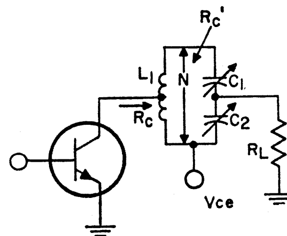
$$(2)\ X_{L1} = \frac{R_C'}{Q_L} = \frac{N^2 R_C}{Q_L}$$

$$(3)\ X_{C2} = R_L\sqrt{\frac{N^2 R_C}{R_L} - 1}$$

$$(4)\ X_{C1} = \frac{N^2 R_C}{Q_L}\cdot\frac{1}{\left(1 - \frac{X_{C2}}{Q_L R_L}\right)}$$

Figure 29. Tuned-circuit output coupling method and design equations in which output is transferred to load by a series coupling capacitor.

circuit. The collector electrode of the transistor is tapped down on the output coil. Capacitor C_1



FOR N:1 TURN RATIO

$$(1)\ R_C = \frac{V_{ce}^2}{2P_O}\ \text{(FOR CLASS C)}$$

$$(2)\ X_{L1} = \frac{N^2 R_C}{Q_L}$$

$$(3)\ X_{C1} = \frac{N^2 R_C Q_L}{(Q_L^2 + 1)}\left[1 - \frac{R_L}{Q_L X_{C2}}\right]$$

$$(4)\ X_{C2} = \frac{R_L}{\sqrt{\frac{(Q_L^2 + 1)R_L}{N^2 R_C} - 1}}$$

Figure 30. Tuned-circuit output coupling method and design equations in which output to the load is obtained from a capacitive voltage divider.

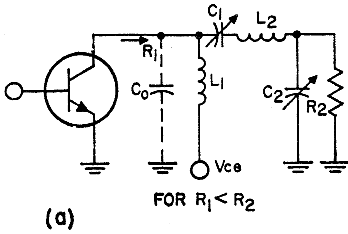
provides tuning for the fundamental frequency, and capacitor C_2 provides load matching of R_L to the tuned circuit. The transformed R_L across the entire tuned circuit is stepped down to match the collector by the proper turns ratio of the coil L_1. If the value of the inductance L_1 is chosen properly and the portion of the output-coil inductance between the collector and ground is sufficiently high, the harmonic portion of the collector current in the tuned circuit is small. Therefore, the contribution of the harmonic current to the dc component of current in the circuit is minimized. The use of a tapped-down connection of the collector to the coil maintains the loaded Q of the circuit and minimizes variation in the bandwidth of the output circuit with changes in the output capacitance of the transistor.

Although the circuits shown in Figs. 29 and 30 provide coupling of the load to the collector circuit with good harmonic-current suppression, the tuned-circuit networks have a serious limitation at very high frequencies. Because of the poor coefficient of coupling in coils at very high frequencies, the tap position is usually established empirically so that proper collector loading is achieved. Fig. 31 shows several suitable output coupling networks that provide the required collector loading and also suppress the circulation of collector harmonic currents. These networks are not dependent upon coupling coefficient for load-impedance transformation.

The collector output capacitance for the networks shown in Fig. 31 is included in the design equations. The collector output capacitance of a transistor varies considerably with the large dynamic swing of the collector-to-emitter voltage and is dependent upon both the collector supply voltage and the power output.

Input Circuit

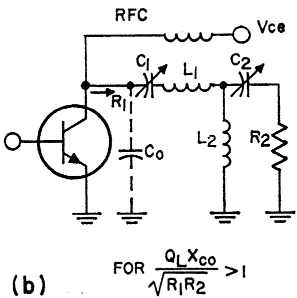
The input circuit of most transistors can be represented by a resistor r_{bb}' in series with a capacitor C_{in}. The input network must tune out the capacitance C_{in} and provide a purely resistive load to the collector of the driver stage. Fig. 32 shows several networks capable of coupling the base to the output of the driver stage and tuning out the input capacitance C_{in}. In the event that the transistor used has an inductive input, the reactance X_{ci} is made equal to zero, and the base inductance is included as part of inductor L_1 for networks such as that shown in Fig. 32(a) and is included as part of L_2 for networks of the type shown in Fig. 32(c). In Fig. 32(a), the input circuit is formed by the T network consisting of C_1, C_2, and L_1. If the value of the inductance L_1 is chosen so that its reactance is much greater than that of C_{in}, series tuning of the base-to-emitter circuit is obtained by L_1 and the parallel combination of C_2 and $(C_1 + C_o)$. Capacitors C_1 and C_o provide the impedance matching of the resultant input resistance r_{bb}' to the collector of the driving stage. Fig. 32(b) shows a T network in which the location of L_1 and C_2 is chosen so that the reactance of the capacitor is much greater than that of C_{in}; C_2 can then be used to step up r_{bb}' to an appropriate value across L_1. The resultant par-



(1) $X_{C1} = Q_L R_1$ (2) $X_{C2} = \dfrac{R_2}{\sqrt{\dfrac{R_2(Q_L^2+1)}{R_1 Q_L^2} - 1}}$

(3) $X_{L1} = \dfrac{Q_L R_1}{\left[\dfrac{Q_L R_1}{X_{co}} + 1\right]}$

(4) $X_{L2} = Q_L R_1 \left[1 + \dfrac{R_2}{Q_L X_{C2}}\right]$

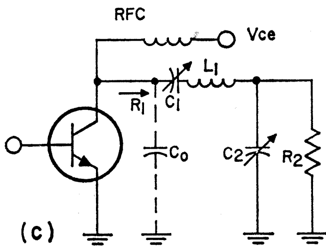


(1) $X_{L1} = \dfrac{Q_L X_{co}^2}{R_1}\left[1 - \dfrac{\sqrt{R_1 R_2}}{Q_2 X_{co}}\right]$

(2) $X_{L2} = X_{co}\sqrt{R_2/R_1}$

(3) $X_{C1} = \dfrac{Q_L X_{co}^2}{R_1}\left[1 - \dfrac{R_1}{Q_L X_{co}}\right]$

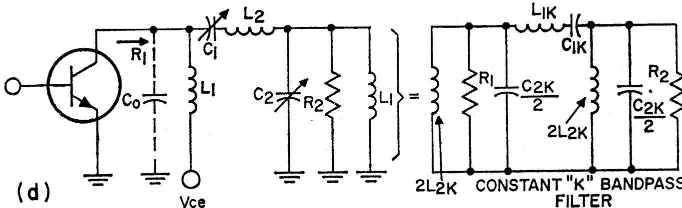
(4) $X_{C2} = \dfrac{R_2}{Q_L}\left[\dfrac{Q_L X_{co}}{\sqrt{R_1 R_2}} - 1\right]$



(1) $X_{C1} = \dfrac{Q_L X_{co}^2}{R_1}\left[1 - \dfrac{R_1}{Q_L X_{co}}\right]$

(2) $X_{C2} = \dfrac{R_2}{\sqrt{\dfrac{(Q_L^2+1)}{Q_L^2}\,\dfrac{R_1 R_2}{X_{co}^2} - 1}}$

(3) $X_{L1} = \dfrac{Q_L X_{co}^2}{R_1}\left[1 + \dfrac{R_2}{Q_L X_{C2}}\right]$

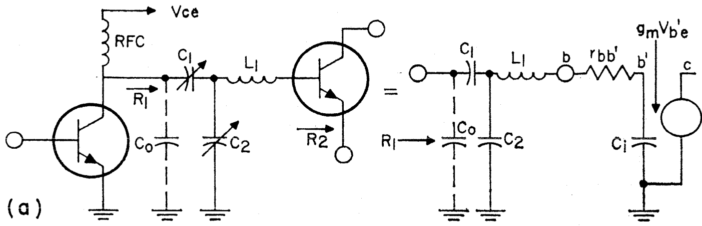


LET $C_{2K} = 2\,C_{OUT}$; $R_1 = R_2$; f_1=LOW FREQ. CUTOFF; f_2=HI-FREQ. CUTOFF

(1) $(f_2 - f_1) = \dfrac{1}{2\pi C_o R_L}$ (2) $L_2 = L_{1K}\,\dfrac{R_1}{\pi(f_2 - f_1)}$ (3) $L_1 = 2L_{2K} = \dfrac{(f_2 - f_1)R_1}{2\pi f_1 f_2}$

(4) $C_1 = C_{1K} = \dfrac{f_2 - f_1}{4\pi f_1 f_2 R_1}$ (5) $C_2 = C_o = \dfrac{C_{2K}}{2}$

Figure 31. Additional transistor output-coupling networks including transistor output capacitance.

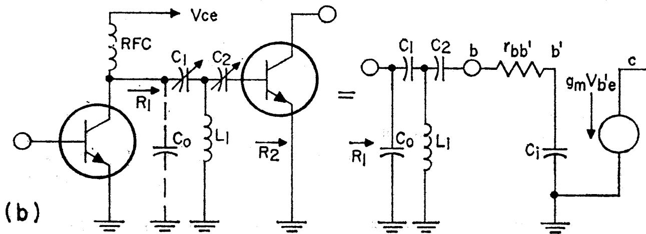


FOR $X_{L_1} \gg X_{c_i}$; $R_1 > R_2 = r_{bb'}$

(1) $X_{L_1} = Q_L R_2 = Q_L r_{bb'}$

(2) $X_{c_1} = X_{co}\left[\sqrt{\frac{(Q_L^2+1)\, r_{bb'}}{R_1}} - 1\right]$

(3) $X_{c_2} = \frac{r_{bb'}(Q_L^2+1)}{Q_L} \cdot \frac{1}{\left[1-\sqrt{\frac{R_1 r_{bb'}(Q_L^2+1)}{X_{co}^2 Q_L^2}}\right]}$

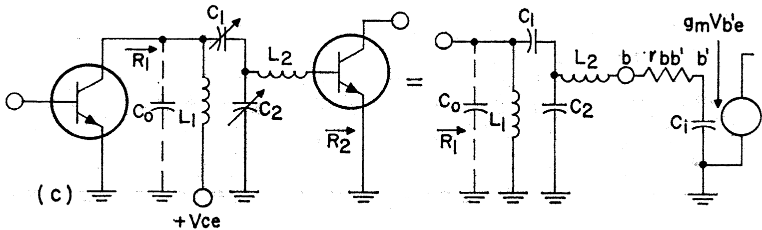


FOR $X_{c_2} \gg X_{c_i}$; $R_1 > R_2 = r_{bb'}$

(1) $X_{L_1} = \frac{R_2(Q_L^2+1)}{Q_L} \cdot \frac{1}{\left[1+\sqrt{\frac{R_1 R_2}{X_{co}^2}\cdot\frac{(Q_L^2+1)}{Q_L^2}}\right]} = \frac{r_{bb'}(Q_L^2+1)}{Q_L} \cdot \frac{1}{\left[1+\sqrt{\frac{R_1 r_{bb'}}{X_{co}^2}\cdot\frac{(Q_L^2+1)}{Q_L^2}}\right]}$

(2) $X_{c_1} = X_{co}\left[\sqrt{\frac{r_{bb'}(Q_L^2+1)}{R_1}} - 1\right]$

(3) $X_{c_2} = Q_L R_2 = Q_L r_{bb'}$



FOR $R_1 > R_2$; $R_2 = r_{bb'}$; $X_{L_2} \gg X_{c_i}$

(1) $X_{L_1} = \frac{R_1}{Q_L}$

(2) $X_{L_2} = \frac{R_2}{Q_L} \cdot \frac{\left[\sqrt{\frac{R_1}{R_2}} - 1\right]}{\left[1 - \frac{R_1}{Q_L X_{co}}\right]}$ (3) $X_{c_1} = \frac{R_1}{Q_L} \cdot \frac{\left[1 - \sqrt{\frac{R_2}{R_1}}\right]}{1 - \frac{R_1}{Q_L X_{co}}}$ (4) $X_{c_2} = \frac{R_1}{Q_L} \cdot \frac{\sqrt{\frac{R_2}{R_1}}}{\left[1 - \frac{R_1}{Q_L X_{co}}\right]}$

Figure 32. Transistor input-circuit coupling networks.

allel resistance across L_1 is transformed to the required collector load value by capacitors C_1 and C_o. Parallel resonance of the circuit is obtained by L_1 and the parallel combination ($C_1 + C_o$) and C_2.

The circuits shown in Fig. 32(a) and 32(c) require the collector of the driving transistor to be shunt-fed by a high-impedance rf choke. Fig. 32(c) shows a coupling network that eliminates the need for a choke. In this circuit, the collector of the driving transistor is parallel tuned, and the base-to-emitter junction of the output transistor is series tuned. Fig. 33 shows several other forms of coupling networks that can be used in rf power-amplifier designs.

USE OF IMPEDANCE-ADMITTANCE CHARTS IN NETWORK DESIGN

One of the most useful tools for designing matching networks is the impedance-admittance chart. This chart can be described simply as the plane of reflection coefficient for admittances, and provides an easier and faster method of circuit analysis than that offered by rectangular admittance or impedance charts. The chart displays graphically all ladder-type matching networks, and shows the applicable tuning ranges for variable components. Lumped-component values for a given frequency may be determined directly from the chart in normalized values. The chart can be used for idealized equivalent circuits, as well as for circuits that employ transformers or tapped coils.

Fig. 34 shows the basic layout of the chart. Shunt elements in a ladder follow the admittance circles (shown dotted). Values of shunt elements correspond to values on the intersection arcs. Series elements follow the impedance circles; corresponding values are read from corresponding intersection arcs.

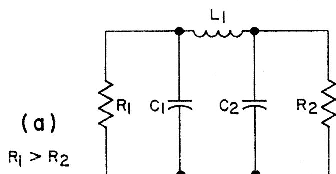
Rules for Plotting Networks and Components

When a single component L, C, or R is added to a known impedance, one of the following parameters does not change: resistance (R), reactance (X), conductance (G), or susceptance (B). (Non-ideal components must be divided into two separate ideal components; e.g., a lossy inductor into separate L and R components.) Therefore, the component follows that constant-parameter curve. For example, an inductor added in series with the circuit does not change the series resistance curve. The procedure for each type of component is listed in Table IV, which, together with Fig. 35, indicates the direction of travel along the curve and makes it unnecessary to determine the plus or minus sign on the reactances and susceptances.

Quality Factor, Q

The operating Q must be specified, together with the input and output impedances, in the design of a matching network. The magnitude of the operating Q is a compromise between efficiency and harmonic rejection.

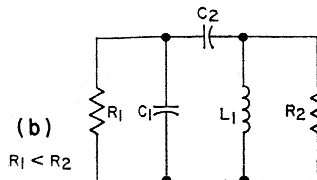
Unfortunately, the exact operating Q of a complex circuit cannot always be determined by calculations at a single frequency. When circuit design equations are used, this problem is circumvented by defining an operating Q which



(1) $X_{C1} = \frac{R1}{Q_L}$

(2) $X_{C2} = \frac{R2}{\sqrt{\frac{R2}{R1}(Q_L{}^2 + 1) - 1}}$

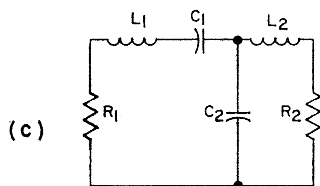
(3) $X_{L1} = \frac{Q_L R1}{Q_L{}^2 + 1}\left[1 + \frac{R2}{Q_L X_{C2}}\right]$



(1) $X_{C1} = \frac{R1}{Q_L}$

(2) $X_{C2} = \frac{Q_L R1}{(Q_L{}^2 + 1)}\left[\frac{R2}{Q_L X_{L1}} - 1\right]$

(3) $X_{L1} = \frac{R2}{\sqrt{\frac{R2}{R1}(Q_L{}^2 + 1) - 1}}$

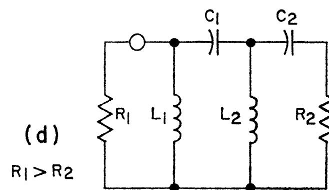


(1) $X_{L1} = Q_L R1$

(2) $X_{L2} = \frac{R2}{Q_L}\left[\sqrt{\frac{R1(Q_L{}^2 + 1)}{R2}} - 1\right]$

(3) $X_{C1} = \frac{R1(Q_L{}^2 + 1)}{Q_L}\left[1 - \sqrt{\frac{R2}{R1(Q_L{}^2 + 1)}}\right]$

(4) $X_{C2} = \frac{R1}{Q_L}\sqrt{\frac{R2(Q_L{}^2 + 1)}{R1}}$

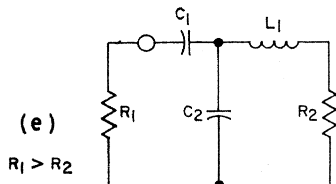


(1) $X_{L1} = \frac{Q_L R1}{\left[1 + \sqrt{\frac{R1}{R2}}\right]}$

(2) $X_{L2} = \frac{Q_L R2}{\left[1 + \sqrt{\frac{R2}{R1}}\right]}$

(3) $X_{C1} = Q_L R1\sqrt{\frac{R2}{R1}}$

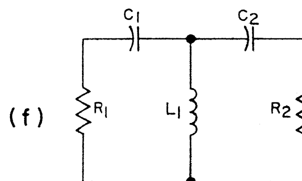
(4) $X_{C2} = Q_L R2$



(1) $X_{L1} = Q_L R2$

(2) $X_{C1} = R1\sqrt{\frac{R2(Q_L{}^2 + 1)}{R1} - 1}$

(3) $X_{C2} = \frac{R2(Q_L{}^2 + 1)}{Q_L} \cdot \frac{1}{\left[1 - \frac{X_{C1}}{Q_L R1}\right]}$



(1) $X_{C1} = R1\sqrt{\frac{R2(Q_L{}^2 + 1)}{R1} - 1}$

(2) $X_{C2} = Q_L R2$

(3) $X_{L1} = \frac{R2(Q_L{}^2 + 1)}{Q_L} \cdot \frac{1}{\left[1 + \frac{X_{C1}}{Q_L R1}\right]}$

Figure 33. Other suitable rf-amplifier networks for maximum power transfer.

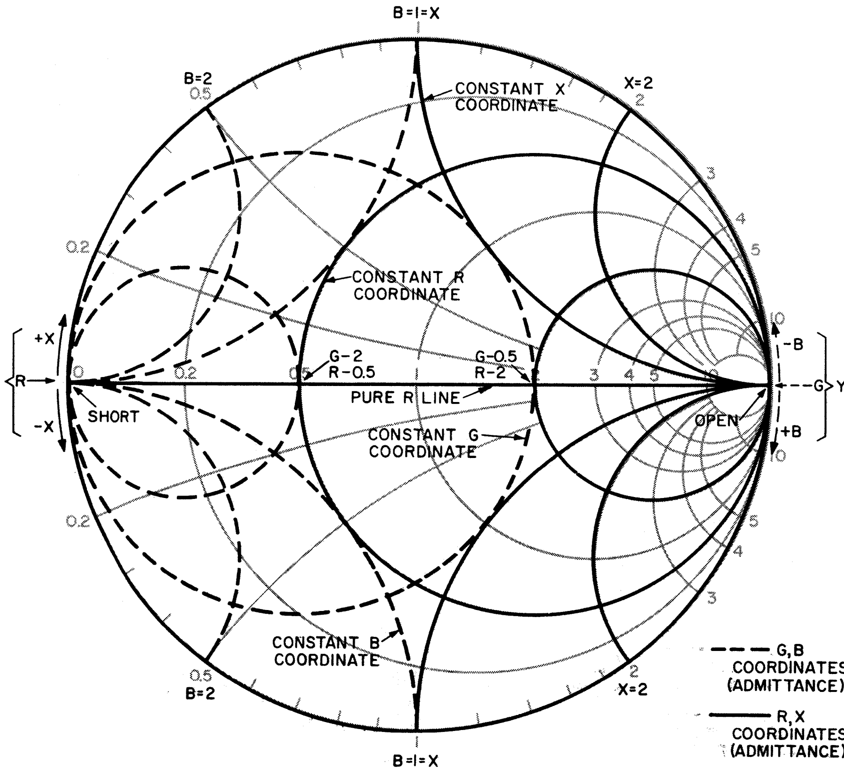


Figure 34. Impedance-admittance chart.

Table IV—Procedure for Plotting Component Values on Impedance-Admittance Chart

o Add	Use Chart	Follow a Curve of	Direction	Component Value
eries L	Z	Constant Series R	CW	$X_L = X_f - X_i$
eries C	Z	Constant Series R	CCW	$X_C = X_f - X_i$
eries R	Z	Constant X	toward open	$R_S = R_f - R_i$
hunt + L	Y	Constant Parallel R (G)	CCW	$B_L = B_f - B_i$
hunt + C	Y	Constant Parallel R (G)	CW	$B_C = B_f - B_i$
hunt + R	Y	Constant B	toward short	$1/R_p = G_f - G_i$

alculate the change in X, B, G and R by disregarding the + and — signs of the points on the hart. However, be sure to measure the entire change in X, R, B, or G. For example, a series apacitor which changes $X_i = 0.4$ inductive (above pure R line) to $X_i = 0.3$ capacitive (below ure R line) has a value of 0.7.

lote: Shunt refers to components with one terminal grounded and in parallel with the rest of the network.

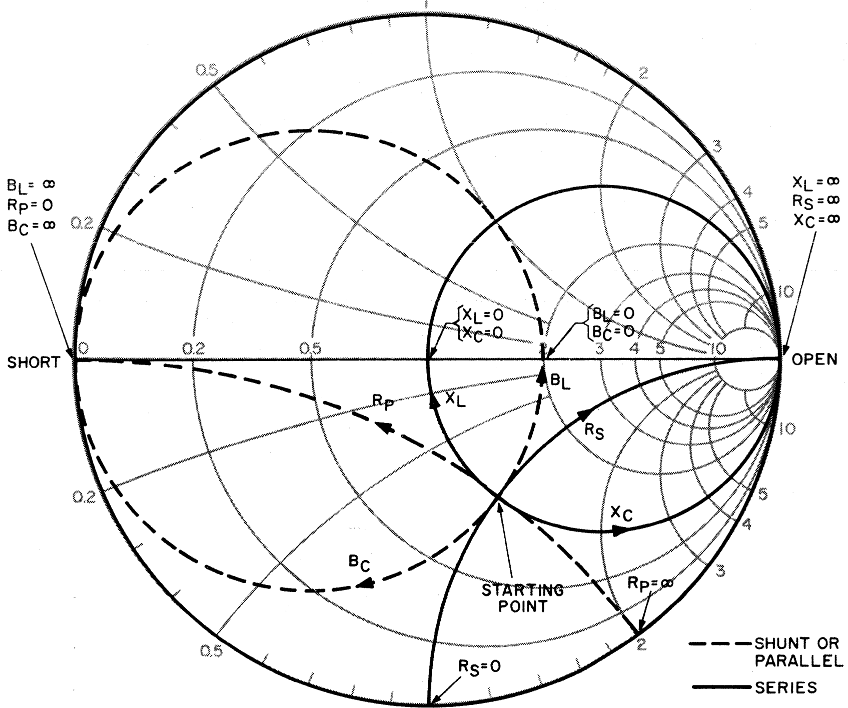


Figure 35. Method used to trace constant-parameter curves for matching-network components.

is easily calculated and which approximates the actual Q. The graphical technique uses the same type of approximation, but more simply and more visibly. The Q of each node of the circuit plot is determined by the constant-Q curves shown in Fig. 36. The node that has the highest Q dominates; this Q is then defined as the operating Q of the circuit.

Normalized Values

Impedance charts use normalized values. This graphical technique requires that normalized impedance and admittance values be consistent. The examples use $1\Omega = 1\mho = [50\Omega]$ (for impedances) $= [(1/50\mho]$ (for admittances). (Note: Brackets are used here, and in succeeding text and illustrations, to indicate the actual impedances or admittances represented by the normalized values.) The ohm (Ω) and mho ($\mho$) symbols are retained on the chart to distinguish between impedance and susceptance. A 50-ohm normalizing factor is used because this value represents a common rf-amplifier load impedance.

Mapping Technique

The matching network can be designed or analyzed by use of a network map, which is prepared by plotting each component (includ-

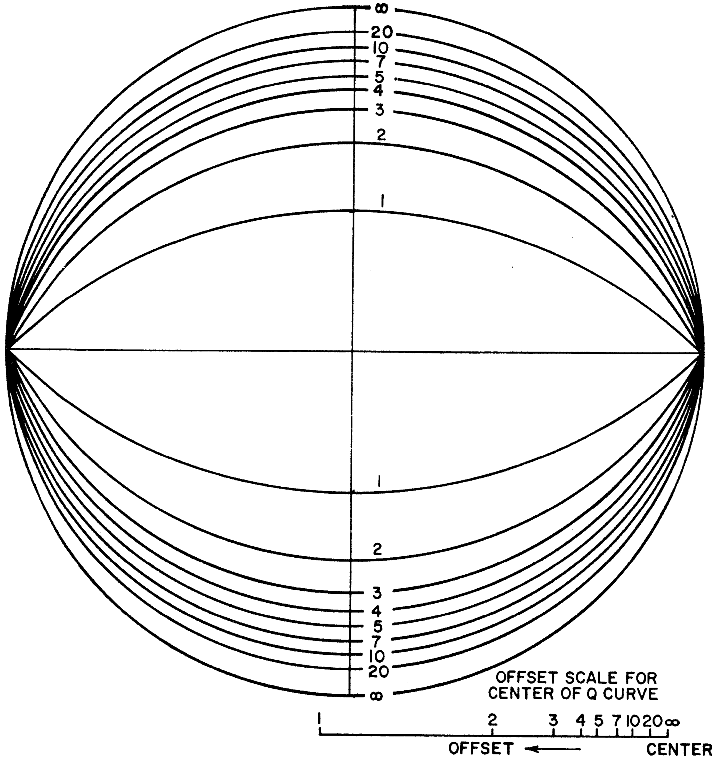


Figure 36. Chart of constant-Q curves.

ing input and output impedances) n the impedance chart. Dual im-edance-admittance charts, such as hat shown in Fig. 34, are avail-ble for this purpose, but the nany curves required make these harts difficult to read. A more ractical network map is prepared n tracing paper. The tracing pa-er is placed over either an imped-nce or an admittance chart to race curves, read values, and com-are impedances. Figs. 37 and 8 show simplified versions of a tandard impedance chart and a tandard admittance chart, respec-ively. The admittance coordi-ates have the same shape as the mpedance coordinates except that they have been rotated 180 degrees around the chart. This statement can be easily verified by superposition of Fig. 37 on Fig. 38 with the short and open points of one chart aligned with the open and short points, respectively, of the other chart.

The first step in the preparation of a network map is to trace the perimeter of the impedance chart (line of pure R) and those standard R, G, X, and B curves which are absolutely necessary. The "open" and the "short" points (or the pure R line) should also be marked to assist in accurate realignment of the tracing paper. The values of components may

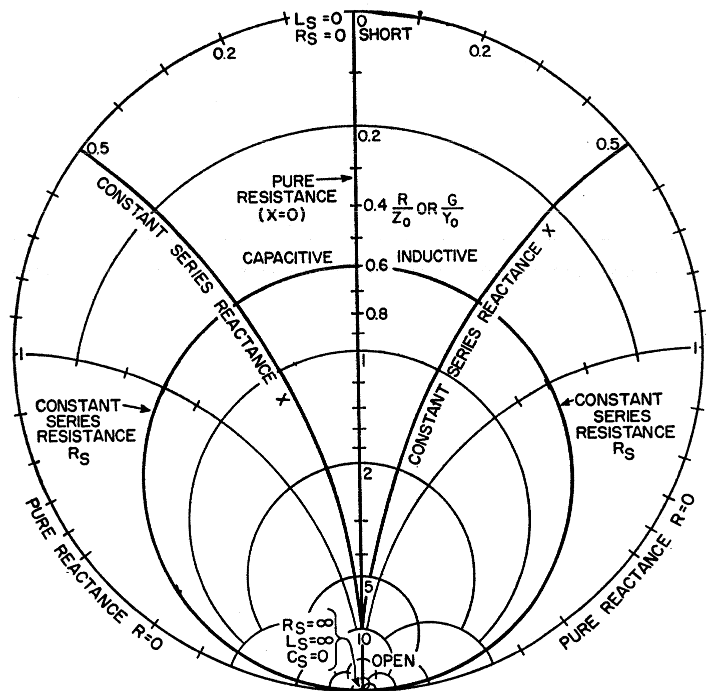


Figure 37. Impedance chart.

be determined with sufficient accuracy from curves that are traced from impedance or admittance charts placed under the tracing paper.

The following numerical examples illustrate the use of the mapping technique in the determination of the parameters of various types of matching networks.

Determination of Input Impedance—Fig. 40 shows a typical output matching network, together with the network map used to determine the required input impedance for this network. The reactance of each component in the matching network is known (from component values and operating frequency) and the input impedance is to be determined. The component reactances are plotted by use of curves traced from the Z, Y, and Q charts as follows: The output load impedance, 50 ohms normalized to 1 ohm, is located on the Z chart. Next, the series C_2 curve is plotted on a constant-R curve through 1 ohm, as indicated in Fig. 40. The series C_2 curve must change the reactance by its given value, 100 ohms normalized to 2 ohms, and the required normalized constant $X = 2$ ohms curve is traced from the Z chart. The C_2 curve ends at point A, where $B = 0.04$ mho, and the shunt L curve begins at this point. The shunt inductor has a normalized susceptance of 1 mho. The curve of this inductor follows the constant R_p admittance coordinate, passes through point A, and ends at $B_f = B_L - B_i = 1 - 0.4 = 0.6$ mho. (Table IV

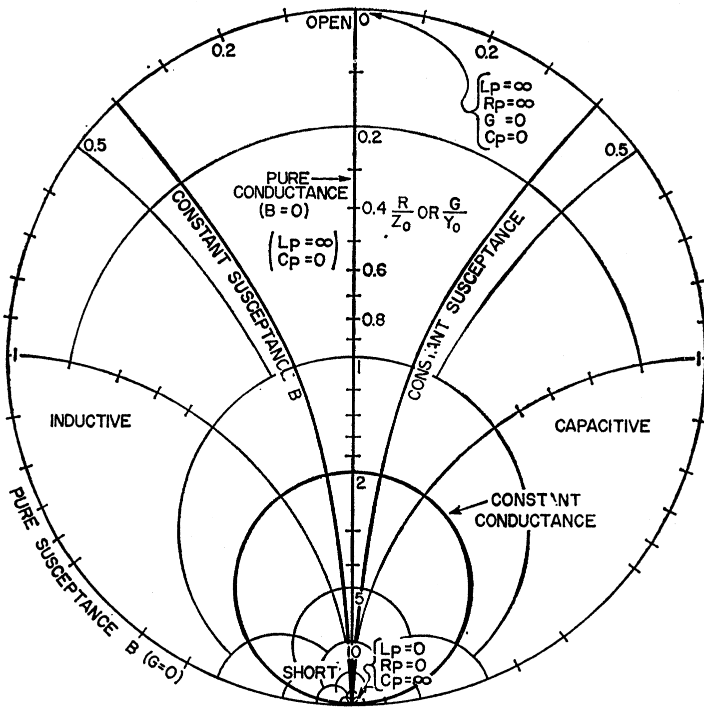


Figure 38. Admittance chart.

summarizes this procedure. It should be remembered that, in this case, the curve crosses the pure R axis.) This point is labeled point B on the network map shown in Fig. 40. Similarly, the series C_1 curve is plotted along the constant R_S coordinate, passes through point B, and ends at $X_f = X_c - X_i = 1.5 - 1.5 = 0$ ohms. The normalized value of input impedance is read on the Z chart as 0.5 ohm. The Q of the matching network is read from the Q curves at both points A and B. The Q is 2 at point A and 3 at point B. The higher value 3 is taken as most representative of the network Q.

Determination of Network Components—In the following examples, the graphical procedures used in the design of four different types of matching network are given. For the first type, a detailed explanation of the graphical procedure is given. For the other types, a tabular list of the steps required is considered sufficient because of the basic similarity of the graphical processes. The network maps for these examples show only the curves that are required to determine network parameters; all other curves are omitted for clarity. (In the examples, component curves are plotted as described in

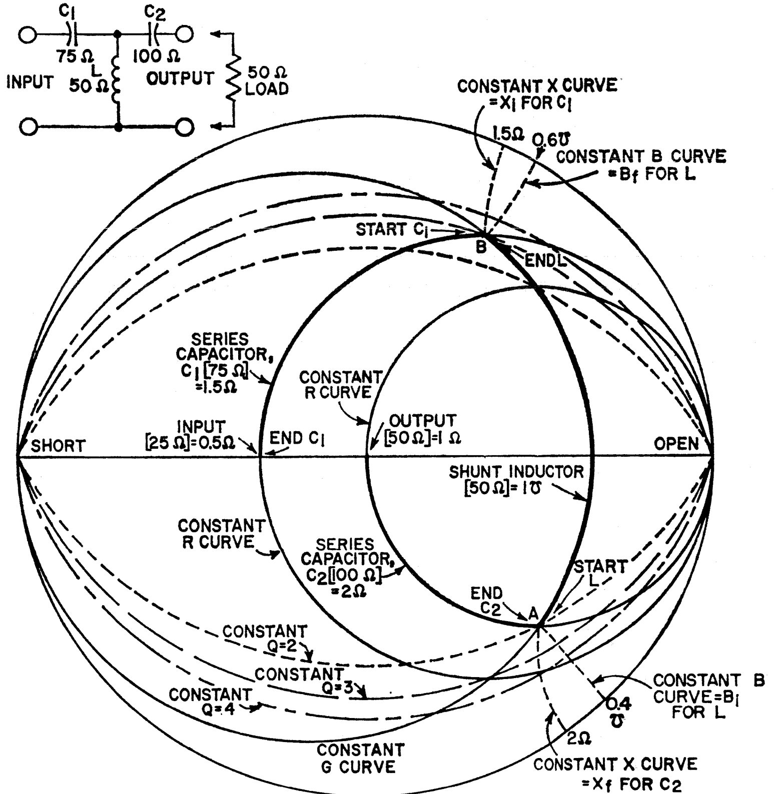


Figure 40. Typical output matching network and network map used to determine required input impedance for this network.

Table IV.)

1. Design of tapped-C network. Fig. 41 shows the circuit configuration and the network map used to determine the component values for a tapped-C matching network that is required to transform 50 ohms to 20 ohms with a Q of 6. The procedures used to prepare the network map are as follows:

(a) The normalized input and output points (i.e., points 1 + j0Ω and point 0.4 + j0Ω are located on the impedance-chart coordinates.

(b) The Q = 6 curve is traced from the Q chart, Fig. 36.

(c) The curve for the shunt L is traced along the constant R_p = 1.0℧ curve (from the admittance chart) from the termination point 1 + j0Ω to the intersection of this curve with the Q = 6 curve. This intersection is labeled A for further reference.

(d) A constant R_S curve for the series C_1 is traced from the impedance chart. The starting point for this curve is point A.

(e) The curve for the shunt C_2 is then traced between the

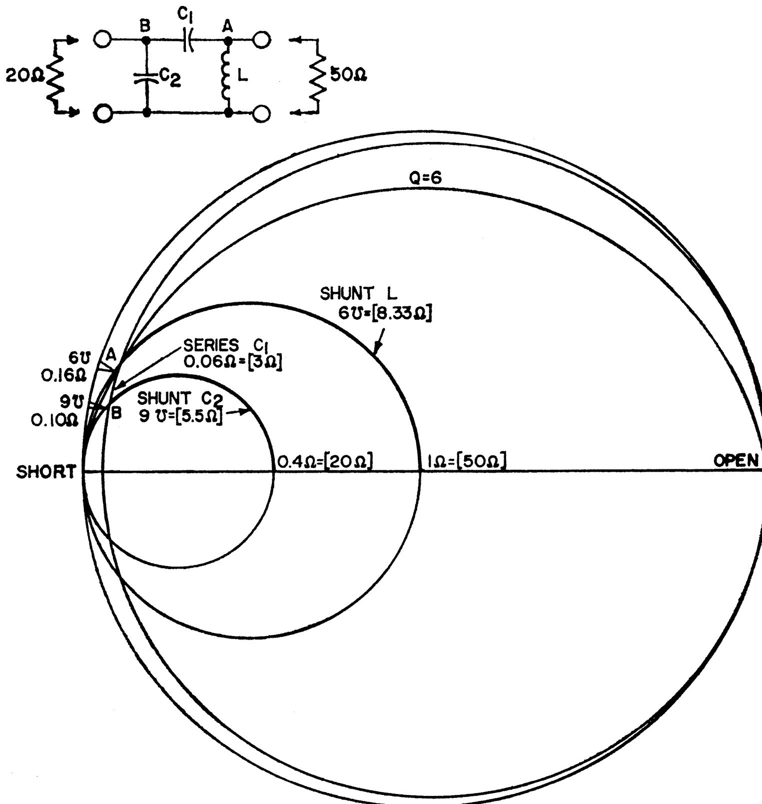


Figure 41. Circuit configuration and network map used to determine the component values for a tapped-C matching network.

termination point 0.4 + j0Ω and the intersection of this curve with that for the series C_1. The intersection of the C_1 and C_2 curves is labeled point B. (Although the intersection B is determined after the curve for the shunt C_2 is traced, this intersection is considered as the starting point for the shunt C_2 curve.)

(f) As a routine matter, the reactance X and the susceptance B values for the intersection points A and B are determined by means of series and parallel charts.

For point A, X = 0.16 ohm, and B = 6 mho.

For point B, X = 0.10 ohm, and B = 9 mho.

(g) As indicated in Table IV, normalized reactance values for the shunt inductor L, the series capacitor C_1, and the shunt capacitor C_2 are determined by subtraction of the values at the starting point of the curves for these components from the values at the end point of these curves. The following values are obtained:

$\Delta B_L = B_{(at\,A)} - B_{(at\,1+j0\Omega)}$
$= 6\mho - 0 = 6$ mhos

$\Delta X_{c1} = X_{(at\,B)} - X_{(at\,A)}$
$= 0.10 - 0.16 = 0.06$ ohm

$\Delta B_{C2} = B_{(at\,0.4+j0\Omega)} - B_{(at\,B)}$
$= 0 - 9\mho = 9$ mhos

(h) The actual reactance values for L, C, and C_2 can then be determined as follows:

$X_L = 50/\Delta B_L = 50/6$
$= 8.3$ ohms

$X_{C1} = 50(\Delta X_{C1}) = 50(0.06)$
$= 3$ ohms

$X_{C2} = 50/\Delta B_{C2} = 50/9$
$= 5.5$ ohms

(i) The component values for the filter can then be calculated on the basis of the reactances and the operating frequency.

The detailed step-by-step procedure given above is summarized in Table V. For one familiar with the basic graphical processes, this table provides sufficient information for the design of the filter network.

An intuitive analysis of the tapped-C network indicates that the shunt inductor L reduces the 50-ohm output impedance to the value represented by point A on the network map and that the shunt capacitor C_2 reduces the impedance of the 20-ohm input to a nearly equal value, as represented by point B. The series capacitance C_1 makes up the difference in the reactance of the two impedance points A and B and provides resonance. The values of both capacitors C_1 and C_2 must be changed together to maintain resonance when the input impedance is changed. The Q is determined by inductor L and the 50-ohm load impedance. At the input side, the transformation ratio is smaller, and the Q must be smaller.

2. **Design of pi network.** Fig. 42 shows the circuit configuration and the network map for a pi-type matching network required to transform 50 ohms to

Table V—Procedure for Determining Component Values for Tapped-C Matching Network

Component Curves

Step No.	COMPONENT	INITIAL POINT	FINAL POINT AND HOW IT IS DETERMINED
1	$Q = 6$		
2	Shunt L	$(1 + j0)\Omega$	A, determined by intersection of L and $Q = 6$ curves
3	Series C_1	A	B, determined by intersection of C_1 and C_2 curves
4	Shunt C_2	B	$0.4\Omega + j0\Omega = [20\Omega]$

Intersection Parameter Values from X and Y Charts

Step No.	INTERSECTION	REACTANCE X	SUSCEPTANCE B
5	A	0.16Ω	$6\mho$
6	B	0.10Ω	$9\mho$

Computing Component Values

Step No.	COMPONENT	INITIAL POINT	FINAL POINT	PARAMETER CHANGE (NORMALIZED VALUES)	COMPONENT VALUE
7	Shunt L	$(1 + j0)\Omega$	A	$\Delta B = 6 - 0 = 6\mho$	$X_L = 50/\Delta B = 8.33\Omega$
8	Series C_1	A	B	$\Delta X = 0.1 - 0.16 = 0.06\Omega$	$X_{C1} = \Delta X(50) = 3\Omega$
9	Shunt C_2	B	$(0.4 + j0)\Omega$	$\Delta B = 0 - 9 = 9\mho$	$X_{C2} = 50/\Delta B = 5.5\Omega$

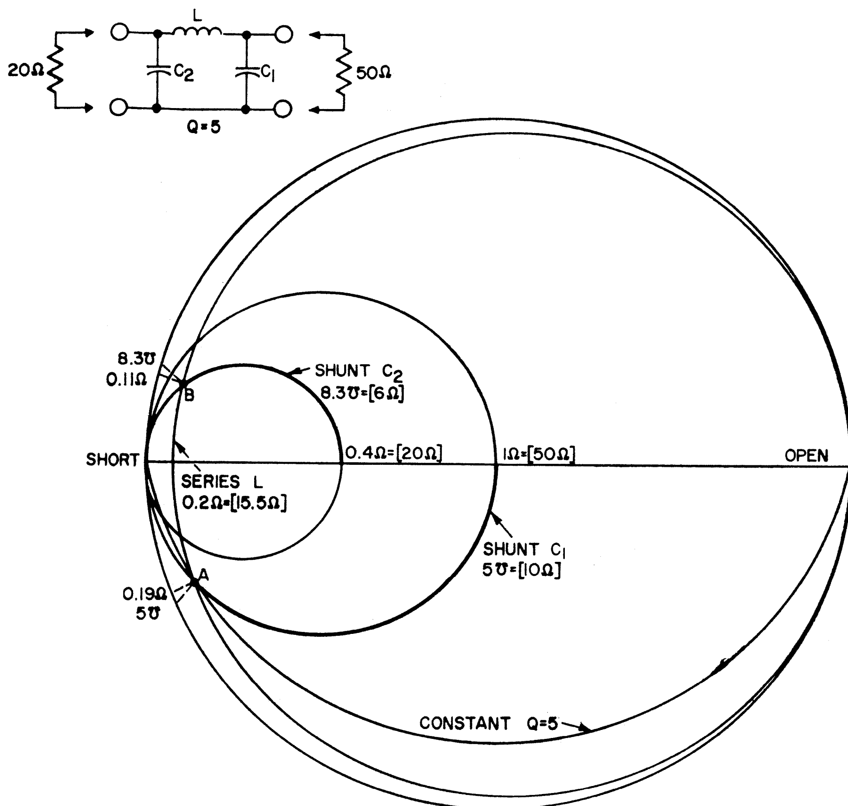


Figure 42. Circuit configuration and the network map used to determine the component values for a pi matching network.

20 ohms with a Q of 5. The network-mapping procedures used to determine the component values for the pi network are given in Table VI.

In the pi matching network, the shunt C across the 50-ohm output reduces the output impedance to the value represented by point A on the network map. The shunt capacitor across the 20-ohm input reduces the input impedance to the nearly equal value represented by point B. The Q at the input is smaller because the change in impedance is less. The series inductor connects the input and output and cancels the reactances of the two capacitors. The impedance transformation is determined by the difference in the input and output Q.

3. **Design of lossy-L network.** Fig. 43 shows the circuit configuration and network map for a lossy-L matching network required to transform 50 ohms to 10 ohms with a Q of 5. Table VII gives the graphical procedure used to determine the component values for this network.

