R-F AMPLIFIERS

R-F
AMPLIFIERS

Edited by
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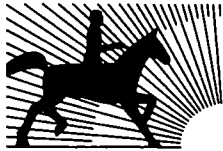
PREFACE

The circuit and design considerations pertaining to radio-frequency amplifiers are an important aspect of electronic theory. Considerations relating to such amplifiers are reflected in many aspects of electronic equipment design and maintenance procedure. For this reason it is necessary that those concerned with electronics possess a working knowledge of the essential relationships pertaining to r-f amplifiers.

The intent of this book is to discuss and evaluate amplifiers that operate in the r-f portion of the electromagnetic spectrum, particularly the low and medium ranges of the r-f band. Thus, fundamental concepts of the subject of r-f amplifiers are presented. The mathematical treatment has been kept simple, but the analyses are sufficiently extensive to permit the interested technician or student to develop a full comprehension of the pertinent theory. To ensure this aim, adequate information is given relating to broad concepts and information designed for ready use; detailed descriptions of a small number of selected major topics are presented, rather than a larger body of less important material; and, through presentatior of practical situations, equipment, and problems, the reader is afforded an opportunity to apply the principles he has learned.

Specific attention is given to the properties of tuned circuits; general considerations of capacitors and coils; the tuning inductance; coil quality; factors controlling Q; excitation of tuned circuits; problems relating to tuned circuits; r-f voltage amplifiers; their performance criteria; the gain of impedance-coupled amplifier; bandwidth of a single-tuned r-f amplifier; formulas and problems relating thereto; the gain of a transformer-coupled amplifier; the bandwidth of a transformer-coupled amplifier; double-tuned r-f amplifiers; stagger-

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tuned bandpass amplifiers; oscillation prevention; r-f power amplifiers; their general considerations, voltage and current relations, required driving power, power and efficiency, class-C power amplifier operating equations, biasing methods, grid excitation of class-C amplifiers; underdrive; saturation drive; overdrive; plate tank circuit functions and efficiency; the tank-circuit L-C ratio; factors and problems in tank-circuit design; coupling tank circuit to load; neutralization requirements and circuits; neutralization adjustments; frequency multipliers; linear-tuned class-B amplifiers; push-pull class-C amplifier considerations; grounded-grid amplifiers; cathode followers; cathode-coupled r-f amplifier; transistor r-f amplifier; factors governing transistor frequency response; effect of transit time on alpha cutoff frequency; alloy-junction transistors; surface-barrier transistor; the tetrode transistor; high-frequency transistor-amplifier circuit considerations; converters; i-f amplifiers; and automatic gain control of transistor r-f and i-f amplifiers. Through these topics, a foundation is provided upon which more advanced concepts can be built.

Grateful acknowledgment is made to the staff of the New York Institute of Technology for its assistance in the preparation of the manuscript of this book.

August, 1959   A. S.
New York, New York

CONTENTS

PROPERTIES OF TUNED CIRCUITS

1. General Considerations for Capacitors and Coils

The line of demarcation between low- and high-frequency amplifiers is not at all distinct. In general, high-frequency amplifiers utilize tuned or resonant circuits as grid loads, plate loads, or both, while low-frequency amplifiers are seldom equipped with peaking circuits except where special effects are desired. We shall be concerned specifically with amplifiers that operate in the r-f portion of the electromagnetic spectrum. We shall confine our discussions to the low and medium ranges of the r-f band of frequencies since very-high- (vhf) and ultra-high-frequency (uhf) amplifiers require a somewhat different approach and are outside the intended coverage of this book. In addition, in many cases the differences between high-frequency amplifiers and oscillators may be merely a matter of adjustment. Thus, much of the descriptive and analytical material applies to oscillators as well as amplifiers.

The tuning capacitor — In all except the ultra-high frequencies, the tuning capacitance is provided by an actual capacitor. The criteria for selecting a tuning capacitor are always based upon the lowest possible loss and the smallest possible inductance. It is usually possible to design a capacitor with extremely small resistance, but, under certain conditions, the inherent inductance may become an important factor. Table 1 shows the comparative behavior of various types of capacitors in r-f resonant circuits.

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The frequencies given in Table 1 for the various capacitor types apply particularly to receiver r-f amplifier and tuned circuits. Although the performance in higher-powered transmitters and oscillators is very much the same, the tendency has been to use fixed capacitors and variable inductors in these applications.

The tuning inductance — An inductor that approaches an ideal form is much more difficult to design and construct than a capacitor with these attributes. Consider a coil having lumped inductance,

TABLE 1

Type	*Usage*
Electrolytic	Not suited for tuned circuits at any frequency
Standard paper dielectric	These capacitors have considerable inductance and should not be used at any frequency above 800 cps
Noninductive paper	Due to reduction of inductance, these capacitors may be used in resonant circuits up to 5000 cps
Mica, standard	Suitable up to approximately 1 mc
Mica, silvered	Usually recommended up to 1 mc
Ceramic	Depends upon manufacturer. Refer to manufacturer's literature
Air dielectric, Bakelite insulation, variable	Good up to 3 mc
Air dielectric, steatite, or similar insulation, variable	Up to 100 mc
Air or vacuum dielectric, special high-frequency design	500 mc

which also displays resistance and capacitance effects. Its performance as a pure inductance is hampered by its ohmic resistance, the inductance of its leads, and its distributed capacitance. Figure 1 is an approximate representation of these factors.
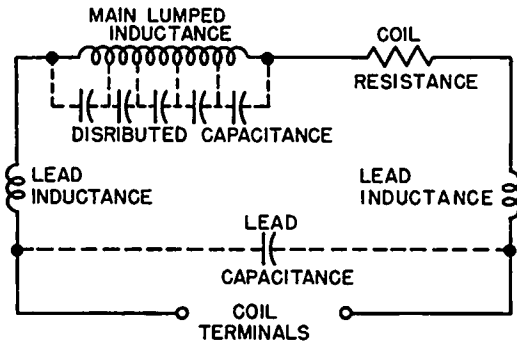
MAIN LUMPED
INDUCTANCE   COIL

DISRIBUTED CAPACITANCE

RESISTANCE

LEAD
INDUCTANCE

LEAD
INDUCTANCE

LEAD
CAPACITANCE

COIL
TERMINALS

Fig. 1. Approximate equiv-
alent circuit of a coil having
lumped inductance.

Single-layer solenoids do not suffer from distributed capacitance to the same degree as multilayer coils. Single-layer coils are practical at about 500 kc and above, because distributed capacitance effects are not so troublesome in the medium- to medium-high-frequency ranges. As the frequency rises, however, stray capacitance between coil leads and other components in the resonant circuit begin to take on a more important aspect and must be included in tuned-circuit calculations.

The a-c resistance of a coil is not a simple quantity. In addition to the pure resistance of the wire, an inductor operating in a high-frequency circuit is subject to *skin effect* and *eddy current* losses. The former is a condition in which the conducting body becomes increasingly limited to the surface of the conductor due to the effects of the high-frequency fields that surround the wire. Eddy currents represent losses due to random self-induction in the conducting wire due to the same fields. The use of special stranded wire, such as the familiar Litz type, is of some help up to several mc, but beyond this range the efficacy of Litz wire drops rapidly. Fortunately, the number of turns needed to establish resonance with an optimum value of capacitance at these frequencies becomes small enough to permit the use of heavy solid wire or tubing to minimize the a-c resistance.