In the lossy-L network, the series inductor increases the im-

Table VI—Procedure for Determining Component Values for Pi Matching Network

Component Curves

Step No.	COMPONENT	INITIAL POINT	FINAL POINT AND HOW IT IS DETERMINED
1	$Q = 5$		
2	Shunt C_1	$(1 + j0)\Omega$	A, intersection of C_1 and $Q = 5$
3	Series L	A	B, intersection of C_1 and C_2 (step 4)
4	Shunt C_2	B	$(0.4 + j0)\Omega$

Intersection Parameter Values

Step No.	INTERSECTION	SUSCEPTANCE B	REACTANCE X
5	A	5℧	0.19Ω
6	B	8.3℧	0.11Ω

Compute Component Values

Step No.	COMPONENT	INITIAL POINT	FINAL POINT	PARAMETER CHANGE (NORMALIZED VALUES)	COMPONENT VALUE
7	Shunt C_1	$(1 + j0)\Omega$	A	$\Delta B = 5 - 0 = 5$℧	$X_{C1} = 50/\Delta B = 10\Omega$
8	Series L	A	B	$\Delta X = 0.11 + 0.19 = 0.3\Omega$	$X_L = 50(\Delta X) = 15\Omega$
9	Shunt C_2	B	$(0.4 + j0)\Omega$	$\Delta B = 0 - 8.33 = 8.33$℧	$X_{C2} = 50/\Delta B = 6\Omega$

pedance of the 10-ohm input and determines the operating Q of the network. The series capacitor increases the impedance of the 50-ohm output, and the shunt capacitor tunes out the surplus reactance. In spite of the large impedance transformation (10 to 50 ohms), all component values have nearly equal impedances (56, 90, and 100 ohms). These relatively large values make the components quite practical, and are particularly advantageous for matching into the base of a transistor in which the impedance is only a few ohms.

4. **Design of network containing four unspecified components.** Fig. 44 shows the circuit configuration and network map for a matching section required to transform a 50-ohm load impedance to 12.5 ohms for the collector load impedance of a transistor amplifier that has a Q of 5. The transistor collector has a parallel output capacitive reactance of 250 ohms. This network has four unspecified components (L_1, L_2, C_1, and C_2) and three required conditions. The values for only three of the components can be determined by the graphical techniques; the value of the fourth component must be arbitrarily assigned. The value of L_1 is normally selected so that this component is nearly resonant with C_{in} at the operating frequency. In this example, however, the value selected for L_1 is small to demonstrate the flexibility in the choice.

The first step in the preparation of the map is to plot all known and assigned values. The three remaining components are then plotted and calculated as in examples 1, 2, and 3. The graphical procedures are outlined in Table VIII.

This network provides the best separation of impedance transformation, resonance adjustment, and operating Q. The components L_1 and C_{in} nearly resonate; however, perfect resonance is not required. The circuit tunes well even for large errors in C_{in} or L_1. The capacitor C_2 reduces the 50-ohm output impedance to the series re-

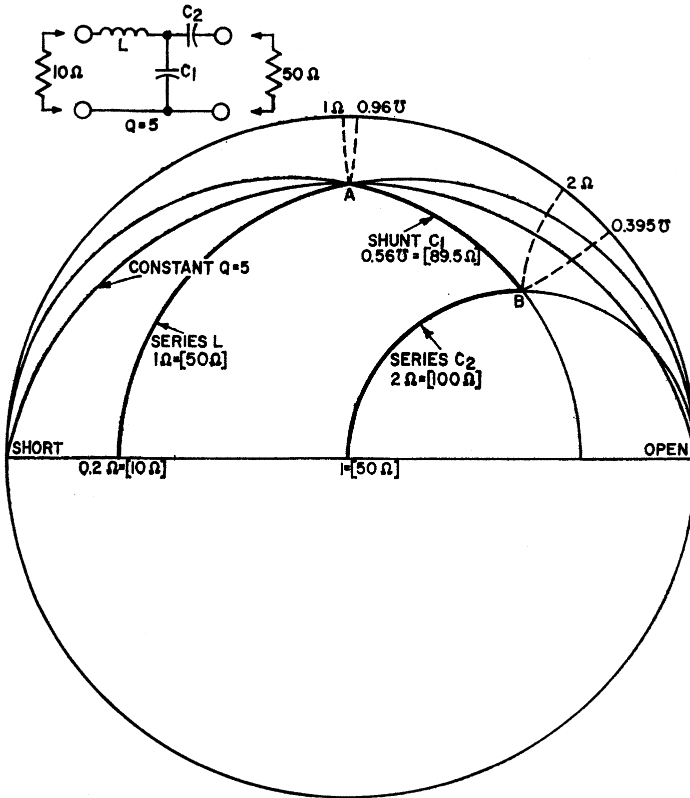


Figure 43. Circuit configuration and the network map used to determine the component values for a "lossy"-L matching network.

Table VII—Procedure for Determining Component Values for Lossy-L Matching Network

Component Curves

Step No.	COMPONENT	INITIAL POINT	FINAL POINT AND HOW IT IS DETERMINED
1	$Q = 5$		
2	Series L	0.2Ω	A, intersection of L and Q = 5 curves
3	Shunt C_1	A	B, intersection of C_1 and C_2 curves
4	Series C_2	B	1.0Ω

Determine Component Values

COMPONENT	INITIAL POINT	FINAL POINT	PARAMETER CHANGE (NORMALIZED VALUES)	COMPONENT VALUE
L	0.2Ω	$X = 1.0$	$\Delta X = 1.0 - 0 = 1.0\Omega$	$X_L = 50\Omega$
C_1	$A,B = 0.96$	$B, B = 0.395$	$\Delta B = 0.395 - 0.96 = 0.56℧$	$X_{CL} = 89.5\Omega$
C_2	$B,X = 2.0$	1.0	$\Delta X = 0 - 2.0 = 2.0\Omega$	$X_{C2} = 100\Omega$

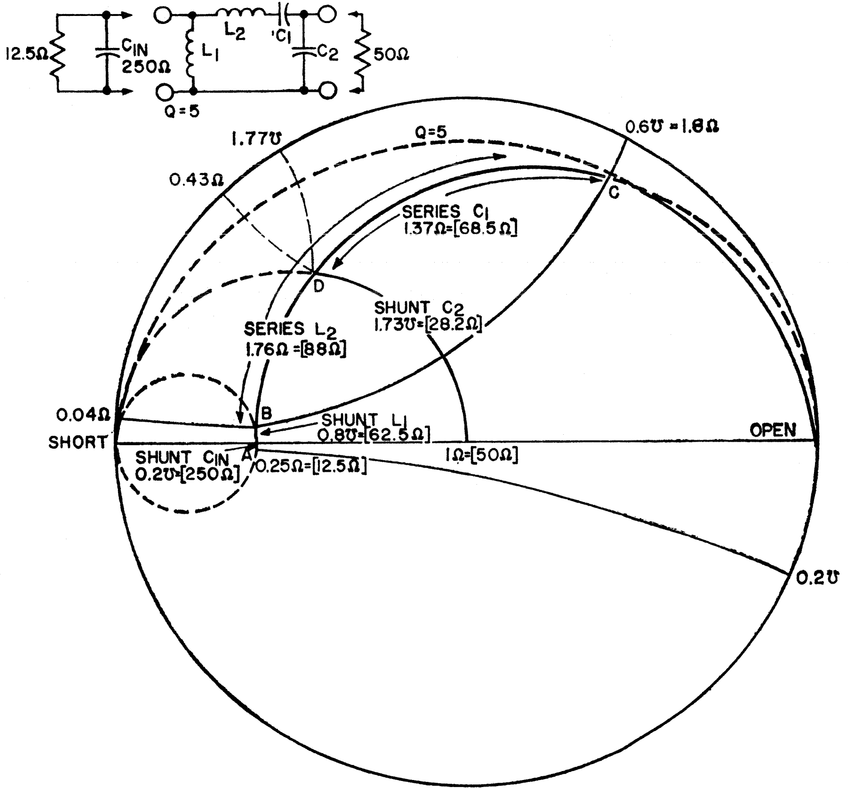


Figure 44. Circuit configuration and network map used to determine component values for matching network that includes four unspecified components.

Table VIII—Procedure for Determining Component Values for Matching Network in Which Four Unspecified Components Are Used

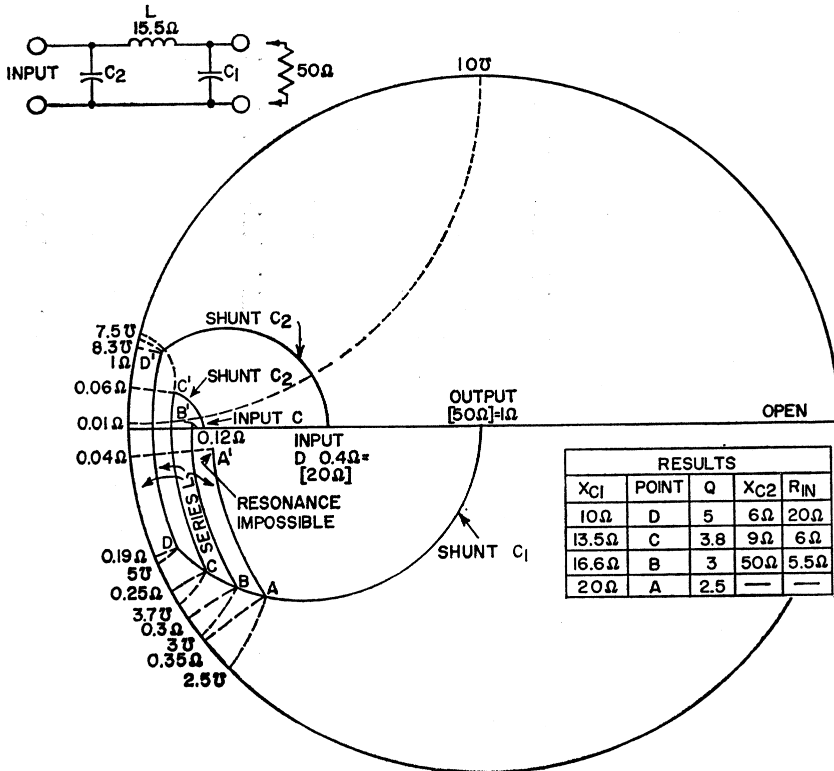
Step No.	COMPONENT	INITIAL POINT	FINAL POINT AND HOW IT IS DETERMINED
1	Shunt C_{in}	$(0.25 + j0)\Omega$	A, determined by given value for XC_{in}
2	Shunt L_1	A	B, determined by assigned value for X_{L1}
3	$Q = 5$	—	—
4	Series L_2	B	C, determined by intersection of L_2 and $Q = 5$
5	Series C_1	C	D, determined by intersection of C_1 and C_2 (step 6)
6	Shunt C_2	D	$(1 + j0)\Omega$

COMPONENT	INITIAL POINT	FINAL POINT	PARAMETER CHANGE (NORMALIZED VALUES)	REACTANCE VALUE
C_{in}	0.25Ω	$A,B = 0.2\mho$	$\Delta B = 0.20 = 0.2\mho$	$X_{Cin} = 250\Omega$
L_1	$A,B = 0.2\mho$	$B,B = 0.6\mho$	$\Delta B = 0.6 + 0.2 = 0.8\mho$	$X_{L1} = 62.5\Omega$
L_2	$B,X = 0.04\Omega$	$C,X = 1.8\Omega$	$\Delta X = 1.8 - 0.04 = 1.76\Omega$	$X_{L2} = 88\Omega$
C_1	$C,X = 1.8\Omega$	$D,X = 0.43\Omega$	$\Delta X = 0.43 - 1.8 = 1.37\Omega$	$X_{C1} = 68.5\Omega$
C_2	$D,B = 1.77\mho$	1.0Ω	$\Delta B = 0 - 1.77 = 1.77\mho$	$X_{C2} = 28.2\Omega$

sistance required at the input. The capacitor C_2, therefore, is the principal loading adjustment for the amplifier. The components L_2 and C_1 form a series resonant circuit which compensates for the differences in input and output reactances. Inductor L_1 and the 12.5-ohm input determine the operating Q relatively independent of resonance. The Q, therefore, is rather tightly controlled. Capacitor C_1 compensates for the additional inductor L_2 needed to provide the proper Q but not needed to match the input to the output. Therefore, C_1 is the resonance adjustment, and C_2 is the loading adjustment.

Effect of Component Changes in Network Design

A particular advantage of graphical network design is that changes in component values can be easily evaluated. The pi network in example 2 (Fig. 42) was designed for an input impedance of 20 ohms, but it may be changed by means of variable components. Two components must be varied to (1) change the impedance, and (2) maintain resonance. For this pi network, X_{C1} is increased in steps and L is kept constant, as shown in Fig. 45. Also shown is the X_{C2} required to produce



X_{C1}	POINT	Q	X_{C2}	R_{IN}
10Ω	D	5	6Ω	20Ω
13.5Ω	C	3.8	9Ω	6Ω
16.6Ω	B	3	50Ω	5.5Ω
20Ω	A	2.5	—	—

Figure 45. Circuit configuration for pi matching network and network map showing tuning range for variable components used in this type of matching network.

resonance, and the resulting Q and input impedance.

It should be noted that the first step increase in X_{C1} (35%) changes the Q and R_{in} greatly, but requires little change in X_{C2}. The second step increase in X_{C1} (23%) changes the Q slightly, changes the input very little (8%), but requires a large change in X_{C2}. Any further change in X_{C2} makes resonance impossible.

TRANSMISSION-LINE MATCHING TECHNIQUES

The network-design techniques discussed in the previous sections apply largely to lumped-constant circuits operated in the vhf and uhf ranges. At uhf and higher frequencies it may be more desirable to use short sections of transmission line to provide the reactive elements needed in the previous discussions. The Smith Chart is generally used in these determinations also. There are many special-case conditions which the circuit designer can use without resorting to the general transmission-line equation or the graphical method of the Smith Chart. A few of the more useful expressions are presented in this section.

Half-Wave Line Sections

Sections of uniform transmission lines which are electrically an integral number of half-wavelengths ($\lambda/2$) in length are useful in transferring an impedance from one point to another, i.e., the terminating impedances on the line are equal, or $Z_S = Z_L$.

Quarter-Wave Line Sections

Sections of uniform transmission lines which are electrically a quarter of a wavelength ($\lambda/4$) in length have a number of interesting and useful properties. A quarter-wave line which is short-circuited at one end provides a very high impedance at the open end. This property can be used to provide high-resistance stub supports for rf structures as well as to provide rf-choke action for dc bias circuits.

The quarter-wave lines are also useful as an impedance transformer between real impedances. The characteristic impedance can be determined as follows:

$$Z_o = (R_S \times R_L)^{1/2} \qquad (18)$$

where R_S is the source or input impedance and R_L is the load or output impedance.

If quarter-wave transformers are used to match a real impedance to an active device, as shown in Fig. 46, the reactive component of the complex impedance (the admittance) of the active device must be tuned out. For example, in the input circuit of a power-transistor amplifier circuit, the quarter-wave transformer matches the resistive com-

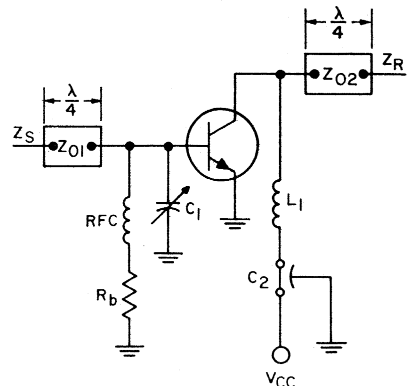


Figure 46. Quarter-wave transformers for rf power-transistor amplifiers.

ponent of the complex admittance of the device. An external capacitance C_0 or a stub provides the necessary susceptance needed to cancel the reactive component of the device. In the output portion of the circuit, a stub or a lumped element at the collector is used to bring the impedance to a real value and then to a quarter-wave line that goes to the actual load.

Direct transformation between the transistor (complex impedance) and a given source or load (real resistance) is also possible. The characteristic impedance Z_0 and length l of the transmission line required to provide direct transformation from a pure resistance R_1 to an impedance $Z_2 = R_2 + jX_2$ can be determined by use of the following equations:

$$Z_o = \sqrt{R_1 R_2} \times \sqrt{1 - \frac{X_2^2}{R_2 (R_1 - R_2)}} \quad (19)$$

$$\tan \beta l = Z_o \left(\frac{R_1 - R_2}{R_1 X_2} \right) \quad (20)$$

If the impedance Z_2 is a resistance (i.e., $X_2 = 0$), the expression for Z_0 reduces to the quarter-wave transformer equation, and $l = \lambda/4$.

Eighth-Wave Line Sections

Eighth-wave ($\lambda/8$) sections of uniform line have additional useful properties. If the eighth-wave section of line is terminated in a pure resistance, the input impedance will have a magnitude equal to the characteristic impedance Z_o of the line. Conversely, an eighth-wave section of line which is terminated in an impedance whose magnitude is equal to Z_o must have a real input impedance. Therefore, for an eighth-wave line section, Z_L is real if the following relationship is valid:

$$Z_o = |Z_s| = (R_S^2 + X_S^2)^{1/2} \quad (21)$$

Fig. 47 shows the use of eighth-wave transformers in a typical rf power-amplifier circuit.

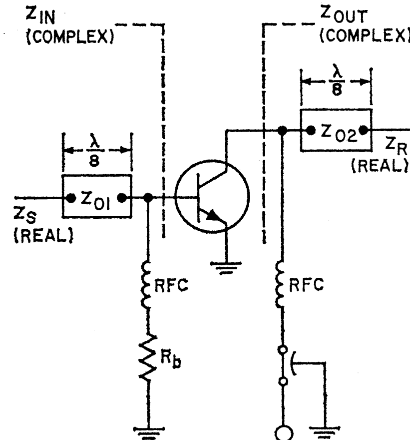


Figure 47. Eighth-wave transformers in a typical rf power-amplifier circuit.

The real impedance Z_L can be determined from the Smith Chart or by use of the following equation:

$$Z_L = \frac{R}{1 - \frac{X}{|Z_S|}} \quad (22)$$

where R and X are the real and imaginary parts, respectively, of the complex impedance Z_S.

Tapered Line Section

Quarter-wave or eighth-wave line sections in which the impedance changes exponentially (or hyperbolically) have additional properties useful to the circuit designer. These line sections can be tapered directly to a desired real impedance rather than a

predetermined impedance as was the case for a uniform line. In addition, because of the nature of the TEM mode of propagation in these tapered lines, substantial reductions in effective line lengths and increased transformation bandwidths are possible. The design of an amplifier circuit at a frequency of 2 GHz is described as an example. The dynamic input impedance Z_{in} and collector load impedance Z_{CL} of the transistor to be used are as follows:

$$Z_{in} = 7.5 + j\,8.0 \text{ ohms} \quad (23)$$

$$Z_{CL} = 6.5 + j\,33 \text{ ohms} \quad (24)$$

The amplifier uses an air-line input circuit and an air-line output circuit similar to that shown in Fig. 48.

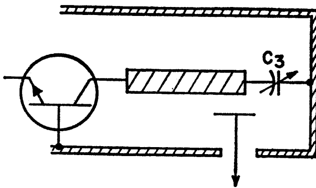


Figure 48. Capacitive-probe-coupled output cavity.

For the design of the output circuit, the optimum characteristic impedance Z_o of the output line is calculated from Eq. (21) to be 31 ohms. The collector load impedance Z_{CL} is normalized as follows:

$$Z_{CL}' = Z_{CL}/Z_o = 0.18 + j0.915 \quad (25)$$

Point Z_{CL}' is then located on the Smith Chart shown in Fig. 49 and rotated about the constant-VSWR circle toward the load. The intersection of the VSWR circle and the 1.39 constant-resistance circle is denoted as point Z_L' (the load resistance is assumed to be 50 ohms and the normalized load resistance is, therefore, 1.39 ohms). At point Z_L', the normalized impedance is given by

$$Z_L' = 1.39 - j\,3.3 \quad (26)$$

The load impedance Z_L is then equal to

$$Z_L = Z_oZ_L' = 36(1.39 - j\,3.3)$$

$$= 50 - j\,119 \text{ ohms} \quad (27)$$

The line length required to transform the transistor collector load impedance from 0.5 ohm to a load impedance of 50 ohms is determined from Fig. 49 to be equal to 0.33λ (where λ is the wavelength in air). At 2 GHz, λ is equal to 5.9 inches, and the length of output line is calculated to be 1.95 inches. A capacitive reactance component with a value equal to 119 ohms is needed to complete the output circuit, as shown in Fig. 50(a).

For the design of the input circuit, a characteristic impedance Z_o of 11 ohms is calculated from Eq. (21). The input impedance Z_{in} is normalized as follows:

$$Z_{in}' = (Z_{in}/Z_o) = 0.68 + 0.725 \quad (28)$$

Point Z_{in}' is then located on the Smith Chart shown in Fig. 51 and rotated about the constant-VSWR circle toward the generator to locate the intersection between the VSWR circle and the 4.55-ohm constant-resistance circle. (The driving-source impedance is assumed to be 50 ohms and the normalized source impedance is, therefore, 4.55 ohms.) However, Fig. 51 shows that such an intersection is not possible and that a more sophisticated input circuit is needed. One

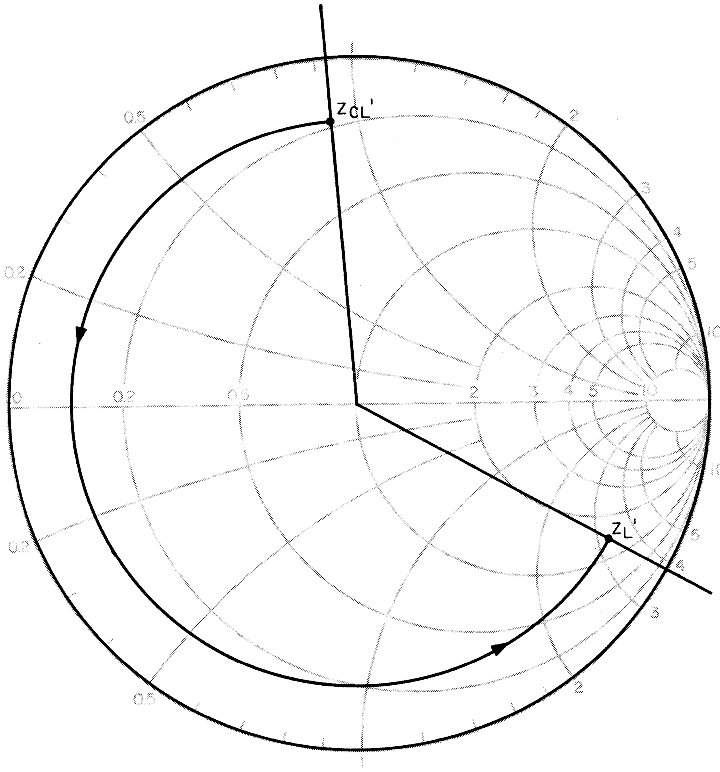


Figure 49. Smith-chart admittance plot for design of output transmission-line matching section.

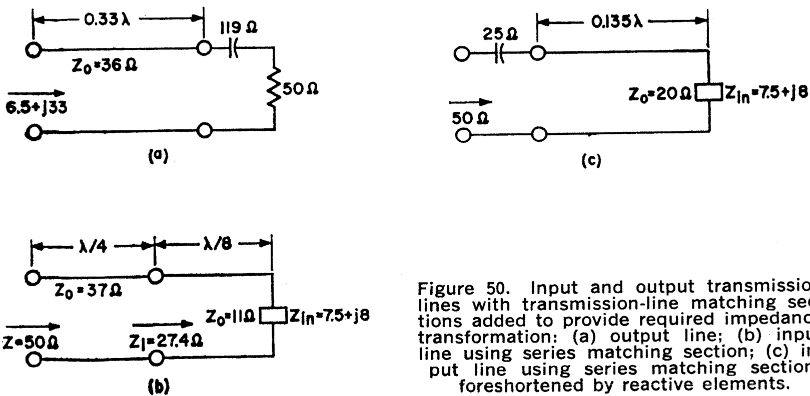


Figure 50. Input and output transmission lines with transmission-line matching sections added to provide required impedance transformation: (a) output line; (b) input line using series matching section; (c) input line using series matching section foreshortened by reactive elements.

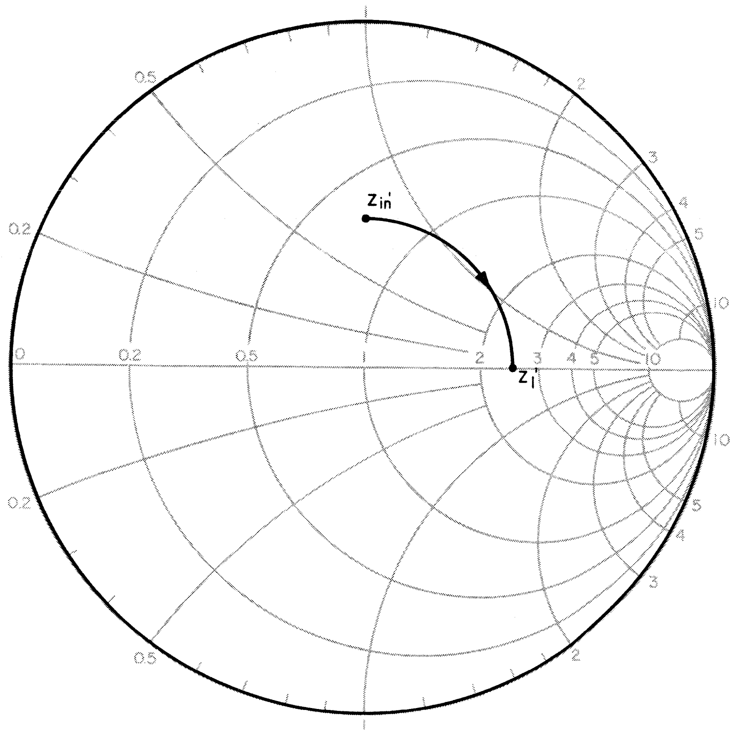


Figure 51. Smith-chart admittance plot used for design of input transmission-line matching sections.

possible circuit employs another section of line. For minimum VSWR in the added section of line, the line length for the 11-ohm line must be $\lambda/8$ or 0.75 inch, as discussed previously. This point, denoted as Z_1' in Fig. 51, is equal to 2.5 ohms; the impedance Z_1 is then 27.4 ohms. The characteristic impedance of the added line section required to transform a resistance of 27.4 ohms to a 50-ohm source is calculated from Eq. (18) to be 37 ohms. Such an input circuit is shown in Fig. 50(b). Another possible input circuit uses added reactive elements, as shown in Fig. 50(c), to foreshorten the additional line section.

The design of microstripline circuits is the same as that described for air-line circuits, except that the wavelength of the line must be modified by a factor of $1/\sqrt{\epsilon}$, where ϵ is the dielectric constant of the insulator material of the stripline.

Single-Sideband Systems

SINGLE-SIDEBAND communication systems have many advantages over AM and FM systems. In areas where reliability of transmission as well as power conservation are of prime concern, SSB transmitters are usually employed. The main advantages of SSB operation include reduced power consumption for effective transmission, reduced channel width to permit more transmitters to be operated within a given frequency range, and improved signal-to-noise ratio.

In a conventional 100-per-cent modulated AM transmitter, two-thirds of the total power delivered by the power amplifier is at the carrier frequency, and contributes nothing to the transmission of intelligence. The remaining third of the total radiated power is distributed equally between the two sidebands. Because both sidebands are identical in intelligence content, the transmission of one sideband would be sufficient. In AM, therefore, only one-sixth of the total rf power is fully utilized. In an SSB system, no power is transmitted in the suppressed sideband, and power in the carrier is greatly reduced or eliminated; as a result, the dc power requirement is substantially reduced. In other words, for the same dc input power, the peak useful output power of an SSB transmitter, in which the carrier is completely suppressed, is theoretically six times that of a conventional AM transmitter.

Another advantage of SSB transmission is that elimination of one sideband reduces the channel width required for transmission to one-half that required for AM transmission. Theoretically, therefore, two SSB transmitters can be operated within a frequency spectrum that is normally required for one AM transmitter.

In a single-sideband system, the signal-to-noise power ratio is eight times as great as that of a fully modulated double-sideband system for the same peak power.

ANALYSIS OF SSB SIGNAL

A single-sideband signal is usually generated at low level and then amplified through a chain of linear amplifiers to the desired power. The two most commonly used methods of generating sideband signals are with the filter-

type generator and the phasing-type generator.

It can be shown mathematically that a single-sideband signal is derived from an amplitude-modulated wave. If an rf carrier frequency is modulated by an audio frequency, the resulting AM wave can be expressed by the following equation:

$$e = E_o (1 + m \cos 2\pi f_m t) \sin 2\pi f_c t$$

$$e = E_o \sin 2\pi f_c t + m E_o \cos 2\pi f_m t \sin 2\pi f_c t \quad (29)$$

in which f_m is the audio modulating frequency, f_c is the carrier frequency, and m is the per-cent-modulation factor. Expansion of the last term of this equation into functions of sum and difference angles by the usual trigonometric formulas results in the following expression:

$$e = E_o \sin 2\pi f_c t + (m E_o/2) \sin 2\pi (f_c + f_m) t + (m E_o/2) \sin 2\pi (f_c - f_m) t \quad (30)$$

This equation contains three components, each of which represents a wave. The first wave, represented by the term $E_o \sin 2\pi f_c t$, is called the carrier. It is present with or without modulation and maintains a constant average amplitude at a frequency f_c. The other two components of the equation represent waves that have equal amplitude, but frequencies above and below the carrier frequency by the amount of the modulating frequency. These components contain identical intelligence and are called sideband frequencies. The amplitude of the sideband frequencies depends on the degree of modulation (m). The higher the m factor, the greater the "talk power." Because only the sidebands transmit intelligence and because each sideband is a mirror image of the other, it is reasonable to assume that if the carrier and one sideband are eliminated, the remaining sideband is adequate for transmission of intelligence. This technique is applied in single-sideband transmission.

As mentioned previously, the elimination of one sideband reduces the bandwidth required by one half. This advantage is not fully realized unless the transmitter has the capability to amplify a signal linearly without introducing distortion products. Excessive distortion nullifies the advantage of reduced bandwidth in SSB transmission by generating unwanted frequencies which occupy segments of the spectrum that are allocated for other transmitters. The main objection to this distortion is not that it seriously affects intelligibility of the signal in the passband, but that it radiates rf energy on both sides of the passband and interferes with adjacent channels.

LINEARITY TEST

For an amplifier to be linear, a relationship must exist such that the output voltage is directly proportional to the input voltage for all signal amplitudes. Because a single-frequency signal in a perfectly linear single-sideband system remains unchanged at all points in the signal path, the signal cannot be distinguished from a cw signal or from an unmodulated carrier of an AM transmitter. To measure the linearity of

an amplifier, it is necessary to use a signal that varies in amplitude. In the method commonly used to measure nonlinear distortion, two sine-wave voltages of different frequencies are applied to the amplifier input simultaneously, and the sum, difference, and various combination frequencies that are produced by nonlinearities of the amplifier are observed. A frequency difference of 1 to 2 kHz is used widely for this purpose. A typical two-tone signal without distortion, as displayed on a spectrum analyzer, is shown in Fig. 52. The resultant signal envelope

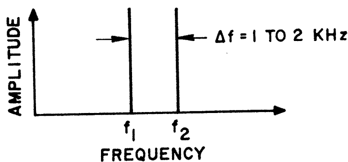


Figure 52. Frequency spectrum for a typical two-tone signal without distortion.

varies continuously between zero and maximum at an audio-frequency rate. When the signals are in phase, the peak of the two-frequency envelope is limited by the voltage and current ratings of the transistor to the same power rating as that for the single-frequency case. Because the amplitude of each two-tone frequency is equal to one-half the cw amplitude under peak power condition, the average power of one tone of a two-tone signal is one-fourth the single-frequency power. For two tones, conversely, the peak envelope power (PEP) rating of a single-sideband system is two times the average power rating.

INTERMODULATION DISTORTION

Nonlinearities in an amplifier generate intermodulation distortion IMD. The important IMD products are those close to the desired output frequency, which occur within the pass band and cannot be filtered out by normal tuned circuits. If f_1 and f_2 are the two desired output signals, third-order IM products take the form $2f_1 - f_2$ and $2f_2 - f_1$. The matching third-order terms are $2f_1 + f_2$ and $2f_2 + f_1$, but these matching terms correspond to frequencies near the third harmonic output of the amplifier and are greatly attenuated by tuned circuits. It is important to note that only odd-order distortion products appear near the fundamental frequency. The frequency spectrum shown in Fig. 53 illustrates the frequency relationship of some distortion products to the test signals f_1 and f_2. All such products are either in the difference-frequency region or in the harmonic regions of the original frequencies. Tuned circuits or filters following the nonlinear elements can effectively remove all products generated by the

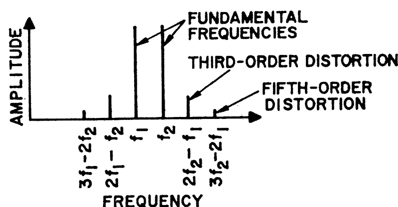


Figure 53. Frequency spectrum showing the frequency relationship of some distortion products to two test signals f_1 and f_2.

even-order components of curvature. Therefore, the second-order component that produces the second harmonic does not produce any distortion in a narrow-band SSB linear amplifier. This factor explains why class AB and class B rf amplifiers can be used as linear amplifiers in SSB equip-

ment even though the collector-current pulses contain large amounts of second-harmonic current. In a wideband linear application, however, it is possible for harmonics of the operating frequency to occur within the pass band of the output circuit. Biasing the output transistor further into class AB can greatly reduce the undesired harmonics. Operation of two transistors in the push-pull configuration can also result in cancellation of even harmonics in the output.

The IMD ratio (in dB) is the ratio of the amplitude of one test frequency to the amplitude of the strongest distortion product. A signal-to-distortion specification of −30 dB means that no distortion product will exceed this value for a two-tone signal level up to the PEP rating of the amplifier. A typical presentation of IMD for an RCA-2N6093 transistor at various output-power levels is shown in Fig. 54.

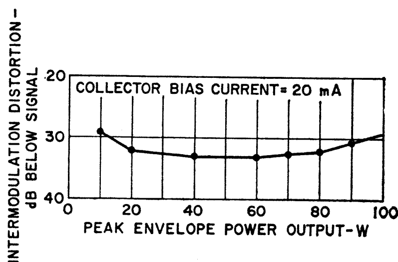


Figure 54. Typical intermodulation distortion in an RCA-2N6093 transistor at various output power levels.

TRANSISTOR REQUIREMENTS

Most high-frequency power transistors are designed for class C operation. Forward biasing of such devices for class AB operation places them in a region where second breakdown may occur. The susceptibility of a transistor to second breakdown is frequency-dependent. Experimental results indicate that the higher the frequency response of a transistor, the more severe the second-breakdown limitation becomes. For an rf power transistor, the second-breakdown energy level at high voltage (greater than 20 volts) becomes a small fraction of its rated maximum power dissipation. This behavior is one of the reasons that vacuum tubes have traditionally been used in single-sideband applications.

A power transistor designed especially for use as a linear amplifier is required to perform satisfactorily when forward-biased for class AB operation, as well as to exhibit the desired high-frequency response. Table IX lists transistors characterized for single-sideband applications.

The ability of the transistor to withstand second breakdown is improved by subdividing the emitter into many small sites and resistively ballasting the individual sites. An RCA-2N6093 transistor designed specifically for linear-amplifier service in SSB applications has an overlay structure with 540 parallel emitter sites, interconnected with metal fingers. Current-limiting resistors are placed in series with each emitter site between the metalizing and the emitter-to-base junction. The SSB RCA-2N6093 transistor has a high emitter-periphery-to-collector-area and a high emitter-periphery-to-emitter-area ratio, and thereby combines good high-current performance with low capacitance.

Physically, second breakdown is a local thermal-runaway effect induced by severe current concen-

Table IX—RCA Transistor Types for Single-Sideband (2-to-30-MHz) Applications

Type	Operating Frequency M(Hz)	Output Power (Min.) (W)	Collector Supply Voltage (V)	Power Gain (Min.) (dB)	Package Type	Primary Application
2N5070	30	25 (PEP)	28	13	TO-60	Linear Amplifier
2N6093	30	75 (PEP)	28	13	TO-217AA	Linear Amplifier

trations. The evidence of the random distribution of hot spots over the surface of the unit indicates that second breakdown may occur anywhere in the transistor. When a ballast resistor is used in each emitter site, current concentration is minimized. Fig. 55 is a schematic representation of the transistor showing the separate emitters with resistors in series with

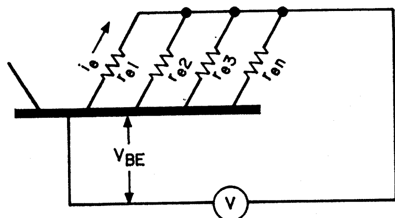


Figure 55. Schematic representation of the separate emitter sites in an "overlay" transistor with a resistor connected in series with each site.

each site. The voltage drop across each site is expressed by the following equation:

$$V = V_{BE} + i_e r_e \qquad (31)$$

Changes in V_{BE} have an exponential effect on the emitter current i_e, as follows:

$$i_e = i_s [e^{(q/kT) V_{BE}} - 1]$$
$$= i_s [e^{(q/kT) (V - i_e r_e)} - 1] \qquad (32)$$

This equation indicates that when a constant voltage V is applied across the emitter-to-base junction and resistor network, an increase in i_e at any one site causes a rise in the $i_e r_e$ voltage drop which, in turn, results in a decrease in the current to that site, i.e., the exponential term of the equation diminishes as the quantity $(V - i_e r_e)$ decreases. This condition effectively stabilizes that region. The addition of resistance to the emitters of the transistor has a degenerative effect on the device performance. However, if a large number of sites are connected in parallel, high-value individual resistors (r_e) can be sustained while a small total resistance (R_t) is still maintained at the input of the transistor, as indicated by the following relationship:

$$(1/R_t) = (1/r_{e1}) + (1/r_{e2}) + (1/r_{e3}) + \ldots + (1/r_{en}) \qquad (33)$$

A relatively large value of ballast resistance is desirable for prevention of second breakdown and for improvement of thermal stability and linearity of transfer characteristics. However, because ballast resistors are in series with the load, excessive ballasting can seriously degrade the rf performance of the transistor. Therefore, in a high-frequency power amplifier with low supply voltage, the impedance of the emitter resis-

tance can become an appreciable portion of the reflected load presented to the collector and, as a result, can limit the power output. In determining the proper emitter-resistance value, a compromise must be made empirically so that sufficient second-breakdown protection is provided without serious effects on rf performance.

The adverse effect of high ballast resistance, besides reduced rf output power, is the increase in saturation voltage. Viewed externally, the total saturation voltage also includes the voltage drop across the ballast resistance. This additional voltage makes the "soft" output characteristics of a transistor at high-current even softer. As a result, the available linear region through which the signal can swing is limited.

Examination of the relationship of intermodulation distortion to power output reveals that third-order distortion increases at both high and low output levels, as shown in Fig. 54. The inherent decrease in beta at high current, which causes variation in gain over a large portion of the collector dynamic characteristics, introduces additional distortion. The additional distortion is indicated by flattening of the peak of the sinusoidal swing.

The operation of a transistor near the saturation region has a pronounced effect on third-order distortion. All higher odd-order distortion products do not seem to be affected greatly by transistor operating conditions. The increase in distortion below 20 watts PEP can be attributed to lack of sufficient collector quiescent current. Nonlinearity caused by the voltage-current characteristic of the base-to-emitter junction affects distortion at low power levels. Third-order distortion is improved by use of a higher bias current, as shown in Fig. 56.

If collector bias current is set too high initially in an attempt to improve linearity at low power-output levels, the linear region of the collector characteristic is reduced. As a result, distortion because of saturation occurs much sooner. The controlling factor in determining the proper bias-current level is usually the maximum distortion that can be tolerated at a given power output. For a given transistor type, the bias point that yields the best compromise between linear performance and good collector efficiency must be determined experimentally. A

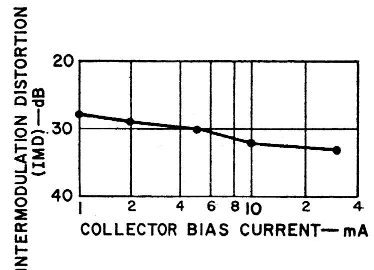


Figure 56. Intermodulation distortion as a function of collector bias current for the RCA-2N6093 SSB transistor.

collector bias current of from 2 to 20 milliamperes for the RCA-2N6093 SSB transistor is adequate to deliver 85 watts PEP. Fig. 57 shows a curve of power output as a function of supply voltage with distortion maintained at −30 dB.

Bias Control

Operation of the transistor in a class AB amplifier to improve linearity requires the use of a positive base voltage for an n-p-n silicon transistor. The magnitude

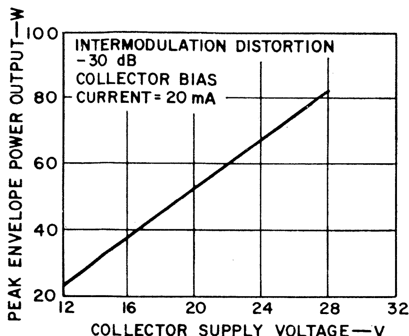


Figure 57. Peak envelope power as a function of collector supply voltage for the RCA-2N6093 SSB transistor.

of the positive voltage must be large enough to bias the transistor to a point slightly beyond the threshold of collector-current conduction. The class AB bias condition must be maintained over a wide temperature range to prevent an increase in idling current to the level at which the transistor can be destroyed as a result of thermal runaway and to minimize distortion that results from a shift in the quiescent point. Investigations of transistors that fail reveal that these devices exhibit a maximum V_{BE} and then go into a negative-resistance region as shown in Fig. 58. The onset of

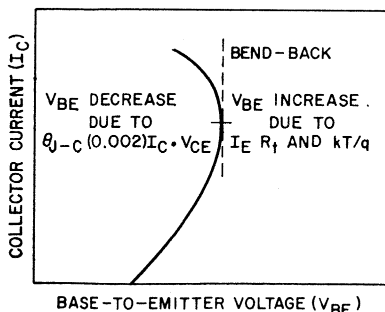


Figure 58. The bend-back phenomenon.

negative resistance, called bend-back, results in a runaway condition that ultimately destroys the transistor.

In most linear applications where the operating point of the device is biased with a voltage source, this I_C-V_{BE} curve becomes an accurate means of predicting device stability. It is difficult to maintain a stable quiescent point of a transistor with low bend-back. Laboratory results indicate that a minimum bend-back current of 1 ampere at 22 volts is needed for a transistor to operate safely at 40-per-cent efficiency with approximately 50 watts of dissipation.

Bend-back occurs when the increase of V_{BE} with collector current is just balanced by the decrease in V_{BE} caused by junction-temperature rise. Therefore, at bend back,

$$kT/q + I_E R_t = \theta_{J-C}(0.002V/°C\ I_C V_{CE} \quad (34)$$

where

$kT/q = 0.032$ volt at 100°C

R_t = total ballast resistance

θ_{jc} = junction-to-case thermal resistance

$0.002V/°C$ = base-to-emitter junction temperature coefficient

I_E = emitter current

I_C = collector current

V_{CE} = collector-to-emitter voltage

If $I_C = I_E$, Eq. (34) can be solved to find I_E at bend-back:

$$I_E = \frac{-kT/q}{R_t - \theta_{J-C}(0.002V/°C)V_{CE}} \quad (35)$$

Thermal runaway can be attributed to the fact that the base-to-emitter junction of a transistor has a negative temperature coefficient. For example, the RCA-

2N6093 transistor is forward-biased by 0.65 volt to produce a quiescent collector current of about 20 milliamperes at $V_{CC} =$ 28 volts. This operating point is shown at point A in Fig. 59 When rf drive is applied, the collector current increases to 3 amperes. If the efficiency is 40 per cent, the power dissipated in the transistor is given by

$$P_{diss.} = 28 \times 3\ (1 - 0.40) = 50 \text{ watts}$$

If the ambient temperature is 25°C, the case temperature is 50°C, and the thermal resistance is 1.5°C per watt, the junction temperature is given by

$$T_J = T_C + P_{diss.}\ \theta_{J-C} = 50 + 50 \times 1.5 = 125°C. \quad (36)$$

The junction temperature is thus 100°C above ambient temperature. At this junction temperature, the V_{BE} required to maintain a collector current of 20 milliamperes is only $0.65 - 100 \times 0.002 =$ 0.45 volt, as shown at point B. If the bias voltage is fixed at 0.65 volt, however, and the drive is removed instantaneously, the quiescent current will no longer be 20 milliamperes. Instead, the collector current will move to point C, where the operating point falls outside of the safe area of Fig. 60. Therefore, catastrophe failure will occur as a result of thermal runaway.

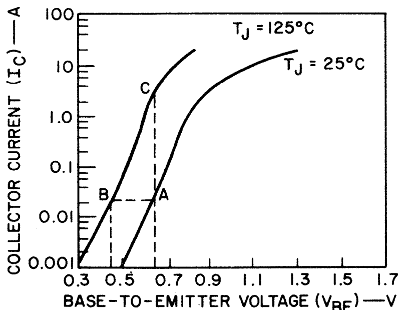


Figure 59. Collector current as a function of base-to-emitter voltage in the RCA-2N6093 for two values of junction temperature.

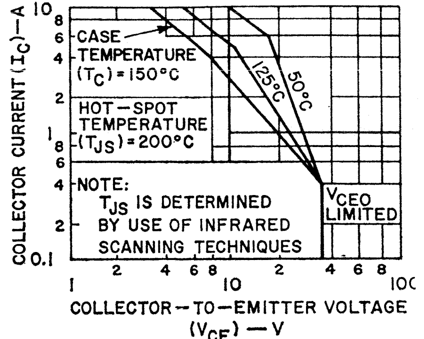


Figure 60. Safe area for dc operation of the RCA-2N6093.

Compensating Diode

To provide a bias voltage that varies with temperature in the same manner as V_{BE} of the transistor, the 2N6093 incorporates a compensating diode as shown in Fig. 61(a). To assure fast thermal response time, this diode is mounted on the same beryllia disc as the transistor chip. The diode, forward-biased through R_{Bias}, serves as a temperature-sensing element. The voltage developed across the diode is amplified to provide a "stiff" bias-voltage source. Fig. 61(b) shows the block diagram of a temperature-compensated 30-MHz linear power amplifier that uses this transistor.

NARROW-BAND LINEAR AMPLIFIER

A bias-compensation circuit is included in the 30-MHz, 75-watt (PEP) amplifier shown in Fig. 62. The current amplifier uses Q1 and Q2 in a differential-amplifier arrangement so that the out-

Figure 61. Block diagram of a temperature-compensated 30-MHz linear amplifier that uses the RCA-2N6093 transistor.

L_1 = 3 turns of No. 14 wire, 1/4-inch inner diameter, 1/2 inch long
L_2 = 3 turns of No. 10 wire, 1/2-inch inner diameter, 3/8 inch long
RFC = Ferroxcube No. VK200-01-3B or equiv.

Figure 62. Use of the RCA-2N6093 in a 30-MHz, 75-watt (PEP) amplifier with temperature compensation.

put voltage is independent of ambient-temperature variations. Q3 and Q4 provide the necessary current amplification. The bias current in rf transistor Q5 can be adjusted by varying R_1.

As shown in Fig. 63, with no rf signal the forward-biased transistor is statically stable up to a case temperature of 160°C. The

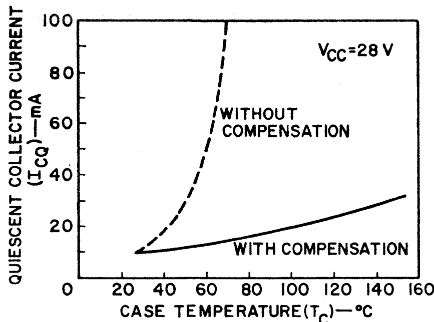


Figure 63. Quiescent collector current in the RCA-2N6093 as a function of case temperature with and without temperature compensation.

dashed line in Fig. 63 shows that without temperature compensation the transistor tends to thermal runaway around 80°C. To further show the effectiveness of compensation, the third-order distortion and output power are plotted as a function of case temperture in Fig. 64. The decrease in output power at high temperatures is caused by a drop in high-frequency gain and an increase in rf saturation voltage. The decrease in h_{fe} produces a soft saturation knee that increases distortion.

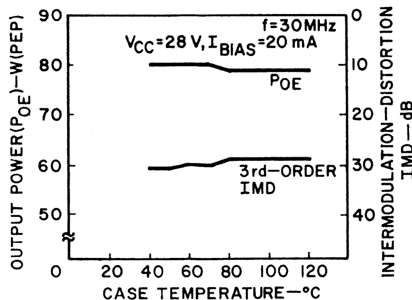


Figure 64. Output power and intermodulation distortion as a function of case temperature for the RCA-2N6093 amplifier shown in Fig. 62

The common-emitter configuration should be used for a linear power amplifier because of its stability and high power gain. Tuning is less critical, and the amplifier is less sensitive to variations in parameters among transistors. The class AB mode is used to obtain low intermodulation distortion. Neither resistive loading nor neutralization is used to improve linearity because of the resulting drastic reduction in power gain; furthermore, neutralization is difficult for large signals because parameters such as output capacitance and output and input impedances vary nonlinearly over the limits of signal swing.

Fig. 62 shows a schematic diagram of a narrow-band, high-power, 30-MHz amplifier. The amplifier provides an output power of 75 watts PEP from a 28-volt power supply. The impedance of the base-to-emitter junction of the RCA-2N6093 SSB transistor in this circuit is transformed to 50 ohms to match the impedance of the drive source. The input circuit to the transistor can be represented as a resistance ($r_{bb'}$) in series with a capacitance C_i. The input network must tune out the capacitance C_i and must present a pure resistive load to the driver. The input network is formed by the T-network consisting of capacitors C_1 and C_2 and inductor L_1. The value of L_1 is chosen so that the inductive reactance is much greater than the reactance of C_i. Series tuning of the base-to-emitter circuit is ob-

tained by L_1 and the parallel combination of C_2 and C_1, together with the capacitance of the driver stage.

Inductor L_2 in the output circuit is selected to resonate with the transistor output capacitance. Capacitors C_3 and C_4 and inductor L_3 provide the proper impedance transformation from 50 ohms to 3.13 ohms at the resonant frequency. Base-bias voltage is obtained from the output of the compensating circuit. If the bias voltage is not temperature-compensated, both linearity and collecter efficiency can be affected. When an rf signal is applied to the amplifier under high-power conditions, the rectifying property of the base-to-emitter junction charges any capacitance present in the base circuit of the transistor. This charge can alter the bias point and reduce the angle of conduction; the amplifier then operates more toward class C, and distortion and efficiency are both increased. Fig. 63 shows the effect of temperature compensation on the collector current of the 2N6093, and Fig. 64 shows the output and intermodulation-distortion characteristics of the 30-MHz amplifier as a function of temperature.

BROADBAND LINEAR AMPLIFIER

Fig. 65 shows a 2-to-30-MHz wide-band linear amplifier that uses an RCA rf power transistor other than the diode-compensated 2N6093. This amplifier uses an RCA-2N5070 to develop a power output of 5 watts PEP. At this power level, IMD products are more than 40 dB below one tone of a two-tone signal. Power gain is greater than 40 dB.

In low-power linear amplifiers, the use of temperature-compensating circuits is sometimes not necessary if the transistor output power is less than 50 per cent of

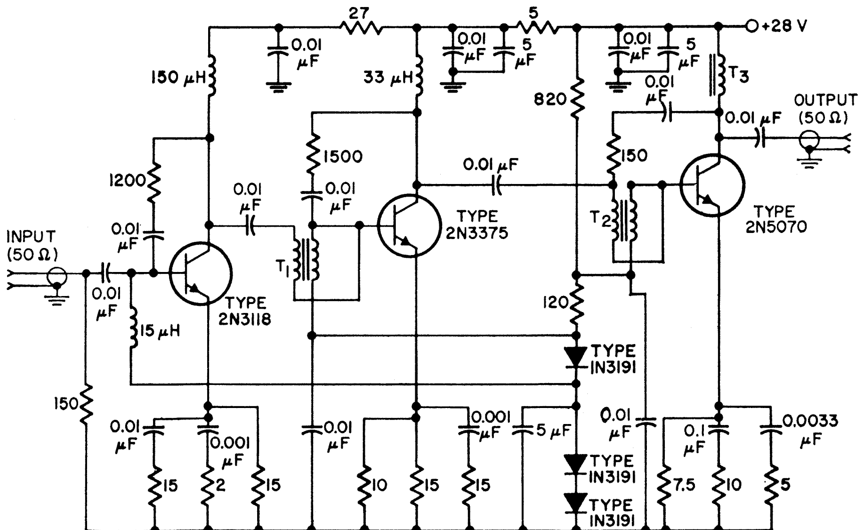


Figure 65. A 5-watt PEP 2-to-30-MHz linear power amplifier.

its maximum cw power rating. The RCA-2N5070 transistor is useful in such applications. This transistor is specified for SSB applications without temperature compensation as follows:

Frequency = 30 MHz
P_o (PEP) at 28 V = 25 W
Power Gain = 13 dB (min.)
Collector Efficiency = 40 % (min.)

HIGHER-POWER BROADBAND LINEAR AMPLIFIERS

Before any circuit can be designed, the transistor input impedance and the collector load impedance over the required frequency band and at the desired levels of output power, IMD, case temperature, and collector supply voltage must be known or measured. The circuit designer must also know the transistor power gain over the same band. Curves of these characteristics for the RCA-2N6093 are shown in Figs. 66, 67, and 68. A broadband transistor should be selected for minimal impedance variation and low input Q across the frequency band. A transistor that has an f_T well above the highest operating frequency, if available, can provide constant gain under broadband operation; such a transistor eliminates the need for additional gain-leveling circuitry. Because circuit optimization becomes more difficult with high-power broadband operation, the need for thermal stability becomes more acute, and the necessity of diode compensation at high output powers becomes greater. To provide this stability, the transistor should have an internally mounted compensating diode.

The advantages which especially suit the 2N6093 for broadbanding are its low input Q and its internally mounted compensating diode. Its main disadvan-

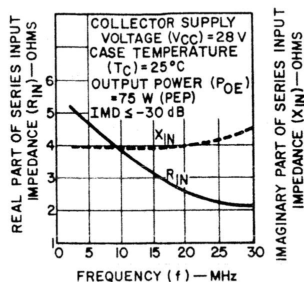


Figure 67. Typical large-signal series input impedance (R_{in} + jX_{in}) as a function of frequency for the RCA-2N6093.

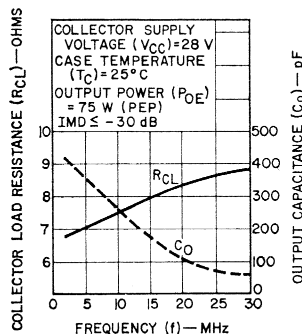


Figure 68. Typical large-signal parallel collector load resistance and parallel output capacitance as a function of frequency for the RCA-2N6093.

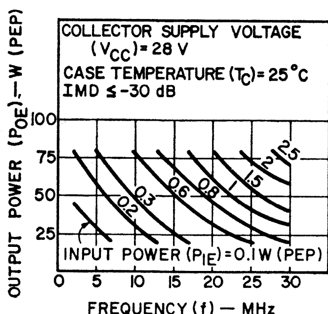


Figure 66. Typical output power as a function of frequency for the RCA-2N6093.

tage is a 15-dB gain decrease from 2 to 30 MHz that results from operation on a power-gain slope of 6 dB per octave.

Design Techniques

After selection of the transistor and measurement of its broadband parameters, the next step is to select the circuit approach. The most practical broadbanding method to provide an effective impedance transformation over four octaves (2 to 30 MHz) is a transmission-line-transformer/ferrite-core combination. The major disadvantage of a transmission line transformer is the limited number of impedance transformations available: 1:1, 4:1, 9:1, etc. The two fundamental configurations are the 1:1 reversing transformer and the 4:1 impedance transformer shown in Fig. 69.

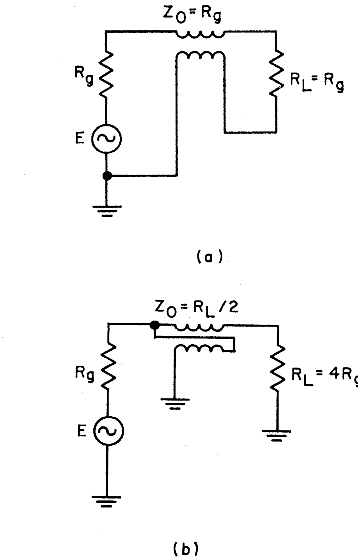


Figure 69. Transmission-line transformers: (a) 1:1 reversing/isolating transformer; (b) 4:1 impedance transformer.

Ferrite Cores—At low frequencies, a high primary reactance can be obtained with a few turns of transmission line on a high-permeability ferrite core. At high frequencies where length becomes critical, the permeability of the core decreases, thereby maintaining approximately the same levels of reactance with a short length of transmission line. Ferramic-Q core material is available in three high-frequency grades; a tabulation of their useful properties is given in Table X. Because the transformer performance is less dependent on core material at the higher-frequency end of its useful range, the poor intrinsic Q of Q-1 material above 20 MHz does not degrade the transformer operation at 30 MHz. Q-2 material, having lower permeability, requires more turns for operation at the lower frequencies.

Hybrid Combiner/Dividers—Hybrid combiner/dividers can be made by use of combinations of the 1:1 and 4:1 transformers on

Table X—Permeability and Frequency Dependence of Ferramic-Q Materials

Material	Permeability	Approximate Frequency at which core losses increase by a factor of 10 (MHz)
Q-1	125	10
Q-2	40	90
Q-3	16	225

ferrite cores to provide high impedance-transformation ratios. As an example, Fig. 70 shows a 180-degree-phase hybrid divider that matches a 50-ohm source to a 3.12-ohm push-pull configuration. Two 1:1 transformers are used to make the 4:1 transformation,

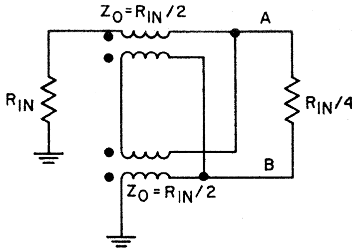


Figure 70. A 4:1 broadband transformation network that uses two 1:1 transformers to provide a balanced output.

rather than one 4:1 transformer, to provide the balanced output needed for a push-pull configuration. An equivalent transformation also can be made with one 1:1 transformer and one 4:1 transformer, as shown in Fig. 71.

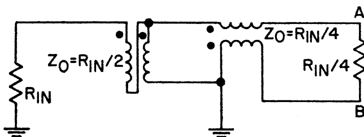


Figure 71. A 4:1 broadband transformation network that uses a 1:1 transformer and a 4:1 transformer to provide a balanced output.

Fig. 72 shows a 16:1 broadband transformation network for a push-pull configuration. The circuitry to the left of V_2 is the same as in Fig. 70; to the right of V_2 an extra transformer and dissipating resistor have been added. Points A and B are transistor base inputs, R_2 represents the resistive input to a conducting transistor, and R_3 is a resistor much larger than R_2 that is connected in shunt with each base-to-emitter junction. (Thus A-to-ground represents a conducting transistor, while B-to-ground represents a cut-off transistor, in Fig. 72.) R_1 dissipates any imbalances in power or phasing.

To find the input resistance to the network of Fig. 72, the network equations are written as follows:

$$
\begin{aligned}
&I_1 = I_2 = I_3 = I_4 && V_2 - V_4 = V_4 - V_3 \\
&I_5 = I_6 = 2I_1 && V_1 = 2(V_2 - V_3) \\
&I_7 = I_5 - I_8 && V_4 = R_1 I_{10} \\
&I_8 = I_9 && V_2 = I_7 R_2 \\
&I_{11} = I_9 + I_6 && V_3 = R_3 I_{11} \\
&I_{10} = I_8 + I_9 &&
\end{aligned} \quad (37)
$$

These equations yield V_1/I_1 as a function of R_1, R_2, and R_3:

$$
R_{IN} = \frac{V_1}{I_1} = 16\left(\frac{R_1 R_2 + R_1 R_3 + R_2 R_3}{4R_1 + R_2 + R_3}\right) \quad (38)
$$

If $R_1 = 1/2R_2$ and $R_3 = 5R_2$, $R_{IN} = 16\ R_2$. Thus the 3.12-ohm-transistor resistance is transformed to 50 ohms.

Because of symmetrical loading, the same hybrid configura-

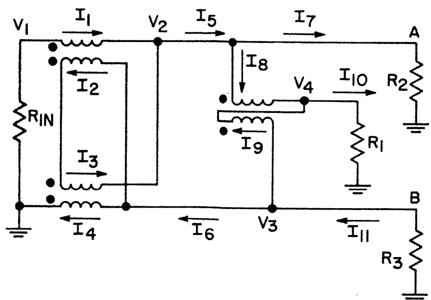


Figure 72. A 16:1 broadband transformation network with balanced output.

tion provides an 8:1 impedance transformation when used as a 180-degree-phase power combiner at the transistor collectors. This combiner operation of the network is shown in Fig. 73; the output resistance is given by

$$R_{OUT} = \frac{V_{OUT}}{I_{OUT}} = 16\left(\frac{R_1 R_2 + R_1 R_3 + R_2 R_3}{4R_1 + R_2 + R_3}\right) \quad (39)$$

If the collector load-line resistance is R_L, and if $R_1 = \frac{1}{2}R_L$ and $R_2 = R_3 = R_L$, then

$$R_{OUT} = 8R_L \quad (40)$$

Thus each collector is provided with a 6.25-ohm load line for $R_{OUT} = 50$ ohms. The inductance of the transmission line and its connectors is utilized to tune out both input and output negative reactances.

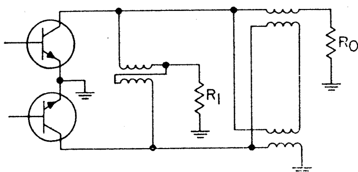


Figure 73. The network of Fig. 72 used as a 180-degree-phase power combiner.

2-to-30-MHz Circuit Design

The push-pull configuration is used not only because the 180-degree-phase hybrids provide a high transformation ratio, but also because this configuration suppresses second harmonics and thus minimizes filter requirements at the output. If the output power level and the input and output impedance values at that power level are known, the circuit designer can use a combination of 180-degree-phase hybrids, hybrid resistance values, and additional transmission-line transformers to complete the proper transformation at the input and output. After the transformation closest to optimum match at the highest operating frequency has been selected, individual transformers are wound and measured over the desired frequency band. The HP 4815A vector impedance meter, RX Boonton Meter, or a similar instrument can be used for these measurements.

A 120-watt (PEP) linear amplifier for the 2-to-30-MHz frequencey range has been built with a pair of RCA-2N6093 transistors in push-pull, 180-degree-phase hybrid power combiner/dividers, and single-ended 4:1 transformers. The block diagram of this amplifier is shown in Fig. 74, and the circuit diagram is shown in Fig. 75.

Typical performance of this amplifier across the hf band is shown in Fig. 76. The power gain exhibits the same 6-dB-per-octave slope at mid-band low-frequency roll-off noted in the narrowband measurements (Fig. 66). Total gain variation is approximately 15 dB.

The intermodulation distortion exceeds —30 dB at frequencies below 6 MHz. The circuit is capable of —35 dB IMD over a good portion of the band if operated at the reduced output power of 100 to 110 watts PEP. If the same circuit components and transformation networks are utilized, the efficiency is somewhat reduced at the reduced power level because the collector circuit is optimized for higher power.

The efficiency of the amplifier is 40 to 50 per cent across the

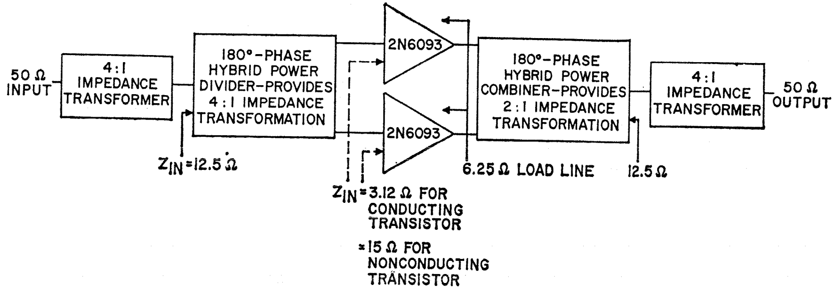
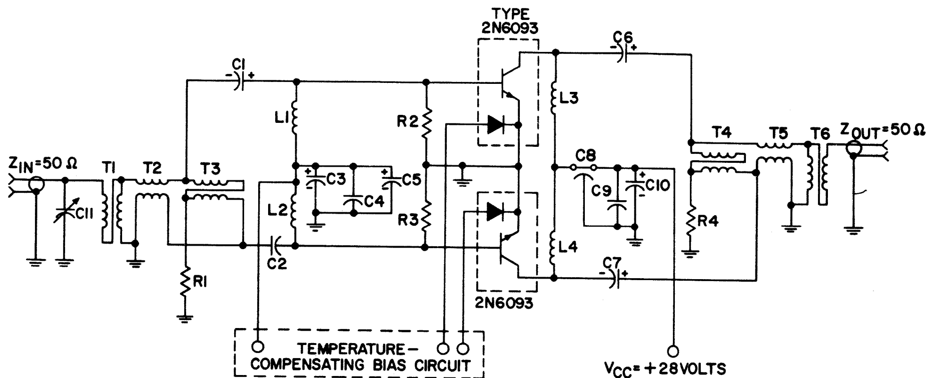


Figure 74. Block diagram of a push-pull linear amplifier that provides 120 watts PEP.



C_1, C_2 = 0.15 μF, electrolytic
C_3, C_9 = 0.04 μF, ceramic
C_4 = 0.0027 μF, ceramic
C_5 = 100 μF, electrolytic
C_6 C_7 = 0.1 μF, electrolytic
C_8 = 1000 pF, feedthrough, Allen-Bradley FA5C or equiv.
C_{10} = 5 μF, electrolytic
C_{11} = variable capacitor, ARCO 249 or equiv.
L_1, L_2 = rf choke, 10 μH
L_3, L_4 = 15 turns of No. 20 wire on Q1 CF-108 Indiana General ferrite core or equiv.

R_1 = 3.3 ohms, 0.5 watt
R_2, R_3 = 30 ohms, 0.5 watt in parallel with 30 ohms, 0.5 watt
R_4 = 5.6 ohms, 0.5 watt in parallel with 5.6 ohms, 0.5 watt
T_1, T_6 = two twistel pairs (9 turns per inch) of No. 26 wire in parallel; five turns on Q1 CF-108 Indiana General ferrite core or equiv.
T_2, T_3, T_4, T_5 = six twisted pairs (9 turns per inch) of No. 28 wire in parallel; five turns on Q1 CF-108 Indiana General ferrite core or equiv.

Figure 75. 120-watt PEP, 2-to-30-MHz push-pull linear amplifier.

band. When operated at 120 watts PEP with a V_{CC} of 28 volts, the amplifier becomes current limited at frequencies below 3 MHz. The increase in VSWR is related to the increase in the real part of the transistor input impedance (see Fig. 67).

Fig. 77 shows the performance of the 120-watt PEP amplifier as a function of case temperature at 30 MHz.

The main advantages of this type of circuit are its simplicity and compactness. The disadvantages are lack of gain leveling and low efficiency at lower frequencies because of increased VSWR.

Because the real value of the transistor input impedance increases with decreasing frequency, which affects both VSWR and IMD, a resistance-inductance

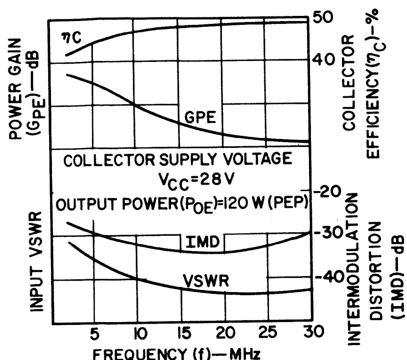


Figure 76. Typical performance of the broadband 120-watt (PEP) push-pull linear amplifier.

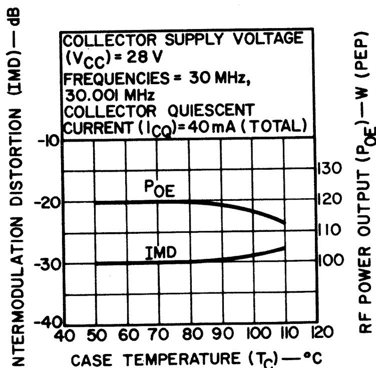


Figure 77. Performance of the 120-watt (PEP) linear amplifier as a function of case temperature at 30 MHz.

series combination placed in parallel with the 50-ohm input or placed from base to base aids the transformation network in making a practical match at low frequencies. The impedance match is improved and some input power is absorbed at low frequencies; therefore, the VSWR improves and some gain leveling occurs. Other methods of gain leveling include collector-to-base feedback and loop feedback; for high-power circuits, the loop feedback system shown in Fig. 78 would be the most effective. In this system, input and output signals are compared, and gain differences are compensated by commensurate increases in input attenuation.

For higher powers, modules of push-pull pairs can be pyramided by the same hybrid-combining techniques.

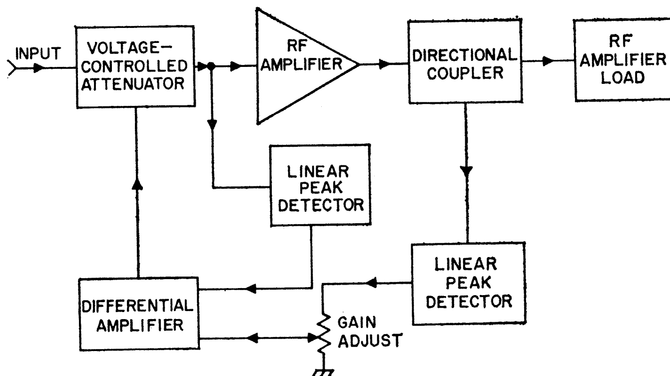


Figure 78. A loop feedback system for gain-leveling.

RF Amplifiers for Military Communications Systems

THE ruggedness, compactness, reliability, and efficiency of transistors make them especially useful in military environments. Aircraft communication equipment, sonobuoy transmitters, and air-rescue beacons are typical military applications of rf power transistors.

MILITARY AIRCRAFT COMMUNICATIONS

The frequency range from 225 to 400 MHz is used in a large variety of relatively-high-power military communication systems. Equipments are usually amplitude-modulated and used for voice-communication purposes. The circuits discussed in this section are class B and class C amplifiers for use in driver or final output stages that provide power outputs in the range from 5 to 30 watts from a single transistor. Higher power can be obtained from combinations of transistors. Table XI lists RCA transistor types characterized for application in vhf/uhf broadband military communications systems.

These amplifiers make extensive use of power combiners and broadband impedance matching.

Table XI—RCA Transistor Types for Broadband Applications (225 to 400 MHz)

Type	Operating Frequency (MHz)	Output Power (Min.) (W)	Collector-Supply Voltage (V)	Power Gain (Min.) (dB)	Package Type	Primary Application
2N3866	400	1	28	10	TO-39	VHF-UHF Broadband Military Communications
2N5916	400	2	28	10	TO-216AA	
2N5917	400	2	28	10	Studless TO-216AA (HF-31)	
2N3375	400	3	28	5	TO-60	
2N5918	400	10	28	8	TO-216AA	
2N5016	400	15	28	5	TO-60	
2N5919A	400	16	28	6	TO-216AA	
2N6105	400	30	28	5	TO-216AA	

In an amplifier chain, all stages are designed to operate from a 50-ohm source into a 50-ohm load. For effective cascading, the input VSWR to any of the amplifiers in the chain must be as low as possible over the entire frequency band. Various techniques for broadbanding and for reducing input VSWR are discussed in following sections.

The lumped-constant circuit shown in Fig. 79 uses low pass, LC ladder networks for impedance transformation. (The values given for the various components in the circuit diagram are measured at 400 MHz and include parasitic elements.) The output network transforms the 50-ohm load down to 20 ohms for the collector load. The dynamic output capacitance of the transistor provides the first shunt capacitor in the output network, and capacitor C_C provides dc blocking. Similarly, the base input inductance of the 2N5016 transistor serves as the last series inductor of the input network, and capacitor C_C is again used for dc blocking. The input match of the lumped-constant circuit is optimized at 400 MHz. The m-derived end section (L_1 and C_1) helps to provide the proper amount of mismatch at frequencies below 400 MHz to compensate for the gain characteristic of the transistor. With 6 watts of drive, this circuit provides approximately 17 watts of output power across the 225-to-400-MHz frequency band with a total output variation of 0.5 dB, as shown in Fig. 80.

Fig. 81 shows a schematic diagram of an amplifier that uses the RCA-2N5919A. The circuit utilizes a lumped-element approach to broadband design. Typical amplifier performance is shown in Fig. 82. For a constant power output of 16 watts, response is fairly flat; the gain variation is within 1 dB across the band. Maximum input VSWR is 2:1.

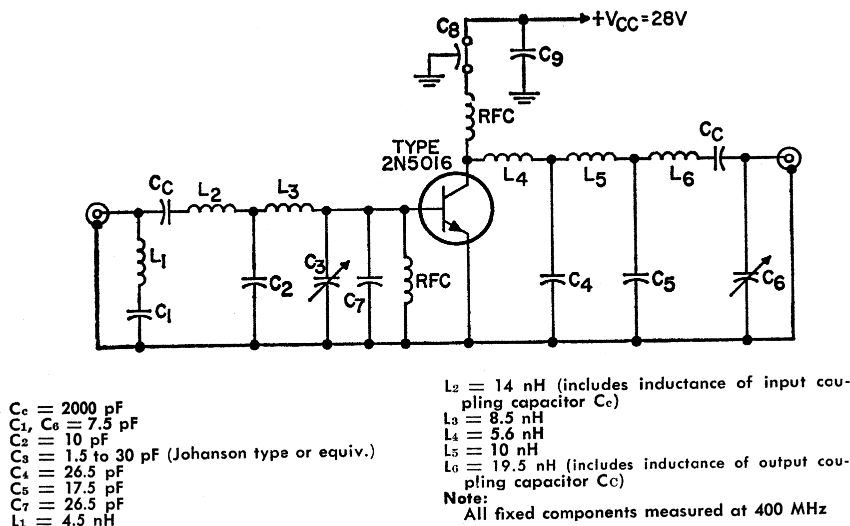


Figure 79. Lumped-constant 255-to-400-MHz power amplifier.

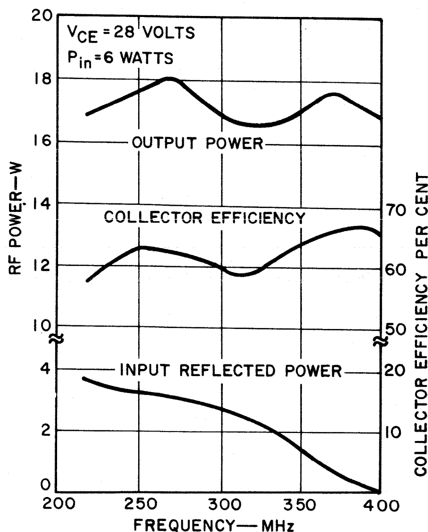
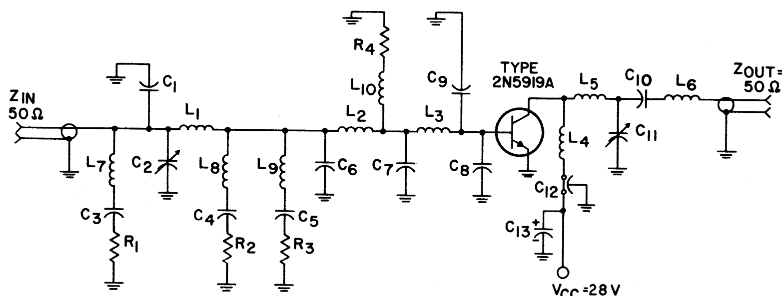


Figure 80. RF power output, input reflected power, and collector efficiency of the RCA-2N5016 transistor as functions of frequency.

Such flatness of response and low input VSWR are obtained by designing for the best possible match across the band and then dissipating some of the power at the low end of the band through dissipative RLC networks. The effectiveness of this technique can be evaluated by comparison of the gain and input VSWR curves in Fig. 82 (a) with those in Fig. 82 (b). The flatter the response, the smaller the dynamic range required in the output leveling system. Low input VSWR is necessary for effective cascading and protection of the driving stage in a cascade connection. The collector efficiency is not constant, but has a minimum value of about 63 per cent. The second harmonic of the 225-MHz signal is 12 dB down and that of the 400-MHz signal is 30 dB down from the



C_1 = 10 pF silver mica
C_2 = 0.8-10 pF, Johanson 3957*
C_3 = 2.2 pF, Quality Components type 10% QC, "gimmick"*
C_4 = 1.0 pF, Quality Components type 10% QC, "gimmick"*
C_5 = 1.5 pF, Quality Components type 10% QC, "gimmick"*
C_6 = 36 pF, ATC-100*
C_7 = 51 pF, ATC-100*
C_8 = 47 pF, ATC-100*
C_9 = 68 pF, ATC-100*
C_{10} = 12 pF, silver mica
C_{11} = 0.8-20 pF, Johanson 4802*
C_{12} = 1000 pF feedthrough type, Allen-Bradley FA5C*
C_{13} = 1 μF electrolytic
L_1 = 1½ turns▲
L_2 = Copper strip ⅝ in. (15.875 mm) L; 5/32 in. (3.95 mm) W
L_3 = Transistor base lead, 3/6 in. (4.74 mm) L
L_4, L_6 = 3 turns▲
L_5 = 2 turns▲
L_7, L_8, L_9 = 0.18 μH RFC, Nytronics, P.#DD-0.18
L_{10} = 0.1 μH RFC, Nytronics, P.#DD-0.10
R_1 = 100 Ω, 1 W, carbon
R_2, R_3 = 100 Ω, ½ W, carbon
R_4 = 5.1 Ω, ½ W, carbon
* Or equivalent
▲ All coils are 5/32 in. (3.96 mm) I. D., # 18 wire, 12 turns per inch.
Allen-Bradley Co., Milwaukee, Wis.
American Technical Ceramics, Huntington Station, N. Y. 11746
Johanson Mfg. Corp., Boonton, N. J. 07005
Nytronics, Inc., Berkeley Heights, N. J.

Figure 81. 16-watt broadband amplifier circuit using the RCA-2N5919A.

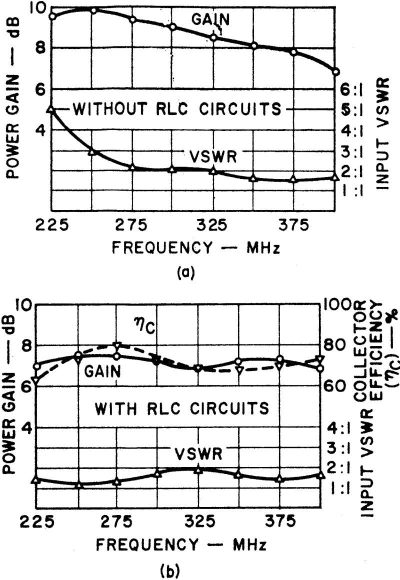
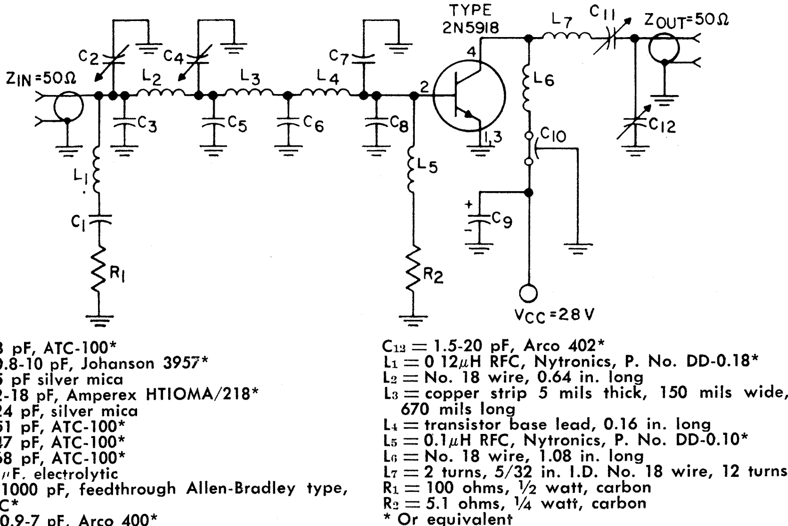


Figure 82. Typical performance of circuit of Fig. 81 from 225 to 400 MHz.

fundamental. Further reduction of the second harmonic of the 225-MHz signal is difficult to obtain because the amplifier bandwidth covers almost an octave.

In a cascade arrangement, a lower-power transistor, the 2N5918, is used to drive the 2N5919A. The output circuit for the driver is modified to accommodate a higher collector load. The input circuit remains essentially the same as for the 2N5919A. The 2N5918 amplifier schematic is shown in Fig. 83, and the performance of the two amplifiers connected in cascade is shown in Fig 84. When the two stages are connected together, the broadband characteristiss of the amplifiers minimize the number of adjustments required.

Fig. 85 shows the RCA-2N6105 high-power transistor in a 225-to-400-MHz broadband amplifier, and Fig. 86 shows the performance of the amplifier. This circuit also utilizes lumped-circuit-element broadbanding. No special care is evident in this circuit to reduce the input VSWR. Two of these amplifiers can be combined by quadrature combiners, as shown in Fig. 87, to obtain



C_1 = 3 pF, ATC-100*
C_2 = 0.8-10 pF, Johanson 3957*
C_3 = 5 pF silver mica
C_4 = 2-18 pF, Amperex HT1OMA/218*
C_5 = 24 pF, silver mica
C_6 = 51 pF, ATC-100*
C_7 = 47 pF, ATC-100*
C_8 = 68 pF, ATC-100*
C_9 = 1 μF. electrolytic
C_{10} = 1000 pF, feedthrough Allen-Bradley type, FA5C*
C_{11} = 0.9-7 pF, Arco 400*
C_{12} = 1.5-20 pF, Arco 402*
L_1 = 0 12μH RFC, Nytronics, P. No. DD-0.18*
L_2 = No. 18 wire, 0.64 in. long
L_3 = copper strip 5 mils thick, 150 mils wide, 670 mils long
L_4 = transistor base lead, 0.16 in. long
L_5 = 0.1μH RFC, Nytronics, P. No. DD-0.10*
L_6 = No. 18 wire, 1.08 in. long
L_7 = 2 turns, 5/32 in. I.D. No. 18 wire, 12 turns
R_1 = 100 ohms, ½ watt, carbon
R_2 = 5.1 ohms, ¼ watt, carbon
* Or equivalent

Figure 83. Driver amplifier using the RCA-2N5918.

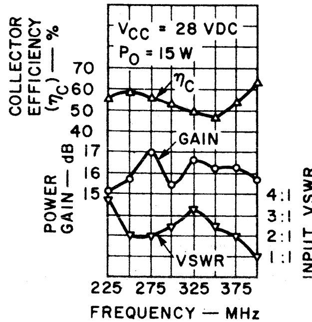
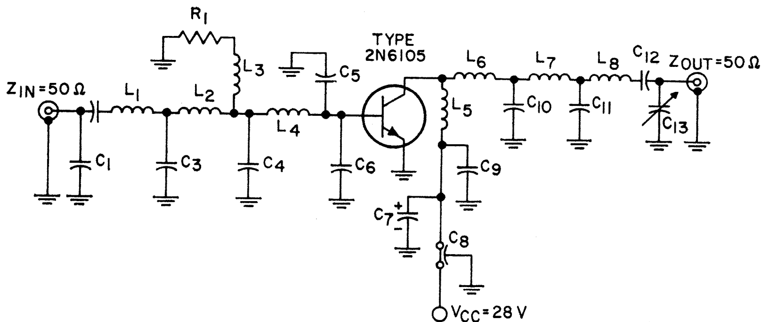


Figure 84. Performance characteristics of amplifiers shown in Fig. 81 and 83 connected in cascade.

higher output power. The input VSWR of the individual amplifier is not important in such a combination because of the high isolation characteristics of quadrature combiners; reflected power is dissipated in ports terminated with 50-ohm resistors. The performance of the 2N6105's combined by this method is shown in Fig. 88.

Another effective way to combine transistors is push-pull operation utilizing transmission-line techniques. The advantage of low second harmonic in the push-pull configuration is especially important in the 225-to-400 MHz frequency band; because the second harmonic of the low frequency falls just outside the band, filtering it out presents considerable difficulty. The use of transmission lines results in a compact, relatively simple structure. The input VSWR in a push-pull amplifier is very high at the low end of the band, so this type of circuit is especially suitable for use with quadrature combiners. Fig. 89 shows details of an individual push-pull amplifier using two RCA-2N6105 transistors. Fig 90 shows two of these amplifiers combined by quadrature combiners to make up a 100-watt broadband module. This ap-



C_1 = 8.2 pF chip, Allen-Bradley*
C_2 = 18 pF silver mica
C_3 = 33 pF chip, Allen-Bradley*
C_4 = 47 pF chip, Allen-Bradley*
C_5 = 68 pF chip, ATC-100*
C_6 = 62 pF chip, ATC-100*
C_7, C_8 = 1000 pF, Feedthrough long
C_9, C_{12} = 1000 pF chip, Allen-Bradley*
C_{10} = 22 pF chip, Allen-Bradley*
C_{11} = 6.9 pF chip, Allen-Bradley*
C_{13} = 0.8-10 pF variable air, Johanson No. 3957*

L_1 = 2 turns, 5/32-in. I.D. coil
L_2 = 17/32-in. long wire
L_3 = RFC, 0.1 μH, Nytronics*
L_4 = 5/32-in. long transistor base lead
L_5, L_7 = 13/16-in. long wire
L_6 = 9/16-in. long wire
L_8 = 7/8-in. long wire
R_1 = 5.0 Ω, 1/4 W
All wire is No. 20 AWG

* Or equivalent

Figure 85. 225-to-400-MHz broadband amplifier using RCA-2N6105.

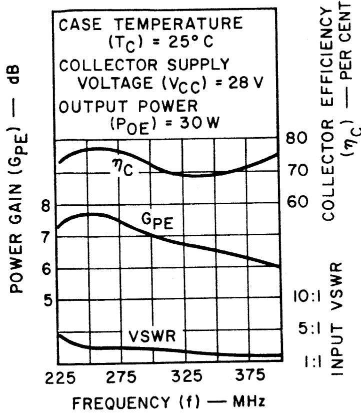


Figure 86. Typical performance of a 225-to-400-MHz broadband amplifier using RCA-2N6105 at V_{CC} = 28V.

proach results in a saving of four combiners, at least two of which are quadratures. The performance of this module for 100-watt constant output is shown in Fig. 90

SONOBUOY TRANSMITTERS

A sonobuoy is a floating submarine-detecting device that incorporates an underwater sound detector (hydrophone). The audio signals received are converted to a frequency-modulated rf signal which is transmitted to patrolling aircraft or surface vessels. The buoy is battery-operated and is designed to have a very limited active life.