Coil insulating materials must be carefully selected to avoid further r-f energy losses of significant magnitude. Bakelite is not satisfactory above about 3 mc and must be replaced by glazed ceramics such as steatite and isolantite or with special plastics such as polystyrene and lucite.

Improved powdered-iron and ferrite core materials have done much to raise the efficiency of r-f inductors in recent years. There is

an increasing tendency, particularly in receiver circuitry, to make the cores movable so that part or all of the resonant circuit tuning is accomplished by this means.

2. Coil Quality

Since no coil can be constructed with zero a-c resistance, the merit of a practical coil can be judged on the basis of the ratio of its inductive reactance to its a-c resistance. Obviously, the best coils will have a very small resistance compared to the inductive reactance. The "figure of merit," quality factor, or Q of a coil is thus defined:

$$Q = \frac{2 \pi fL}{R_{ac}} \tag{1}$$

where R_{ac} takes into account the skin effect and eddy current losses as well as the pure ohmic resistance of the coil wire. A statement of the value of Q means little unless accompanied by the frequency at which the Q is measured, since this quantity is frequency dependent.

A more fundamental interpretation of this simple definition of coil Q can be obtained by considering the following example. As-
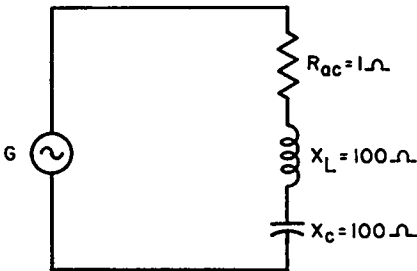

Fig. 2. Schematic showing a coil and a capacitor connected in series with an a-c generator.

sume a coil whose inductive reactance becomes exactly 100 ohms at a frequency of 15 kc, connected in series with a generator and capacitance as illustrated in Fig. 2. The assumed value of its a-c resistance R_{ac} is 1 ohm, and the capacitor has been selected so that it resonates with the coil at 15 kc. Since X_L and X_c are equal, the net reactance in the circuit may be taken as zero, making the total impedance equal to R_{ac} or 1 ohm.

To cause 1 ampere to flow in this circuit, the generator would have to develop an emf of 1 volt. With this current flowing through the coil, the voltage appearing across the coil will be 100 volts since

X_L is 100 ohms. Thus, an applied emf of 1 volt has resulted in a coil voltage of 100 volts; at the same time, it will be recognized that the Q of this coil is 100 since $2 \pi fL/R_{ac} = 100$. Hence, in a simple series-resonant circuit containing a real coil and a theoretically perfect capacitor, the Q of the coil is not only the ratio of inductive reactance to a-c resistance but is also the ratio of resonant coil voltage to the applied emf. Again, assuming a perfect capacitor, the same voltage that develops across the coil will also appear across the capacitor.

It might seem from a cursory examination of equation (1) that the Q of a coil increases in direct proportion to increasing frequency and that we could double the Q of a given inductor merely by doubling the frequency. However, the a-c resistance of a coil is also a function of the frequency since skin effect and eddy current losses increase with rising frequency. Over *moderate* ranges of frequency, Q remains *relatively* constant. For instance, the Q of well-designed radio receiver coils stays the same over the whole broadcast band (550 kc to 1600 kc). The average value for the Q of such coils is in the order of 250 to 300.

3. Factors Controlling Q

When an attempt is made to increase the Q of a coil by introducing an ordinary transformer iron core into the winding, the rise in hysteresis losses more than offsets the increase in inductive reactance. Iron losses contribute to the a-c resistance as well as skin effect, eddy-current losses, and ordinary I^2R power dissipation in the copper of the coil. Since Q is then largely determined by these various losses, we might investigate the relationship between the merit of a circuit component and the a-c power factor of the circuit. Such a step suggests itself because power factor also provides information concerning various types of losses.

Since $Q = 2 \pi fL/R$ and power factor is defined as the cosine of the angle between current and voltage in an a-c circuit, it follows that:

$$\cos \phi = \frac{R}{Z} = \frac{R}{\sqrt{R^2 + (2 \pi fL)^2}} \qquad (2)$$

Both sides of equation (2) may now be squared and written in reciprocal form as follows:

$$\frac{1}{\cos^2 \phi} = \frac{(2 \pi fL)^2 + R^2}{R^2} \qquad (3)$$

or

$$\frac{1}{\cos^2 \phi} = Q^2 + 1 \qquad (4)$$

Solving for Q:

$$Q = \sqrt{\frac{1}{\cos^2 \phi} - 1} \qquad (5)$$

In any well-designed tuned circuit used at radio frequencies, the power factor is quite small since the coil has comparatively few turns (especially at high frequencies), so that the resistance is always small with respect to the total impedance. If, for example, the power factor is of the order of .05, the square becomes a still smaller quantity, and the reciprocal is correspondingly large. It is a good approximation, therefore, to ignore the unity factor under the radical. This permits simplification of equation (5) thus:

$$Q = \frac{1}{\cos \phi} \qquad (6)$$

From this relation we may state that the Q of a coil is inversely proportional to its power factor. The significance of this is evident when we consider that a coil having a laminated iron core has a power factor of the order of 5%. Substituting .05 in equation (6) gives an answer $Q = 20$. Since the magnitude of the voltage developed across the L or C component in a resonant circuit is a function of Q, this low value of Q suggests that voltage multiplication in low-frequency circuits is more difficult to obtain than in high-frequency circuits. For example, a heavy-wire or copper-tubing transmitter tank coil may be given an extremely high Q because it can be designed with a very low power factor. Hence, the r-f voltage that can appear across it for a given power input is also relatively high.

4. Excitation of Tuned Circuits

Tuned circuits may be excited by introducing signal voltage or signal current at the frequency of resonance into the tuned system. The method of transferring the a-c power from the generator to the resonant circuit determines whether current or voltage will be the principal agent. This will be clarified by the discussion that follows.
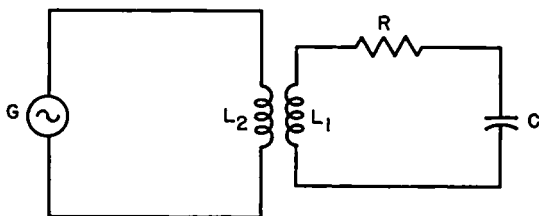
Fig. 3. Coupling the generator to the tuned circuit with a mutual inductance.

A basic series resonant circuit as illustrated in Fig. 2 contains the actual generator within the string of series components. A similar effect may be obtained by coupling the signal generator to the coil by the mutual inductance between the generator's coil and that of the tuned circuit (Fig. 3). The alternating voltage that appears across L_1 by induction is a function of the ac flowing in L_2, its frequency, and the mutual inductance between the two coils. The magnitude of this induced voltage E_1 is given by the expression:

$$E_1 = 2 \pi fMI_2 \tag{7}$$

in which E_1 is the induced voltage, f is the frequency of the current in L_2, M is the mutual inductance in henries, and I_2 is the current in L_2. The relationship of values is perhaps best illustrated by a numerical example.

Problem 1. The mutual inductance between the generator coil and the tuned circuit coil is 0.5 mh. If a current of 0.02 ampere flows in L_2 at a frequency of 20 kc, what voltage will be induced in L_1?