Typical requirements for the rf-transmitter section of the sonobuoy are as follows:

Frequency = 165 MHz
Supply Voltage = 8 to 15 volts
CW Output = 0.25 to 1.5 watts
Over-all Efficiency = 50 per cent
Harmonic Output = 40 dB down from carrier

Fig. 92 shows a diagram of an experimental sonobuoy transmitter designed to produce a power

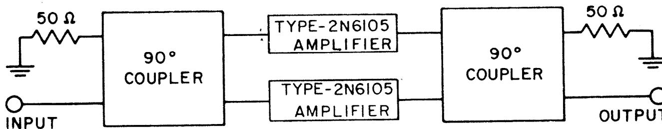


Figure 87. Two RCA-2N6105 amplifiers connected in parallel by use of quadrature couplers.

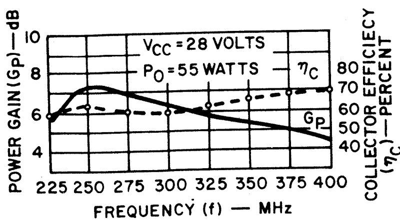
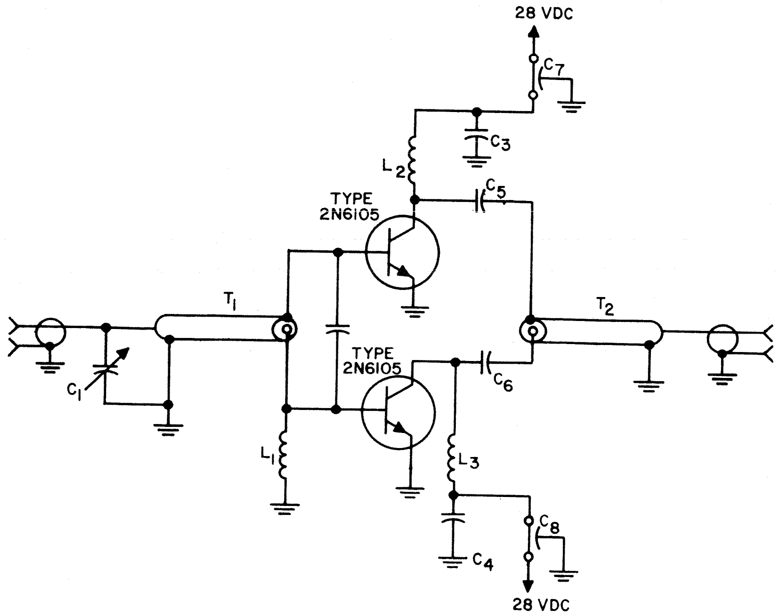


Figure 88. Power gain and efficiency as function of frequency for the broadband module shown in Fig. 87.

output of 2 watts at 160 MHz. Only three stages, including the crystal-controlled oscillator section, are required. Efficiency is greater than 50 per cent (over-all) with a battery supply of 12 to 15 volts.

The 2N3866 or 2N4427 transistor can be used in a class A oscillator-quadrupler circuit which is capable of delivering 40 milliwatts of rf power at 80 MHz.



C_1 = 0.8 to 10 pF, piston trimmer
C_2 = 56-pF chip, ATC-100 or equiv.
C_3, C_4, C_5, C_6 = 1000-pF chip, Allen-Bradley type or equiv.
C_7, C_8 = 1000 pF, feedthrough

L_1 = 0.18 μH, RFC, Nytronics type or equiv.
L_2, L_3 = 3/4-inch-long No. 20 wire
T_1 = coaxial line, Z_o = 25 ohms, 3 3/4 inches long
T_2 = coaxial line, Z_o = 25 ohms, 4 1/2 inches long

Figure 89. 225-to-400-MHz broadband push-pull amplifier using two 2N6105's.

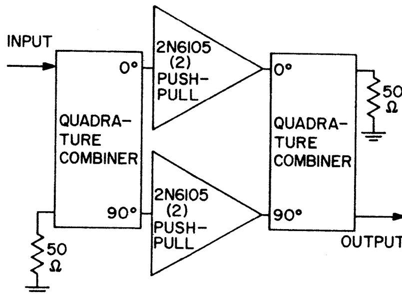


Figure 90. 100-watt 225-to-400-MHz broadband module.

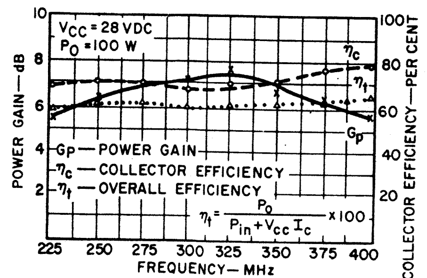


Figure 91. 100-watt, 225-to-400-MHz broadband module performance.

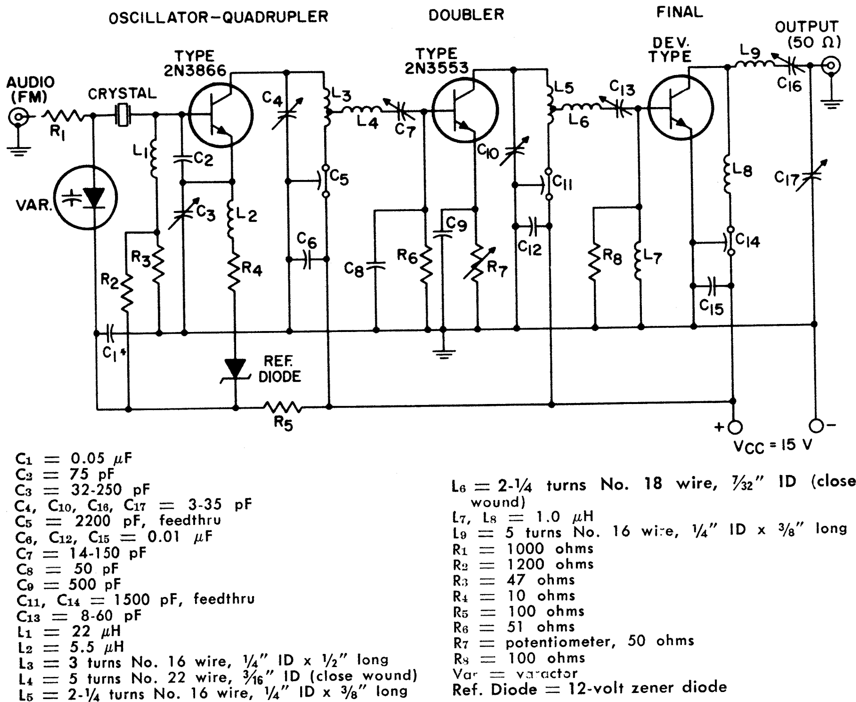


Figure 92. 1.5-watt (rf power output) sonobuoy transmitter.

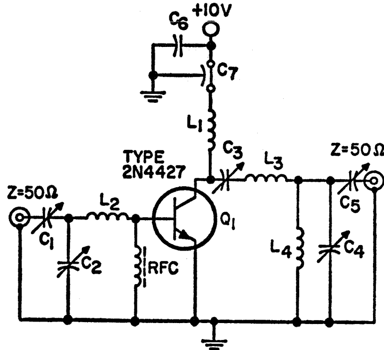
Narrow-band frequency modulation is accomplished by "pulling" of the crystal oscillator. The crystal is operated in its fundamental mode at 20 MHz. The oscillator is broadly tuned to 20 MHz in the emitter circuit and is sharply tuned to 80 MHz in the collector circuit. The supply voltage to the oscillator section is regulated at 12 volts by means of a zener diode. Spectrum-analyzer tests indicate that this stage is highly stable even though rather high operating levels are used.

The oscillator-quadrupler section is followed by a 2N3553 class C doubler stage. This stage delivers a power output of 250 milliwatts at 160 MHz from a 12- to 15-volt supply. The over-all output of the sonobuoy can be adjusted by varying the emitter resistance of this stage.

The final power output is developed by an RCA developmental transistor which operates as a straight-through class C amplifier at 160 MHz. A pi network matches this output to the 50-ohm line. The spurious output (measured directly at the output port) is more than 35 dB down from the carrier. This suppression is achieved by means of series resonant trap circuits between stages and the use of the pi network in the output.

Many sonobuoy systems require power outputs in the range of only 0.25 to 0.5 watt, preferably with a supply voltage of 8 to 12 volts. The 2N4427 transistor

is suitable for use as the doubler and also the final output device in such low-power applications. Fig. 93 shows a diagram of an output stage which uses the 2N4427 as a straight-through 175-MHz class C amplifier. This



C_1, C_2, C_4, C_5 = 7-to-100 pF Arco 423 or equiv.
C_3 = 14-to-150 pF, Arco 424, or equiv.
C_6 = 0.01 μF, 50 V
C_7 = 1000 pF, feedthru
L_1 = 0.75 μH
L_2 = 1 turn No. 18 wire, 5/32" ID
L_3 = 1½ turns No. 18 wire, ¼" ID
L_4 = 1¼ turns No. 18 wire, 3/16" ID
RFC = 450 ohms, ferrite

Figure 93. 0.5-watt 175-MHz sonobuoy rf power output stage.

circuit can deliver output power of more than 500 milliwatts with a supply voltage of 10 volts and a drive power of 60 milliwatts.

For the lower power-output requirement at low supply voltages, the oscillator-quadrupler stage should use lower-power transistors such as the 2N1491 or 2N914. Only 10 to 15 milliwatts of fourth harmonic power is required in this case. The bias-network resistors (R_2 and R_3) should be adjusted for reliable oscillator starting conditions at the lower supply voltages.

Sonobuoy circuits, in general, must be reliable, simple, and low in cost. The three-stage transmitter circuit shown in Fig. 92 is intended to be representative of the general design techniques used in these systems. However, four-stage sonobuoy transmitter systems are also in common use at the present time. Typically, a four-stage arrangement consists of an oscillator-tripler stage, a second tripler stage, a buffer stage, and a final amplifier stage. Most present-day sonobuoy applications require cw power output between 0.25 and 1.5 watt.

AIR-RESCUE BEACONS

The air-rescue beacon is intended to aid rescue teams in locating airplane crew members forced down on land or at sea. The beacons are amplitude-modulated or continuous-tone line-of-sight transmitters. They are battery-operated and small enough to be included in survival gear.

Typical requirements for rescue beacons are as follows:

Frequency = 243 MHz (fixed)
Power Output = 300 milliwatts (carrier)
Efficiency = greater than 50 per cent
Supply Voltage = 6 to 12 volts
Modulation = AM, up to ±100 per cent

The 2N4427 transistor is especially suited for this service. A general circuit for the driver and output stages is shown in Fig. 94. Collector modulation, as well as some driver modulation, is used to achieve good down-modulation of the final amplifier. Conventional transformer-coupled modulation is used; however, a separate power supply and resistor network in the

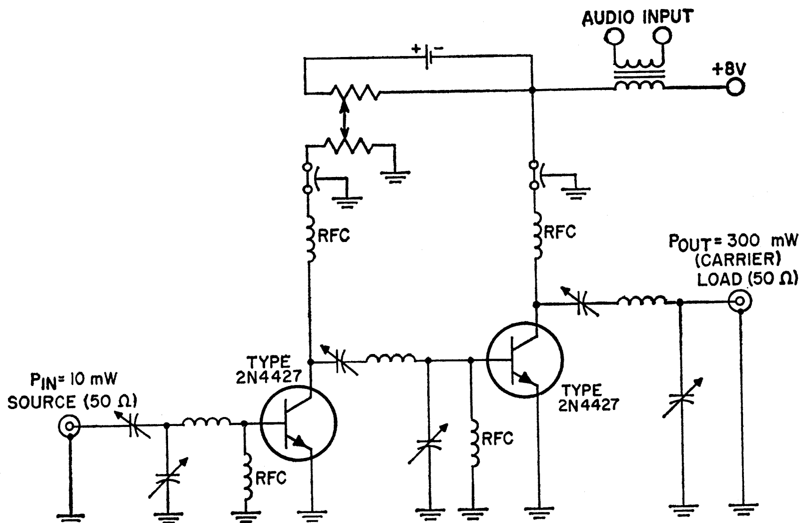


Figure 94. Driver and output stage for a 243-MHz beacon transmitter.

driver circuit are provided to adjust the modulation level of this stage independently of the output stage.

The rf-amplifier design is conventional; pi- and T-matching networks are used; simpler circuits (e.g., device-resonated tapped coils), however, could be used. The T-matching network at the driver input is used to match the amplifier to a 50-ohm source for test purposes. A 10-to-20-milliwatt input signal is needed to develop a 300-to-400-milliwatt carrier output level.

MINIATURIZED LOW-POWER OSCILLATORS

Low-power transistor oscillators are used as transmitters for telemetering or signal use in such devices as radiosondes, military fuses, beacons, and other remote sensing devices. Many of these units currently operate in the uhf range at output levels of about 0.25 to 1 watt. Battery supplies are normally used.

The 2N3866 and 2N4427 transistors are ideally suited for low-power oscillator service. Fig. 95 shows a simple microstripline circuit in which these transistors can provide power outputs of up to 1 watt in the frequency range of 400 to 600 MHz. The frequency of oscillation is primarily determined by capaci-

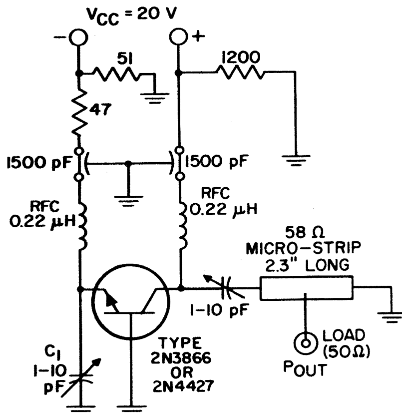


Figure 95. 1-watt, 500-MHz microstripline oscillator using the RCA-2N3866 or 2N4427 transistor.

tor C and the parasitic emitter-lead inductance. The microstrip-line output circuit can be matched to a wide range of loads by use of taps along the line length.

Fig. 96 shows a very simple lumped-constant oscillator circuit for operation in the 700-to-1000-MHz frequency range. The parasitic emitter- and base-lead inductances are tuned directly with high-Q air dielectric capacitors, and no other external inductances are required for this frequency range. The collector is grounded directly to the ground plane for best dissipation of transistor heat. Capacitor C_1 primarily determines the oscillator frequency, and the output capacitors are used primarily for impedance matching. The 2N3866 is used for operation at supply voltages of 20 to 28 volts, and the 2N4427 is preferred for supply voltages of 15 to 20 volts. Power outputs in the order of 500 to 1000 milliwatts into a 50-ohm load can be developed by this simple circuit.

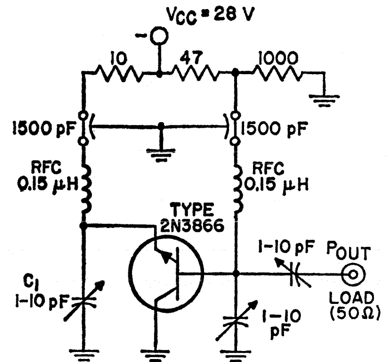


Figure 96. 0.5-watt, 1000-MHz lumped-constant oscillator using the RCA-2N3866 transistor.

Mobile and Marine Radio

In the United States, three frequency bands have been assigned to two-way mobile radio communications by the Federal Communications Commission. These frequency bands are 25 to 50 MHz, 148 to 174MHz, and 450 to 470 MHz. The low-frequency band for overseas mobile communications is 66 to 88 MHz.

Frequency modulation (FM) is used for mobile radio communications in the United States and most overseas countries. The modulation is achieved by phase-modulation of the oscillator frequencies (usually the 12th or 18th submultiple of the operating frequency). In vhf bands, the frequency deviation is ±5 kHz and channel spacing is 25 kHz. In uhf bands, at present, the modulation deviation is ±15 kHz and channel spacing is 50 kHz. In the United Kingdom, AM as well as FM is used in mobile communications.

The minimum mobile-transmitter power-output levels in the United States are 50 watts in the 50-MHz band, 30 watts in the 174-MHz band, and 15 watts in the 470-MHz band. Some of the transmitters used in the United States have power-output ratings well in excess of 100 watts. Overseas power requirements are more moderate and are often regulated by law; the most common power output level is 12 watts at the antenna.

TRANSISTOR REQUIREMENTS

The transistors in the rf power stages are the heart of every solid-state transmitter. In fact, the present power, load mismatch, and frequency capabilities constitute major design limitations for new mobile transmitters; the extension of the present limits are often achieved through transistor design tradeoffs. For instance, a high f_T is necessary for gain optimization; however, a transistor with high f_T is highly susceptible to mismatch in the load condition. Therefore a lower, flatter current-gain curve, which would allow good rf current gain at high currents, is a better choice. An indication of the frequency capability is the ratio of emitter periphery to base area, called the "design ratio;" a design ratio of 2 to 3 provides adequate gain and rugged performance. Table XII lists several RCA transistor types characterized

Table XII—RCA Transistor Types for Mobile-Radio Applications

Type	Operating Frequency (MHz)	Output Power (Min.) (W)	Collector-Supply Voltage (V)	Power Gain (Min.) (dB)	Package Type	Primary Application
			TYPES FOR 450-TO-470-MHz FREQUENCY BAND			
2N5914	470	2	12.5	7	TO-216AA	
2N5915	470	6	12.5	4.8	TO-216AA	Broadband Amplifier
40893	470	15	12.5	5.2	HF-36	
			TYPES FOR 148-TO-175-MHz FREQUENCY BAND			
40637	156	0.1	12	—	TO-52	78-to-156-MHz Doubler
40280	175	1	13.5	9	TO-39	
2N4427	175	1	12	10	TO-39	
2N5913	175	1.75	12.5	12.4	TO-39	
40281	175	4	13.5	6	TO-60	
40282	175	12	13.5	4.8	TO-60	Broadband Amplifier
2N5995	175	7	12.5	9.7	TO-216AA	
2N5996	175	15	12.5	4.5	TO-216AA	
			TYPES FOR 50-TO-88 MHz FREQUENCY BAND			
2N4932	88	12	13.5	5.3	TO-60	
2N4933	88	20	24	7.5	TO-60	
40340	50	25	13.5	7	TO-60	Broadband Amplifier
40341	50	30	24	10	TO-60	
2N5992	88	7	12.5	10	TO-216AA	
2N5993	88	18	12.5	10	TO-216AA	

for operation in mobile-radio applications.

PACKAGE CONSIDERATIONS

The rf package design plays an important role in determining mobile transistor requirements. The minimization of emitter lead inductances, for instance, can decrease degeneration and significantly increase power gain; the elimination of basic parasitics can have a dramatic bandwidth widening effect. The thermal capability of the package determines the junction temperature at any given output power; a lower package $R_{\Theta J\text{-}C}$ can increase high power performance (and mismatch capability) by allowing the transistor chip to operate at lower effective temperatures. Reliability can be greatly improved by both proper package-chip interfaces and hermetically isolating the chip from environmental changes.

Because of low device impedances, especially in 12 volt devices, all package losses must be eliminated; bond wires must be kept as short as possible.

DC OPERATING VOLTAGES

All-solid-state mobile transmitters can be divided into two basic types: transmitters that operate from 24-to-28-volt collector supply voltages, obtained from dc-to-

dc converters, and transmitters that operate directly from the 12-volt electrical system of a vehicle.

Both types have advantages and disadvantages. The advantages of 24-to-28-volt operation include higher power gains per stage, good transient suppression, and fairly simple current and voltage limiting. The disadvantages are the additional cost of dc-to-dc converters and the somewhat higher power consumption and increased size of the radio. Direct operation from a 12-volt system permits savings in cost and size, as well as higher efficiency. Because 12-volt operation produces less gain per stage, however, additional rf stages are often needed. Transient suppression and voltage and current limiting are also somewhat more difficult. However, the savings in cost and size make 12-volt systems somewhat more desirable. For this reason most mobile transmitters are designed for 12-volt operation.

Because of the two discrete voltage ranges used for mobile radios, the transistor must be designed specifically for either 24-to-28-volt operation or 12-volt operation. Devices designed for 24-to-28-volt operation have substantially higher collector-breakdown voltages. Devices designed for the 12-volt radio must have substantially higher current-handling characteristics.

MATCHING NETWORKS

The design of high-power, high-frequency transistor amplifiers presents unique problems. Low operating voltages and relatively high power levels result in impedances that become very small and circulating rf currents that become very large. For example, if an rf power output of 60 watts is required from an amplifier operating directly from a 12-volt supply, the collector load impedance to the final amplifier must be approximately 1 ohm. Under these conditions, the peak current can be as high as 20 amperes. At the same time, the series input impedance will be substantially below 1 ohm. Similar conditions often hold for powers as low as 10 watts. Because of the small magnitudes, all matching elements must have as high a Q as possible, line lengths must be minimized, and all stray inductances must be eliminated.

Microstrip circuit elements can provide high Q at low cost; Fig. 97 shows tunable and fixed-tuned microstrip circuits.

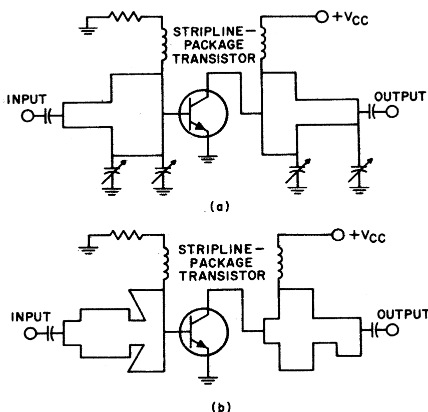


Figure 97. (a) Tunable and (b) fixed-frequency microstripline circuits.

INSTABILITIES IN VHF/UHF TRANSISTOR POWER AMPLIFIERS

In vhf/uhf transistor power amplifiers, the most common instabilities occur at frequencies

far below operating frequencies because the gain of the transistor increases at a rate of approximately 6 dB per octave as the frequency decreases. For example, a device that has a power gain of 5 dB at 174 MHz may have a gain of as much as 30 dB at 10 MHz. With such high gain, any kind of stray low-frequency resonant circuit can set the circuit into violent oscillation and even cause destruction of the transistor.

These low-frequency oscillations can be prevented by means of the following simple precautions, as indicated in Fig. 98:

(1) Because the base-emitter junction is highly capacitive at low frequencies, a resonant circuit can be easily formed with this capacitance and the choke RFC.

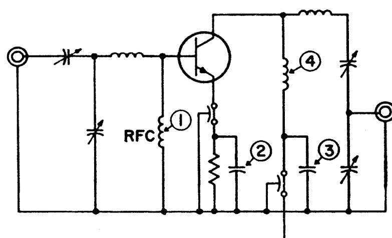


Figure 98. Circuit indicating areas where simple precautions prevent low-frequency oscillations.

This low-frequency resonant circuit can be avoided by the replacement of RFC with a low-Q, ferrite-type choke or even a wire-wound resistor.

In uhf circuits, a 0.22-microhenry choke in series with a 0.24-to-1.5-ohm resistor provides both a stable dc return and some base bias for better efficiency.

(2) The emitter bypassing should be effective not only at operating frequencies, but also at low frequencies; thus, two bypassing capacitors should be used. One of these capacitors should be effective at the operating frequency, and the other at low frequencies.

(3) DC-power wiring should have adequate bypassing both at operating and low frequencies to shunt out stray inductances in the wiring.

(4) Output-matching networks should make use of a coil as an integral part of matching for feeding dc to the collector. As a rule, the inductance of these coils is much smaller than that of self-resonant rf chokes, and thus the reactances are lower at low frequencies.

In higher-power uhf circuits, however, it is often desirable to series-tune the collector capacitance; this arrangement can result in better harmonic supression, and better transistor efficiency.

RELIABILITY

Mobile radio applications place severe requirements on transistor operation. At full rated input and output power, the device must consistently survive load mismatches from short to open circuit; this condition often occurs simultaneously with an increase in V_{CE}. The ability of a device to both survive such conditions and to avoid any permanent degradation can be realized only by uncompromising transistor design procedures coupled with the best available engineering judgment. Emitter ballasting, which forces transistor current to be equally shared through all active transistor emitter regions, is critical for ruggedness as are proper frequency and power tradeoffs. Because of the powers involved, thermal resistance must be minimized to avoid excessive junction

temperatures. Long-term reliability can be enhanced by protection of critical junctions from shorting caused by migration, and by use of a hermetically sealed package to protect the sealed pellet.

The mobile transmitter manufacturer also contributes to reliability by proper heat-sink design and by building in voltage regulation and VSWR protection.

CIRCUIT DESIGNS

The following paragraphs describe the design details for several rf power-amplifier circuits intended for use in mobile- or marine-radio applications.

66-to-88-MHz Band

Transmitters that operate directly from vehicle batteries in the 66-to-88-MHz frequency band can use either AM or FM.

In an AM transmitter, the peak envelope power P_p of a collector-modulated transistor is

$$P_p = P_c(1 + m)^2 \qquad (41)$$

where P_c is the carrier power and m is the modulation index. If m = 1, the peak envelope power is four times the unmodulated carrier power. The peak rf voltage on the collector under 100-per-cent modulation is at least four times the supply voltage for the AM transistor. A suitable transistor for AM operation therefore must have high current handling capability for good upward modulation and high voltage ratings to prevent second breakdown.

The 2N5992 provides 7 watts of carrier power with a minimum power gain of 10 dB at 88 MHz. Modulation higher than 90 per cent can easily be achieved with the driver slightly modulated. Emitter-site-ballasting employed in this device not only improves the modulation capability, but also enables the device to withstand high VSWR caused by antenna mismatch.

For FM applications, the 2N5993 can provide 18 watts of cw power with a power gain greater than 10 dB in the 66-to-88-MHz band. To insure device ruggedness, the 2N5992 and 2N5993 transistors are 100-per-cent tested under infinite VSWR through all phases at rated output power. For the AM type, the test is performed under full modulation. A typical circuit arrangement for testing output power, power gain, modulation index, and load-mismatch capability is shown in Fig. 100.

175-MHz Band

The circuit shown in Fig. 101 provides 25 watts for FM mobile transmission. It operates from a 12.5-volt supply and requires a drive power of 0.1 watt at 175 MHz. Transformers are used for interstage coupling, with matching networks that allow the circuit to be tuned for optimum performance.

Some of the design considerations discussed above for AM transmitters also apply to the design of FM circuits. In an FM transmitter, the transistor requirements are less stringent because the collector breakdown voltage needs to be only approximately twice the supply voltage, and the transistor can be driven very hard into saturation. In the absence of collector modulation,

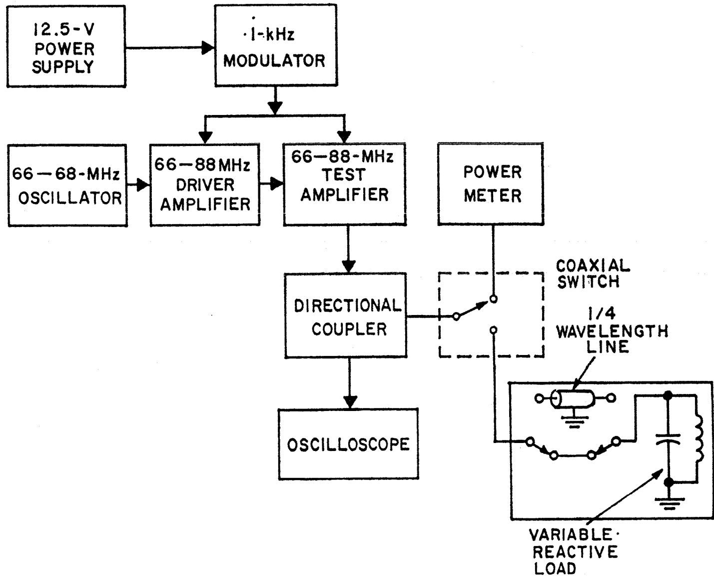
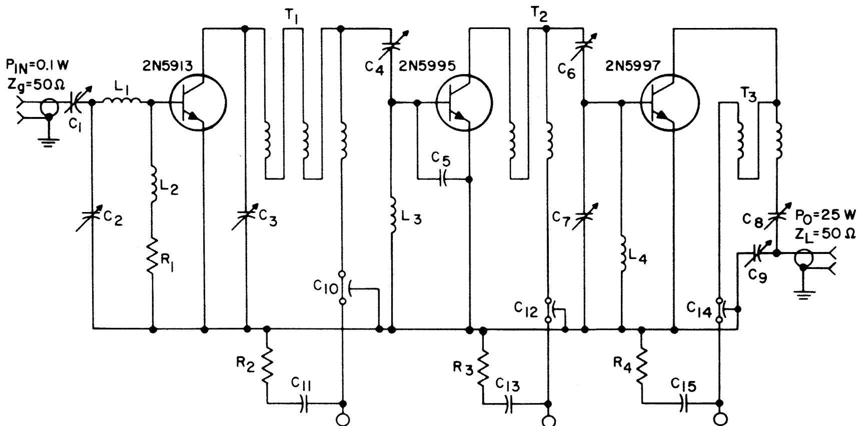


Figure 100. Test setup for testing output power, power gain, modulation index, and load-mismatch capability.



C_1, C_2 = 1.5 to 20 pF, Arco No. 402 or equiv.
C_3 = 10 pF, mica
C_4, C_8, C_9 = 14 to 150 pF, Arco No. 424 or equiv.
C_5 = 56 pF, mica
C_6, C_7 = 7 to 100 pF, Arco No. 423 or equiv.
C_{10}, C_{12}, C_{14} = 1000 pF, feedthrough
C_{11}, C_{13}, C_{15} = 0.01 μF, ceramic
R_1 = 33 ohms, ½ watt
R_2, R_3, R_6 = 10 ohms, ¼ watt

L_1 = 3 turns of No. 22 enamel wire, ¼-inch inner diameter, close wound
L_2, L_3, L_4 = No. 22 wire thread through Ferroxcube bead No. 56-590-65-4A or equiv.
T_1 = 3 twisted No. 22 wires, approximately 14 turns per inch, formed into a loop; ⅜-inch inner diameter; cross connected (i.e., end of one wire connected to beginning of the other)
T_2, T_3 = same as T, except only 2 twisted wires

Figure 101. 25-watt amplifier for mobile FM transmission.

parasitic oscillations in a FM amplifier are also minimized.

156-MHz Marine Band

The increasing number of boating enthusiasts has generated a demand for low cost reliable communication equipment operating in the 156-MHz marine band. The amplifier chain shown in Fig. 102 incorporates the techniques discussed for calculating matching networks and avoiding low-frequency oscillations. The amplifier operates from a 12.5-volt collector supply and delivers 32 watts at 156 MHz with an input power of 200 milliwatts. The design is quite conservative; the output stage can withstand any load mismatch at its rated output power.

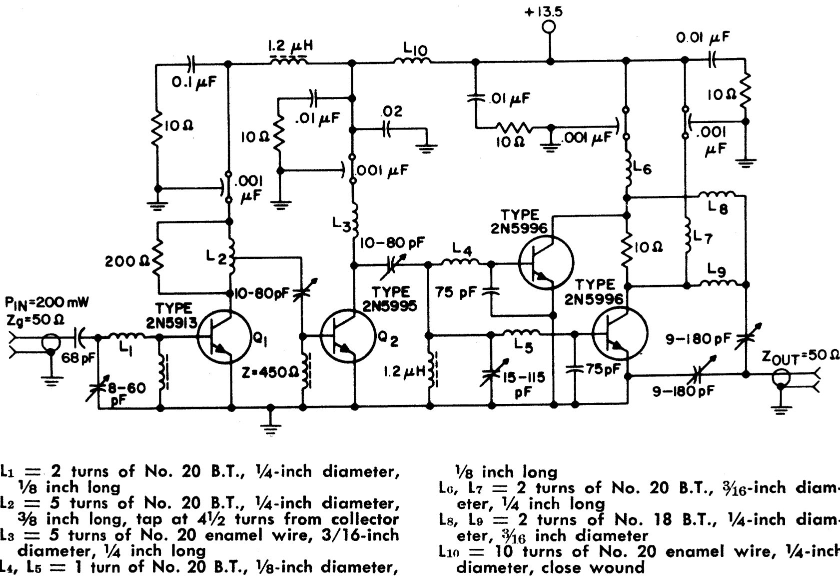


Figure 102. 32-watt, 156-MHz marine-band amplifier.

Commercial-Aircraft Radio

THE aircraft radios discussed in this section are of the type used for communication between the pilot and the airport tower. The transmitter operates in an AM mode on specific channels between 118 and 136 MHz. Radios of this type are regulated by both the FCC (Federal Communications Commission) and the FAA (Federal Aeronautics Administration). The FCC assigns frequencies to airports and places some requirements on the transmitters, particularly as regards spurious radiation and interference. The FAA sets minimum requirements on radio performance which are based on the maximum authorized altitudes for the plane, whether paying passengers are carried, and on the authorization for instrument flying. The FAA gives a desirable TSO certification to radio equipment that satisfies their standards of airworthiness.

The FCC checks aircraft-radio transmitter designs for interference and other electrical characteristics (as it does all transmitters). Additional requirements are specified for radios intended for use by scheduled airlines by a corporation supported by the airlines themselves. The name of this corporation is ARINC (Aeronautical Radio, Inc., 2551 Riva Road, Annapolis, Maryland 21401).

All these specifications combine to generate radio-transmitter requirements for different types of aircraft, as indicated in Table XIII.

DESIRABLE FEATURES

Because multiple channel use is necessary, it is desirable that aircraft radios have all 360 channels. These channels are spaced every 50 kHz from 118 to 136 MHz, and are assigned to specific airports. Each must be crystal-controlled. Synthesizer techniques are used to reduce the number of crystals required.

Simple, foolproof operation is necessary because the pilot has little time to spare and little interest in adjustments to the radio equipment. The frequency settings are made by switches that provide a digital read-out. "Squelch," volume, and on/off controls are added.

Size is important because the instrument panel is crowded. On large aircraft, the transmitter is operated by remote control by means of a set of switches on the

Table XIII—Four Popular Aircraft-Radio Transmitters (Designs by Aircraft Type*)

TYPICAL OWNER	NO. OF ENGINES IN AIRCRAFT	FAA & ARINC CLASS	VOLTAGE AVAIL-ABLE	TRANS-MITTER POWER (MIN.)	TYPICAL POWER RANGE		TRANSMITTER FEATURES
Private Planes	1	I	13 V	1 W	>1.5 W	Type #1	Low cost, few channels, may be portable
Owner/Pilot	1	I	13 V	4 W	>6.0 W	Type #2	Panel mounted, 90 or 360 channels.
Private/Business	2	II	28 V	4 W	6 to > 20 W	Type #2	
Chartered & Cargo	2-4	III	28 V	16 W	>20 W	Type #3	Remote Operation, 360 channels.
Scheduled Air Lines	2-4 Jets	III & ARINC	28 V	25 W	30 W	Type #4	Maximum reliability

* This chart is not complete or exact and is not intended to show actual requirements, but merely what is typical. Consult FAA for complete requirements.

panel. Weight and power drain are secondary considerations.

A primary consideration in all aircraft equipment is reliability. Spare radios are common in private aircraft, and are universal in aircraft equipped for instrument flying. The inherent reliability of transistorized equipment is a major advantage in aircraft radios.

DESIGN REQUIREMENTS

Amplitude modulation is an important design consideration for all transistor power amplifiers (as explained in the general section on AM). Amplitude-modulation requirements are set by the TSO at a minimum of 85 per cent, which corresponds to a PEP of 3 times the carrier power. Careful design is required to meet this specification because many factors tend to limit the PEP, including the decrease in transistor gain at high currents, transistor rf V_{CE}(sat), and modulator losses.

The owners and pilots of aircraft require reliable, foolproof operation of their radio equipment. Unfortunately, they are not often technically trained and do not appreciate the importance of proper maintenance of the antenna and the transmission line. These vulnerable items directly affect the performance of the radio because optimum performance is achieved only when the transmission line VSWR is unity. With a mismatch (i.e., VSWR greater than 1), the power output may be low, and there may be spurious or distorted output. Even more important is the fact that antenna and transmission-line faults stress the transmitter output stage with high voltage-current products and/or high power dissipation. These characteristics can overstress and destroy a weak transistor. The likelihood and the drastic effects of a load mismatch make the transmitter output transistor a primary influence on equipment reliability and make mandatory the selection of a transistor rugged enough to withstand

the possible stresses. Table XIV list several RCA transistor types characterized for operation in aircraft-radio applications.

Table XIV—RCA Transistor Types for Aircraft-Radio Applications (108 to 156 MHz)

40290	136	2	12.5	6	TO-39	
40291	136	2	12.5	6	TO-60	
40292	136	6	12.5	4.8	TO-60	AM Broadband Amplifier
2N5102	136	15	24	4	TO-60	
2N5994	118	15	12.5	7	TO-216AA	

Aircraft radios must cover the entire frequency range from 118 to 136 MHz. The mcre expensive radios cover all 360 channels. This 18-MHz bandwidth is a major design challenge which may be met by use of either a narrow-band step-tuned transmitter or a broad-band transmitter.

A broad-band transceiver design is possible with transistors. In a power transistor, the input may be considered to consist of the base-lead inductance L_b in series with r_{bb}'. If L_b is minimized, the Q is reduced, and broad-band operation is possible. The output-circuit Q is less of a problem than that of the input. The Q is formed by C_{ob} in parallel with the load impedance presented to the collector by the tuned circuit. Broad-band matching circuits between amplifier stages commonly use ferrite-core transformers of the transmission-line type (balun).

The use of broad-band amplifiers permits the largest portion of the transmitter to be remotely located without the need for expensive and complex servo tuning mechanisms. This feature is a great advantage in larger aircraft.

One problem encountered with a broad-band amplifier is reduction of harmonic output. Harmonics originate in the class C operating mode because of the nonlinear characteristics of transistors. These nonlinear characteristics, particularly the voltage sensitivity of C_{bc}, cause subharmonic-frequency generation as well as harmonic-frequency generation. The wide-band gain also increases the possibility of oscillation if any feedback exists. This condition is further intensified by the use of high-gain transistors or by excessive over-all gain.

The amount of harmonic output and transmitted interference permitted is rigidly specified by the FCC. A broad-band, band-pass filter should be added, therefore, after the transmitter.

POWER AND MODULATION

Because the only useful power in an AM transmitter is sideband power, it is reasonable to use this power as a reference in evaluation of the transmitter. When a single-tone sinusoidal modulating signal is used, the total sideband power P_{SB} in a modulated wave is given by

$$P_{SB} = P_{AV}\left(\frac{m^2}{2 + m^2}\right) \qquad (42)$$

where P_{AV} is the average power and m is the modulation index.

This relationship is convenient to use because P_{AV} is easy to measure and

$$P_{SB} = \frac{P_{AV}}{3} \quad (43)$$

for 100-per-cent modulation.

The performance of an AM transmitter can also be expressed in terms of peak envelope power PEP. The peak envelope power is equal to 2.66 P_{AV} in a 100-per-cent modulated wave. The value of PEP indicates the ultimate peak power-handling capabilities of the transistors being used.

It is unfortunate that carrier power is sometimes used as a reference in evaluation of the performance of AM transmitters, especially transistorized transmitters. Unlike the sideband power P_{SB}, the carrier power P_C does not always have a definite relationship to P_{AV} and PEP. When the carrier is used for a reference, "center shift" and "upward modulation" must be considered. Use of these terms in conjunction with P_C to define transmitter modulation only complicates the definition of per-cent amplitude modulation. For example, Fig. 103 shows an ampli-

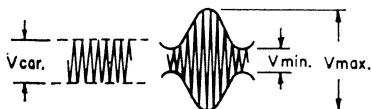


Figure 103. The amplitude modulated wave; V_{car} is the amplitude of carrier before modulation.

tude-modulated wave. The amplitude modulation AM in per cent is defined as follows:

$$AM = \left(\frac{V_{max} - V_{min}}{V_{max} + V_{min}}\right) \times 100 \quad (44)$$

Use of this equation indicates that when $V_{min} = 0$, the wave is 100-per-cent modulated without reference to the carrier. The following expressions are based on carrier amplitude V_{car} or carrier power P_C:

$$AM = \left(\frac{V_{max}}{V_{car}} - 1\right) \times 100 \quad (45)$$

$$P_{AV} = P_C\left(1 + \frac{m^2}{2}\right) \quad (46)$$

These expressions contain the tacit assumption that carrier level must not vary from the unmodulated state, which may not be the case. If the modulation is adjusted to 100 per cent by the use of Eq. (401) and P_{AV} is measured, values can easily be computed for P_{SB}, PEP, and even P_C.

DESIGN TECHNIQUES

The need for wideband performance in aircraft transmitters precludes the use of sharply tuned circuits to reduce harmonic power in the output; instead, low-pass filters are used. Any configuration of active devices that reduces the harmonic content in the output helps to ease the requirements placed upon these filters. One such configuration is a push-pull amplifier, which inherently has low even harmonics in the output. The higher input impedance of a push-pull stage as compared to a single-ended parallel combination of two transistors is also advantageous for obtaining wider bandwidths because only one-half as much current is injected into the input of push-pull transistors as into parallel devices during one-half cycle.

The coupling circuits in the amplifier of Fig. 104 are basically double-tuned interstage circuits, as shown in Fig. 105. R_1 and C_1 represent the collector output resistance and the collector output capacitance of the driver transistor. L_i and R_i represent the input series inductance and the input series resistance of a transistor. (For simplicity, coil resistances are omitted.) Q values for the two circuits shown in Fig. 105 are expressed as follows:

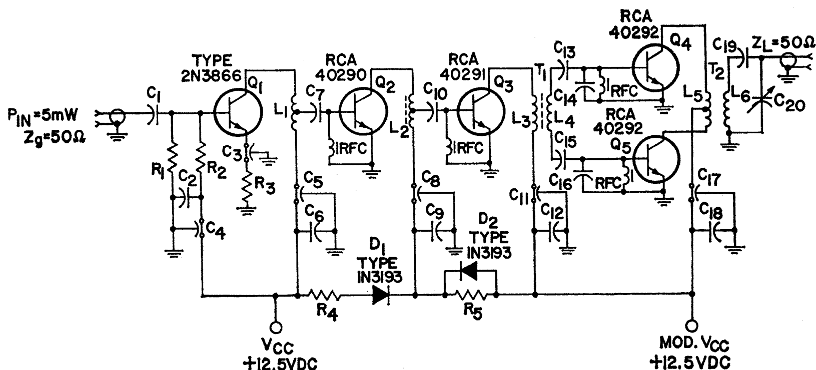
$$Q_1 = \frac{R_1}{\omega L_1} \qquad (47)$$

$$Q_2 = \frac{\omega (L_2 + L_i)}{R_i} \qquad (48)$$

For large bandwidths, it is desirable that Q_1 be much larger than Q_2. L_2, C_2, and L_i are series resonant at some frequency f_o within the bandwidth; L_1 and C_1 can then be determined as follows:

$$L_1 C_1 = \frac{1}{(\omega_o)^2} \qquad (49)$$

In practice, the resonant frequency f_o may not be exactly the center frequency of the passband, but may tend toward the high end of the bandwidth to compensate for degradation of the frequency response of the transistor itself. Normally, there is no problem obtaining relatively high values of Q_1 because transistors have large collector output resistance R_1. However, it is more difficult to obtain a low value of Q_2 in a transis-



C_1 = 300 pF, silver mica, Arco, or equiv.
C_2 = 0.005 μF, ceramic
C_3 C_4 C_5 C_8 C_{11} C_{17} = 1000 pF, feedthrough
C_6 C_9 C_{12} C_{18} = 0.5 μF, ceramic
C_7 = 50 pF, silver mica, Arco, or equiv
C_{10} C_{13} C_{15} = 82 pF, silver mica, Arco, or equiv.
C_{14} C_{16} C_{19} = 150 pF, silver mica, Arco, or equiv.
C_{20} = 8 to 60 pF, Arco #404, or equiv.
R_1 = 470 ohms, 0.5 W
R_2 = 1500 ohms, 0.5 W
R_3 = 47 ohms, 0.5 W
R_4 = 15 ohms, 0.5 W
R_5 = 33 ohms, 0.5 W

L_1 = 7 turns of No. 22 wire, 13/64" dia. 9/19" L. tap 1.5 T.
L_2 = 5.5 turns of No. 22 wire, 13/64" dia., closely wound, tap 2.0 turns
L_3 = 6 turns of No. 22 wire, 13/64" dia , interwind with L4
L_4 = 4 turns of No. 22 wire, 13/64" dia., interwind with L3
(L_2–L_4: Cambion IRN-9 core mat'l or equiv.)
L_5 = 5 turns of No. 22 wire, 13/64" dia. C.T., interwind with L6
L_6 = 5 turns of No. 22 wire, 13/64" dia., interwind with L5
R.F.C. = 1 turn of No. 28 wire on ferrite bead, Ferroxcube #56-590-65/4B or equiv.

Figure 104. A 118-to-136-MHz 40-watt PEP transistor amplifier.

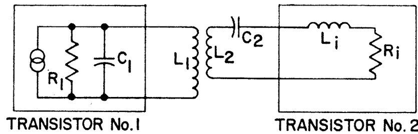


Figure 105. A double-tuned interstage.

tor double-tuned interstage circuit because high-power transistors have low series input resistance R_i. The contribution of the inductive series input reactance L_i may be sufficient to raise the value of Q_i to undesirable levels and thereby limit the obtainable bandwidth.

This problem can be solved by use of an L-section and its transforming properties. The inductive input impedance of a transistor may be represented by the solid lines of Fig. 106.

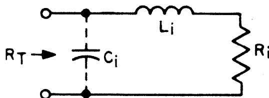


Figure 106. Transistor input as an L-section.

The definite Q value associated with this input impedance may be represented as Q_i. If a capacitor C_i is added to the transistor input of Fig. 561, as shown by the dotted line, the resistance R_i can be transformed up by the L-section to a new value R_T, as follows:

$$R_T = R_i\,(Q_i^2 + 1) \qquad (50)$$

The value of the capacitor C_i is calculated as follows:

$$C_i = \frac{1}{\omega R_T}\sqrt{\frac{R_T}{R_i} - 1} = \frac{L_i}{\omega^2 L_i^2 + R_i^2} \qquad (51)$$

When an L-section is used in conjunction with a double-tuned interstage circuit, the value Q_2 of the second circuit is given by

$$Q'_2 = \frac{\omega L_2}{R_T} \qquad (52)$$

This value is, of course, lower than that shown in Eq. (405). Consequently, an L-section can be used to match resistances of not-too-different magnitudes and at the same time maintain low values of Q. The value of L_i in the circuit is given by

$$L_i = \frac{R_i}{\omega}\sqrt{\frac{R_T}{R_i} - 1} \qquad (53)$$

There are limits to the results that can be accomplished with this type of transformation. For some combination of L_i and R_i, the required value of C_i may be too large to be practically realizable. In addition, R_T is a frequency-dependent parameter. For very low values of Q_i, the capacitor C_i loses its effectiveness because R_T becomes very nearly equal to R_i.

Double-tuned interstage coupling circuits are used throughout the amplifier shown in Fig. 559. When it is necessary to use a two-winding transformer, as in the case of T_1 and T_2, bifiliar windings are employed for tighter coupling. In other cases, autotransformers with their high coefficient of coupling are used quite successfully. Eq. (49) is used as the starting point for determination of the inductances in the primaries of the double-tuned interstages; the collector to base capacitance C_{CB} of the transistor

is substituted for C_1. Turn ratios are determined by the impedance levels to be transformed. The load resistance R_L for each stage is determined as follows:

$$R_L = \frac{(V_{CC})^2}{2P_o} \qquad (54)$$

where V_{CC} is the collector supply voltage and P_o is the power output. The collector-to-emitter saturation voltage is omitted for simplicity.

A single 40292 transistor is capable of delivering 6 watts of output power with an input of 2 watts and a supply of 12.5 volts dc at 135 MHz. For these conditions, the load resistance R_L is given by

$$R_L = \frac{(12.5)^2}{12} = 13 \text{ ohms}$$

This value of 13 ohms from one-half of the primary winding of T_2 is transformed to 50 ohms in the secondary winding. This impedance level allows the use of a 1:1 transformer, which is convenient for bifilar winding. For 40292 transistors, R_i is approximately 6 ohms and X_{Li} is about 3 ohms. An L-section is used in the inputs to the 40292 transistors in the push-pull amplifier. To maintain a low value of Q_i, the leads on the base-to-emitter capacitors (C_{14} and C_{16}) are kept short, and the capacitors are placed as close to the base and the emitter as possible. The values of C_{14} and C_{16} of Fig. 104 are determined empirically. The effective capacitances may differ appreciably from the nominal value of 150 picofarads shown.

Drive power of about 3 to 3.5 watts is required for the push-pull amplifier. This power is provided by the 40291 driver transistor operating into a 24-ohm load

$$\left[R_L = \frac{(V_{CE})^2}{2P_o} = (12.5)^2/6.5 \right]$$

Because the input resistance to the driver is sufficiently high (12 ohms), no L-section is used. The load resistance for the 40290 predriver transistor is selected to provide the required input to the driver of about 0.6 watt. The 100-milliwatt input required for the pre-driver stage is supplied by the 2N3866 class A input stage. Again, a double-tuned interstage circuit is used for coupling. The class A amplifier is biased to a quiescent current of 40 milliamperes for maximum gain, and has a load line of approximately 300 ohms, which is computed from

$$R_{\text{load line}} = \frac{V_{CC}}{I_C} \qquad (55)$$

An autotransformer is used to transform the 300-ohm load down to about 12 ohms at the predriver. The input of the 2N3866 stage is matched to the 50-ohm source. This stage has a gain of about 13 dB which increases the power from the 5-milliwatt input. The problem of subharmonic generation is solved by use of cores in the interstage transformers. Stable operation is obtained if the stages are kept 1.25 inches apart.

The final amplifier and the driver are modulated symmetrically about the carrier level. The predriver is modulated more in a positive direction as a result of the resistor-diode arrangement (R_4, R_5, D_1, and D_2).

Several precautions should be taken to avoid conditions which may lead to the destruction of transistors. For example, over-modulation should not be allowed

to occur because excessive negative excursions of the collector voltage may forward-bias the collector-to-base junction to a destructive point. Also, when a transmitter is keyed off, a steady-state current flow in the order of 2 amperes is suddenly interrupted in the modulation transformer. The resulting transient voltages may easily exceed the transistor breakdown ratings. Use of a zener diode rated at twice the supply voltage in the collector circuit provides protection from this type of transient.

PERFORMANCE AND ADJUSTMENT

The curves of Fig. 107 show typical values of average modu-

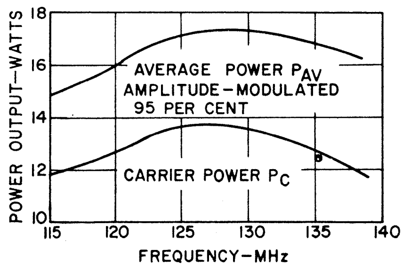


Figure 107. Typical output power as a function of frequency.

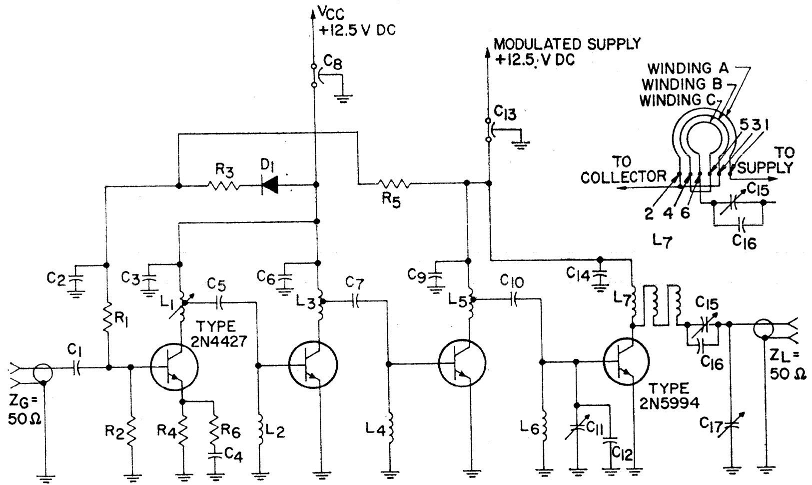
lated power P_{AV} at an amplitude modulation of 95 per cent, and carrier power P_C, as measured by a bolometer-type power meter. The peak envelope power PEP is computed as follows:

$$PEP = P_{AV} + \frac{(1 + m^2)}{1 + \frac{m^2}{2}} \qquad (56)$$

Output-power variation across the aircraft band is about 0.5 dB for both curves shown in Fig. 107. For this performance, the coil L_1 was stretched or compressed for maximum power output at 136 MHz and optimum bandwidth, and the trimmer C_{20} was adjusted for the best combination of output flatness and efficiency. Efficiency is somewhat better at higher frequencies than at lower frequencies; harmonic rejection is better at lower frequencies, and may be as good as 20 dB. A spectrum analyzer is required for detection of subharmonics when the slugs in L_2 and T_1 are adjusted.

15-WATT AMPLIFIER

The schematic diagram of a 15-watt single-ended broadband amplifier is shown in Fig. 108. The output transistor, a 2N5994, is completely tested for load mismatch capability at 118 MHz with a VSWR of infinity through all phases under full modulation. A minimum modulation of 85 per cent can be achieved in the 118-to-136-MHz band. Typical rf performance of this amplifier is shown in Fig. 109.



C_1 = 91 pF, silver mica
C_2, C_3, C_9, C_{14} = 0.01 μF, ceramic
C_4 = 200 pF, silver mica
C_5 = 50 pF, silver mica
C_6 = 0.001 μF, ceramic
C_7 = 56 pF, silver mica
C_8, C_{13} = 1000 pF, feedthrough
C_{10}, C_{12} = 68 pF, silver mica
C_{11}, C_{15} = 8-60 pF, Arco 404 or equiv.
C_{16} = 10 pF, silver mica
C_{17} = 3-35 pF, Arco 403 or equiv.
D_1 = 1N3193
R_1 = 2700 ohms, ½ W
R_2 = 470 ohms, ½ W
R_3, R_5 = 5.1 ohms, ½ W
R_4 = 51 ohms
R_6 = 20 ohms, ¼ W
L_1 = 6 turns No. 18 wire, 0.225-in. dia., 0.4-in. long; tapped at 4½ turns from collector
L_2, L_4, L_6 = RFC 1 turn No. 28 wire, ferrite bead, Ferroxcube No. 56-590-65/4B or equivalent
L_3 = 7½ turns No. 18 wire, 0.15-in. dia., 0.55-in. long; tapped at 5½ turns from collector
L_5 = 5 turns No. 18 wire, 0.225-in. dia., 0.35-in. long; tapped at 2½ turns from collector
L_7 = 3 strands No. 20 enameled wire twisted at 6 turns/in., formed in a loop of ⅜-in. dia., then cross-connected

Figure 108. 15-watt amplitude-modulated amplifier for 118-to-136 MHz operation.

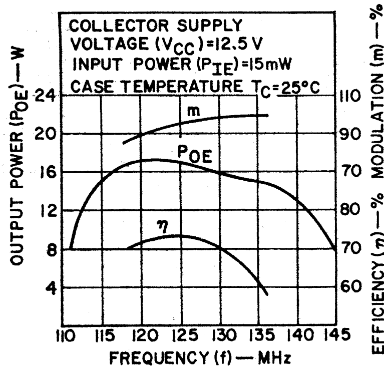


Figure 109. Typical broadband performance of the 118-to-136-MHz amplifier circuit shown in Fig. 108.

Community-Antenna Television

COMMUNITY-ANTENNA television (CATV) systems have experienced rapid growth in the last decade. These systems serve areas in which conventional antennas do not provide adequate television reception. The basic equipment consists of a "head-end" that picks up the signals, and a distribution system that delivers the signals to the subscriber's television receiver. The central antenna is erected at the most advantageous site for best reception; in remote locations, program reception is usually accomplished by means of microwave relays.

The distribution system has two major parts: the main transmission or "trunk" line, and the distribution or "feeder" line. The main trunk line consists of low-loss coaxial cable with main trunk amplifiers spaced along the cable. Bridger amplifiers are used to provide several outputs to the feeder lines from which signals are tapped off to individual subscribers. The backbone of the distribution system is the wide-band amplifier.

SYSTEM OPERATION

Fig. 110 shows a simplified block diagram of a CATV system in which the TV signals are received directly off the air (no microwave relay). Elaborate arrays of stacked antenna elements in conjunction with narrow-band preamplifiers are used to receive signals in each channel; the signals are then fed into a combining network. The combined multichannel signal is then fed into the main trunk line, which brings the signal from the antenna into the community. The trunk line consists of wide-band amplifiers spaced along a 75-ohm coaxial cable. The gain of each amplifier is adjusted to compensate for cable losses and attenuation characteristics. Typical trunk-line amplifier spacing is in the order of 2500 feet. At various points along the trunk line, signals are supplied to the feeder lines by **bridger amplifiers.** A bridger amplifier provides several outputs to the feeder lines from which signals are tapped off to in-

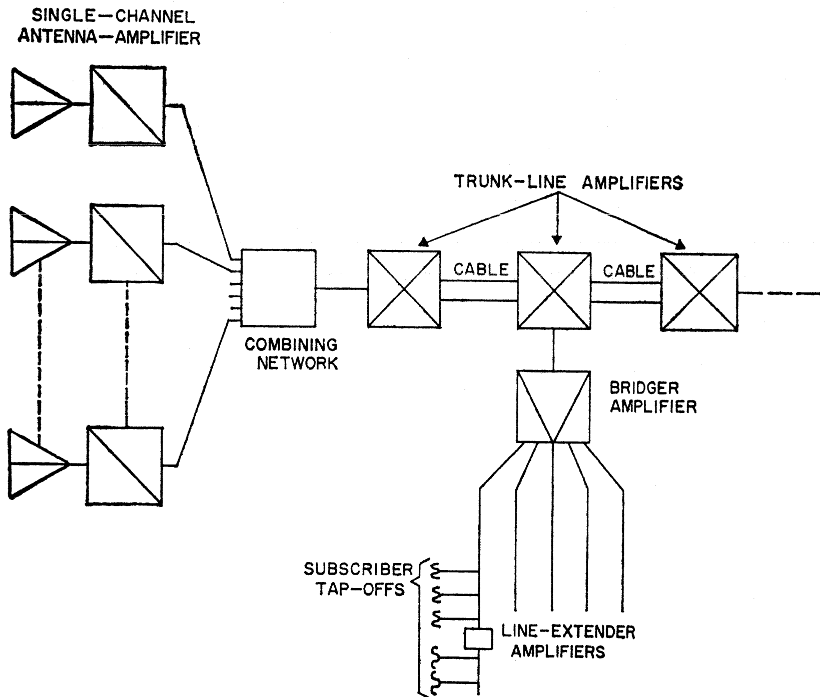


Figure 110. Simplified community-antenna television (CATV) system.

dividual subscribers. One or more line-extender amplifiers may be placed along each feeder line, depending upon its length and the number of subscribers.

AMPLIFIER REQUIREMENTS

The first requirement for CATV wide-band amplifiers is large bandwidth. The amplifiers should be able to cover a band of frequencies from 50 MHz to 300 MHz.

The next major consideration for a CATV wide-band amplifier is the required gain. The attenuation characteristic of a coaxial cable is a function of frequency; the cable losses increase logarithmically, as shown in Fig. 111. Typical loss is 0.4 dB per 100 feet at channel 2, and 1 dB per 100 feet at channel 13. The loss between trunk amplifiers is typically 25 dB at channel 13. The gain of the trunk amplifier operating at this spacing, therefore, should be 25 dB at 216 MHz. In addition,

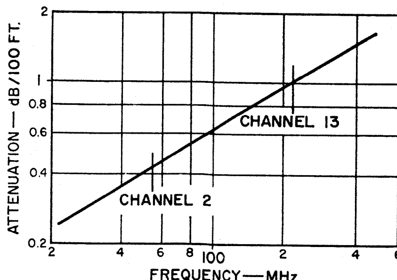


Figure 111. Attenuation characteristics of a coaxial cable as a function of frequency.

such an amplifier must be compensated for cable-attenuation differences at each channel by controllable "slope" or "tilt." The amplifier gain must be higher at the high end of the band than at the low end.

The final requirement is for output power or voltage, which is determined by the distortion and signal-to-noise-ratio requirements. If the level of power or voltage is too high, overloading and interference between channels occur; if the level is too low, the signal-to-noise ratio decreases. The most serious distortion is cross-modulation, which produces a "windshield-wiper" effect. Cross-modulation results when several channels are passing through a wide-band amplifier. The modulation of undesired interfering signals appears as modulation of the desired signal. The permissible cross-modulation level is 57 dB below the operating output-voltage level in an all-band CATV amplifier, at the end of the cable system.

"Snowy" pictures can be avoided if the signal at any point in a system is maintained at a level high enough to over-ride the noise. This relation is expressed by the signal-to-noise-ratio. The required ratios for various grades of picture quality have been determined as follows: 45 dB for excellent picture (no perceptible snow), 36 dB for fine picture (snow just perceptible), and 29 dB for passable picture (snow definitely perceptible but not objectionable). The signal-to-noise ratio always decreases when a signal passes through an amplifier. The difference between the input signal-to-noise ratio in dB and the output signal-to-noise ratio in dB is defined as the **noise figure** in dB. Noise figure, therefore, is the measure of degradation of signal-to-noise ratio in an amplifier. The noise figure in a CATV cascaded system increases 3 dB each time the length of the system is doubled; the signal-to-noise ratio decreases 3 dB under the same condition.

Typical requirements for trunk-line amplifiers to be used in a CATV cascaded system are as follows:

Frequency Band
50 MHz to 300 MHz
Input Operating Level
= 10 dBmV
Output Operating Level
= 32 dBmV
Maximum Output Capability
= 50 to 55 dBmV
Gain
= 22 dBmV
Response
±0.5 dB over the band
Noise Figure
= 12 dB at channel 13
= 8 dB at channel 2
Tilt
= 12 dB over the frequency range

These performance specifications must be met in outdoor temperatures ranging from −40 to 140°F. The following paragraphs discuss in more detail the basic considerations for the design of single-stage amplifiers suitable for use in CATV trunk lines and distribution amplifiers.

DESIGN RELATIONSHIPS FOR WIDE-BAND AMPLIFIERS

The gain-bandwidth product of a transistor connected in a common-emitter configuration is equal to f_T. Thus, the bandwidth of an

uncompensated common-emitter amplifier stage may be expressed as follows:

$$BW = f_T/h_{fe} = f_T\,(1-\alpha_o)/\alpha_o \quad (57)$$

where h_{fe} is the low-frequency common-emitter current gain of the transistor and α_o is the low-frequency common-base current gain. Eq. (57) dictates the bandwidth of a transistor amplifier stage if the source impedance is large and if the load impedance is small compared to the output impedance of the transistor. In practice, the load and source impedance are such that the bandwidth of the actual amplifier is smaller than the value determined from Eq. (57). Fig. 112

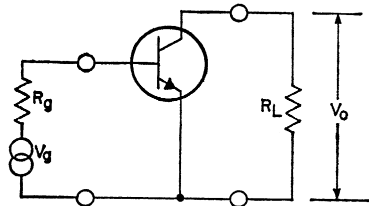


Figure 112. Common-emitter transistor amplifier.

shows a common-emitter transistor amplifier in which R_L is the load resistance and R_g the source resistance. The transistor can be represented by its hybrid-pi equivalent circuit, shown in Fig. 113(a), in which parasitics are not included. One difficulty with the circuit of Fig. 113(a) is the capacitance C_c, which prevents the circuit from being unilateral. The effect of C_c may be approximated by connecting a "Miller-effect" capacitance C_{eq} equal in value to $C_c\ (1 + \alpha_o\ R_L/r_e')$ from point b′ to ground and omitting the capacitance C_c entirely. The resulting equivalent circuit is shown in Fig. 113 (b). Because, in general, $1/\omega_T r_e' >> C_c$ and $\alpha_o \approx 1$, the value for C_{eq} is conveniently approximated by

$$C_{eq} \approx (1/\omega_T r_e')\ (1 + \omega_T\ C_C\ R_L) \quad (58)$$

With the aid of the simplified circuit of Fig. 113(b), the following equations for the gain and bandwidth of the amplifier are derived:

$$\text{Gain} = \frac{V_o}{V_g} = \left[\frac{\alpha_o\, RL}{(R_g+r_b')\,(1-\alpha_o) + r_e'}\right] \left[\frac{1}{1 + \dfrac{p\,(R_g+r_b')\,r_e'\,C_{eq}}{(1-\alpha_o)\,(R_g+r_b') + r_e'}}\right] \quad (59)$$

$$BW = \frac{(1-\alpha_o)\,(R_g+r_b') + r_e'}{(R_g+r_b')\,r_e'\,C_{eq}} = \frac{W_t}{(1 + W_t\,C_c\,R_L)} \left[1 - \alpha_o + \frac{r_e'}{(R_g+r_b')}\right] \quad (60)$$

Eq. (60) shows that the bandwidth is decreased by an increase in the load resistance R_L and increased by a reduction in the source resistance R_g. For a given R_g and R_L, the bandwidth is also increased by an increase in ω_T and by a decrease in r_b' and C_c. Thus, a transistor suitable for wide-band operation should have high f_T (or ω_T), a low collector capacitance C_c, and a low base resistance r_b'. If a transistor which has an f_T of 1.5 GHz and a C_c of 1.5 picofarads is used, the bandwidth calculated from Eq. (59)

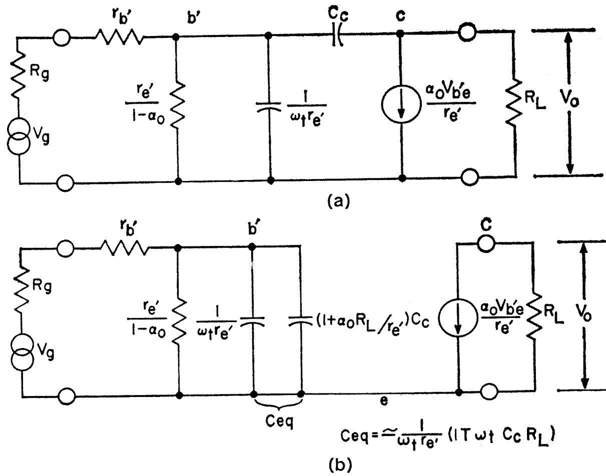


Figure 113. Equivalent circuits for common-emitter amplifier shown in Fig. 112: (a) without parasitic elements; (b) simplified equivalent circuit.

is 8 MHz for R_g = 75 ohms, R_L = 300 ohms, and I_C = 50 milliamperes. The corresponding voltage gain is 140. To obtain the bandwidth required in CATV, it is necessary to use compensation techniques that permit the trade of gain for increased bandwidth.

Transistor amplifiers cannot be designed to permit a gain-for-bandwidth trade in a 1:1 ratio. The voltage gain of a common-emitter amplifier stage, as can be determined from Eqs. (59) and (60), is not inversely proportional to the bandwidth. One of the important criteria of a wide-band transistor amplifier, therefore, is its ability to trade gain for bandwidth. Another way of stating this criterion is that degradation in gain-bandwidth should be small.

Collector-to-Base Shunt Feedback

One common method for trading gain for bandwidth in a common-emitter amplifier is by use of shunt resistance-inductance feedback, as shown in Fig. 114.

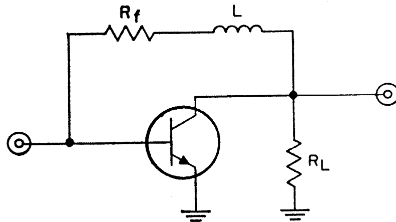


Figure 114. Common-emitter amplifier that uses shunt RL feedback to increase circuit bandwidth.

Simple feedback is provided from collector to base for trading gain and bandwidth. The current gain at low frequencies is approximately equal to the ratio of R_f to R_L. The feedback resistance R_f should be in the range $R_L < R_f < R_L/(1-\alpha_o)$. Without the inductance L, this technique is not an efficient way to obtain wide bandwidths because the feedback resistance becomes so low that it loads both the input

and output circuits and thus reduces the gain-bandwidth product. The effective input impedance is the input impedance of the transistor in parallel with a resistance equal to the feedback resistance R_f divided by $(1+K)$, where K is the voltage gain. The load presented to the collector of the transistor consists of the feedback resistance R_f in parallel with resistance R_L. However, if an inductance L is connected in series with the feedback resistance R_f, the gain-bandwidth product can be restored to its value without feedback. The inductance tends to remove the feedback resistance from the circuit at frequencies above $(1-\alpha_o)\, f_T$, and thus eliminates its effect on the high-frequency current amplification. The approximate expression for determining the value of the inductance is as follows:

$$L = (R_f + r_b' + R_L)\, R_f/2\pi\, f_T\, R_L. \qquad (61)$$

Emitter Degeneration

An-

Another common method of trading gain for bandwidth is to use emitter regeneration or emitter peaking, as shown in Fig. 115. Simple resistance-capacitance

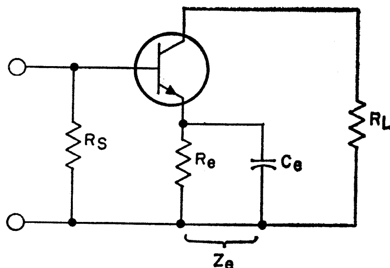


Figure 115. Transistor amplifier that uses emitter degeneration to increase circuit bandwidth.

feedback is provided from the amplifier to ground for trading gain for bandwidth. The effect of the resistance R_e is to reduce the gain at low frequencies. The equations for the voltage gain and the bandwidth of this amplifier, when $Z_e = R_e$, are as follows:

$$\text{Gain} = \frac{V_o}{V_g} = \left[\frac{\alpha_o R_L}{\alpha_o R_e + r_e' + (1-\alpha_o)(R_s + r_b' + R_e)}\right] \left[\frac{1}{1 + \dfrac{pC_{eq}\, r_e'\,(R_s + r_b' + R_e)}{\alpha_o R_e + r_e' + (1-\alpha_o)(R_s + r_b' + R_e)}}\right] \qquad (62)$$

$$BW = \frac{\alpha_o R_e + r_e' + (1-\alpha_o)(R_s + r_b' + R_e)}{r_e'\,(R_s + r_b' + R_e)\, C_{eq}} \qquad (63)$$

Comparison of Eqs. (60) and (63) shows that increased bandwidth is possible if $R_s + r_b' > R_e$. The effect of the capacitance C_e in shunt with the emitter resistance R_e is to decrease the degeneration at high frequencies. The required value of capacitance C_e is approximately equal to $1/(15 f_T R_e)$.

Fig. 116 shows a typical amplifier that might be used as a single stage in a CATV line amplifier. Generally, four or five such stages are cascaded to provide the gain of 20 to 30 dB required for a trunk-line or extender amplifier. This amplifier uses both shunt and series feedback to achieve the simultaneous requirements of good input and output match, broadband performance, satisfactory gain, and good linearity.

To realize optimum linearity, most transistors require a load impedance other than the cable-

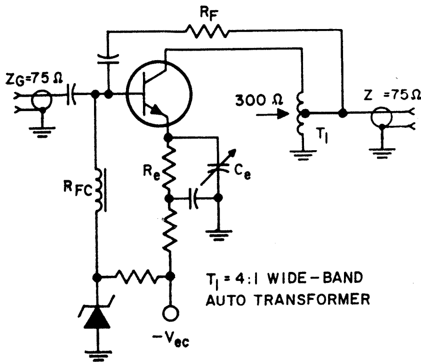


Figure 116. Typical CATV single-stage amplifier.

line impedance of 75 ohms. A typical optimum impedance, as shown in Fig. 116, is about 300 ohms. To achieve this impedance at the transistor collector terminals, a transformer such as that shown in Fig. 117 is often used.

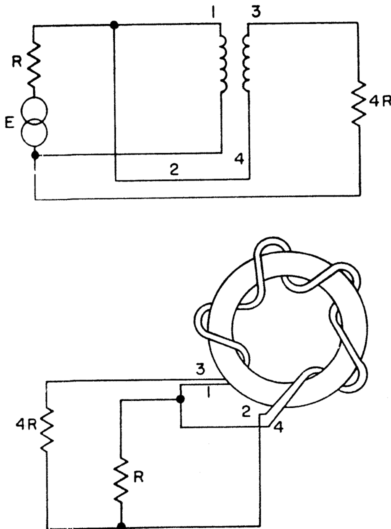


Figure 117. Wide band transformer. This transformer may be used to provide a 4:1 impedance ratio, as indicated in top diagram. The transformer is basically a twisted-pair transmission line wound about a ferrite torroid, as shown in lower diagram.

This transformer consists of a ferrite toroid around which a twisted pair of wires are wound. It is a transmission-line type and has excellent bandwidth. The transmission lines take the form of twisted pairs of wires. The coils are arranged so that the interwinding capacitance is a component of the characteristic impedance of the line, and forms no resonances which seriously limit the bandwidth, as in the case of a conventional transformer. For this reason, the windings can be spaced closely together to assure good coupling. Transformers of this type can provide good high-frequency response (this response is determined by the length of the windings).

The low-frequency response, on the other hand, is determined by the permeability of the core. The greater the core permeability, the fewer the turns required for a given low-frequency response and the larger the bandwidths. Thus, a good core material is desirable. Ferrite toroids have been found very satsfactory. The permeability of some ferrites is very high at low frequencies and decreases at higher frequencies. Large reactance, therefore, can be obtained with few turns at low frequencies. When the permeability decreases, the reactance is maintained by the increase in frequency, and good response is obtained over a large frequency range. It is important that coupling be high at all frequencies, or the transformer action fails.

The transformer shown in Fig. 117has an impedance ratio of 4:1. The high-frequency response of this transformer may be calculated by use of the following equation:

$$\frac{\text{Power Available}}{\text{Power Output}} = \frac{(1+3\cos B\,l)^2 + 4\sin^2 B\,l}{4\,(1+\cos B\,l)^2} \quad (64)$$

where B is the phase constant of the line and l is the length of the line. The response is down 1 dB when the line length is $\lambda/4$; the response is zero at $\lambda/2$. For wide-band response, therefore, this transformer must be made small.

TRANSISTOR CONSIDERATIONS

In selection of a transistor for CATV amplifier applications, the following performance criteria must be considered: maximum cross-modulation at a given output level, maximum IMD for a given level, noise figure, dc operating conditions, and dissipation. Obviously, optimization of all of these characteristics cannot be achieved simultaneously, so trade-offs must be made, taking into consideration the stages in which the transistor is to be used. For instance, low noise figure and moderate output levels are associated with the input stage of a line amplifier, while extremely low distortion is required for the output stage. Table XV lists several RCA transistor types characterized for operation in CATV and low-noise applications.

CROSS-MODULATION

One of the more serious types of nonlinearities associated with a CATV amplifier is cross-modulation. Cross-modulation is the transfer of modulation for one AM signal to another within an amplifier. Because CATV line amplifiers process many signals simultaneously, this exchange of signal information is highly undersirable. Cross-modulation generally sets the upper limit on the output signal level at which an amplifir may operate.

Cross-modulation is caused by odd- (predominantly 3rd-) order nonlinearities in the amplifier's transfer characteristic. All the amplifier nonlinearities are, of course, attributable to the transistor. Nonlinearities within the transistor may be separated into three major areas: (1) emitter-

Table XV—RCA Transistor Types for CATV/MATV and Small-Signal, Low-Noise Applications

Type	Operating Frequency (MHz)	Power Gain (Min.) (dB)	Noise Figure (dB)	Collector-to-Emitter Voltage, (V)	Package Type	Primary Application
2N918	60	13	6	6	TO-72	Low-Noise Amplifier/Oscillator
2N2857	450	12.5	4.5	6	TO-72	Low-Noise Linear Amplifier
2N3600	200	17	4.5	15	TO-72	Low-Noise Amplifier/Oscillator
2N3839	450	12.5	3.9	6	TO-72	Low-Noise Amplifier/Oscillator
2N5109	200	11	3	15	TO-39	Low-Noise Linear Amplifier (CATV)
2N5179	200	15	4.5	6	TO-72	Low-NoiseAmplifier/Converter
40294	450	12.5	4.5	6	TO-72	High-Reliability Version of 2N2857
40608	200	11	3	15	TO-39	Low-Noise Linear Amplifier (CATV)

base diode nonlinearities, (2) current gain (h_{fe}/Ic) nonlinearities, and (3) collector-base depletion-layer capacitance variations.

Through proper selection of bias point and amplifier circuit (primarily collector load line), a device can be optimized for minimum cross-modulation.

Because cross-modulation is a transfer of modulation from one channel to another, it can be measured by determining the degree of modulation produced on an unmodulated carrier by various combinations of interfering signals. The basic test for cross-modulation is shown in Fig. 118. unmodulated signal is substituted for the picture signal on the viewing channel. This technique provides a white screen which does not change during the test, and allows more consistent and critical observations.

While the "white-screen" test is simple, the resulting output ratings depend somewhat on the judgment of the person carrying out the test. For more accurate results a method that gives repeatable readings is needed.

Fig. 119 shows a block diagram of a standard test set-up. Any combination of the twelve crystal-controlled carriers is available.

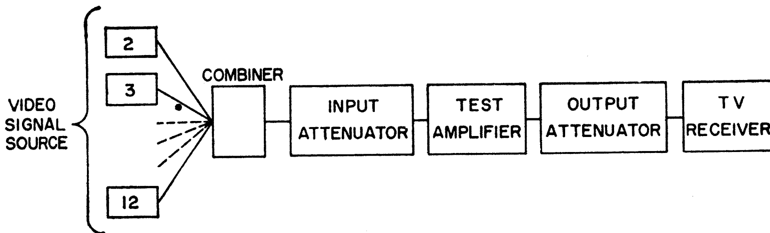


Figure 118. Block diagram of basic test setup used for cross-modulation measurements.

A number of clean TV (modulated) signals at the various channel frequencies are combined and fed through the amplifier under test. The output signal is viewed on a good television receiver, and the output levels are increased until "windshield-wiper" (cross-modulation) effects are just visible in the picture. The level at which this condition occurs is called the maximum usable output of the amplifier.

This test is not really conclusive, however, because TV "windshield-wiper" effects can be seen much more readily on some pictures than on others. The accuracy of the test is greatly increased if an Each one is 100-per-cent modulated with a 15-kHz square wave, except the one to which the receiver is tuned; this carrier is unmodulated, but its amplitude is the same as the peak amplitudes of the modulated carriers.

100-per-cent modulation is applied momentarily to the modulated channel and a 100-per-cent modulation reference level is noted on the 15-kHz VTVM. Modulation is then removed from the test channel. The cross-modulation is then read on the 15-kHz VTVM as a fraction of the 100-per-cent modulated value. The receiver is tuned to each of these channels successively and the cross-modulation

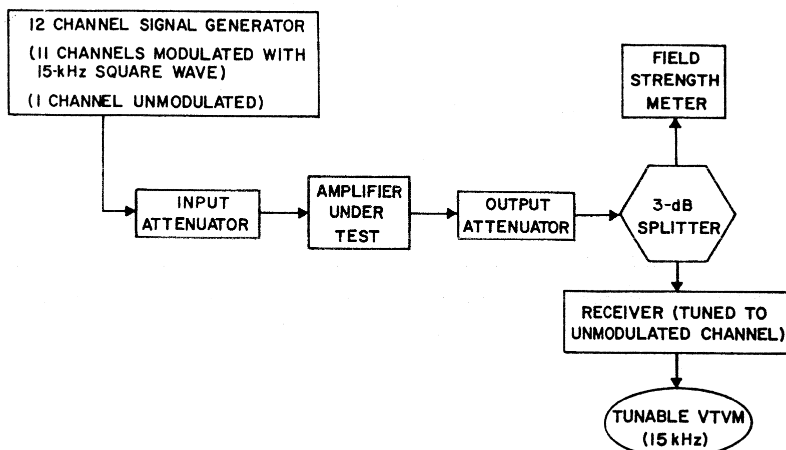


Figure 119. Block diagram of a 12-channel modulation test set.

readings are recorded. The amplifier rating is based on that channel which shows the greatest cross-modulation. When an amplifier is being tested, its "behavior" is determined by changing the settings of the two attenuators to alter the output levels in, for example, 2-dB steps to find if the cross-modulation follows the "two-for-one" law expected of a "well-behaved" amplifier. If the amplifier does not change "two-for-one," there is a likelihood that some form of cancellation of nonlinearities is taking place.

ANALYSIS OF TRANSISTOR CHARACTERIZED FOR CATV APPLICATIONS

The RCA-2N5109 transistor, packaged in a TO-39 case, is designed to provide large dynamic range, low distortion, and low noise, and is well suited for use in a wideband amplifier in CATV applications.

Characteristics

The 2N5109 is an epitaxial silicon overlay transistor that features low r_b' and C_C and high and relatively flat f_T with current level. Fig. 120 shows the f_T of a typical 2N5109 as a function of

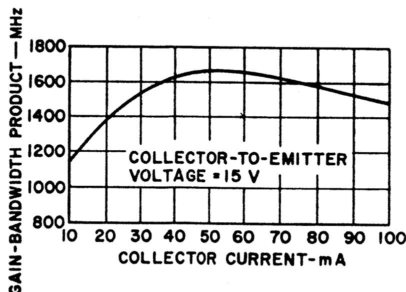


Figure 120. Gain-bandwidth product as a function of collector current for a typical RCA-2N5109 transistor.

collector current at a V_{CE} of 15 volts. The f_T measured at a collector current of 50 milliamperes is 1.5 GHz; f_T is within 20 per cent of its maximum value from 25 to 100 milliamperes. Fig. 121 shows the f_T of a typical 2N5109

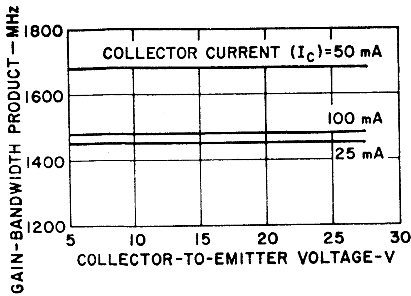


Figure 121. Gain-bandwidth product as a function of collector voltage for a typical RCA-2N5109 transistor.

as a function of V_{CE} at collector currents of 25, 50, and 100 milliamperes, respectively. The electrical characteristics of this transistor are summarized in Table XVI.

Fig. 122 shows the noise figure of a typical 2N5109 as a function of collector current. The noise figure is measured with the 2N5109 operating as a narrow-band 200-MHz amplifier at a V_{CE} of 15 volts. The best noise figure occurs at a collector current of less than 10 milliamperes.

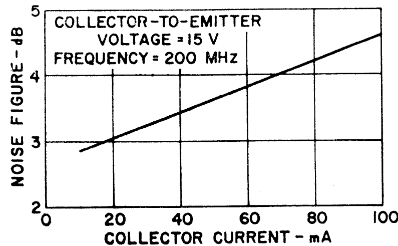


Figure 122. Noise figure as a function of collector current for a typical RCA-2N5109 transistor.

Table XVI—Electrical Characteristics of RCA-2N5109 Overlay Transistor (Case Temperature = 25°C)

CHARACTERISTIC	SYMBOL	TEST CONDITIONS: DC COLLECTOR VOLTS V_{CB}	V_{CE}	DC CURRENT (mA) I_B	I_C	LIMITS Min.	Max.	UNITS
Collector-Cutoff Current	I_{CEO}		15				20	μA
Collector-to-Base Breakdown Voltage	BV_{CBO}				0.1	40		V
Collector-to-Emitter Voltage (Sustaining)	V_{CER}(sus)*				5	40		
	V_{CEO}(sus)				5	20		V
Emitter-to-Base Breakdown Voltage	$V_{(BR)EBO}$			0.1	0	3		V
Collector-to-Emitter Saturation Voltage	V_{CE}(sat)			10	100		0.5	V
Collector-to-Base Capacitance (Measured at 1 MHz)	C_{ob}	15					3.5	pF
Small-Signal Common-Emitter Forward-Current Transfer Ratio (Measured at 200 MHz)	h_{fe}		15		20	4.8		
			15		50	6.0		
			15		100	4.8		
Voltage Gain (Wideband, 50 to 216 MHz)	V.G.		15		50	11		dB
Cross Modulation at 54-dBmV Output	C.M.		15		50		—57 (typ)	dB
Power Gain (Narrow-band, Measured at 200 MHz; P_{in} = —10 dBmV)	P.G.		15		10	11		dB
Noise Figure (Measured at 200 MHz)	N.F.		15		10		3 (typ)	dB

* With external base-to-emitter resistance R_{BE} = 10 ohms.

Choice of Operating Conditions

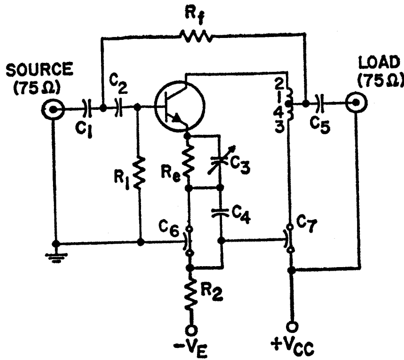
The most important parameter in the input stage of a CATV system is the noise figure. Distortion is not usually important in the input stage because the voltage and current swings of the transistor are small. The dc bias of the transistor should be chosen for mimimum noise figure. An RCA-2N5109 used in the first stage should be biased at a collector current I_C of 10 milliamperes and a collector-to-emitter voltage V_{CE} of 10 to 15 volts. The noise figure of a typical 2N5109 measured in a CATV amplifier is 8 dB at channel 13.

The final stage, on the other hand, should be biased so that maximum power output can be obtained with minimum cross-modulation distortion. In addition, the bias condition should be within the dissipation capability of the transistor. For example, if it is assumed that 1 volt rms (60 dBmV) is required across a load of 75 ohms, the peak-to-peak voltage swing across the 75-ohm load is 2.83 volts, and the corresponding current swing is 36.4 milliamperes. The 75-ohm load must be transformed into the collector load with the use of a wideband transformer. If a 1:1 impedance transformer is used, the collector voltage and current swings are the same as those of the 75-ohm load. The collector bias current I_C can then be selected from Fig. 120 for minimum change in f_T within the 36.4-milliampere current swing. The value of I_C that satisfies this condition is approximately 60 milliamperes. The V_{CE} corresponding to a collector load of 75 ohms is therefore 4.5 volts. Figs. 120 and 121 show that large current swings (rather than voltage swings) result in a change in f_T and, therefore, in large distortion.

If a 4:1 impedance transformer is used, the collector load becomes 300 ohms and the collector voltage and current swings become 4.66 volts and 18.2 milliamperes, respectively. From Fig. 120, the value of I_C can be chosen as 55 milliamperes for a minimum change of f_T within this current swing. The V_{CE} value corresponding to the collector load of 300 ohms is 16.5 volts. Fig. 121 shows that the f_T is substantially constant within the 4.66-volt swing around 16.5 volts. The power dissipation is 0.9 watt, which is within the limit of the 2N5109. The power output for a typical 2N5109 operated at 16.5 volts and 55 milliamperes in a 12-channel system is 52 dBm with −57 dB cross modulation.

WIDE-BAND AMPLIFIER

A typical single-stage wideband amplifier circuit is shown in Fig. 123. This common-emitter class A amplifier uses the RCA-2N5109 transistor and is designed for 75-ohm source and load resistances; it is suitable for the CATV application. A ferrite-toroid wideband transformer that has an impedance ratio of 4:1 is used in the output to transform a 75-ohm load into a 300-ohm collector load. Both shunt feedback and emitter degeneration are employed through R_f and R_e, respectively. Emitter peaking is accomplished by the use of C_3. Two dc power supplies are used; one supply provides the collector reverse bias, and the other supply provides the emitter-to-base forward bias through resistances R_1 and R_2.