Solution. Substituting in equation (7):

$$E_1 = 6.28 \times 0.5 \times 10^{-3} \times 0.02 \times 2 \times 10^4$$

$$= 1.3 \text{ volts}$$

Further analysis of the mutual-inductance method of exciting a tuned circuit demonstrates that the current that flows in L_1 is identical to the current that would flow *if the induced voltage (E_1) were applied in series with L_1 in the absence of any primary (L_2) at all.* Thus, this excitation method is the equivalent of the circuit in Fig. 2; for this reason it is often termed *series excitation.*

An important characteristic of series excitation is that the signal source furnishes *current* rather than *voltage* and that the voltage it does supply is small compared to the potentials that exist across the capacitive and inductive portions of the tuned circuit. The circuit loss (I^2R) is exactly the power that the source must supply.

An alternative method of supplying the required power to a resonant circuit consists in applying the exciting voltage in parallel with
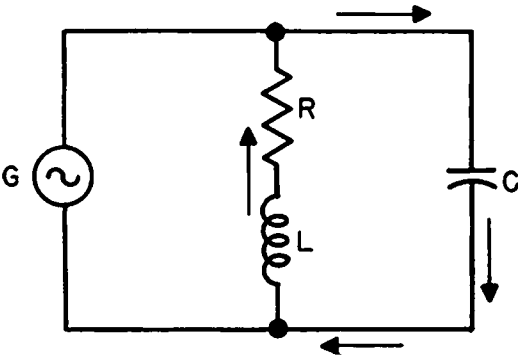
Fig. 4. Parallel excitation of a resonant circuit.

the tuning elements (Fig. 4). Since the connection is a parallel one, the same voltage is applied to both branches of the circuit. The current in the capacitive branch has a leading phase angle of 90°, while that in the inductive branch has a lagging phase angle of *nearly* 90° if the resistance factor is small. When resonance is approached, the current magnitudes in the branches also approach equality with a net phase angle of 180° between them. That is, the instantaneous current in C at resonance is equal to but opposite in direction to the instantaneous current in L. This must be the case, or very nearly the case, at resonance when the reactances of the two branches are almost identical. With equal voltages applied to equal reactances, the current *magnitude* must also be equal. If the arrows are taken as instantaneous current, it may be seen from Fig. 4 that large *circulating* currents may exist in the R-L-C circuit if anything approaching ideal conditions is realized. The net current taken from the source is therefore theoretically zero (if both inductance and capacitance are perfect). In practice, with well-constructed components, the net current is very small.

Thus, in the parallel excitation process, the source furnishes very little current but must provide a potential equal to the voltage that appears across either of the reactive components. Note the contrast between parallel and series excitation methods.

5. The Apparent Resistance of a Parallel Resonant Circuit

When inductance and capacitance are properly selected with reference to a specific frequency, the resulting parallel resonant circuit exhibits an impedance which is purely resistive. From previous con-

siderations, it would appear that a relationship must exist between the a-c resistance (R) of the coil (due to ohmic resistance, skin effect, eddy currents, and hysteresis) and the apparent resistance R_z. This conclusion is justified by the fact that, in a perfect parallel resonant circuit, the source need supply no current at all; thus the L-C combination behaves as though it had infinite resistance at the resonant frequency.

If the circuit is not perfect, that is if the a-c resistance of the coil is not zero, then the phase shift in the coil is not exactly 90° so that absolute cancellation of the inductive and capacitive currents does not take place. In this case, the source must supply some current, small as it may be. Thus, the apparent resistance R_z of the tuned circuit drops from an infinite value to a finite one due to the presence of R in the coil. The exact and approximate relationships between R_z and R are not difficult to derive; certainly the final equations are most valuable in any study of tuned circuits and r-f amplifiers.

Assuming a practical, imperfect resonant circuit, we shall first determine how much current the signal source must supply. Our discussion will be based upon the vector diagram presented in Fig. 5.

Vector BE represents the exciting voltage applied by the external generator parallel to the resonant circuit of Fig. 4. A current I_c leading the applied voltage by 90° appears in the capacitive component and is represented by vector BA in Fig. 5. The applied voltage also gives rise to a current in the inductive branch which lags the voltage by an angle ϕ *less* than 90° due to the resistance present in this element (the a-c resistance R). This inductive-branch current I_L is given by vector BC.

The current through the coil produces alternating voltage drops across its own a-c resistance R and across its inductive reactance X_L. These are symbolized, respectively, as $I_L R$ and $I_L X_L$ and are given by vectors BF and FE. Since these voltages must sum vectorially to the applied voltage, they must form the two arms of a right triangle (BFE), of which the applied voltage (BE) is the hypotenuse. Vector BF is shown coincident with vector BC, because the voltage drop in R must be in phase with the current flowing through it. FE is drawn at right angles to I_L because the voltage developed by the current flowing through the coil must be 90° ahead of the current.

The current to be supplied by the source must be the vector sum of I_c and I_L, so that vectors BA and BC must be added vectorially. This is done by means of the paralellogram ABCD in which the

diagonal BD must represent the total source current. Since the apparent resistance R_z of a parallel resonant circuit behaves as a true pure resistance, then the total current I must be in phase with the total applied voltage BE. For this reason the total current vector

Fig. 5. Vector diagram containing the elements necessary to calculate the current taken from the source by a resonant circuit in which the a-c resistance R is finite in value.
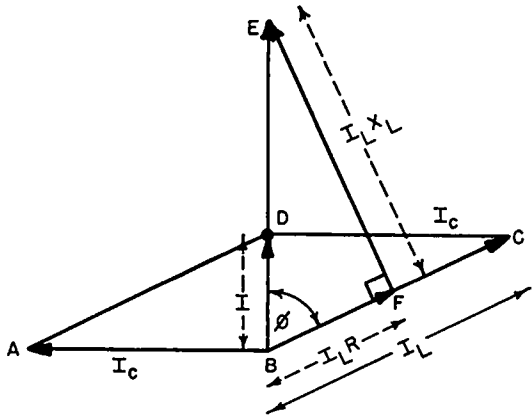
BD must lie along the same line as the applied voltage vector BE. Finally, the phase angle ϕ is represented as the angle between the applied voltage BE and the current through the coil I_L.

The relationship between the apparent resonant resistance R_z and the a-c resistance R of the tuned circuit may now be derived with the aid of the vector diagram.