C_1, C_2, C_5 = 0.002 μF
C_3 = 8-60 pF, Arco 404 or equiv.
C_4 = 0.03 μF
C_6, C_7 = 1500 pF
R_1 = 390 ohms, ½ watt
R_2 = 330 ohms, 1 watt
R_e = 6.8 ohms, ½ watt
R_f = 200 ohms, ½ watt
T = 4-turn bifilar winding, 3/16" ID, No. 30 wire; Core: G.I. material Q1 or equiv.

Figure 123. Single-stage wideband amplifier using the RCA-2N5109 transistor to provide a gain of 12 dB in the frequency band from 54 to 216 MHz.

The amplifier of Fig. 123 uses a 2N5109 operated at an I_C of 55 milliamperes and a V_{CE} of 16.5 volts, and can provide a minimum gain of 12 dB within the band of 54 to 216 MHz.

LOW-NOISE AMPLIFIERS

When a signal is processed by a system, a certain amount of extraneous noise is added, which degrades the original input signal-to-noise ratio.

The output signal-to-noise ratio of an amplifier with a noise figure F is 1/F times less than the input signal-to-noise ratio. The following equation expresses this relationship:

$$\frac{P_{SO}}{P_{NO}} = \frac{1}{F}\left(\frac{P_{Si}}{P_{Ni}}\right) \quad (65)$$

Eq. (65) is valid only when the input noise (P_{NI}) associated with the signal is equal to kT_oB; here k is Boltzmann's constant, $T_o = 290°$ K, and B is the system bandwidth. If the equivalent noise temperature is not 290°K, the effect of the 1/F factor may be increased or decreased, as shown by the following more general relationship:

$$\frac{P_{SO}}{P_{NO}} = \frac{GP_{Si}}{(GP_{Ni}) + (F-1)Gk\,T_oB} \quad (66)$$

where G is the system gain.

From Eq. (66) it can be seen that the value of P_{SO}/P_{NO} depends upon the level of input noise, P_{NI}. For high levels of P_{NI} and low values of F there is little degradation in signal-to-noise ratio of the incoming signal as it is amplified:

$$\frac{P_{SO}}{P_{NO}} \approx \frac{P_{Si}}{P_{Ni}}$$

If several amplifiers are cascaded, the total noise figure, F_{total}, is related to the noise figures and gains of the individual stages by the following equation:

$$F_{TOTAL} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1G_2} + \ldots + \frac{F_N - 1}{G_1G_2 \ldots G_{N-1}} \quad (67)$$

In Eq. (67), G_1 and F_1 are the gain and noise figure of the first stage of amplification. If G_1 is large, the noise figure of the entire amplifier chain is nearly equal to F_1.

The noise figure of a single-stage transistor amplifier is a function of frequency and transistor parameters, as shown by

the following equation:

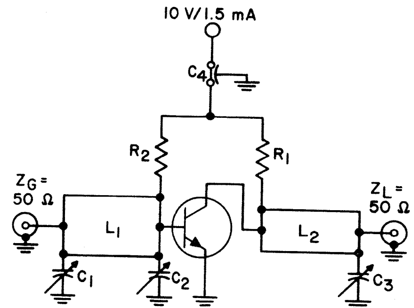
$$NF = 1 + \frac{r_b'}{r_g} + \frac{r_e'}{2\,R_g} + \frac{(r_b'+r_e'+R_g)^2}{2\,\alpha_o\,R_g\,r_e'} \left[\frac{I_{co}}{I_E} + \frac{1-\alpha_o}{\alpha_o} + \left(\frac{f}{f_T}\right)\right] \qquad (68)$$

At frequencies below approximately 0.1 f_T, the noise figure is constant with frequency and is primarily determined by R_g, r_b', r_e, and α_o. The resistance r_e' is inversely related to the dc emitter current ($r_e' = 26/I_e$); therefore, there is a value of I_e that corresponds to a minimum noise figure. At higher frequencies, the $(f/f_T)^2$ term in the noise-figure equation becomes predominant, with the result that the noise figure asymptotically approaches a 6-dB-per-octave slope. From the viewpoint of noise considerations, the r_b' and I_{co} of the transistor should be low, and f_T should be high. Eq. (68) shows that the noise figure is also a function of the source resistance R_g and, therefore, can be minimized by proper selection of R_g. The optimum source resistance can be determined if Eq. (68) is differentiated and the result is set equal to zero and solved for R_g. The following equation for R_g is then obtained:

$$R_g\,(\text{optimum}) = \left[(r_e'+r_b')^2 + \frac{\alpha_o\,r_e'\,(2\,r_b'+r_e')}{\frac{1-\alpha_o}{\alpha_o} + \left(\frac{f}{f_T}\right)^2 + \frac{I_{co}}{I_E}}\right]^{\frac{1}{2}} \qquad (69)$$

At low frequencies, where $(f/f_T)^2$ is small, a transistor that has high dc current gain requires a high source resistance R_g for best noise performance. As the frequency approaches f_T, the second term of Eq. (69) becomes small, and the optimum source resistance approaches $(r_b' + r_e')$.

The circuit shown in Fig. 124 is a narrow-band 1-GHz amplifier with a gain of 10 dB and a noise figure of 3.0 dB. Microstrip construction with variable tuning allows this circuit to be used for noise-figure testing or as an amplifier in a 50-ohm system.

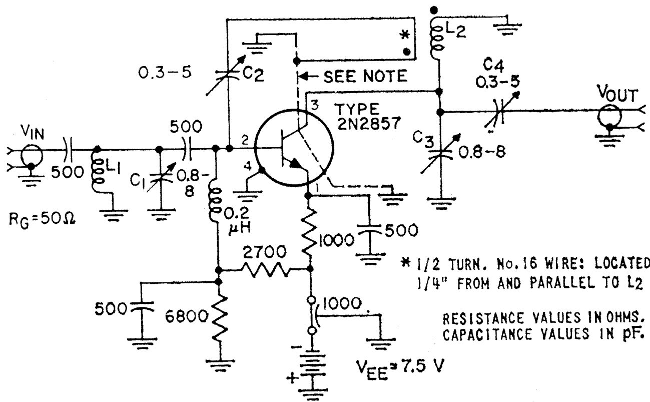


C_1, C_2, C_3 = 1 to 10 pF, Johanson No. 2954 or equiv.
C_4 = 1000 pF, feedthrough, Allen-Bradley type FAC5 or equiv.
R_1 = 2.7 ohms, 1/8 W, carbon
R_2 = 120K, 1/8 W, carbon
L_1 = microstrip transmission line, length = $\lambda/4$, Z_o = 30 ohms
L_2 = microstrip transmission line, length = $\lambda/4$, Z_o = 90 ohms

Figure 124. Low-noise 1-GHz amplifier.

Fig. 125 shows a circuit diagram of a typical low-noise 450-MHz amplifier that uses an RCA-2N2857 transistor. This amplifier provides a gain of 12.5 dB with a noise figure of 4.5 dB.

Fig. 126 shows a typical test arrangement for measurement of transistor noise figure.



L_1, L_2 = silver-plated brass rod, 1½ inch long-by-¼ inch diameter; install at least ½ inch from nearest vertical chassis surface

Note: External interlead shield to isolate the collector lead from emitter and base leads.

Figure 125. Neutralized amplifier circuit used to measure 450-MHz power gain and noise figure for type 2N2857.

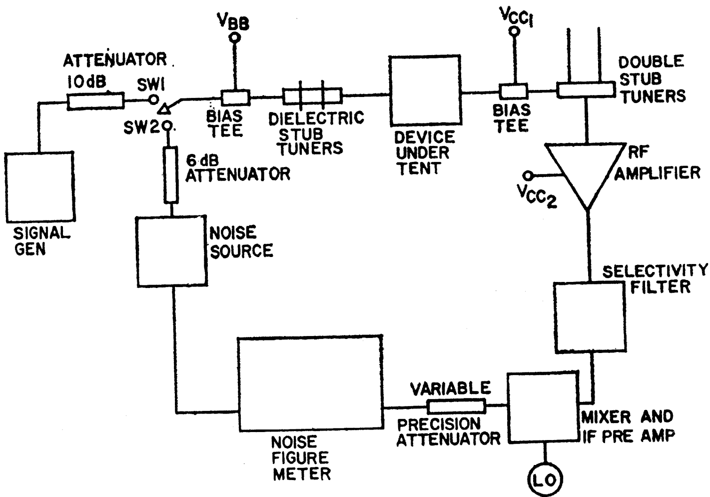


Figure 126. Typical noise-figure test set.

Microwave Power Amplifiers

THE considerations of construction, geometry, and ratings that were discussed previously for rf power transistors are particularly important at microwave frequencies.

TRANSISTOR SELECTION

The selection of the proper transistor for a specific microwave application is determined by the required power output, gain, and circuit preference. RCA offers the circuit designer a wide variety of microwave power transistors from which to choose the optimum type for his application. Such transistors are available in either coaxial or stripline packages. Units that are optimized for operation from either 22 or 28 volts and that are packaged specifically for either amplifier or oscillator applications are also available. Table XVII lists RCA transistor types characterized for operation in amplifier or oscillator applications at microwave frequencies.

At a frequency of 2 GHz, a single microwave power transistor can supply an output power of more than 10 watts with a gain of 7 dB. If higher output power is required, two or more transistors can be connected in parallel, either directly or by use of hybrid combiners. Higher gain can be obtained by connection of several stages in cascade. The transistor supply voltage is usually dictated by the application. For example, the supply voltage in telemetry systems and microwave relay links normally ranges from 20 to 23 volts; most other systems use 28 volts. The type of package selected, either coaxial or stripline, usually depends on the type of circuit in which the designer plans to use the transistor.

PACKAGE DESIGN

The output power, power gain, efficiency, and bandwidth are all strongly affected by package parasitic elements, which become increasingly significant at

Table XVII—RCA Microwave Power Transistors

Type No.	Collector Efficiency (%)	Collector-to-Emitter Voltage (volts)	Frequency (GHz)	Output Power (W)	Power Gain (dB)	Package Type
			Stripline Types			
2N6265	33	28	2.0	2	8.2	HF-28
2N6266	33	28	2.0	5	7	HF-28
2N6267	35	28	2.0	10	7	HF-28
2N6268	32	22	2.3	2	7	HF-28
2N6269	35	22	2.3	6.5	5	HF-28
			Coaxial Types			
2N5470	30	28	2.0	1	5	TO-215AA
2N5920	40	28	2.0	2	10	TO-215AA
2N5921	40	28	2.0	5	7	TO-201AA
40898	35	22	2.3	2.0	7.0	TO-215AA
40899	35	22	2.3	6.5	6.5	TO-201AA
40908	45	22	1.7	7.6	8.8	TO-201AA
			Oscillator Types			
40836	20	21	2.0	0.5		TO-215AA
40837	20	28	2.0	1.25		TO-215AA
40909	20	25	2.0	2.0		TO-201AA

higher frequencies. Therefore, the package design is an especially important feature of microwave transistors.

A suitable high-power transistor package for microwave applications must have low common-lead inductance and low shunt and feedthrough capacitance, as well as good thermal properties.

Packages such as the TO-39 and TO-60 are useful in applications above 1 GHz, but superior performance can be obtained with stripline and coaxial packages specifically designed for microwave frequencies. Fig. 23 in the section on **Special features of High-Frequency Power Transistors** showed a stripline package (HF-28) and two coaxial packages (HF-11 and HF-21) that are used at frequencies well into S-band. Best microwave-frequency amplifier performance is provided by common-base configurations in these packages. In the stripline package, the base is connected directly to the flange, and in the coaxial package, the base is connected to the flange separating the ceramics for amplifier units. Coaxial-package transistors in which the emitter is connected to the flange to enhance feedback are also available for some oscillator applications. In both types of packages, the base can be rf-grounded with very low parasitic lead inductance; this grounded-base configuration minimizes out-put-to-input feedback and, therefore, facilitates stable operation.

CIRCUIT DESIGN TECHNIQUES

In designing transistor microwave power circuits, several fundamental considerations must be taken into account: the type of circuit to be used, e.g. microstrip, coaxial, lumped element, or a combination thereof; the type of package to be used; the required size and type of heat sink; and the power output, gain, efficiency, and bandwidth required. All the above items are interrelated; for example, various heat sinks can be used with the coaxial-package devices, depending upon the circuit technique chosen.

Fig. 127 shows a coaxial package jig that uses a standard beryllium oxide ring to conduct heat from the center conductor to the outside conductor of an air-dielectric line section. This type of arrangement is useful for power dissipation of 5 watts or less. A more efficient heat sink is obtained by use of a boron nitride cylinder that makes intimate contact between the coaxial line conductors over the entire length of the cavity. This arrangement results in much improved heat conduction and, therefore, is more suitable for high-power microwave transistors. In addition, the boron nitride, which has electrical and thermal properties similar to those of beryllium oxide, is readily machineable and is nontoxic. Coaxial line lengths are also substantially reduced.

Fig. 128 shows a circuit mounting arrangement for coaxial-package transistors in microstripline and lumped-element circuits. The transistor is mounted vertically through a hole in the metal block which serves as both a heat sink and ground for the device. The bottom side of the metal block is counter-bored so that the base flange of the transistor is level with the surface of the block. The hole through the metal block has a somewhat larger diameter than that of the

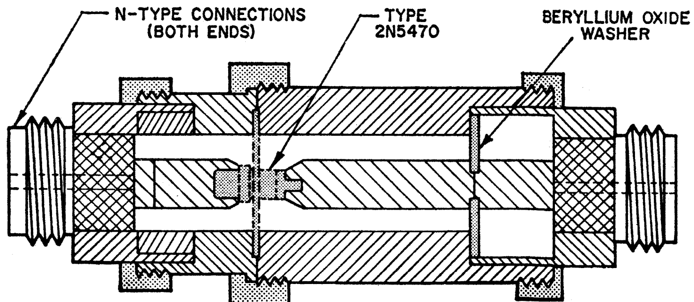


Figure 127. Heat sink for use with coaxial transistor package.

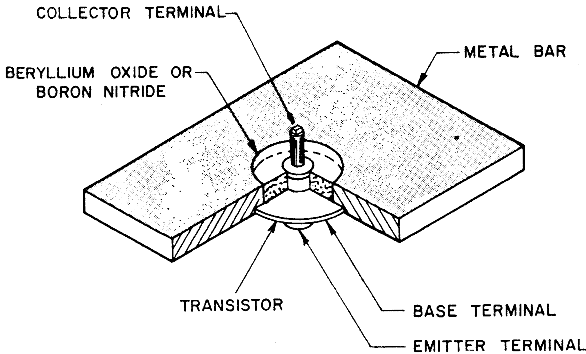


Figure 128. Mounting arrangement for a coaxial-package transistor in a microstripline circuit.

ceramic portion of the transistor which separates the base flange and the collector stud. This larger diameter permits insertion of a press-fit cylindrical sleeve of beryllium oxide or boron nitride between the transistor and the metal block to provide a heat-conducting path from the collector stud to the block. The diameters of the hole through the metal block and the cylinder of beryllium oxide (or boron nitride) are determined by the desired characteristic impedance of the short coaxial-line section which is formed by this mounting technique. Beryllium oxide and boron nitride have excellent heat conductivity and low electrical losses and thus provide satisfactory heat dissipation from the coaxial transistor without adversely affecting the rf performance.

The circuit arrangement shown in Fig. 128 is excellent for isolation of the input and output circuits. The output circuit is constructed on the top portion of the metal block and the input circuit on the bottom portion. Fig. 129 shows the construction of the microstripline circuit. The output circuit is constructed of standard microstripline mounted to the top surface of the metal block. The input circuit is con-

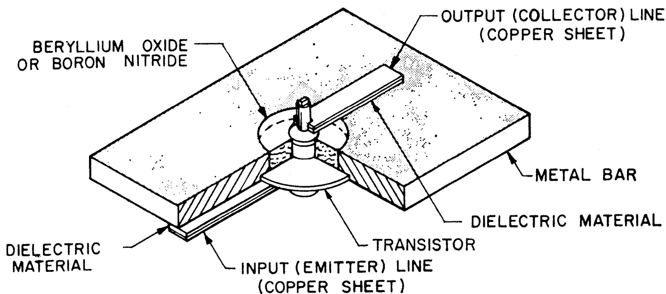


Figure 129. Construction of the microstripline circuit.

structed of another microstripline placed directly over the bottom surface of the metal block. A stripline circuit can be formed by placing another strip of dielectric material and ground plane above the conductor strips of Fig. 129.

The design of transistor microwave power circuits involves two steps: (1) the determination of load and input impedances under dynamic operating conditions, and (2) the design of properly distributed filtering and matching networks required for optimum circuit performance. For design of the input circuit, the input impedance at the emitter-to-base terminals of the packaged transistor at the drive-power frequency under operating conditions must be known. For design of the output circuit, the load impedance presented to the collector terminal at the fundamental frequency must be known. These dynamic impedances are difficult to calculate at microwave frequencies because transistor parameters vary considerably under large-signal operation from small-signal values, and also change with power level. Small-signal equations that might serve as useful guides for transistor design cannot be applied rigorously to large-signal circuits, although it has been determined empirically that some small-signal parameters at the 10-volt level correspond rather closely with the large-signal values at 28 volts. Because practical large-signal representation of microwave transistors has not yet been developed, transistor dynamic impedances are best determined experimentally by use of slotted-line or vector-voltmeter measurement techniques.

The system required to determine transistor impedances under operating conditions is shown in Fig. 130. This system consists of a well-padded power signal generator, a directional coupler (or reflectometer) for monitoring the input reflected power, an input triple-stub tuner, an input low-

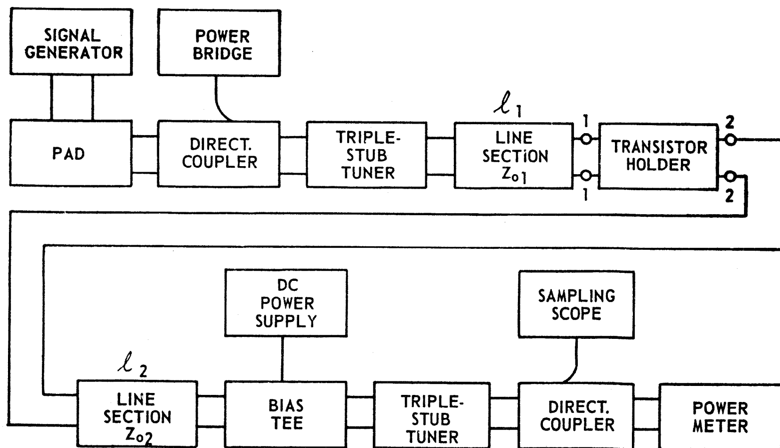


Figure 130. Block diagram of test setup used to determine input and output impedances of transistors.

impedance line section, the transistor holder (or test jig), an output line section, a bias tee, an output triple-stub tuner, another directional coupler for monitoring the output waveform or frequency, and an output power meter. At a given frequency and input-power level, the input and output tuners are adjusted for maximum power output and minimum input reflection power. When the system has been tuned properly, the impedance across terminals 1-1, without the transistor in the system, is measured at the same frequency in a slotted-line set-up or with a vector voltmeter. The conjugate of this impedance equals the dynamic input impedance of the transistor. Similarly, the impedance across terminals 2-2, without the transistor in the system, is the load impedance presented to the collector of the transistor. Such measurements are performed at each frequency and power level.

In addition to determining dynamic input impedance and load impedance, the system shown in Fig. 130 is useful for determination of the performance capability of the transistor. Power output, power gain, and efficiency are readily determined. For optimum performance of the test system, careful consideration must be given to the selection of the line length and the characteristic impedance Z_0 of the input and output line sections l_1 and l_2, respectively). Eighth-wavelength ($\lambda/8$) line sections are preferred for l_1 and l_2 because such sections exhibit the lowest VSWR and the smallest line losses.

An alternative method of determining the dynamic input impedance is shown in Fig. 131. This method uses a well-padded, high-power signal generator connected in series with a slotted-line setup.

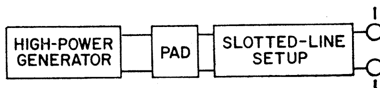


Figure 131. Block diagram of dynamic-impedance test setup that may also be used to test transistor performance capability.

The setup beyond terminals 1-1 is identical to that of Fig. 130. The high-power generator is adjusted until a desired power output is obtained. The input impedance under this condition can be measured simultaneously in a slotted-line setup. In this case, the test fixture must contain a short line section (a $\lambda/8$ section is preferred for smallest line losses) to provide a connection to the transistor.

When the dynamic input impedance and the load impedance of a packaged transistor have been established, either by direct measurement (as described in the preceding paragraph) or from published data, the input and output circuits can be designed.

Some simple designs are shown in Fig. 132. Although coaxial-line configurations are shown, the design procedures are similar for the other forms of TEM-mode distributed line sections. For the circuit shown in Fig. 132(a), the line section l transforms the small input impedance of the transistor to a value closer to that of the driving-source resistance (such as a 50-ohm generator). If line section l is made an eighth-wavelength long and its characteristic impedance Z_0 is properly de-

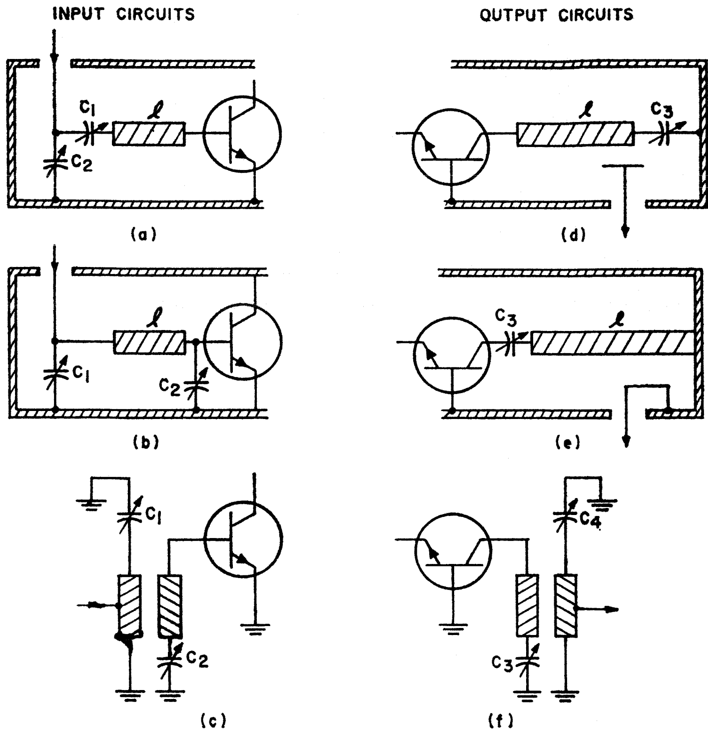


Figure 132. Transistor input and output coupling circuits suitable for use at microwave frequencies: (a) direct-coupled input network using series tuning capacitor; (b) direct-coupled input network using shunt tuning capacitor; (c) resonant-line input circuit; (d) capacitive-probe-coupled output cavity; (e) inductive-probe-coupled coaxial output cavity; (f) resonant-line output circuit.

termined, then the complex input impedance is transformed to a real value at the other end of this line, and the VSWR on the line section is a minimum. Capacitors C_1 and C_2, together with some lead inductance, are used as reactive dividers to step up or step down the impedance, depending on the value of the real impedance compared to the 50-ohm source. Transformation directly to 50 ohms or some other desired real impedance is also possible with this configuration. The length of the line section l is less than a quarter-wavelength when the dynamic input impedance is inductive and greater than a quarter-wavelength for capacitive inputs. In this type of application, capacitors C_1 and C_2 serve to tune out imaginary components, modify imaginary components, or adjust the values of real components, depending on the frequency and the characteristics of the line section l.

The input circuit shown in Fig. 132(b) can be used effectively when the dynamic input impedance of the transistor is inductive. Capacitor C_2 is used to tune

out the inductive component of the input impedance. A quarter-wave line of the proper characteristic impedance is then used for the impedance transformation between the small input resistance of the transistor and the driving-source resistance. Capacitor C_1 may be used to adjust for minor differences between transistors.

The output circuit shown in Fig. 132(d) is a capacitive-loaded, foreshortened quarter-wave coaxial-line cavity. A capacitive probe is used to match the output to the desired real load impedance. In the design of the circuit, the line section *l*, the capacitance C_3, and the dynamic output capacitance of the transistor must be resonant at the desired frequency.

The ouput circuit shown in Fig. 132(e) is similar to that shown in Fig. 132(d) except that inductive loop coupling is used. Again, the design of the coupling loop is empirical. In general, the inductive loop is placed near the ground (high-current) end of the line; in fact, it may be tapped directly to the center conductor. Conversely, capacitive probes are generally located near the high-voltage end of the line.

The coupling networks shown in Fig. 132(a) through Fig. 132(e) can apply to either input or output circuits, and the specific illustrations are used for discussion only. The circuits shown in Figs. 132(c) and 132(f) make use of inductive coupling and are particularly suitable for stripline circuits.

CIRCUITS

In general, transistor amplifiers are operated in the common-base configuration at microwave frequencies. As was mentioned in the previous section, either coaxial or microstrip circuit techniques can be utilized.

Coaxial Amplifier Circuits

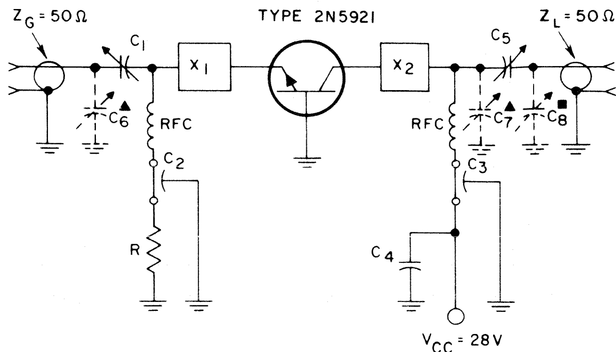
Fig. 133 shows a typical 2N5921 coaxial-circuit amplifier for 1.2- or 2-GHz operation and Fig. 134 shows the mechanical construction of the circuit. Figs. 135 and 136 show typical performance that can be obtained with this transistor.

The performance data show that the 2N5921 can provide an output of 5 watts at 2 GHz with a gain of 7 dB and a collector efficiency greater than 40 per cent.

In the design of an amplifier, such as shown in Fig. 134, the input and output impedances of the transistor must be known. Fig. 137 shows typical large-signal input impedance and collector load impedance values for the RCA-2N5921 as functions of frequency. These data are particularly useful for the design of broadband circuits.

Fig. 138 shows construction details of the lower-power amplifier that uses the 2N5920, and Figs. 139 and 140 show performance characteristics of this transistor when used in the common-base configuration. The 2N5920 is particularly well suited for use as a driver for the 2N5921.

For even lower powers, the RCA-2N5470 can be utilized. Circuit and performance details of an RCA-2N5470 common-base amplifier are shown in Figs. 141 through 143.



▲ Use only in the 2-GHz coaxial-line power amplifier circuit.
■ Use only in the 1.2-GHz coaxial-line test circuit.

CIRCUIT	C_1 pF	C_2 pF	C_3 pF	C_4 μF	C_5 pF	C_6 pF	C_7 pF	C_8 pF	R ohm
1.2 GHz (Test Circuit)	1-10	1000	1000	0.01	1-10	—	—	0.3-3.5	0.75
2 GHz (Test Circuit)	1-10	470	470	0.01	1-10	—	—	—	0.43
2 GHz (Amplifier)	1-10	470	470	0.01	0.3-3.5	0.3-3.5	0.3-3.5	—	0.43

C_1, C_5 = 1 to 10 pF, Johanson 4581, or equivalent*
C_5, C_6, C_7, C_8 = 0.3 to 3.5 pF, Johanson 4700, or equivalent*
RFC (For 2-GHz Circuits) = 3 turns No. 32 wire 1/16 in. (1.59 mm) ID, 3/16 in. (4.76 mm) long.
(For 1.2-GHz Circuit) = 6 turns No. 32 wire 1/16 in. (1.59 mm) ID, 3/16 in. (4.76 mm) long.
X_1, X_2: Coaxial-line circuits.

* Johanson Mfg. Corp., Boonton, N.J. 07005

Figure 133. 1.2- or 2-GHz coaxial-line amplifier circuit.

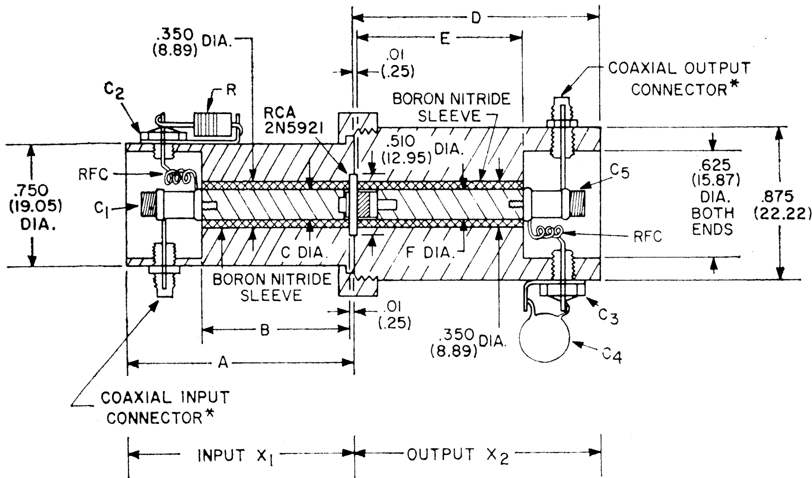
Typical Microstripline Amplifier Circuits

The circuits in Figs. 133, 134, 138, and 141 employ coaxial resonators and matching circuits. Microstripline circuitry can also be used with excellent results. Fig. 144 shows a typical 2-GHz microstripline circuit used with the RCA-2N5921. Construction details of this circuit are shown in Fig. 145.

As discussed previously, microwave transistors are also available in stripline packages. Figs. 146 and 147 show a circuit and suggested construction for a 1- or 2-GHz amplifier that uses an RCA stripline transistor. Typical performance of this transistor is shown in Fig. 148.

The RCA-2N6266 stripline transistor used in the circuit shown in Fig. 146 can provide 5 watts of power output at 2 GHz with a gain of 7 dB and a collector efficiency of 40 per cent.

Another RCA stripline transistor, the 2N6265, can provide 2 watts of output power at 2 GHz with a minimum gain of 8.2 dB and collector efficiency greater than 35 per cent.



DIMENSIONS OF COAXIAL LINES X_1 AND X_2

CIRCUIT	INPUT (X_1)				OUTPUT (X_2)			
	A	B	C	Center Conductor	D	E	F	Center Conductor
1.2-GHz (Test Circuit)	1.385 (35.18)	0.875 (22.22)	0.282 (7.16)	0.825 (20.95)	1.778 (45.16)	1.268 (32.21)	0.213 (5.41)	1.05 (26.67)
2-GHz (Test Circuit)	0.940 (23.88)	0.430 (10.92)	0.266 (6.76)	0.380 (9.65)	1.04 (26.42)	0.530 (13.46)	0.266 (6.76)	0.370 (9.39)
2-GHz (Amplifier)	0.860 (21.84)	0.350 (8.89)	0.265 (6.73)	0.300 (7.62)	1.06 (26.92)	0.550 (13.97)	0.270 (6.86)	0.385 (9.78)

Dimensions in inches and millimeters
Dimensions in parentheses are in millimeters and are derived from the basic inch dimensions as indicated.
MATERIAL: Center conductor—copper
Outer conductor for input and output—brass

* Conhex 50-045-0000 Sealectro Corp., or equiv.

Figure 134. Constructional details for 1.2- or 2-GHz coaxial-line circuit.

Large-Signal Narrow-Band Microstripline Amplifier

The design of a large-signal transistor power amplifier circuit is simplified if the approximate large signal device impedances over the frequency range of interest are known. These impedances are generally specified in the published data on RCA microwave power transistors as the device input impedance (Z_{in}) and the preferred device collctor load impedance (Z_{CL}). The specified input impedance for the RCA-2N6267 transistor at 2.3 GHz is approximately 1.8 + j12 ohms, and the collector load impedance is given as 1.6 − j8.0 ohms.

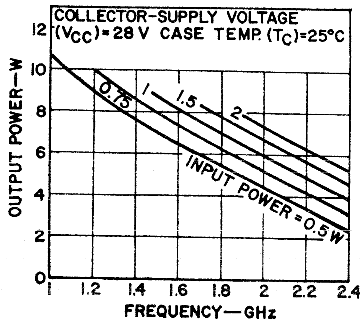


Figure 135. Typical output power as a function of frequency for the 2N5921 transistor.

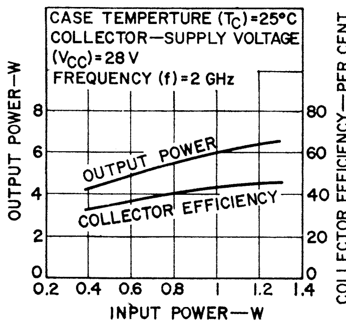


Figure 136. Typical power output or collector efficiency as a function of power input at 2 GHz for circuit show in Fig. 134.

Fig. 149 shows the circuit diagram for a typical 2.3-GHz narrow-band power amplifier that uses the 2N6267 transistor.

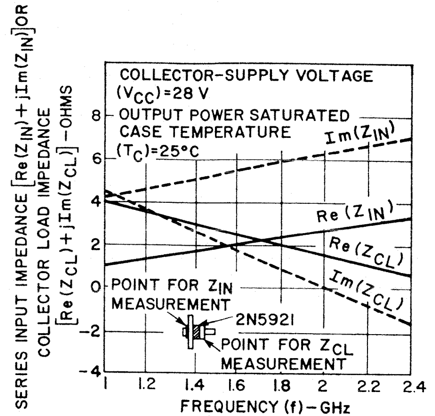


Figure 137. Typical large-signal series input impedance and large-signal collector load impedance as a function of frequency for the RCA-2N5921.

Simple series matching transformers, which are short sections of microstripline, are used at both the input and the output of the device to assure a 1-dB bandwidth of at least 100 MHz. The input impedance is transformed to about 50 + j85 ohms by line section X_1. This line-section transformer is a 0.90-wavelength section of 12.5-ohm microstripline constructed on 1/32-inch-thick Teflon-fiberglass board. The capacitor C_1 is used both to tune out the inductive component of 85 ohms and to provide dc isolation for the bias network.

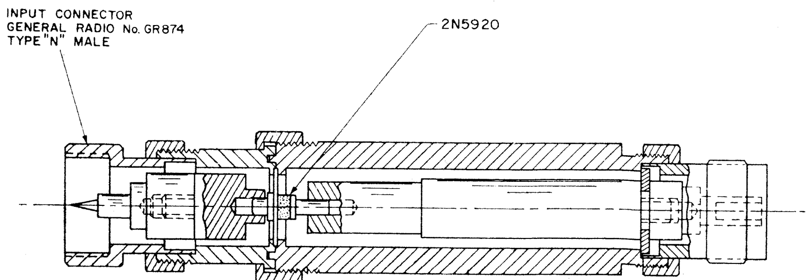


Figure 138. Constructional details of 2-GHz power amplifier.

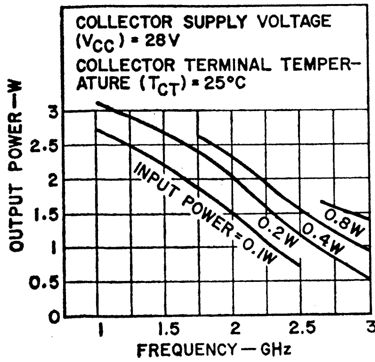


Figure 139. Typical output power as a function of frequency for common-base amplier.

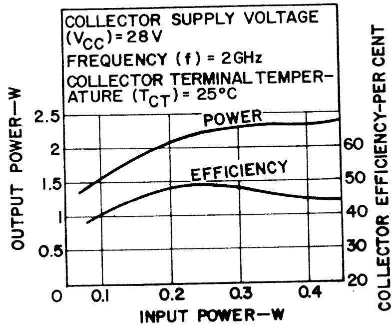
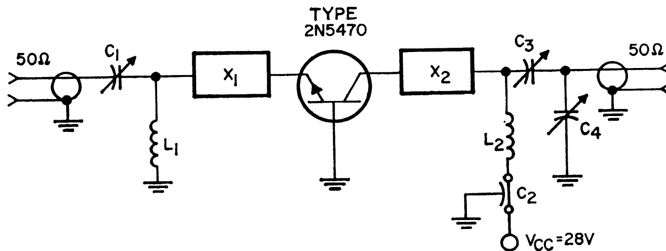
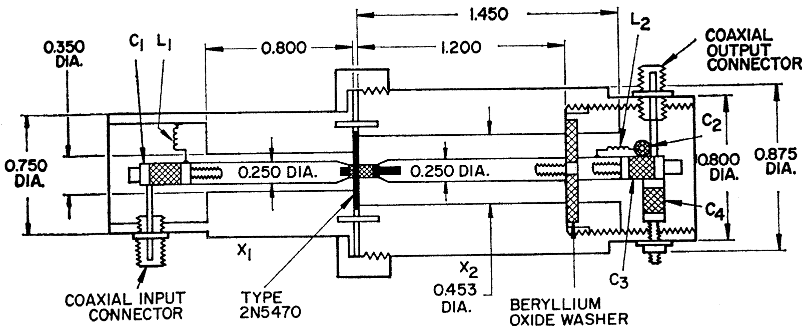


Figure 140. Typical output power and collector efficiency as a function of input power for 2-GHz common-base power amplier.



C_1 = 0.8 to 10 pF, Johnason 4355, or equiv.
C_2 = 1000 pF, feedthru, Allen-Bradley FB2B, or equiv.
C_3 = 0.3 to 3.5 pF, Johnason 4701, or equiv.
C_4 = 0.35 to 3.5 pF, Johnason 4702, or equiv.
L_1, L_2 = RF choke, 3 turns No. 30 wire, 1/16" ID, 3/16" long
X_1, X_2 = Details given in (b)

(a)



DIMENSIONS IN INCHES

(b)

Figure 141. 2-GHz power amplifier using the RCA-2N5470 coaxial transistor: (a) circuit schematic; (b) construction details.

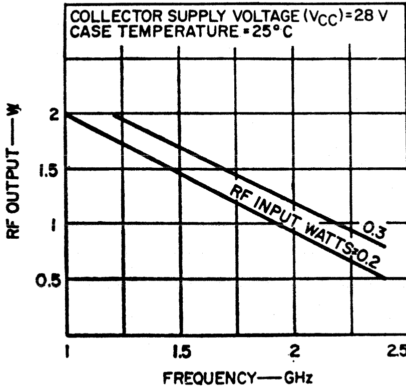


Figure 142. Power output as a function of frequency for the RCA-2N5470 transistor.

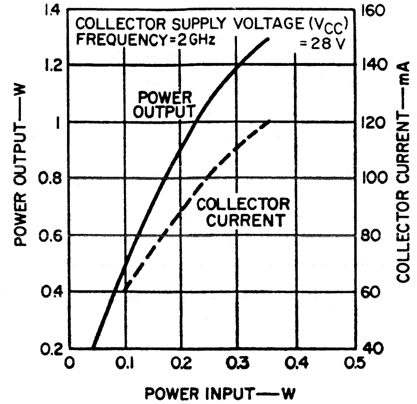
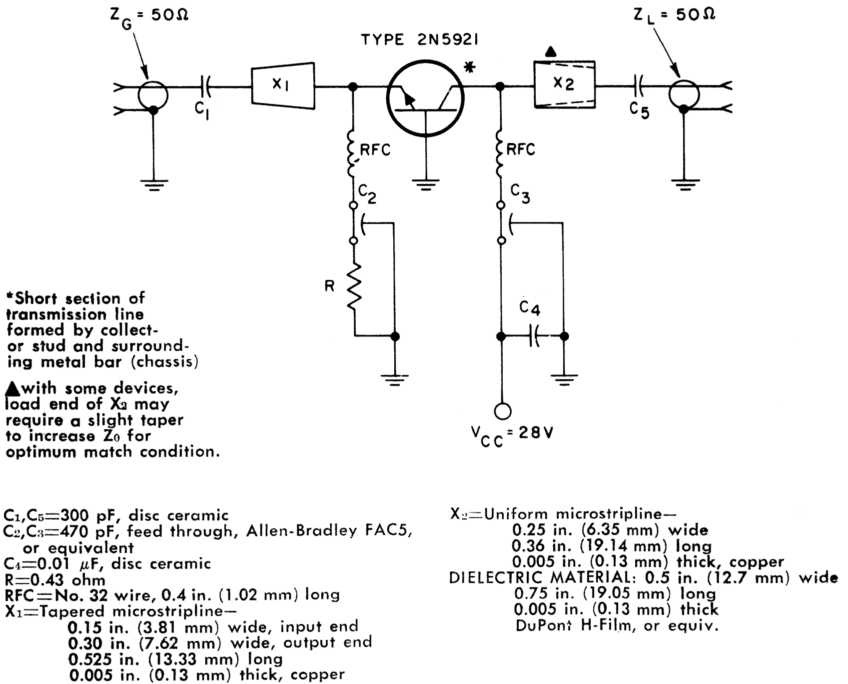


Figure 143. Power output as a function of input for amplier shown in Fig. 141.



C_1,C_5=300 pF, disc ceramic
C_2,C_3=470 pF, feed through, Allen-Bradley FAC5, or equivalent
C_4=0.01 μF, disc ceramic
R=0.43 ohm
RFC=No. 32 wire, 0.4 in. (1.02 mm) long
X_1=Tapered microstripline—
0.15 in. (3.81 mm) wide, input end
0.30 in. (7.62 mm) wide, output end
0.525 in. (13.33 mm) long
0.005 in. (0.13 mm) thick, copper

X_2=Uniform microstripline—
0.25 in. (6.35 mm) wide
0.36 in. (19.14 mm) long
0.005 in. (0.13 mm) thick, copper
DIELECTRIC MATERIAL: 0.5 in. (12.7 mm) wide
0.75 in. (19.05 mm) long
0.005 in. (0.13 mm) thick
DuPont H-Film, or equiv.

Figure 144. Typical circuit for 2-GHz grounded-base microstripline power amplier.

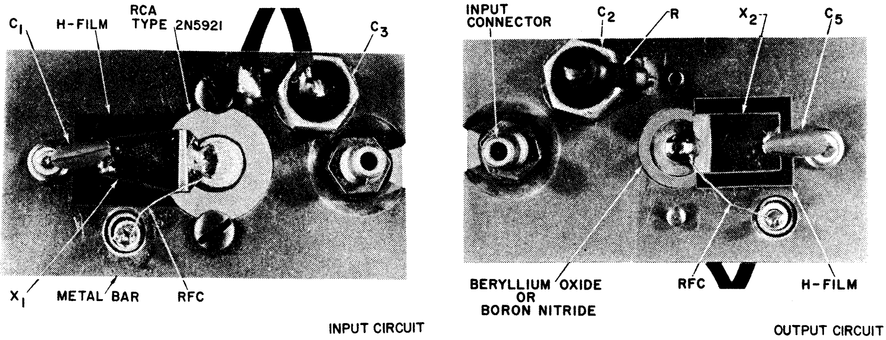


Figure 145. Suggested mounting arrangement of components for 2-GHz microstripline circuit shown in Fig. 144.