The apparent resonant resistance R_z is defined by the vector diagram as:

$$R_z = \text{applied voltage/resulting current}$$

or

$$R_z = BE/BD = E/I \tag{8}$$

Triangles BFE and BDC are similar since they are both right triangles and have angle ϕ in common. Thus:

$$BD/BC = BF/BE \tag{9}$$

The electrical terms that correspond to the geometry of the triangle are:

$$BD = I$$

$$BC = I_L$$

$$BF = I_L R$$

$$BE = \sqrt{(I_L R)^2 + (I_L X_L)^2} = I_L \sqrt{R^2 + X_L^2}$$

In setting up the electrical ratio, the I_L factors on the right cancel out and we have

$$\frac{I}{I_L} = \frac{R}{\sqrt{R^2 + X_L^2}} \tag{10}$$

or

$$I = \frac{I_L R}{\sqrt{R^2 + X_L^2}} \tag{11}$$

The current flowing in the inductance is I_L. Since this current is the ratio of the voltage across this branch to the impedance of the branch, we may write:

$$I_L = E/Z_L \tag{12}$$

The factor Z_L is the root-mean-square sum of the inductive reactance and the resistance of this branch, hence

$$I_L = \frac{E}{\sqrt{R^2 + X_L^2}} \tag{13}$$

Combining equations (11) and (13):

$$I = \frac{ER}{R^2 + X_L^2} \tag{14}$$

Referring back to equation (8), the apparent resonant resistance R_z is therefore given by the equation:

$$R_z = \frac{E}{I} = \frac{R^2 + 4\pi^2 f^2 L^2}{R} \tag{15}$$

This is the basic form of a very important equation frequently used in the solution of parallel resonance problems in practical work. Before analyzing any of its simpler forms, the reader should observe that as the coil is made more nearly perfect, i.e., as R approaches zero, the apparent resonant resistance rises higher and higher. This was predicted from purely qualitative considerations in the paragraph preceding the mathematical analysis.

Equation (15) can be simplified by a justifiable assumption: since the a-c resistance of a good coil is seldom more than 5% of

the inductive reactance, the R^2 factor in the numerator may be ignored and a good approximation obtained by writing:

$$R_z = \frac{(2\ \pi\ fL)^2}{R} \qquad (16)$$

Factoring the numerator, we have:

$$R_z = 2\ \pi\ fL\ \frac{2\ \pi\ fL}{R} \qquad (17)$$

since $2\ \pi\ fL/R$ defines Q, then:

$$R_z = 2\ \pi\ fL\ Q \qquad (18)$$

Equation (18) conveys the information that the apparent resonant resistance of a parallel resonant circuit containing a perfect capacitor and a good coil having a small a-c resistance is equal to the product of the inductive reactance and the Q of the coil.

Another very useful expression may be derived from the foregoing equations by considering that the frequency of an ideal resonant circuit is given by:

$$f = \frac{1}{2\ \pi\ \sqrt{LC}} \qquad (19)$$

from which we can obtain:

$$(2\ \pi\ f)^2 = \frac{1}{LC} \qquad (20)$$

Thus, if $1/LC$ is substituted for $(2\ \pi\ f)^2$ in equation (16), we arrive at:

$$R_z = \frac{L}{RC} \qquad (21)$$

The usefulness of these relationships may be shown by means of an illustrative example.

Problem 2. A parallel resonant circuit contains an inductance of 2 mh and a capacitor of 0.01 μf. The a-c resistance of the coil is 3 ohms. Find (a) the apparent resonant resistance presented to the signal source by the circuit, (b) the resonant frequency, (c) the inductive reactance of the coil at the resonant frequency, (d) the Q of the coil.

Solution. (a) The apparent resonant resistance is most easily obtained from equation (21).

$$R_z = \frac{2 \times 10^{-3}}{3 \times 10^{-8}}$$

$$= 66,000 \text{ ohms}$$

(b) Equation (19) gives the resonant frequency

$$f = \frac{1}{6.28 \sqrt{2 \times 10^{-3} \times 0.01 \times 10^{-6}}}$$

$$= \frac{10^6}{4.46 \times 6.28}$$

$$= 10^6/28 = 35,700 \text{ cps}$$

(c) The inductive reactance is therefore:

$$X_L = 2 \pi fL = 6.28 \times 3.57 \times 10^4 \times 2 \times 10^{-3}$$

$$= 447 \text{ ohms}$$

(d) The Q of the coil is the ratio of X_L to R. Thus,

$$Q = \frac{447}{3} = 149$$

The calculations may be checked by using equation (18).

$$R_z = 2 \pi fL Q$$

$$= 6.28 \times 3.57 \times 10^4 \times 2 \times 10^{-3} \times 149$$

$$= 66,000 \text{ (approx.)}$$

Note that equation (21) permits calculation of the apparent resonant resistance of a tuned circuit without first finding the resonant frequency, thus eliminating several tedious steps.

A rigorous analysis of resonant circuits[1] demonstrates that the frequency of resonance obtained from equation (19), i.e., when the inductive reactance equals the capacitive reactance, is not precisely the same frequency as the one for which the resonant circuit behaves as a pure resistance. Thus, the transition from equation (18) through (21) is not quite accurate. On the other hand, the error produced by this transition becomes insignificantly small when the coil Q exceeds about 20, so that the use of any of the equations in solving r-f amplifier resonant circuits is almost always justified.

6. Review Questions

1. What are the criteria for selecting a tuning capacitor?

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[1] See A. Schure, *Resonant Circuits*, New York: John F. Rider Publisher, Inc., 1957.

2. Explain what is meant by "skin effect." How does it influence the a-c resistance of a coil?

3. Explain why a good powdered-iron core, when inserted in a coil that formerly had an air core, can improve the Q of the coil. Other things being equal, what would the insertion of such a core do to the inductive reactance of the coil?

4. What factors contribute to the a-c resistance of a coil beside the ohmic resistance of its conductor and the skin effect at high frequencies?

5. How else may Q be defined beside saying that it is the ratio of inductive reactance to a-c resistance?

6. What is the Q of a coil whose power factor is 0.004?

7. Two coils are arranged as an r-f transformer having a primary L_2 and a secondary L_1. The mutual inductance between them is 0.08 mh. A current of 50 ma flows in L_2 at a frequency of 100 kc. What voltage will be induced in L_1?

8. Explain why the generator that excites a resonant circuit through mutual inductance must have different characteristics than a generator which excites a parallel resonant circuit directly.

9. Find the Q of a coil that is a part of a parallel tuned circuit that resonates at 16 mc, if the apparent resonant resistance of the circuit is 5000 ohms and the inductive reactance of the coil is 100 ohms.

10. The a-c resistance of a coil is approximately 5 ohms. It resonates in parallel with a 0.1-μf capacitor to produce an apparent resonant resistance of 10,000 ohms. What is the approximate inductance of the coil in henries?

Chapter 2

R-F VOLTAGE AMPLIFIERS

7. Performance Criteria

R-f voltage amplifiers are seldom encountered in transmitting equipment but are found in radio receivers, remote-control equipment, and in radar and pulse amplifier apparatus. In these applications, r-f amplifiers are virtually always operated class A, which provides the highest voltage gain. And, since power output is not required, plate-circuit efficiency is not a factor.

An r-f amplifier is judged on effective gain at its tuned frequency, frequency response to signals lying at each side of the resonant frequency, and ability to discriminate against frequencies immediately adjacent to the passband.

At radio frequencies lying in the broadcast band, for example, it is *not* necessary for the amplifier to handle a range of several octaves with equal gain as is required for audio amplifiers. This makes it possible to take advantage of resonance to realize sufficiently high impedances to match vacuum-tube resistances. In this case it is essential that the r-f amplifier be easily adjustable with respect to the frequency of maximum amplification so that the user may select any portion of the band between 550 kc and 1600 kc for reception. When the adjustment has been made, say, for maximum amplification at 880 kc, it is then necessary for the circuit to amplify equally all frequencies over a band of approximately 10 kc having a center point at 880 kc. Thus, the required bandwidth of this amplifier
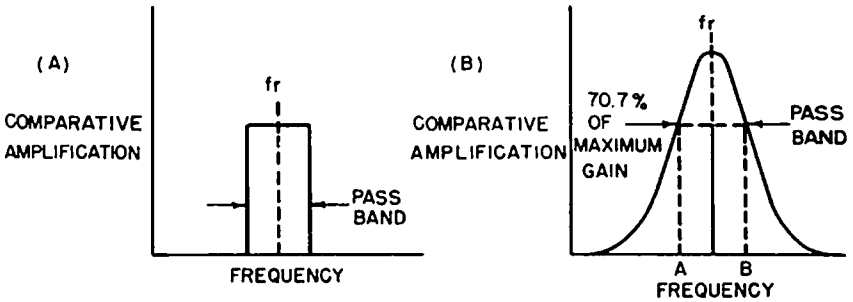
Fig. 6. (A) Resonance curve of an ideal amplifier. (B) Resonance curve of an r-f amplifier using a resonant circuit of reasonably high Q.

would be 10 kc, and the coverage range would stretch from 875 kc to 885 kc. Note that this range is only slightly greater than 1% of a full octave.

An ideal r-f amplifier may be viewed as a bandpass filter having a response curve like that shown in Fig. 6A. Practical resonant circuits cannot be made to yield this order of perfection; when the tuning elements have sufficiently high Q, however, this performance can be approached quite closely. Figure 6B illustrates the frequency response of a reasonably high-Q r-f amplifier. The passband (or bandwidth) of a practical amplifier is defined as the band of frequencies that is amplified not less than 70.7% of the maximum amplification of the input signal at resonance. The bandwidth of the amplifier whose curve is given in Fig. 6B is equal to the bandwidth between A and B; for the broadcast amplifier given above, points A and B would be 875 kc and 885 kc, respectively. As we shall show later, the bandwidth of an amplifier depends upon its resonant frequency and the effective Q of the amplification curve.

The percentage figure (70.7%) previously given as the criterion for determining bandwidth comes directly from the accepted definition of bandwidth. As all normal response curves have more or less sloping sides, it is felt that when the power amplification falls below one-half of that obtained at the resonant frequency, the amplifier can no longer be said to be "passing" the signal. A diminution of power to half-value is a loss of 3 db (−3 db down). As applied to *voltage* amplifiers, −3 db represents a voltage loss of 29.3% of the maximum value, hence a signal is said to be passed only if its voltage is 70.7% (0.707) of maximum or more.

The ability of an amplifier to discriminate against signals that are adjacent to the edges of its passband is also a function of frequency

and effective Q. In general, the narrower the bandwidth of an amplifier, the better is its rejection action for signals lying outside its passband. This is an important consideration in all types of receiver amplifiers where *selectivity* is important.

8. Gain of Impedance-Coupled (Single-Tuned) Amplifier

A single-tuned amplifier is perhaps the simplest form to design and analyze. As illustrated in Fig. 7, a single-tuned stage — often said to be impedance coupled — consists of the amplifying device, which may be either a tube or a transistor, an L-C circuit arranged for parallel resonance, and associated components such as coupling capacitors, bypass capacitors, and resistors. The reader's attention is called to the equivalent circuit and some of its features that warrant examination. To the right of the input generator are r_p, the dynamic plate resistance of the tube; C, the tuning capacitance and any stray
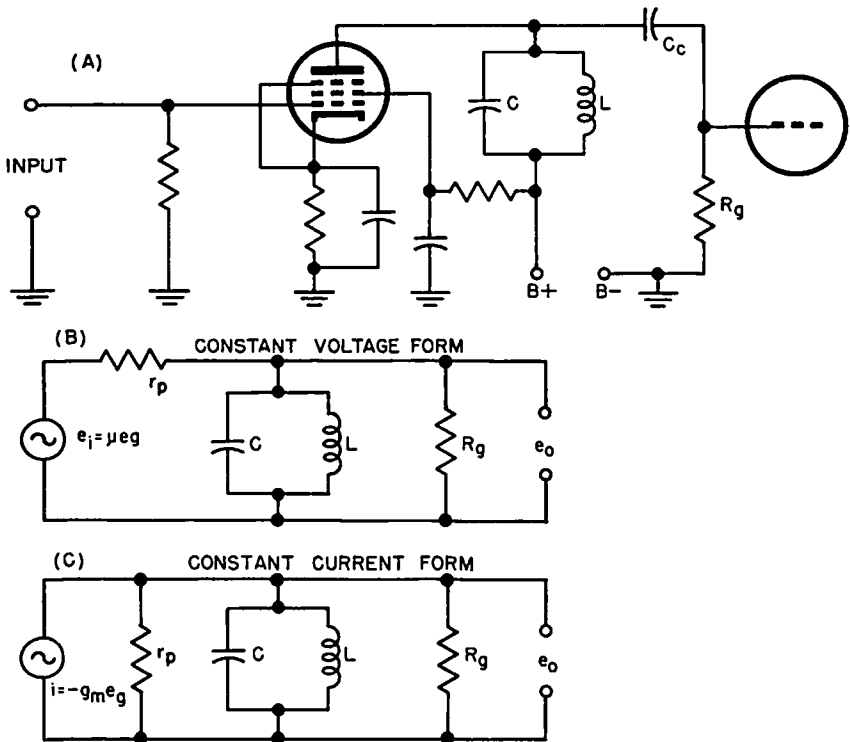
Fig. 7. (A) Single-tuned voltage amplifier schematic circuit. (B) Equivalent circuit.

capacitance that may change the lumped value; L, the total tuning inductance including stray inductances; R_g, the total input resistance to the next stage or the output device; and e_o, the output voltage.

Referring to the generator, it will be seen that this has been labeled as a source that produces either:

(a) a *constant input voltage* to the rest of the equivalent circuit, designated as $e_1 = -\mu e_g$. This is the conventional method of expressing the plate-circuit output of a vacuum tube and takes into account the input grid signal and the amplification factor. This form of the equivalent plate-circuit theorem is very useful for analyzing triodes and triode amplifiers.

(b) a *constant input current* to the rest of the equivalent circuit, expressed as $i = -g_m e_g$. With pentode voltage amplifiers, this form is often more convenient to handle. It applies particularly well to pentodes since external voltage drops do not affect the plate current of a pentode as long as its plate voltage remains higher than the screen voltage. The expression $i = -g_m e_g$ is obtained from the constant voltage form by applying Norton's theorem.* From the constant voltage form:

$$e_1 = -\mu e_g \qquad (22)$$

The current flowing in r_g as a result of e_1 is therefore:

$$i = -\mu e_g / r_p \qquad (23)$$

But the ratio of amplification factor (μ) to plate resistance (r_p) is the transconductance of the tube. Hence:

$$i = -g_m e_g \qquad (24)$$

The transition from equations (22) to (23) is made possible by Norton's theorem, provided that the plate resistance r_p which was taken as a series resistance in the constant voltage equivalent circuit is now considered to be in shunt with the constant current generator (Fig. 7B).

The voltage gain of a pentode may be written as given in equation (25).