STRIPLINE TRANSISTOR TYPE 2N6266

Z=50Ω Z=50Ω X1 X2 C1 C5 RFC RFC C2 C3 C4 *BIAS +VCC

(a)

C_1, C_5=300 pF, ATC-100, or equivalent
C_2, C_3=Filtercon, Allen-Bradley SMFB-A1, or equivalent
C_4=0.001 μF, disc ceramic
RFC=2 GHz - No. 32 wire, 0.4 in. (1.02 mm) long
1 GHz - No. 32 wire, 3 turns, 1/16 in. (1.59 mm) ID, 3/16 in. (4.76 mm) long
X_1, X_2=Microstripline circuits

1 GHz:

INPUT (X_1): 0.085 (2.15); 0.36 (9.14); 0.7 (17.78)

OUTPUT (X_2): 0.225 (5.71); 0.085 (2.15); 1.1 (27.94)

2 GHz:

INPUT (X_1): 0.085 (2.15); 0.35 (8.89); 0.4 (10.16)

OUTPUT (X_2): 0.25 (6.35); 0.085 (2.15); 0.5 (12.70)

Note: All sections are exponentially tapered.●
Dielectric=5-mil DuPont H-Film, or equivalent
Line Material=5-mil copper

* Bias terminals normally have R_E = 0.24 ohm

● Use *sparing* amount of Eastman 910 Adhesive to bond lines and dielectric to circuit board.

(b)

Figure 146. (a) Circuit conguration for 1- or 2-GHz stripline amplier and (b) details of microstripline sections for each frequency.

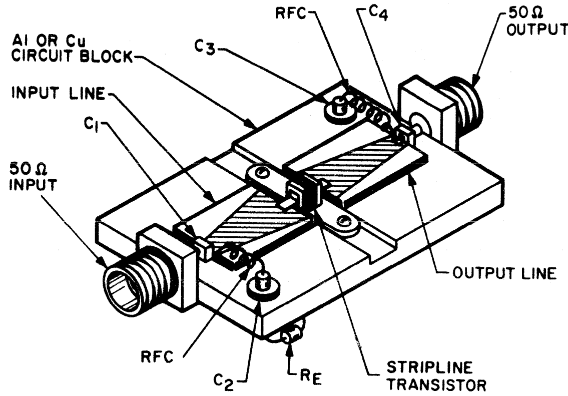


Figure 147. Suggested construction for stripline amplifier shown in Figure 146.

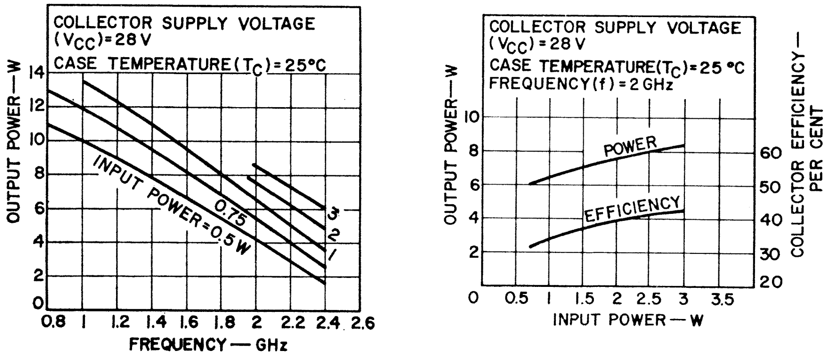


Figure 148. Typical performance characteristics for power amplier shown in Fig. 146: (a) power output as a function of frequency; (b) power output and collector efficiency as a function of input power at 2 GHz.

The collector load impedance is transformed to about 50 + j100 ohms. A shunt susceptance at the collector end of the package is provided by connection of two capacitive stub sections in parallel. This shunt susceptance (which is in the order of 0.04 mho) is used in conjunction with the parasitic inductance of the stripline package as a first-step transformation which effectively increases the real impedance at the input to the line section X_2. Line section X_2, which is a 0.14-wavelength section of 25-ohm microstripline (also constructed on 1/32 inch-thick Teflon fiberglass board) transforms the modified collector load impedance of 5 − j30 ohms to about 50 + j100 ohms. The capacitor C_4 tunes out the inductive component of 100 ohms and also provides the dc isolation for the collector bias supply.

Fig. 150 shows general performance data for this amplifier with the circuit optimized for operation at 2.3 GHz. Power out-

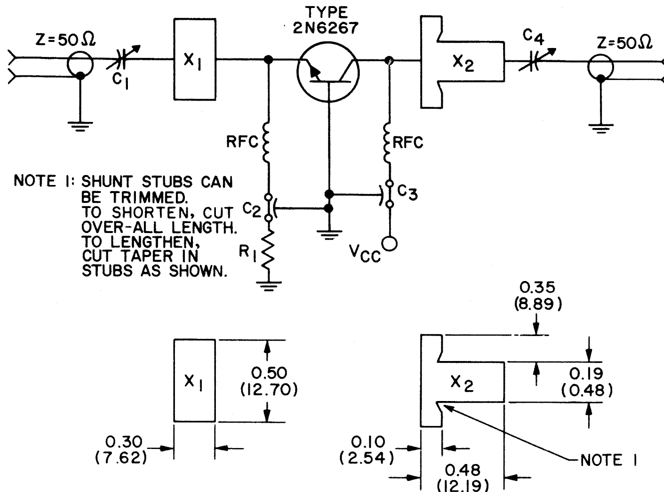


Figure 149. Typical 2.3-GHz power amplifier using the RCA-2N6267 stripline transistor.

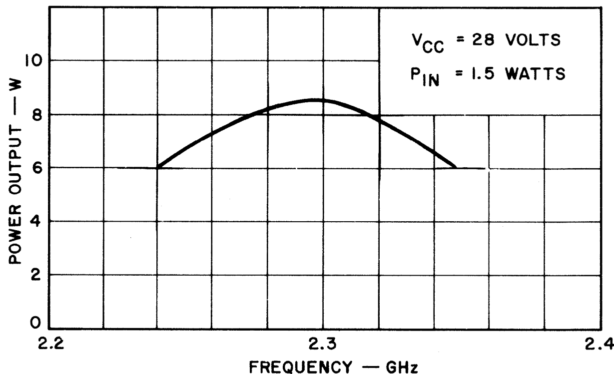
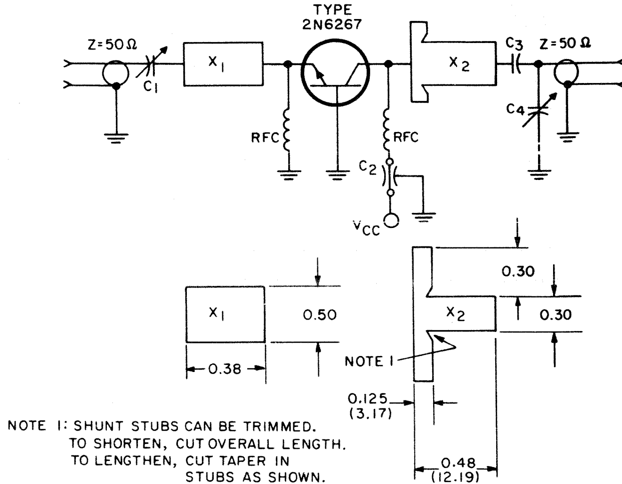


Figure 150. Typical power output as a function of frequency for the 2.3-GHz power amplifier shown in Fig. 149.

put is in the order of 8.5 watts with a power gain of approximately 8.3 dB and a collector efficiency of about 36 per cent. The 1-dB bandwidth is in the order of 100. MHz.

Fig. 151 shows a similar power amplifier that uses an RCA-2N6267 to develop a power output of 10 watts at 2 GHz. This circuit is tunable by adjustment of variable capacitors C_1 and C_4. Circuits of this type may also be designed using the 2N6265 and 2N6266 transistors.



C_1, C_3, C_4 = 0.3 to 3.5 pF, Johanson 1700, or equivalent
C_2 = Filtercon, Allen Bradley SMFB A1, or equivalent
RFC = No. 32 wire, 0.4 in. (1.02 mm) long

Dielectric material = 1/32 in. (0.79 mm) thick Teflon-fiberglass double-clad circuit board. (ε = 2.6). Lines X_1 and X_2 are produced by removing upper copper layer to dimensions shown.

Figure 151. Typical 2-GHz power amplifier using the RCA-2N6267 stripline transistor.

960-MHz Land-Mobile Amplifier

The frequency band from 806 to 960 MHz is being considered for **land-mobile communications** to provide relief to the much congested frequency spectrum currently allocated for two-way land-mobile as well as mobile telephone communications. The 2N6266 and 2N6267 microwave power transistors are highly suitable for this type of application. Because these devices are designed for 28-volt operation, they are extremely rugged when operated from a 12.5volt supply. The 2N6267 can deliver 10 watts of cw power at 960 MHz, as shown in Fig. 161. The 2N6266 can be used as a driver. Typical rf performance of the 2N6266 is shown in Fig. 162.

Fig. 154 show the basic circuit arrangement for an amplifier-multiplier chain that uses a 2N6266 driver and a 2N6267 output stage to develop 10 watts of output power at 960 MHz when operated from a collector supply of 12.5 volts.

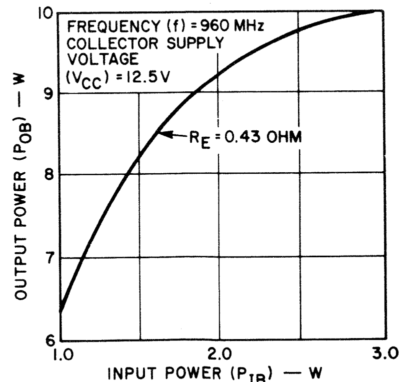


Figure 152. Typical performance data for the RCA-2N6267 stripline power transistor at 960 MHz.

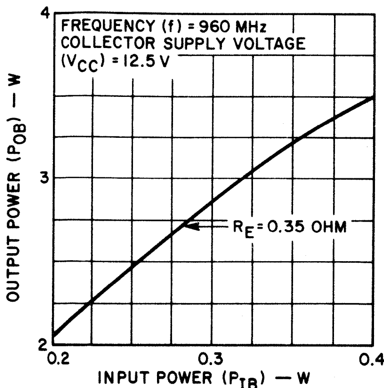


Figure 153. Typical performance data for the RCA-2N6266 stripline transistor at 960 MHz.

Broadband Microstripline Amplifiers

The sequence of steps in the design of any broadband amplifier are (1) determination of the input impedance and optimum collector load impedance across the frequency band for the power output (P_{out}), supply voltage (V_{CC}), power gain (PG), and efficiency (η_C) desired; (2) a review of the required transformation; and (3) selection of the basic transformation designs to be used.

The application of this procedure to the design of three broadband amplifiers is detailed in the following paragraphs.

600-to-950-MHz Power Amplifier—Fig. 155 shows the circuit diagram for a power amplifier designed to operate over the frequency range from 600 to 950 MHz. This amplifier uses an RCA-2N6266 that operates from a supply voltage V_{CC} of 28 volts to generate a cw output of 10 watts. Performance data for this amplifier are shown in Fig. 156.

Initially, the input impedance and the collector load impedance are determined under conditions of optimum efficiency (η_c) and gain (PG) at a power output P_{out} of 10 watts. The Smith chart in Fig. 157 shows the measured impedance characteristics as a function of frequency. Input and output matching circuits are designed on the basis on these impedances. The input circuit consists of a four-stage sixteenth-wavelength ($\lambda/16$) short-step Chebyshev transformer designed for a broadband transformation from 50 ohms to 1.5 ohms. The inductance of the transistor input impedance forms a portion of the last $\lambda/16$ Chebyshev step.

Depending on the desired performance of the amplifier, the last $\lambda/16$ length of the transformer can be varied in length to provide an offset input VSWR that effectively flattens the gain response of the amplifier at either the high or the low end of the frequency band. The input circuit also uses a series resonant trap to flatten a gain peak at the

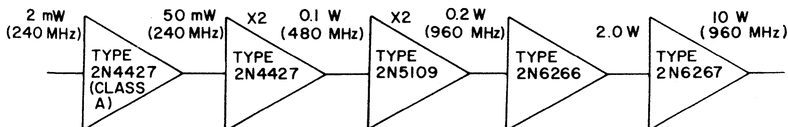
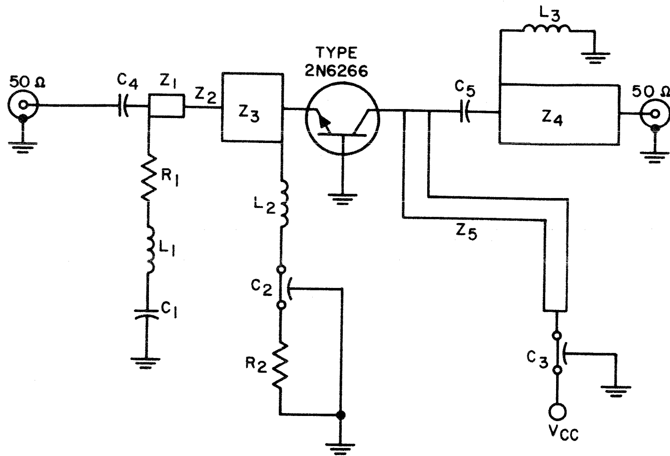


Figure 154. Typical circuit arrangement for a 10-watt, 960-MHz amplifier chain designed to operate from a supply voltage of 12 volts.



$C_1 = 1$ pF, ATC or equivallent
C_2, $C_3 = 1000$ pF, feedthrough
C_4, $C_5 = 1000$ pf, HT ATC or equivalent
$L_1 = 1$ to 2 nH
L_2 = RF choke
L_3 = Lumped element, 0.23 inch long, 0.10 inch wide

$R_1 = 0.472$ ohm, carbon
$R_2 = 0.2$ ohm
$Z_1 = 50$ ohms
$Z_2 = 14.6$ ohms, $\lambda/16$ at 800 MHz
$Z_3 = 57$ ohms, $\lambda/16$ at 800 MHz
$Z_4 = 2.3$ ohms,
$Z_5 = 14.5$ ohms, 0.174λ at 800 MHz
$Z_6 = \lambda/4$ at 950 MHz

Figure 155. 600-to-950-MHz broadband power amplifier using the RCA-2N6266 stripline transistor.

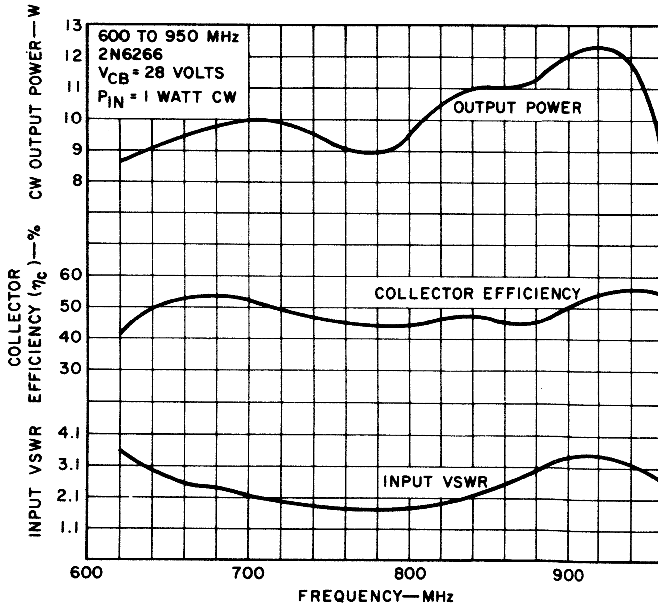


Figure 156. Performance data for the 600-to-950-MHz power amplifier.

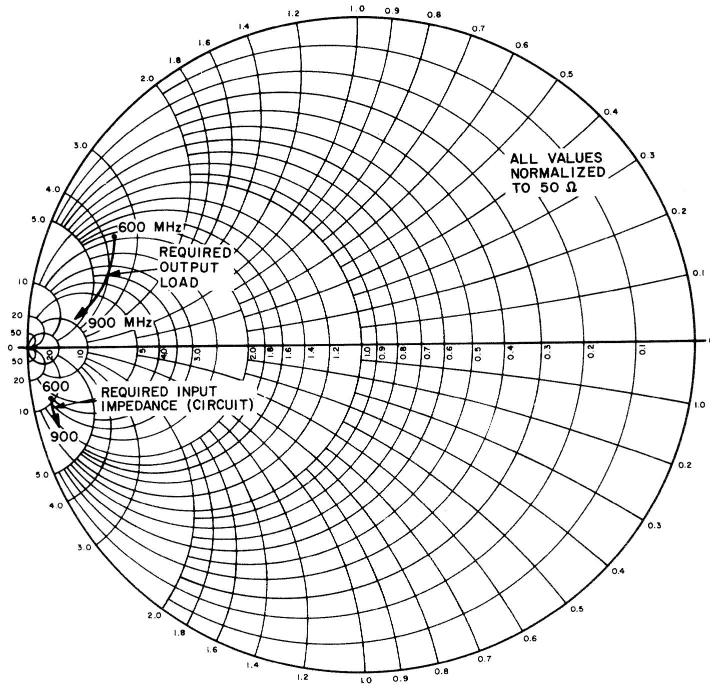
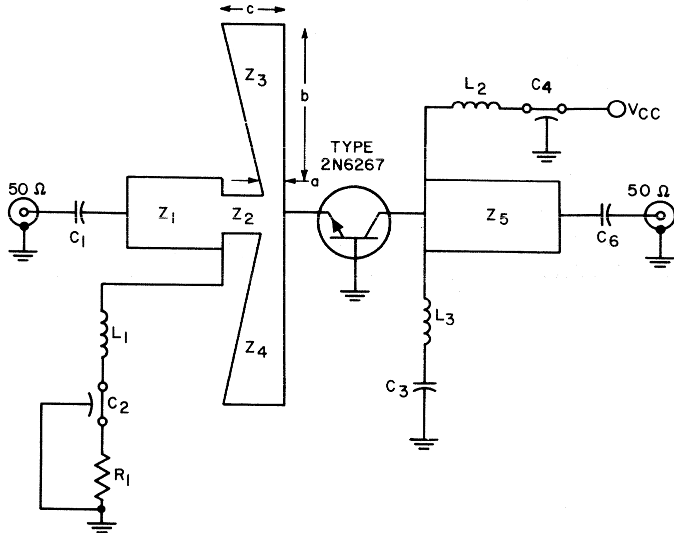


Figure 157. Impedance-admittance chart for 600-to-950-MHz power amplifier.

high-frequency end of the band.

The required collector load impedance is inductive throughout the frequency band. The output circuit consists of a quarter-wavelength ($\lambda/4$) transformer and a shunt inductor at the collector. The values chosen for the shunt inductor and characteristic impedance of the $\lambda/4$ line yield a load impedance across the band which is very close to that required by the transistor.

1.0-to-1.4-GHz Amplifier—Fig. 158 shows the circuit diagram of a power amplifier designed for operation from 1.0 to 1.4 GHz. This amplifier uses an RCA-2N6267 transistor to develop a cw output of 15 watts. The transistor operates from a collector supply of 28 volts. Typical performance data for this amplifier are shown in Fig. 159. The design of the output circuit is obtained by the same techniques as those used for the 600-to-950-MHz circuit. Fig. 160 shows the required transistor load and input impedances. Selection of the proper line characteristic impedance Z_0 and shunt inductance L produces the same type of output matching as that obtained in the 600-to-950-MHz circuit. For the 1.0-to-1.4-GHz amplifier, the input transformation is provided by a short transformation from 50 ohms and a shunt capacitance stub at the transistor.



C_1, C_3, C_4, C_6 = 1000 pF, ATC or equivalent
C_2, C_5 = 1000 pF, feedthrough
L_1, L_2 = RF choke, 5 turns
L_3 = 5-mil lead length
L_4 = 250-mil lead length
Z_1 = rectangular stripline, 240 mils long, 505 mils wide
Z_2 = rectangular stripline, 215 mils long, 235 mils wide
Z_3 = tapered stub, a = 75 mils, b = 400 mils
Z_4 = tapered stub, a = 75 mils, b = 375 mils
Z_5 = rectangular stripline, 1120 mils long, 590 mils wide

Figure 158. 1.0-to-1.4-GHz power amplifier using the RCA-2N6267 stripline transistor.

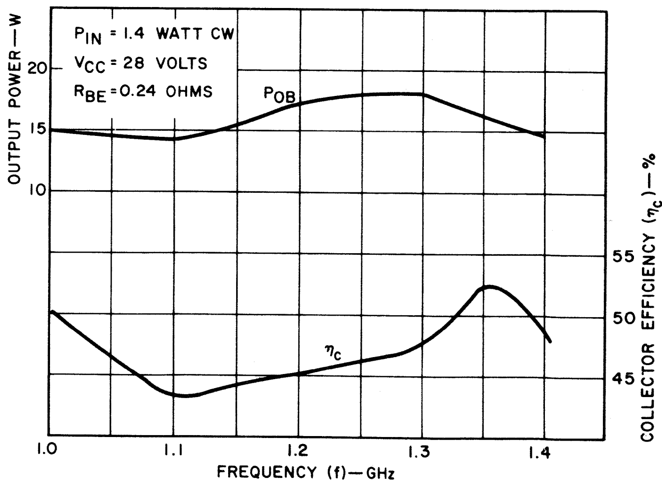


Figure 159. Performance data for the 1.0-to-1.4-GHz power amplifier.

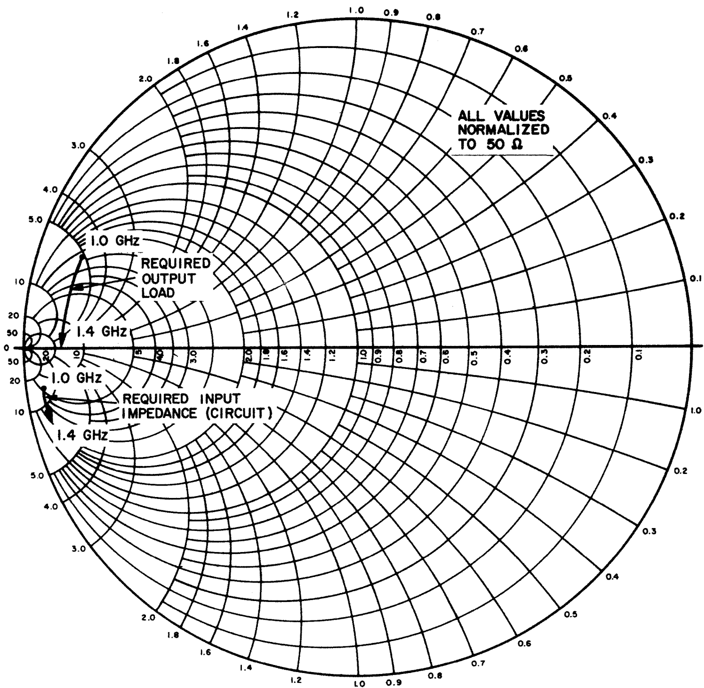
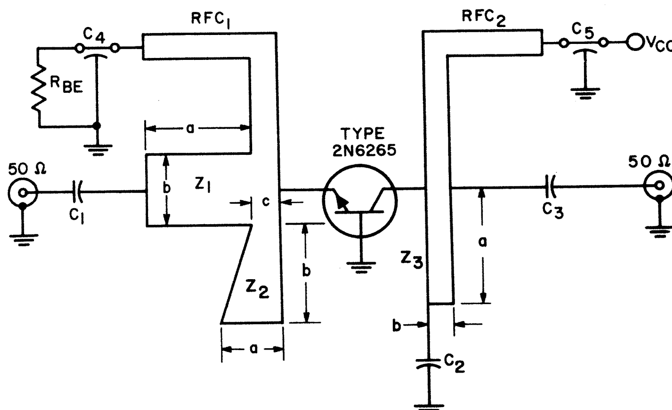


Figure 160. Impedance-admittance chart for the 1.0-to-1.4 GHz power amplifier.



C_1, C_2, C_3 = 1000 pF, ATC or equivallent
C_4, C_5 = 1000 pF, feedthrough, Allen-Bradley or equivalent
R_{BE} = 0.24 ohm

RFC_1, RFC_2 = lumped element, $\lambda/4$ at 1250 MHz
Z_1 = stripline, a = 1060 mils, b = 227 mils
Z_2 = stripline, a = 775 mils, b = 450 mils, c = 80 mils
Z_3 = stripline, a = 500 mils, b = 100 mils

Figure 161. 960-to-1250-MHz power amplifier using the RCA-6265 stripline transistor.

960-to-1250-MHz Amplifier — The amplifier circuit shown in Fig. 161 is designed to operate over the frequency range from 960 to 1250 MHz in TACAN Applications. It uses an RCA-2N6265 transistor to generate a cw power output of 2 watts across the band with a collector voltage of 28 volts. Typical performance data for this circuit are shown in Fig. 162. Because the real part of

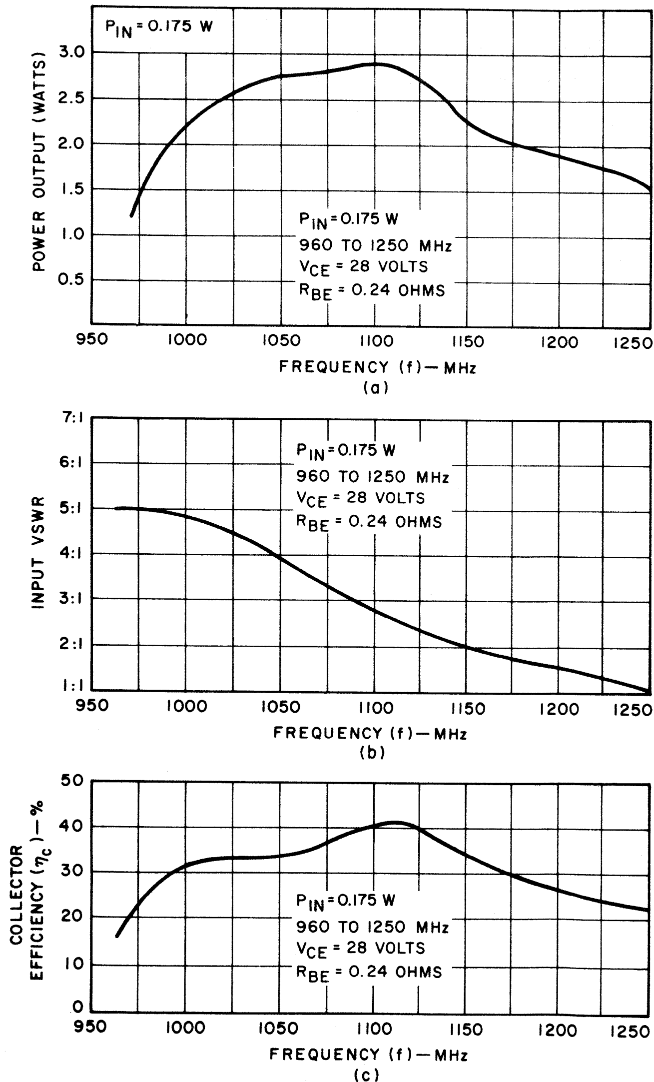


Figure 162. Performance data for the 960-to-1250-MHz power amplifier.

the required collector load is approximately 50 ohms, the only requirement is a shunt inductance L at the collector. The input circuit, which is similar to that of the 1.0-to-1.4-GHz circuit, consists of a short transformation and shunt capacitance C at the transistor.

Amplifier Chains

Most amplifiers include several stages of amplification. Fig. 163 shows a typical 2-GHz power amplifier chain that provides 8 watts of output power with a gain of 26 dB. The power from this chain is limited by the power capability of the output transistor.

Two or more transistors can be combined as shown in Fig. 164 to obtain higher power output. This circuit provides 15 watts at 1.8 GHz by use of quadrature hybrid couplers to parallel two output transistors. These couplers also serve to isolate the driver stage from the output stage. Other combiners, such as the Wilkinson and the simple reactive splitter, can also be used.

If interstage isolation and transistor-to-transistor isolation are not required, the transistors can be paralleled directly, provided that strict attention is given to symmetry of circuit layout.

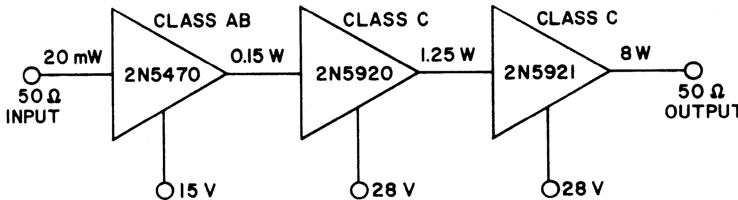


Figure 163. Typical 2-GHz power-amplifier chain.

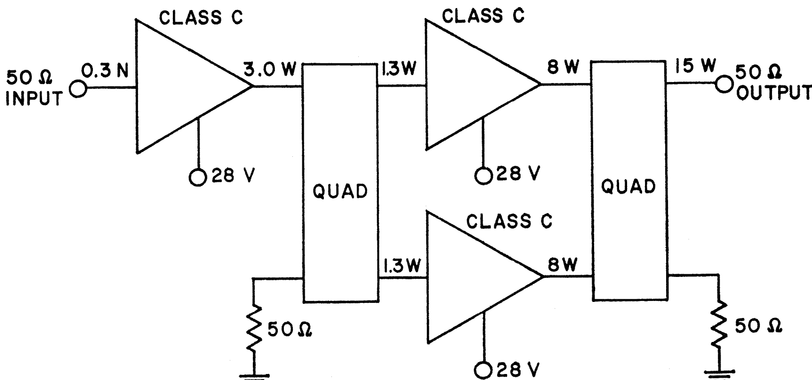


Figure 164. Typical 1.8-GHz power-amplifier chain using quadrature combiners.

Microwave Power Oscillators

TRANSISTORS capable of power amplification are also suitable for power oscillation. The most important part of every oscillator is an element of amplification. It is then necessary only to provide a path that feeds back a part of the power output to the input in the proper phase, together with a source of dc power. The maximum frequency of oscillation, which is related to f_{max} in a small-signal transistor, is usually difficult to define in a uhf or microwave power transistor because of the added parastic elements. The circuit-design approach for an oscillator circuit is similar to that discussed previously for amplier circuits.

BASIC CIRCUIT CONFIGURATION

Fig. 165(a) shows a Colpitts transistor oscillator suitable for microwave applications. The inductance L and the capacitances C_1 and C_2 can be considered as the parasitic elements of the package. The transistor can be grounded at the collector, the base, or the emitter without effect on its performance. For example, a useful oscillator circuit can be derived from the basic Colpitts oscillator by the use of a TO-39 transistor. In Fig. 165(b), the collector of such a transistor is returned to ground through the collector parasitic inductance L. This connection is a convenient method of applying a heat sink to

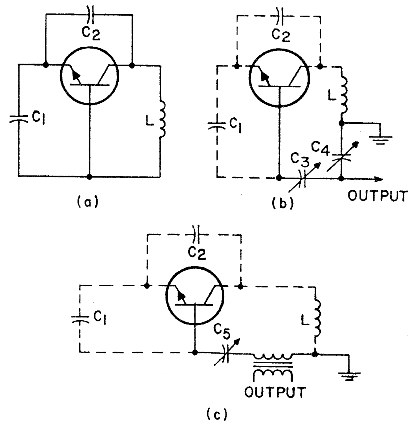
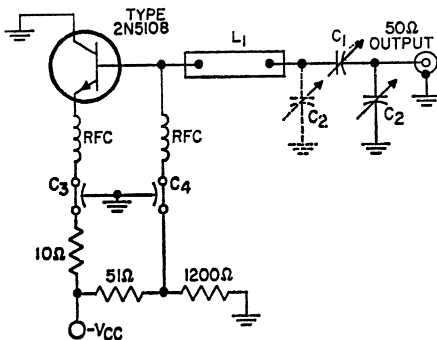


Figure 165. Colpitts oscillator for use at microwave frequencies: (a) basic ac circuit configuration; (b) basic ac circuit with collector returned to ground through parasitic inductance L and the output taken from base through capacitive voltage divider; (c) basic ac circuit with transformer-coupled output.

the collector, which is connected to the case in a TO-39 package. Output power is obtained from the base through capacitances C_3 and C_4. Fig. 165(c) shows another method of coupling power output from the oscillator.

L-BAND OSCILLATORS

Fig. 166 shows the complete circuit diagram of a 1.68-GHz fundamental-frequency oscillator which makes use of the RCA-2N5108 transistor. The collector



C_1, C_2 = Variable capacitor, 0.35 to 3.5 pF, piston type
C_3, C_4 = 470 pF, feedthru
L_1 = Described in text
RFC = 5 turns No. 28 wire, 1/8" ID x 1/2" long

Figure 166. 1.68-GHz fundamental-frequency oscillator using the RCA-2N5108 transistor. shown in Fig. 166.

of this transistor, which is packaged in a TO-39 case, is grounded to the ground plane of a 1/16-inch Teflon fiberglass microstripline board. Power output is taken from the base through a 0.75-inch section of 50-ohm microstripline and the capacitor network C_1 and C_2. This oscillator can supply more than 0.3 watt of power output at 1.68 GHz and has an efficiency of 20 per cent when operated from a supply voltage of 25 volts. Fig. 167 shows the oscillator output power as a function of supply voltage.

This basic oscillator circuit is useful at frequencies from 1 to 2 GHz; only slight modifications in the length of the transmission-line L_1 are required to cover this range. For example, the line length is increased to 0.8 inch

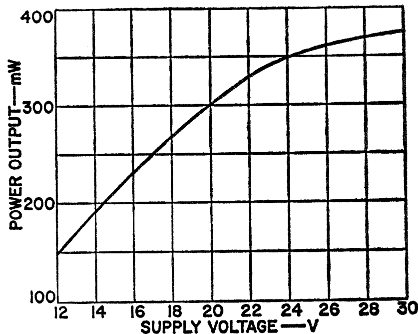
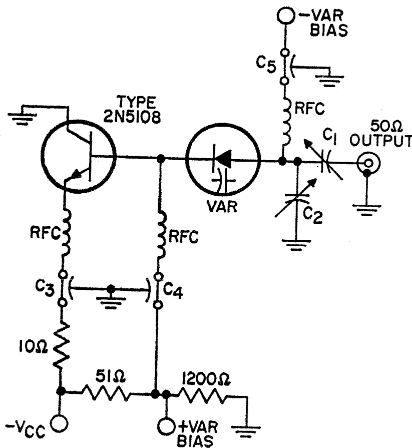


Figure 167. Power output as a function of supply voltage for the 1.68-GHz oscillator shown in Fig. 166.

to obtain optimum circuit operation at 1.5 GHz. An output power of 400 milliwatts (with a 24-volt supply) can be expected at this frequency. Another modification of interest (with the 0.8-inch line) is that optimum operation at 1.25 GHz is achieved simply by movement of capacitor C_2 to the position indicated by the dotted lines. Movement of this capacitor results in an improved output transformation network which can develop more than 800 milliwatts of output power at 1.25 GHz for operation from a 24-volt supply.

The inductive element introduced by the line section L_1 (Fig.166) can be supplied by a high-Q varactor diode operated above its resonant frequency. as shown in Fig. 168. The bias supplied to this varactor, in effect,

electrically varies this inductive component so that broadband oscillator tuning is possible. The output capacitor network, C_1 and C_2, which is used to transform a relatively small load-line impedance to the 50 ohms of the output port, could be replaced with an inductive-reactive divider-network, such as a tapped transmission line or helical coaxial line. Tests in which a cartridge-type silicon microwave varactor is used in this circuit show a relatively constant power output of 600 milliwatts over the range of 1.0 to 1.5 GHz. The bias on this particular varactor ranges between 0 and 22 volts for the specified tuning range, and a transistor collector supply of 28 volts is used.



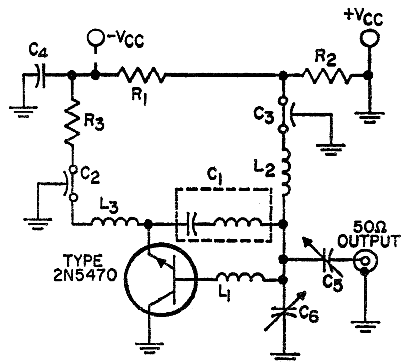
RFC = 0.1 μH
C_1, C_2 = 1 to 7 pF, piston capacitors
C_3, C_4, C_5 = 470 pF, feedthru
Var. = Described in text

Figure 168. Wide-band varactor-tuned L-band oscillator using the RCA-2N5108 transistor.

S-BAND OSCILLATORS

The RCA-2N5470 coaxial transistor, although designed for stable operation at 2.3 GHz in the common-base amplifier mode, can also deliver a power output of 0.3 watt at 2.3 GHz as an oscillator. In oscillator applications of the 2N5470, advantage is taken of the very low parasitic elements in this transistor to simplify circuit requirements, e.g., essentially lumped-constant S-band circuits can be designed around this unit. However, because of the low feedback capacitances of this transistor, external feedback loops are generally needed for stable oscillation at S-band frequencies.

Fig. 169 shows a simple lumped-constant circuit that uses the 2N5470 transistor. The circuit is tunable over the frequency range of 1.8 to 2.3 GHz. Power output at 2 GHz is typically 0.3 watt with a 24-volt supply, and circuit efficiency is in

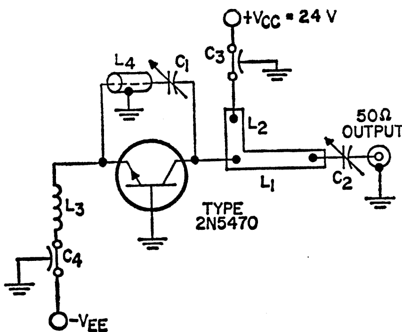


C_1 = 0.82 pF, "gimmick" capacitor (manufactured by Quality Components, Inc. St. Mary's, Pa.)
C_2, C_3 = 100 pF, Allen-Bradley FA5C or equiv.
C_4 = 0.01 pF, disc ceramic
C_5, C_6 = Trimmer capacitor 0.35 to 3.5 pF, Johanson Type 4702 or equiv.
L_1 = 0.05" length of No. 22 wire
L_2, L_3 = 4 turns 7-mil wire, .062" ID x 3/16" long
R_1 = 51 ohms, 1/2 watt
R_2 = 1200 ohms, 1/2 watt
R_3 = 5 to 10 ohms, 1/2 watt

Figure 169. Lumped-constant 2-GHz oscillator circuit using the RCA-2N5470 transistor.

the order of 16 per cent at this frequency. The collector is grounded, and power output is taken from the base circuit. All leads must be kept short for best high-frequency response. The "gimmick" capacitor C_1 forms a necessary part of the feedback loop of the circuit. The circuit is basically a Hartley type of oscillator in that inductor L_1 and the parasitic inductance of C_1 make up a tapped inductor in this feedback loop. Tuning is achieved largely by adjustment of capacitor C_6, and capacitor C_5 is adjusted to maintain the output match over the tuning range.

Fig. 170 shows the use of the 2N5470 transistor in a Colpitts type of microstripline oscillator circuit that operates over the frequency range of 1.8 to 2.2 GHz. In this circuit, the base of the transistor is directly grounded to the ground plane of the stripline board, and collector heat is dissipated to this board



C_1, C_2 = 0.35 to 3.5 pF, Johanson Type 4702 or equiv.
C_3, C_4 = 100 pF, Allen-Bradley Type 5A5C or equiv.
L_1 = microstrip line; ≅ 0.70" long x 0.30" wide strip; mounted on 1/32" Teflon fiberglass board
L_2 = microstrip line; ≅ 0.43" long x 0.080" wide strip; mounted on 1/32" teflon fiberglass board
L_3 = 5 turns 7-mil wire, 0.062" ID x 3/16" long
L_4 = 50-ohm miniature coaxial line, 1.5" long

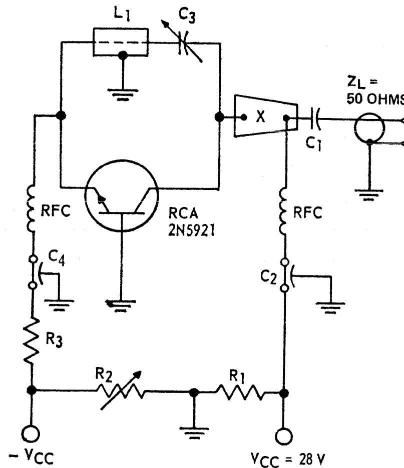
Figure 170. Microstripline 2-GHz oscillator circuit using the RCA-2N5470 transistor.

through a beryllium oxide insulating washer. The necessary feedback is provided by the phase-resonant loop provided by line section L_4 and capacitor C_1. The output line section L_2 makes use of standard microstripline techniques to provide the necessary reactance to tune out the output capacitance; line section L_1 is a quarter-wave transformer which transforms the real part of the collector load impedance to about 50 ohms. This circuit can also provide about 0.3 watt of output power at 2 GHz when operated from a 24-volt supply.

HIGHER-POWER OSCILLATORS

For higher power outputs, the RCA-2N5921 or the RCA-2N5920 coaxial transistors and RCA 2N6267 stripline transistors can be used. Fig. 171 shows the 2N5921 in a power oscillator circuit that tunes from 1.2 to 1.4 GHz. In this circuit, the coaxial transistor is mounted with the collector protruding through a 0.35-inch hole in a metallic circuit board that is 0.20 inch thick. Output line X is constructed on the collector side of this circuit board, while the bias network is on the other side. Coaxial-line section L_1 is used to bring the collector feedback loop through the circuit board to the emitter section of the transistor. The 2N5921 in this circuit can deliver 4 watts of output power at 1.2 to 1.4 GHz. Frequency is tuned by adjustment of C_3.

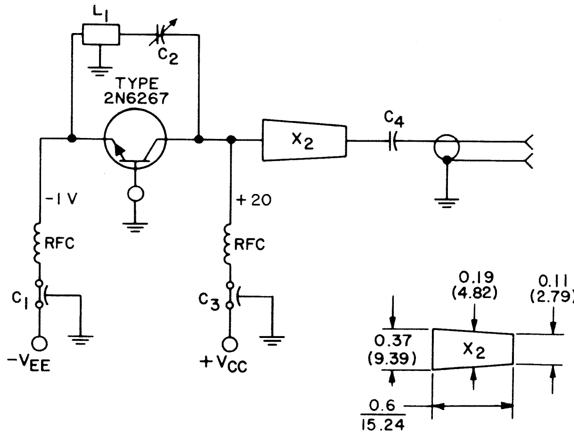
Fig. 172 shows a typical oscil- that uses an RCA-2N6267 microwave power transistor to develop a typical power output of 4.0 watts at 1.7 GHz when operated from a collector supply of 28 volts.



C1=300 pF, disc ceramic
C2, C4=470 pF, feed-through type, Allen-Bradley FA5C, or equivalent
C3=0.3-3.5 pF, Johanson 4702, or equivalent
L1=1.3 in. (33.02 mm) length of 50-ohm coaxial line
R1=1200 ohms
R2=0-250 ohms
R3=5 ohms

RFC=3 turns, No. 29 wire, 0.06 in. (1.59 mm) I.D., 0.18 in. (4.77 mm) lng.
X=Tapered microstripline—
0.1 in. (2.54 mm) wide, input end
0.24 in. (6.09 mm) wide, output end
0.475 in. (12.06 mm) long
0.005 in. (0.13 mm) thick, copper
DIELECTRIC MATERIAL=DuPont 5-MIL Kapton H, or equivalent

Figure 171. Typical circuit for tunable 1.2-to-1.4-GHz, 4-watt microstripline power oscillator.



C1, C3 = Filtercon, Allen Bradley SMFB-A1, or equivalent
C2 = 0.3- 3b pF, Johanson 4700, or equialent
C4 = 300 pF, ATC-100 or equivalent
L1 = 1.0 in. section miniature 50-ohm cable, or microstripline equivalent
RFC = 3 turns, No. 32 wire, 1/16 in. ID, 3/16 in. long

X2 = 13 mil thick Teflon-Kapton double-clad circuit boarl (Grade PE-1243 as supplied by Budd Polychem Division, Newark, Delaware), or equivalent.
Line X2 is exponentially tapered
NOTE: Oscillator is single-screw tunable from 1.6 GHz to 1.8 GHz

Figure 172. Typical 1.7-GHz oscillator circuit.

A power output of 2.3 watts at 1.6 GHz can be obtained when the transistor is operated from a collector supply of 12.5 volts. The collector efficiency in each case is in the order of 37 percent.

In the design of the oscillator, the values of the line section L_1 and the variable capacitor C_2 are chosen so that the resonant frequency of these elements is slightly less than the desired circuit oprating frequency. At frequencies above resonance (e.g., the oscillator operating frequency), the combination of L_1 and C_2, in essence, becomes a variable inductance L. The transistor input and output reactances, the inductance L, and the transistor collector-to-emitter capacitance C_{CE} form a resonant circuit that establishes the oscillation frequency of the over-all circuit and that also determines the correct level of the in-phase signal fed back to the input to sustain oscillation. The operating frequency of the oscillator is controlled by adjustment of the variable capacitor C_2.

The real part of the collector load impedance, Re (Z_{CL}), determined on the basis of large-signal class C load conditions, is transformed to 50 ohms by use of a quarter-wavelength transformer X_2. Because of the wide range of frequency control provided by the variable capacitor C_2, a quarter-wavelength tapered line is used for transformer X_2 to maintain a relatively constant transformation over a 250-MHz bandwidth. The capacitor C_4 included in the output circuit provides dc isolation for the collector bias. Similarly, capacitor C_2 provides dc isolation for the input bias supply.

The oscillator circuit features excellent stability, and the second-harmonic power is more than 45 dB below the fundamental-frequency output. Frequency stability at 1.7 GHz is better than 0.1 per cent for voltage or current excursions of ± 25 per cent. In addition, frequency drift is less than 1 MHz from the cold-start condition to the stabilized condition after 1 hour of operation.

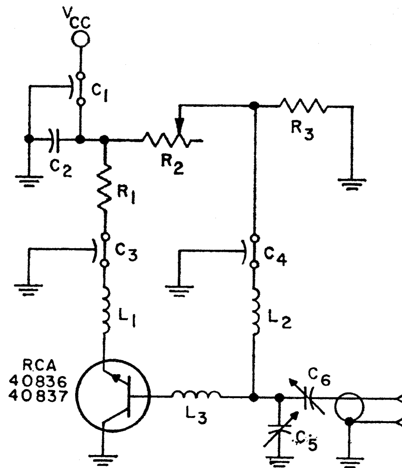
LOW-NOISE OSCILLATORS

The oscillators shown so far have used transistors with the base connected to the flange. For lower-noise oscillator operation it has been found desirable to connect the emitter of the transistor to the flange. The RCA-40836 and RCA-40837 transistors are versions of the 2N5470 and 2N5920, respectively, in which the emitter, rather than the base, is connected to the flange. A simple Colpitts-type power oscillator using essentially lumped-constant circuit elements is shown in Fig. 173 This oscillator can be tuned over the range from 1.8 to 2.1 GHz; minor circuit modifications permit tuning over a 200-MHz bandwidth in the range of 1.3 to 2.6 GHz.

The RCA-40836 can deliver 0.5 watt of output power with a V_{CE} of 21 volts. The RCA-40837 can deliver 1.25 watts at 2.1 GHz with a V_{CE} of 27 volts.

This circuit can operate below 1.8 GHz if the emitter-base capacitance and inductance L_3 are increased. For operation in the range of 2.1 GHz to 2.6 GHz, the emitter-collector capacitance must be increased, and L_3 must be decreased. Capacitors C_5 and C_6 become self-resonant at 2.6

GHz and cannot be used as capacitors above this frequency. The RCA-40837 can be made to operate in the 2.6-to-3.0-GHz range by reduction of L_3 to only the intrinsic inductance of the base cap of the coaxial package, and use of capacitive probes for C_5 and C_6. Electronic tuning of the circuit can be achieved by use of a varactor to vary the effective inductance of L_3.



C_1, C_3, C_4=470-pF, Feed-through Allen-Bradley FA5C, or equivalent
C_2=0.2 µF, disc ceramic
C_5, C_6=0.35 to 3.5-pF, Johanson 4702 or equivalent

L_1, L_2=RF choke, 0.5 in. (12.70 mm) length of No. 32 wire
L_3=Copper strip:
0.005 in (0.127 mm) thick
0.18 in. (0.457 mm) wide
0.3 in (0.76 mm) long
R_1=3 to 10 ohms, ½ W
R_2=0 to 500 ohms, 2 W
R_3=1200 ohms, ½ W

Figure 173. Typical 2-GHz grounded-collector power oscillator.

Microwave Frequency Multipliers

OPERATION of the overlay transistor in the harmonic-frequency mode can extend the upper limit of the frequency range far beyond that possible from the same transistor operating in the fundamental-frequency mode. A further advantage of the harmonic mode of operation is that frequency multiplication and power amplification can be realized simultaneously. An overlay transistor operating in this mode provides power amplification at the fundamental frequency of the input-drive power, and the nonlinear capacitance of the collector-to-base junction, acting as a varactor, generates harmonics of the input-drive frequency. It is possible, therefore, to use a single transistor to replace a transistor power amplifier and a varactor-diode frequency multiplier. In comparison with varactor frequency-multiplier circuits, the transistor multiplier is simpler, less costly, and equally efficient. It is anticipated that this mode of operation will permit extension of the available frequency spectrum for overlay transistors by a factor of two.

TRANSISTOR CONSIDERATIONS

An overlay transistor used in a frequency-multiplier circuit operates simultaneously as a power amplifier to provide gain at the fundamental frequency of the input driving power and as a varactor diode to generate harmonics of the driving power frequency. Thus, two mechanisms provide amplification and frequency multiplication in overlay transistors: one capable of gain at the fundamental frequency, and the other in which the collector-base capacitance serves as a varactor capable of frequency multiplication. Transistors suitable for multiplier applications must be capable of delivering power with gain at the fundamental frequency and of converting the power from the fundamental frequency to a harmonic frequency. A good multiplier transistor, therefore, must first be a good uhf transistor capable of high power output, gain, and efficiency. In addition, its varactor section should have minimum losses to provide maximum conversion efficiency.

The figure of merit for the amplifier portion of the transistor in which parasitic elements are not included is given by the maximum frequency of oscillation f_{max} as follows:

$$f_{max} = (PG)^{\frac{1}{2}} f$$
$$= [(1/8\pi)(1/r_{bb}' C_c \tau_{ec})] \quad (70)$$

where PG is the power gain, f is the fundamental frequency of operation, r_{bb}' is the intrinsic base-spreading resistance, C_C is the collector capacitance, and τ_{ec} is the emitter-to-collector transit or signal-delay time.

The efficiency of the varactor portion formed by the collector-base junction is determined by the cutoff frequency f_{VCB} as follows:

$$f_{VCB} = 1/[2\pi C_{min} (r_b' + r_s)] \quad (71)$$

where C_{min} is the minimum collector-to-base capacitance and r_s is the collector series resistance.

Fig. 175 shows a cross-sectional view of an overlay transistor that indicates the capacitance and loss distributions. Fig. 176 shows how the varactor portion separates from the intrinsic transistor portion. The collector-to-base capacitance consists of two parts. The major part, which comprises the active portion of the varactor, consists of the capacitance formed by the part of the collector-to-base junction that is not opposite emitter sites. This part of the capacitance is called the outer collector capacitance C_{bc}. The second part consists of the part of the collector-to-base junction that is opposite the emitter-to-base junction. This part is called the inner collector capacitance $C_{b'c}$. The outer capacitance C_{bc} is a much more efficient varactor than the inner capacitance $C_{b'c}$ because $C_{b'c}$ has to charge and discharge through both the intrinsic and the extrinsic base-spreading resistance r_{bb}' and r_b', as well as through the series resistance r_{si}, while C_{bc} can charge and discharge through only r_b' and r_{so}. Because the intrinsic base-spreading resistance r_{bb}' is much greater than the extrinsic base-spreading resistance r_b', the cutoff frequency f_{VCB} is much larger in the active varactor portion represented by C_{bc} than in the $C_{b'c}$ portions. The difference in r_b' and r_{bb}' results from the use of different sheet resistances in the two areas. Another unique

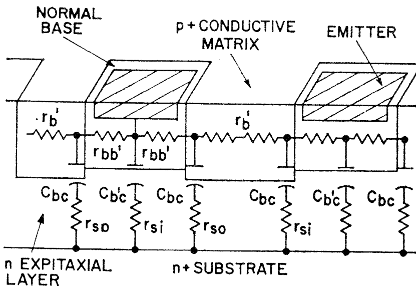


Figure 175. Cross-sectional view of an overlay transistor indicating the capacitance and loss distributions.

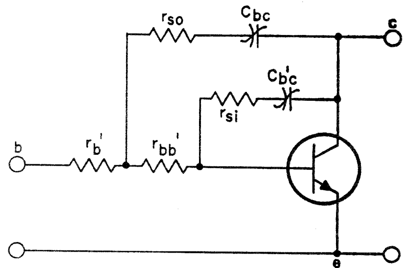


Figure 176. Circuit showing the nonlinear impedance factors that make possible frequency multiplication with overlay transistors.

feature of the overlay transistor is that the emitter area is much smaller than the base area. As a result, the inefficient portion of the varactor formed by the collector-to-base junction opposite the emitter sites is almost negligible because of the reduced emitter area.

The varactor cutoff frequency f_{VCB} is also maximized by use of minimum collector series resistance r_{so}. This resistance is kept to a minimum by the n—n+ epitaxial structure used for the collector region. The n-type epitaxial layer forms the dominant part of the collector series resistance. The thickness of this layer is kept to the minimum value that provides the required collector-to-base breakdown voltage.

Because of the features described above, varactor loss is minimized in overlay transistors and, therefore, high conversion efficiency can be achieved. The inherent varactor frequency-multiplication ability of the collector-to-base junction capacitance, added to the excellent frequency capability of these transistors, has made possible the use of overlay devices as efficient frequency multipliers.

CIRCUIT OPERATION

The outer collector capacitance C_{bc} shown in Fig. 176 varies nonlinearly with the transistor collector voltage in much the same way as the capacitance of a varactor diode varies with the voltage across the diode junction. This variable junction capacitance makes possible harmonic generation in overlay transistor circuits. The nonlinear relationship between the collector-to-base capacitance C_{bc} and the collector bias voltage in overlay transistors may be expressed as follows:

$$C_{bc} = K\,(\phi - V)^{-n} \qquad (72)$$

where K is a constant determined by the area and doping of the junction, ϕ is the contact potential, V is the magnitude of the collector reverse-bias voltage, and the exponent n is a constant determined by the impurity distribution on both sides of the junction.

Fig. 177 shows the variation in the collector-to-base capacitance

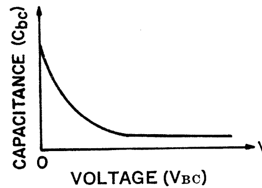


Figure 177. Collector-to-base capacitance C_{bc} as a function of collector bias in overlay transistors.

C_{bc} as a function of the collector bias voltage V_{BC}. However, this form of capacitance-voltage curve is difficult to apply directly in the analysis of high-frequency, high-power transistor circuits. Because power is the product of current and voltage swings in the transistor, the transistor current can be related to the collector-to-base capacitance if the charge Q across the junction is known. Because $dQ/dV = C(V)$, the charge Q can be determined as follows:

$$Q = \int C_{bc}\, dV \qquad (73)$$

Using the relation for capacitance C_{bc} given by Eq. (72), the integration indicated in Eq. (73) can be performed with respect to the voltage V to obtain the charge. The result of this integration, show in Fig. 178, shows the variation in the charge Q as a function of the voltage V_{BC}.

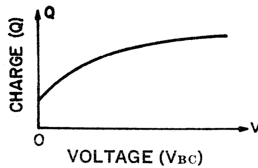


Figure 178. The charge Q in the collector-to-base junction as a function of collector-to-base voltage in an overlay transistor.

If a sinusoidal voltage such as that shown in Fig 179(a) is developed by the amplifier section of the overlay transistor to drive the nonlinear capacitance C_{bc}, a highly distorted charge (or current) waveform is produced because of the nonlinear charge-voltage characteristics of the capacitance. This waveform, shown in Fig. 179(b), contains components of the fundamental frequency and of harmonic frequencies. Power output at the desired harmonic is obtained when suitable selective circuits are coupled to the collector of the transistor. In an actual circuit, the driving voltage developed by the transistor contains both fundamental-frequency and harmonic-frequency components.

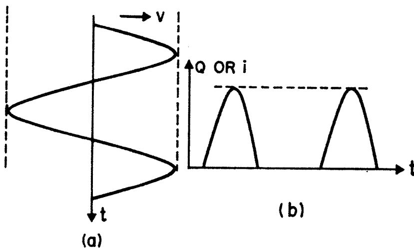


Figure 179. (a) Sinusoidal voltage developed by the amplifier section of an overlay transistor to drive the nonlinear collector-to-base capacitance; (b) distorted charge, or current waveform produced by the nonlinear collector-to-base capacitance of an overlay transistor in the generation of harmonic power.

BASIC CIRCUIT CONFIGURATIONS

Overlay transistors used in frequency multipliers may be connected in either common-base or common-emitter circuit configurations. In the common-base transistor frequency multiplier, harmonic generation is accomplished in essentially the same way as in a shunt-type varactor frequency multiplier because the nonlinear collector-to-base capacitance of the transistor is connected in shunt with the input circuit. In the common-emitter transistor frequency multiplier, the nonlinear capacitance is connected in series with the input; the operation of the transistor circuit is then similar to that of the series-type varactor frequency multiplier.

Fig. 180 shows the basic circuit configuration for the use of an overlay transistor in a common-base frequency doubler. A T matching network, or other type of matching section, must be used in the input of the doubler to set up a conjugate match across the emitter-to-base terminals of the transistor at the fundamental frequency of the input driving power. This conjugate match is required to obtain a maximum transfer of

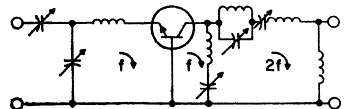


Figure 180. Basic configuration for use of an overlay transistor in a common-base frequency doubler.

power from the driving source to the transistor. Because gain at the fundamental frequency is of primary importance, an idler circuit must be connected between the collector and base of the transistor. The idler loop, which con-

sists of a simple series LC circuit, resonates with the transistor collector-to-base capacitance at the fundamental frequency and thus enhances the flow of fundamental current through the transistor. The idler circuit also develops the driving voltage required by the nonlinear collector-to-base capacitance for the generation of harmonic power. A suitable output circuit, which is series-tuned to select output power at the second harmonic of the input frequency, completes the basic doubler circuit. In some circuits, an output trap must be added to restrict the flow of fundamental-frequency current in the output loop.

Fig. 181 shows the basic circuits for the use of an overlay transistor in the common-base frequency tripler and quadrupler, respectively. These circuits are very similar to the common-base doubler, except that

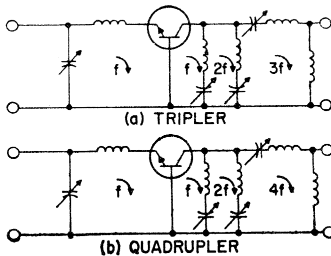


Figure 181. Basic configurations for use of an overlay transistor in a common-base base frequency tripler and (b) a frequency quadrupler.

an additional second-harmonic idler loop is connected in shunt with the transistor collector. The second-harmonic components produced by this idler loop beat with the fundamental-frequency components to generate additional harmonic outputs. In this way, the second-harmonic idler loop enhances the conversion efficiency. When an overlay-transistor frequency multiplier is used in a common-emitter circuit, an additional series resonant circuit must be incorporated in the input. Otherwise, the input, output, and idler circuits of common-emitter multipliers follow the considerations already described for the common-base multipliers.

DESIGN APPROACH

The design of transistor frequency-multiplier circuits generally consists of the selection of a suitable transistor and the design of proper filtering and matching networks for optimum circuit performance.

Transistors suitable for this application must provide the desired output power and gain at the fundamental frequency and must be able to convert the power from the fundamental frequency into power at the desired harmonic frequency. If a lossless circuit were coupled to a lossless nonlinear capacitance C_{bc}, power at the fundamental frequency could be converted into power at any harmonic frequency with 100-percent conversion efficiency. In practice, however, efficiency is limited by the series resistance associated with the nonlinear capacitance and the circuit losses. It can be considered that the harmonic output power of a transistor multiplier circuit, at a given input power level, is equal to the product of the power gain of the transistor at the drive frequency and the conversion efficiency that results from the varactor action of the collector-to-base capacitance C_{bc}. Conversion gain can be obtained only if the power gain of the transistor under consideration at the funda-

mental frequency is larger than the conversion loss.

In the design of such circuits, the input impedance at the fundamental frequency that exists at the emitter-to-base junction of the transistor as well as the load impedance presented to the collector at both the fundamental and harmonic frequencies must be known. Knowledge of the collector load impedance at the harmonic frequency is required for design of the output circuit. Knowledge of the collector impedance at the fundamental frequency is needed to determine the input impedance of the transistor at that frequency so that matching networks can be designed between the driving source and the transistor. The three impedances, of course, are interrelated and are functions of operating power level (i.e., are determined by voltage and current swings). Once the impedances are established, the design of the matching networks is straightforward. For the input circuit, a matching section having low-pass characteristics is preferred; for the output circuit, a matching section having high-pass or band-pass characteristics is preferred. Such arrangements assure good isolation between input and output circuits. As the frequency of operation increases above 800 MHz, the design of transistor multiplier circuits requires the use of distributed circuit techniques.

STABILITY AND BIASING CONSIDERATIONS

In general, the major problem of nonlinear devices is stability. Various types of instabilities can be incurred in transistor frequency-multiplier circuits, including hysteresis, low-frequency oscillations, parametric oscillations, and high-frequency oscillations. These difficulties can be eliminated or minimized by careful design of the bias circuit, by proper location of transistor ground connections, and by the use of packages that have minimum parasitic elements.

Hysteresis refers to discontinuous mode jumps in output power that occur when the input power or frequency is increased or decreased. This effect is caused by the dynamic detuning which results from variation in the average value of the nonlinear capacitance with rf voltage. The tuned circuit has a different resonant frequency for a strong drive input than for a weak drive input. It has been found experimentally that hysteresis effects can be minimized, or sometimes eliminated, when the transistor is used in a common-emitter configuration.

Low-frequency oscillations occur because the gain of the transistor at low frequency is much higher than that at the operating frequency. This effect can be eliminated by use of a small resistance in series with the rf chokes used for the biasing circuit, as shown in Fig. 182.

Parametric oscillations result because spurious low-frequency modulation is added to the harmonic output. This effect can be eliminated by careful selection of the bypass capacitance C_2 in Fig. 182 to provide a low impedance to the spurious component in addition to that provided by the rf bypass capacitance C_1.

High-frequency oscillation is indicated by oscillations that occur at a frequency very close to the output frequency when the input drive power is removed. With a

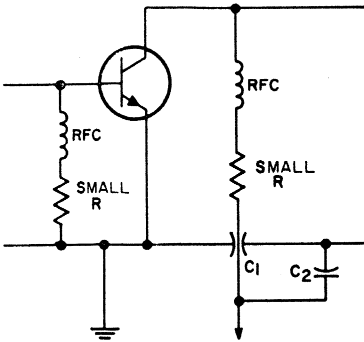


Figure 182. Circuit showing biasing techniques and bypassing capacitances used to eliminate instabilities in common-emitter frequency multipliers.

TO-60 package transistor, common-emitter circuits are found to be less critical in this respect than common-base circuits. The high-frequency oscillations are also found to be strongly related to the input drive frequency. This type of instability can be eliminated if the input frequency is kept below certain values. The input frequency at which stable operation can be obtained seems to depend upon the method of grounding the emitter of the transistor. The highest frequency of operation can be obtained when the emitter has the shortest path to ground.

In practice, stable and reliable operation of transistors in frequency multipliers has been successfully obtained. The circuits discussed in this section are all stable frequency-multiplier circuits.

TRANSISTOR CHARACTERIZED FOR FREQUENCY-MULTIPLIER APPLICATIONS

The 2N4012 power transistor is characterized for frequency-multiplication applications and can provide a minimum power output of 2.5 watts as a frequency tripler at an output frequency of 1 GHz and a collector efficiency of 25 per cent. This overlay transistor is designed to operate in military and industrial communications equipment as a frequency multiplier in the uhf or L-band range. It can be operated as a doubler, tripler, or quadrupler to supply a power output of several watts at frequencies in the low gigahertz range.

Fig. 183 shows the power-output capabilities as a function of output frequency for a typical 2N4012 transistor used in common-emitter circuit configurations for frequency doubling, tripling, and quadrupling. In a common-emitter doubler circuit, the transistor delivers power output of 3.3 watts at 800 MHz with a conversion gain of 5 dB. In a common-emitter tripler circuit, it can supply power output of 2.8 watts at 1 GHz with a conversion gain of 4.5 dB. In a common-emitter quadrupler circuit, it can provide power output of 1.7 watts at 1.2 GHz with a conversion gain of 2.3 dB.

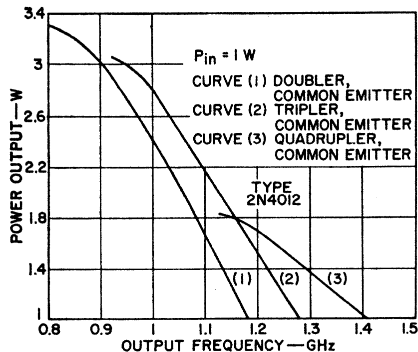


Figure 183. Power output of the RCA-2N4012 overlay transistor as a function of frequency when operated in common-emitter doubler, tripler and quadrupler circuits.

It is of interest that the transistor frequency multipliers provide greater power outputs at higher output frequencies than the unity-gain output obtained from the transistor power amplifier at 700 MHz. When the frequency of operation is low enough so that the transistor can supply rf power with substantial gain, the output capabilities of the transistor frequency multipliers are essentially the same as those of the transistor power amplifier. For operation at the same output frequency and with the same input driving power, approximately equal amounts of power output can be obtained.

Fig. 183 shows that the amount of power output that can be supplied by a transistor frequency multiplier depends upon the order of multiplication. For a given multiplier circuit, the highest output power is obtained at the frequency for which the product of power gain and conversion efficiency has the largest value. When a 2N4012 overlay transistor is used, maximum power output is obtained at 800 MHz from a doubler circuit, at 1 GHz from a tripler circuit, and at 1.3 GHz from a quadrupler circuit.

CIRCUITS

The circuit arrangements and performance data shown in the following paragraphs illustrate several practical frequency-multiplier circuits that use the 2N4012 and other RCA overlay transistors. These circuits include a 400-to-800-MHz doubler, a 150-to-450-MHz tripler, a 367-to-1100-MHz tripler, and a 420-to-1680-MHz quadrupler. As mentioned previously, the design of multiplier circuits that have an output frequency of 800 MHz or higher requires the use of distributed-circuit techniques. All such high-frequency circuits described use coaxial-cavity output circuits. These circuits are discussed first. The low-frequency circuits, which use lumped-element output circuits, are then described.

400-To-800-MHz Doubler

Fig.184 shows the complete circuit diagram of a 400-to-800-MHz doubler that uses the 2N4012 transistor. This circuit uses lumped-element input and idler circuits and a coaxial-cavity output circuit. The transistor is placed inside the cavity with its emitter properly grounded to the chassis. A pi section (C_1, C_2, L_1, L_2, and C_3) is used in the input to match the impedances, at 400 MHz, of the driving source and the base-emitter junction of the transistor. L_2 and C_3 provide the necessary

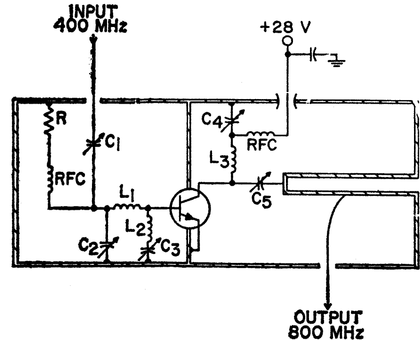


Figure 184. 400-to-800-MHz common-emitter transistor frequency multiplier.

ground return for the nonlinear capacitance of the transistor. L_3 and C_4 form the idler loop for the collector at 400 MHz. The output circuit consists of an open-ended 1¼-inch-square coaxial cavity. A lumped capacitance C_5 is added in series with a ¼-inch hollow-center conductor of the cavity near the

open end to provide adjustment for the electrical length. Power output at 800 MHz is obtained by direct coupling from a point near the shorted end of the cavity. The bias arrangement is the same as that used in the circuit shown in Fig. 182.

Fig. 185 shows the power output at 800 MHz as a function of the power input at 400 MHz for the doubler circuit, which uses a typical 2N4012 operated at a collector supply voltage of 28 volts. The curve is nearly linear at a power output level between 0.9 and 2.7 watts. The power output is 3.3 watts at 800 MHz for an input drive of 1 watt at 400 MHz, and rises to 3.9 watts as the input drive increases to 1.7 watts. The collector efficiency, which is defined as the ratio of the rf power output to the dc power input at a supply voltage of 28 volts, is also shown in Fig 185. The efficiency is 43 per cent measured at an input power of 1 watt. The 3-dB bandwidth of this circuit measured at a power output of 3.3 watts is 2.5 per cent. The fundamental-frequency component measured at a power-output level of 3.3 watts is 22 dB down from the output carrier. Higher attenuations of spurious components can be achieved if more filtering sections are used.

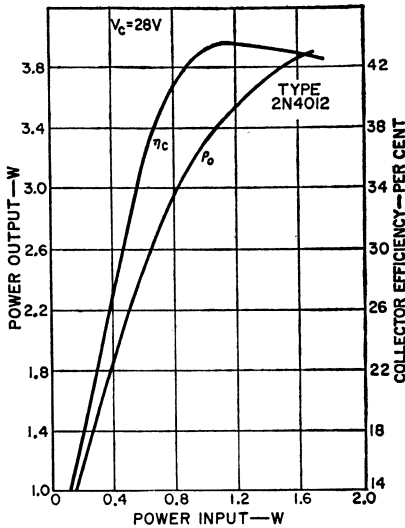


Figure 185. Output power and collector efficiency as a function of input power for the 400-to-800-MHz frequency doubler.

The variation of power output with collector supply voltage at an input drive level of 1 watt is shown in Fig. 186. This curve is obtained with the circuit tuned at 28 volts. The curves of Figs. 185 and 186 indicate that the transistor amplifier-multiplier circuit is capable of amplitude modulation.

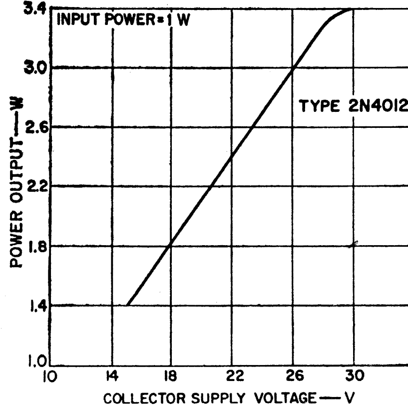


Figure 186. Power output as a function of supply voltage for the 400-to-800-MHz frequency doubler.

367-To-1100-MHz Tripler

The 367-to-1100-MHz tripler shown in Fig. 187 is essentially the same as the doubler shown in Fig. 184 except that an additional idler loop (L_4 and C_6) is added in shunt with the collector of the transistor. This idler loop is resonant with the transistor junction

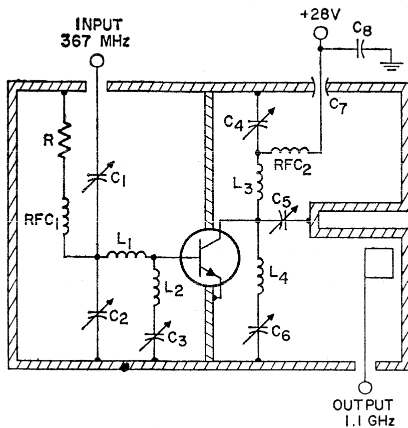


Figure 187. 367-MHz-to-1.1-GHz common-emitter transistor frequency tripler.

capacitance at the second harmonic frequency (734 MHz) of the input drive.

Fig. 188 shows the power output of the tripler at 1.1 GHz as a function of the power input at 367 MHz. This circuit also uses a typical 2N4012 transistor operated at a collector supply voltage

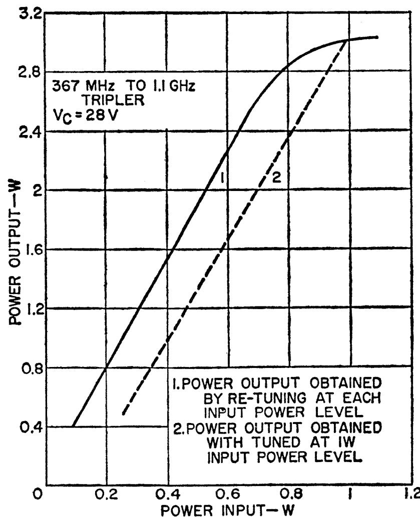


Figure 188. Power output as a function of power input for the 367-MHz-to-1.1-GHz frequency tripler.

of 28 volts. The solid-line curve shows the power output obtained when the circuit is retuned at each power-input level. The dashed-line curve shows the power output obtained with the circuit tuned at the 2.9-watt output level. A power output of 2.9 watts at 1.1 GHz is obtained with drive of 1 watt at 367 MHz. The 3-dB bandwidth measured at this power level is 2.3 per cent. The spurious-frequency components measured at the output are as follows: −22 dB at 340 MHz, −30 dB at 680 MHz, −35 dB at 1360 MHz.

The variation of power output with collector supply voltage at an input drive level of 1 watt is shown in Fig. 189. The variation of collector efficiency is also shown. These curves were obtained with the circuit tuned at 28 volts.

A 367-MHz amplifier that uses the same circuit configuration and components as those of the tripler circuit shown in Fig. 187 was constructed to compare the performance between amplifier and tripler. The conversion efficiency for a large number of tripler units was then measured. The conversion efficiency of the tripler is defined as the 1.1-GHz power obtained from the tripler divided by the 367-MHz power obtained from the amplifier at the same power-input level (1 watt). The efficiency varies between 60 and 75 per cent, and has an average value of 65 per cent; this performance is comparable to that of a good varactor multiplier in this frequency range.

A similar tripler circuit that uses a selected 2N3866 and that is operated from 500 MHz to 1.5 GHz can deliver a power output of 0.5 watt at 1.5 GHz with an input drive of 0.25 watt at 500 MHz.

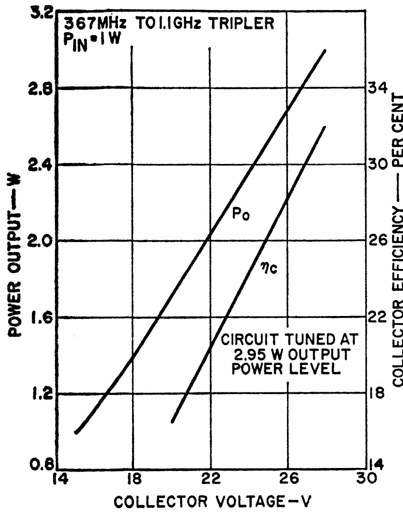


Figure 189. Power output as a function of collector supply voltage for the 367-MHz-to-1.1-GHz frequency tripler.

150-To-450-MHz Tripler Circuit

Fig. 190 illustrates the use of the 2N4012 transistor in a 150-to-450-MHz frequency tripler. The input coupling network is designed to match the driving generator to the base-to-emitter circuit of the transistor. The network formed by C_2 and L_2 provides a ground return for harmonic output current at 450 MHz. The idler network in the collector circuit (L_3, L_4, and C_4) is designed to circulate fundamental and second-harmonic components of current through the voltage-variable collector-to-base capacitance, C_{bc}. The network formed by C_5, C_6, C_7, L_5, and L_6 provides the required collector loading for 450-MHz power output. Fig. 191 shows the 450-MHz

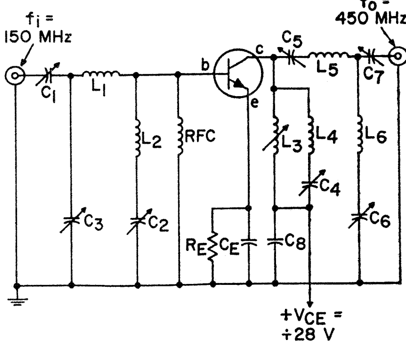


Figure 190. 150-to-450-MHz common-emitter transistor frequency tripler.

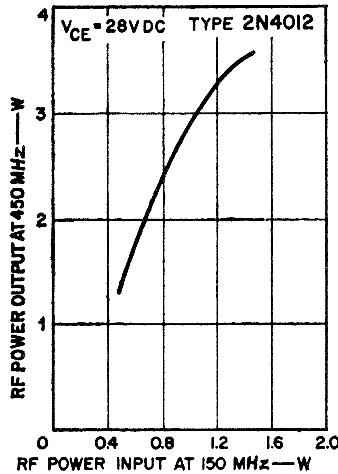


Figure 191. Power output as a function of power input for the 150-to-450-MHz frequency tripler.

power output of the tripler as a function of the 150-MHz power input. For driving power of one watt, power output of 2.8 watts is obtained at 450 MHz. The rejection of fundamental, second, and fourth harmonics was measured as 30 dB below the 2.8-watt, 450-MHz level. The variation of power output with supply voltage is shown in Fig. 192.

Common-Emitter and Common-Base Circuits

The performance data in this section are given for amplifier-

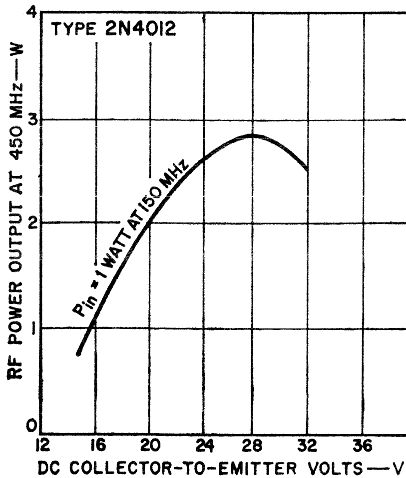


Figure 192. Power output as a function of collector supply voltage for the 150-to-450-MHz frequency tripler.

multipliers in which the transistor is connected in a common-emitter configuration. When transistors are used in common-base circuit configurations, different results are obtained. Fig. 193 shows curves of power output and efficiency for a common-base and a common-emitter tripler circuit using a 2N4012 transistor. At low power levels, the common-base tripler provides higher gain and collector efficiency; at high power levels, higher gain and collector efficiency are provided by the common-emitter circuit. At a power input of 1 watt at 367 MHz, the common-emitter tripler delivers a power output of 2.9 watts at 1.1 GHz and the common-base circuit an output of 2.4 watts. The collector efficiencies for both circuits are approximately the same and are better than 30 per cent. The 3-dB bandwidth measured in the common-emitter tripler is 2.3 per cent, as compared to 2.5 per cent in a common-base tripler. The major difference between the two circuits is that the power output of the common-emitter tripler saturates at a much higher power-input level than that of the common-base circuit. This

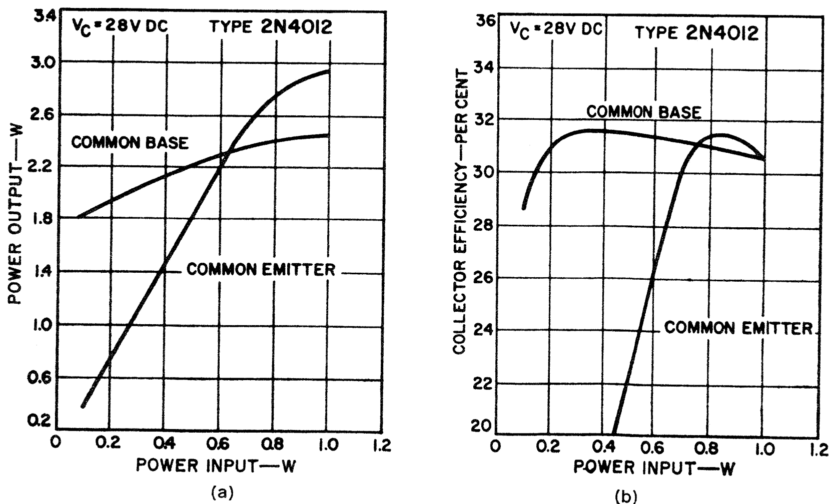


Figure 193. Comparison of performance characteristics of common-base and common-emitter tripler circuits using the RCA-2N4012 transistor: (a) power output as a function of power input; (b) collector efficiency as a function of power input.

effect has also been observed in a straight-through amplifier. In addition, the common-emitter circuit is less sensitive to hysteresis and high-frequency oscillations, as discussed previously.

A 420-MHz-To-1.68-GHz Oscillator-Quadrupler

The inherent varactor frequency-multiplication ability in overlay transistors also permits use of these devices as oscillator-multipliers. Fig 194 shows an oscillator-quadrupler circuit that uses a selected 2N3866 transistor. This circuit can deliver a power output of more than 300 milliwatts at 1.68 GHz. The first two rf chokes and the resistors R_1 and R_2 form the bias circuit. The fundamental frequency of the oscillator is 420 MHz, as determined by C_o, L_1, and C_1. L_2 and C_2 form the second-harmonic idler. The second-harmonic component produced by this idler circuit beats with the fundamental-frequency component to generate additional fourth-harmonic components. A series-tuned circuit consisting of L_3 and C_3 completes the output circuit.

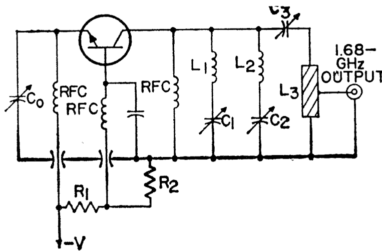


Figure 194. Oscillator-quadrupler circuit.

Microwave Systems

TRANSISTORS that can generate tens of watts of power output at frequencies up and beyond 2.3 GHz are finding applications in a wide variety of new-equipment designs. Some of the major applications for these new transistors are in the following types of equipment.

1. Telemetry
2. Microwave relay links
3. Navigational aid systems (DME, Collision Avoidance, TACAN)
4. Microwave communications
5. Phased-array radars
6. Mobile radio and radio-telephones
7. Instrusion-alarm systems
8. Electronic contermeasures (ECM)

In such equipment, transistors offer the advantages of simplified circuitry, wide bandwidths, and improved reliability, together with reduced size and weight.

MICROWAVE RELAY LINKS

Microwave point-to-point communication links are rapidly replacing wire-line and coaxial-cable networks. Microwave installations have inherently greater signal-handling capability and are less expensive than cable networks. The requirements for commercial relay-link systems are not as stringent as those for the military-oriented systems. However, high efficiency and high reliability are prime requirements because many of these installations are remote and unattended. The links may be used for voice, video, or data transmissions, using AM, FM or various digital modulation methods. The frequency spectrum used in these systems ranges from UHF to K_u-band, with both narrow-band and wide-band requirements. For example, some military telemetry systems operate near 1.5 GHz and 2.25 GHz, while common-carriers and other data link systems operate at C-band and X-band.

Microwave power transistors are finding extensive use in these relay link systems as signal sources, buffer-amplifier stages, and power-amplifier stages which drive either an antenna or a high-frequency multiplier stage. A typical rf power chain for a data-link system is shown in Fig. 195.

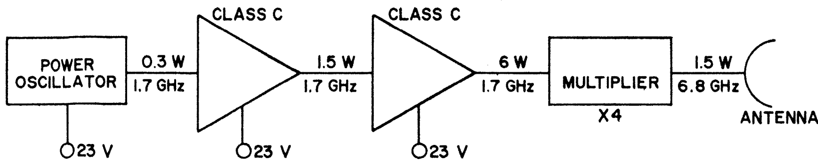
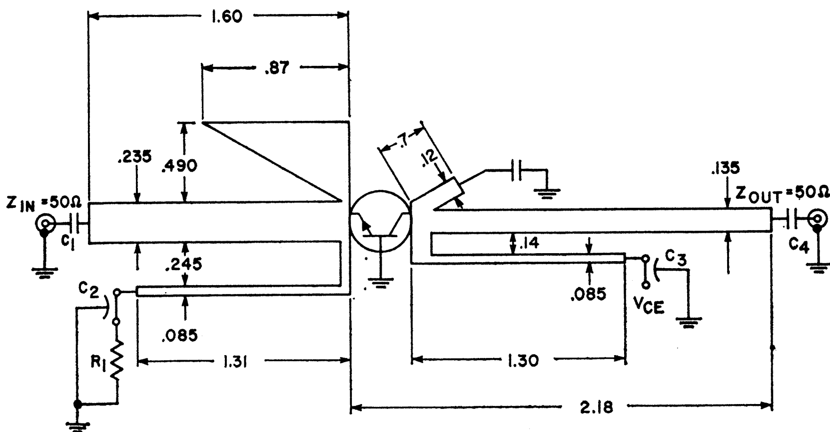


Figure 195. Microwave portion of a typical data link system.

The driver and power-amplifier stages are essentially similar to those already considered in the previous sections. The passive multiplier stage will not be considered here. Because of the close spacing of channels in data-link systems, the AM and FM noise characteristics of the oscillator stage are of prime importance in these systems. These noise requirements are best met with high-Q cavity oscillators, using transistors such as the RCA 40836 or RCA 40837.

AIR-TRAFFIC-CONTROL SYSTEMS

Most commercial and military air traffic control (ATC) systems operate in the frequency range from 960 to 1215 MHz. TACAN (Tactical Air Navigation System) ground stations operate from 960 to 1025 MHz, while airborne DME (Distance Measuring Equipment), TACAN, and transponders operate from 1025 to 1150 MHz. Transistors have great potential usage as pulsed amplifiers or oscillators at 1090 MHz for ATC or IFF (Identification, Friend or Foe) transponders, where peak power outputs of several hundred watts are required with short-pulse, low-duty operation. Microwave transistors are also used extensively for tube driver operation under broadband conditions. Fig. 196 shows a microstrip cir-



C_1, C_4 = 1000 pF, ATC-100 or equivalent
C_2, C_3 = Filtercon, low-pass filter, Allen-Bradley SMFB-A1 or equivalent
R_1 = 0.24 ohms, BMH, wirewound
Dielectric material = 1/32-inch-thick Teflon fiber-glass double-clad circuit board (ε = 2.6)

Figure 196. 2-watt, 960-to-1215-MHz microstrip amplifier.

cuit which provides 2 watts output from 960 to 1215 MHz. Fig. 197 shows the broadband performance of this circuit.

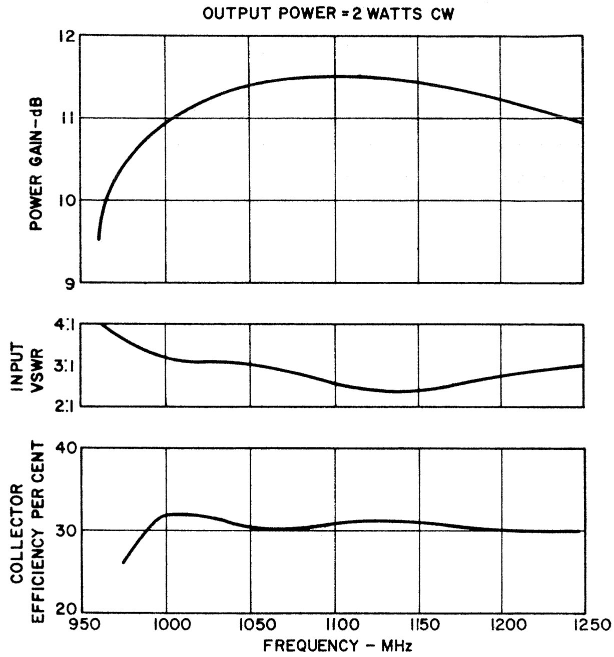


Figure 197. Performance characteristics of amplifier circuit shown in Fig. 196.

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When incorporating RCA Solid State Devices in equipment, it is recommended that the designer refer to "Operating Considerations for RCA Solid State Devices", Form No. 1CE-402, available on request from RCA Solid State Division, Box 3200, Somerville, N.J. 08876.