---

* Dr. E. Norton, Bell Telephone Laboratories, first proposed the theorem. It states that a simple generator delivering a current i may be used to replace any network of complex generators and impedances, provided that the generator is assigned an internal impedance Z such that, when the network terminals are short-circuited and the current i flows in the short-circuiting wires, an impedance Z is seen in the network looking back from the short-circuited terminals.

$$\text{Gain} = g_m \, Z_L \qquad (25)$$

in which g_m is the transconductance of the tube and Z_L is the total load impedance. Using the constant current form, the total load impedance is seen to be the impedance of the parallel resonant circuit consisting of C and L shunted by r_p and R_g. Since shunting impedances change the value of circuit Q, let us designate the effective Q of the shunted resonant circuit as Q'. Now, applying equation (18), we may combine equations (18) and (25) thus:

$$\text{Gain} = 2 \, \pi \, \text{fL} \, Q' \, g_m \qquad (26)$$

The impedance of the resonant circuit has been taken as R_z as in equation (18), while the shunting effects of r_p and R_g have been included in Q'. In most amplifiers, especially those used for narrowband applications in which a sharply peaked, steep-sided resonance curve is desired, both r_p and R_g are very large. Their shunting effect, therefore, is often negligible. In such cases, Q' may be taken as equal to the Q of the resonant circuit.*

Problem 3. A tuned r-f amplifier operating at 1500 kc uses a 6AC7 pentode having a transconductance of 9000 μmhos and a plate resistance of 1 megohm. The coil in its tuned circuit has an inductance of 200 μh and an a-c resistance of 8 ohms at this frequency. The input resistance to the next stage is 1 megohm. What voltage amplification may be expected from this circuit?

Solution. Since the 6AC7 has an extremely high r_p and the following stage a very high R_g, the approximate form of equation (26) may be safely used, and Q' may be assumed to be equal to Q. Since Q equals $2 \, \pi \, \text{fL}/R$ in which f is the resonant frequency and R is the a-c resistance of the coil, then we may write in place of equation (26):

$$\text{Gain} = \frac{g_m \, (2 \, \pi \, \text{fL})^2}{R} \qquad (27)$$

substituting:

$$\text{Gain} = \frac{9000 \times 10^{-6} \, (6.28 \times 1.5 \times 10^6 \times 200 \times 10^{-6})^2}{8}$$

---

* Effective Q is given by the equation

$$Q' = \frac{Q}{1 + \dfrac{R_z Q}{r_p} + \dfrac{R_z Q}{R_g}}$$

As r_p and R_g are made increasingly large in value, it is evident that the fractions with which they are associated approach more closely to zero. Thus, with r_p and R_g very large, these terms lose their significance and Q' becomes approximately equal to Q.

$$\text{Gain} = \frac{9 \times 10^{-3} \times 353 \times 10^4}{8}$$

$$= 3970$$

9. Bandwidth of a Single-Tuned R-F Amplifier

We are now prepared to explain the origin of 70.7% as a factor used in determining amplifier bandwidth. It will be recalled (Section 7) that the bandwidth is measured between the two frequencies adjacent to the resonant frequency where the amplified voltage falls to 70.7% of the maximum resonant amplitude (see Fig. 6B). This percentage is an arbitrary one; as long as the same factor is consistently used in comparing amplifiers, any fraction may be used provided that it is realistic and compatible with our experience in analyzing amplifier performance.

On this basis, sharpness of resonance (or bandwidth) is specified by taking as limiting points those frequencies adjacent to resonance where the reactance of the tuned circuit has risen from its value of zero at resonance to a magnitude equal to the a-c resistance of the inductance and its associated wires. Such an increase of reactance occurs at two frequencies, one above and one below the resonant point. Having taken these frequencies as those for which $X = R$, we may write:

$$Z = R \sqrt{2} \tag{28}$$

This expression evolves directly from the fact that $Z = \sqrt{R^2 + X^2}$. For this condition, the impedance of the resonant circuit is 1.41 times greater than its value at maximum; this, in turn, means that the resonant current is 0.707 times its maximum value and that the power for a given voltage is one-half the resonant magnitude.

Assuming now that we start with an amplifier tuned to resonance, a change of frequency equal to df will cause the capacitive reactance as well as the inductive reactance of the tuned circuits to change. These changes will take place in opposite directions. For example, an increase in frequency df will cause the capacitive reactance to decrease by the same amount as the inductive reactance increases. Thus:

$$dX_L = 2 \pi \, dfL \tag{29}$$

and

$$dX_c = \frac{1}{2 \pi dfC} \qquad (30)$$

Since it can be shown that both reactance changes are equal, the change in net reactance, due to the fact that X_c and X_L change in opposite directions, is given by:

$$dX = 4 \pi dfL \qquad (31)$$

Equation (31) merely says that the change in reactance of a series-resonant circuit in the region of resonance is *twice the change in inductive reactance alone* when a frequency variation of df is permitted to take place.

If the resonant frequency is symbolized by f_r, the upper limiting frequency, where the signal voltage is 0.707 times the maximum value, is symbolized by f_2, and the corresponding lower frequency is designated by f_1, then we obtain the expressions:

$$R = 4 \pi L \ (f_r - f_1) \qquad (32)$$

and

$$R = 4 \pi L \ (f_2 - f_r) \qquad (33)$$

These equations are, of course, based on the initial assumption that the limiting bandwidth frequencies may be designated as those for which $X = R$. In addition, df is given as $f_2 - f_r$ for the frequency difference between resonance and the upper limiting point, while df is similarly $f_r - f_1$ for the lower limiting frequency.

Since bandwidth is specified by the quantity $(f_2 - f_1)$, equation (32) may be solved for f_1, and equation (33) may be solved for f_2. When the resulting expression for f_1 is subtracted from the corresponding expression for f_2, we obtain*:

$$f_2 - f_1 = \frac{R}{2 \pi L} \qquad (34)$$

---

* The algebraic manipulations required to obtain equation (34) are given below.
$4 \pi L \ (f_r - f_1) = R$, when expanded, becomes

$$4 \pi L \ f_r - 4 \pi L \ f_1 = R$$

Solving for f_1, we have:

$$f_1 = \frac{- R + 4 \pi L \ f_r}{4 \pi L}$$

(contd.)

When both sides of equation (34) are divided by the factor f_r, we arrive at a statement which shows the dependence of bandwidth on Q, as follows:

$$\frac{f_2 - f_1}{f_r} = \frac{R}{2 \pi f_r L} \qquad (35)$$

and since the right-hand side is the reciprocal of Q, then

$$Q = \frac{f_r}{f_2 - f_1} \qquad (36)$$

and finally, equation (36) may be solved for the bandwidth of the resonant circuit to yield:

$$f_2 - f_1 = \frac{f_r}{Q} \qquad (37)$$

Note that the factor Q in equation (37) must be replaced by the effective Q (Q') if r_p and R_g are not large. Effective Q may be found from the equation given in the footnote on page 19.

Problem 4. A resonant circuit is to be set up with a bandwidth of 5 kc and a resonant frequency of 2 mc. The tuning capacitor is fixed at 200 μμf. Find the inductance of the coil and the a-c resistance it must have to provide the required bandwidth.

Solution. The inductance required is first determined from the resonant frequency and the value of the tuning capacitor.

$$L = \frac{1}{4 \pi^2 f^2 C}$$

(contd.)

Similarly, when $4 \pi L (f_2 - f_r) = R$ is solved for f_2,

$$f_2 = \frac{R + 4 \pi L f_r}{4 \pi L}$$

Subtracting the upper from the lower:

$$f_2 - f_1 = \frac{R + 4 \pi L f_r}{4 \pi L} - \frac{(- R + 4 \pi L f_r)}{4 \pi L}$$

$$= \frac{R + 4 \pi L f_r + R - 4 \pi L f_r}{4 \pi L}$$

$$= \frac{2 R}{4 \pi L}$$

and finally [equation (34)]:

$$f_2 - f_1 = \frac{R}{2 \pi L}$$

$$L = \frac{1}{39.4 \times 4 \times 10^{12} \times 2 \times 10^{-10}}$$

$$= 3.17 \times 10^{-5} \text{ henries}$$

With the inductance known, the resistance may now be determined with the help of equation (34) as follows:

$$f_2 - f_1 = \frac{R}{2 \pi L}$$

so

$$R = 2 \pi L (f_2 - f_1)$$

$$= 6.28 \times 3.17 \times 10^{-5} \times 5 \times 10^3$$

hence

$$R = 1.0 \text{ ohm}$$

It should be observed from equation (34) that if R is larger than 1 ohm in this example, then $f_2 - f_1$, the bandwidth, will be greater than 5 kc as a consequence of the increase in the numerator of the right-hand member of the equation. With an increase in bandwidth comes a reduction in voltage amplification, however. This follows from equation (26):

$$\text{Gain} = 2 \pi \text{ fL } Q' \text{ } g_m$$

An increase in the resistance term reduces Q' which, in turn, causes the voltage gain to diminish as indicated in the equation above. Conversely, if R is smaller than 1 ohm, the gain may be somewhat greater, but the width of the band covered will be small.

10. The Gain of a Transformer-Coupled Amplifier

It is very often desirable to eliminate the coupling capacitor C_c and the grid resistor R_g in r-f amplifiers (Fig. 7). This is most easily done through transformer coupling, a typical circuit for which is shown in Fig. 8. The transformer in this type of coupling is generally of the air-core variety with both windings on the same cylindrical form. The degree of coupling, or mutual conductance, is a function of the spacing between primary and secondary windings, shape and proportions of the coils, and the relationship between core diameter and overall coil diameter.

Referring to the quivalent circuit, it is seen that the constant-current generator form is again used here because the tube is a pentode. The current from the generator ($i = -g_m e_g$) might be considered to divide between r_p and the series circuit containing the a-c resistance of the primary (R_p) and the primary winding itself. The impedance looking into the terminals of the coil is so much lower
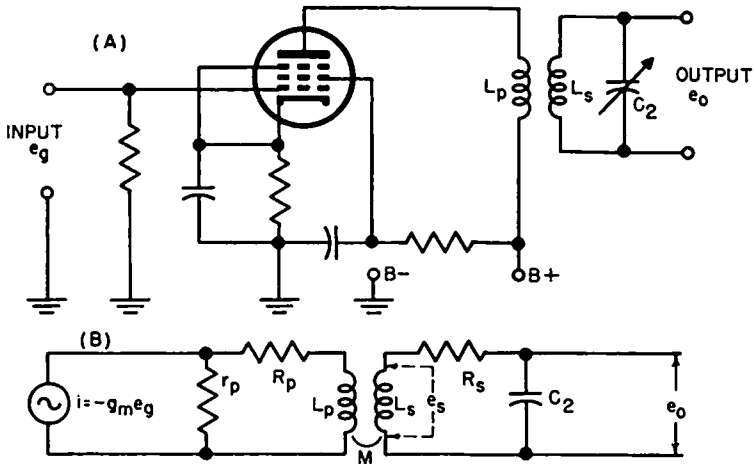
Fig. 8. (A) Schematic diagram of a pentode, transformer-coupled amplifier for high-frequency service. (B) The equivalent circuit of the transformer-coupled r-f service.

than the plate resistance of a pentode, however, that it can be safely assumed that all the current from the generator flows through the coil. Thus, using the general expression:

$$\text{Induced voltage} = 2 \pi \text{fMi} \qquad (38)$$

in which f is the frequency of the sinusoidal current, M the mutual inductance between coils expressed in henries, and i the current in the primary coil in amperes, we can state specifically for the case at hand that

$$e_s = -j \, (2 \pi f) \, Mi \qquad (39)$$

or

$$e_s = +j \, (2 \pi f) \, Mg_m e_g \qquad (40)$$

The sign reversal of the j operator results from the fact that the current from the generator is expressed as $-g_m e_g$.

The induced voltage e_s gives rise to a secondary current i_s. Assuming that the secondary circuit is in resonance, then the total impedance of this circuit is equal to the resistance R_s. Hence:

$$i_s = e_s/R_s \qquad (41)$$

Since the output voltage e_o is equal to the voltage that appears across the capacitor C_2 and since this capacitor has a reactance equal

to $1/2 \pi fC_2$ where f is the resonant frequency, we may write:

$$e_o = i_s \ (-j/2 \ \pi \ fC_2) \qquad (42)$$

Combining equations (40), (41), and (42), we can obtain:

$$e_o = -j \ \frac{1}{2 \ \pi \ fC_2R_s} \times j \ (2 \ \pi \ f) \ Mg_me_g \qquad (43)$$

The factor $1/2 \pi fC_2$ is the capacitive reactance of X_c of C_2. Assuming that a condition of resonance prevails, then the inductive reactance of the secondary coil L_8 must be equal to V_c. We may then replace X_c with X_L in equation (43).

$$e_o = \frac{2 \ \pi \ fL_s}{R_s} \times 2 \ \pi \ fMg_me_g \qquad (44)$$

but $2 \pi fL_s/R_s$ is the Q of the secondary winding, permitting us to rewrite equation (44) as:

$$e_o = 2 \ \pi \ fMg_me_gQ \qquad (45)$$

and since voltage gain is e_o/e_g, then

$$\text{Gain} = 2 \ \pi \ fMg_mQ \qquad (46)$$

Problem 5. A 6AU6 having a transconductance of approximately 4000 μmhos is to serve as an r-f amplifier at 1 mc. Its coupling transformer has a mutual inductance of 50 μh. For what value of Q must the resonant circuit be designed in order to realize a voltage gain of 125?

Solution. Solving equation (46) for Q, we have:

$$Q = \frac{\text{Gain}}{2 \ \pi \ fMg_m}$$

$$= \frac{125}{6.28 \times 10^6 \times 5 \times 10^{-5} \times 4 \times 10^{-3}}$$

$$= 100 \ \text{(approx.)}$$

11. Bandwidth of a Transformer-Coupled Amplifier

Comparing equations (46) and (26), it is evident that the voltage amplification of single-tuned impedance-coupled r-f amplifiers is quite similar to that of single-tuned transformer-coupled circuits. The principal difference between the two equations is the substitution of the mutual inductance M for the inductance L of the single-coil circuit. Since the only resistance present in the transformer-coupled type is the resistance of the secondary winding — usually

very small compared to inductive reactance — we have used Q rather than Q′ throughout the derivation of equation (46). When the same tuned circuit is used, first in the impedance-coupled arrangement and then in the transformer-coupled circuit, there is only one difference in performance: the available voltage gain in the transformer-coupled case is modified by the ratio M/L. When it is remembered that mutual inductance M is:

$$M = k \sqrt{L_p L_s} \qquad (47)$$

in which k is the coefficient of coupling, it is apparent that M may be considerably smaller than L, particularly in cases where the coefficient of coupling is small.

The foregoing statements imply that the bandwidth of a transformer-coupled amplifier may be found by the same expression employed for determining the bandwidth of impedance-coupled amplifiers. This is actually the case. Thus, the bandwidth of this type of amplifier may be found from the relation:

$$f_2 - f_1 = f_r/Q$$

12. Double-Tuned R-F Amplifiers

The general form of a double-tuned r-f amplifier involves a circuit somewhat like that shown in Fig. 9. Two circuits, resonant at the
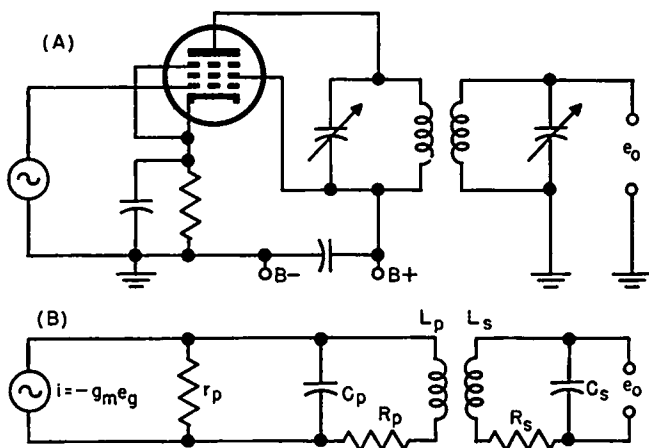
Fig. 9. (A) Double-tuned pentode amplifier, schematic diagram. (B) Double-tuned pentode amplifier, equivalent circuit.
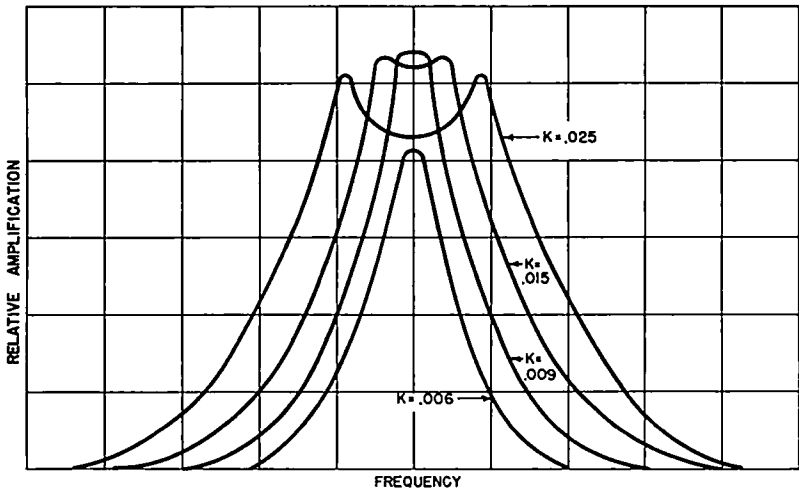
Fig. 10. Frequency-response curves for several values of k.

same frequency, are coupled inductively by the mutual inductance of their coils. Amplifiers of this type are most frequently found in equipment that is intended to handle a *range* of frequencies rather than a single frequency as in i-f amplifiers in radio receivers and television sets. The frequency response of a double-tuned amplifier (or a bandpass amplifier, as it is sometimes called) depends to a substantial degree on the closeness of coupling between the primary and secondary coils. From equation (47), we can obtain:

$$k = \frac{M}{\sqrt{L_p L_s}} \tag{48}$$

where M is the mutual inductance in henries and L_p and L_s are the inductances of primary and secondary windings, respectively; k is the coefficient of coupling.

When the coefficient of coupling is small, the resonant curve due to both tuned circuits is quite narrow and sharp as illustrated in Fig. 10. As the coupling is made closer, the response curves broaden and develop a dip in the vicinity of the resonant frequency. As the coupling is increased from its lowest value ($k = 0.006$, Fig. 10) up-ward, the output voltage across the secondary winding increases somewhat and the curve begins to broaden. At the instant when the resistance reflected into the primary winding by the secondary cir-cuit becomes equal to the actual primary resistance, the curve takes

the form shown for k = 0.009. The coupling required to produce this condition is known as *critical coupling;* in this state, the current in the secondary reaches its maximum possible value at resonance. Beyond the critical-coupling condition, the response curve becomes still broader with double humps, one on each side of the resonant frequency.

It is evident from inspection of the curves that the bandpass of an amplifier is controllable to a definite extent by varying the coupling between the tuned primary and secondary coils. When the coupling is loose (i.e., when k is small), the amplifier displays the greatest ability to discriminate between adjacent frequencies; at critical coupling, the gain is greatest and the selectivity only slightly poorer than before; with overcoupling, the bandpass increases but the amplification at resonance drops. For normal amplifiers in which the required bandpass is a small percentage of the resonant frequency — as in small radio receivers used on the broadcast band — critical coupling assures good gain and adequate selectivity. Should a wider bandpass be required, as in high-fidelity radio receivers, coupling somewhat greater than the critical value is generally employed.

The voltage amplification obtainable from a double-tuned transformer-coupled r-f amplifier is a function of the transconductance of the tube, the frequency at which it operates, the coefficient of coupling, and the inductances and Q's of the primary and secondary coils. The equation that relates these circuit values to amplification is:

$$A = \frac{2\pi f\ g_m k\ \sqrt{L_p L_s}}{k^2 + \dfrac{1}{Q_p Q_s}} \ ^* \tag{49}$$

It is interesting to note that this equation becomes far simpler for the condition of critical coupling, with identical circuits

$$A_c = \frac{2\pi f\ g_m L Q}{2} \tag{50}$$

in the primary and secondary portions of the amplifier. In equation (50), A_c is the amplification for critical coupling, f is the resonant frequency to which both primary and secondary circuits are tuned,

---

* The derivation of this equation is tedious and is not considered sufficiently important to include in a book of this scope.

g_m is the tube transconductance, L is the inductance of either the secondary or primary winding, and Q is the conventional figure of merit for either coil.

Problem 6. A bandpass amplifier is critically coupled and is to be used as an i-f amplifier at a frequency of 455 kc. The inductance of primary coil is 1.2 mh, the coil having a Q 120. The secondary coil has exactly the same inductance and Q, and the tube to be used is a 6BA6 with 100 volts on both its plate and screen grid. Find the voltage amplification.

Solution. The receiving-tube data book provides the information that the transconductance of a 6BA6 is 4300 μmhos under the given conditions of operation. Since the primary and secondary circuits are identical, the simplified equation [equation (50)] may be used to find the voltage gain.

$$A = \frac{6.28 \times 4.55 \times 10^5 \times 4300 \times 10^{-6} \times 1.2 \times 10^{-3} \times 120}{2}$$

$$= 3.14 \times 4.55 \times 4.3 \times 1.2 \times 1.2 \times 10$$

$$= 885$$

Bandwidth of double-tuned amplifiers — It will be recalled that the bandwidth of an amplifier is defined as the range of frequencies over which the voltage amplification does not drop below 70.7% of its maximum resonant value. For a single-tuned circuit, the bandwidth as given previously in equation (37) is:

$$f_2 - f_1 = \frac{f_r}{Q}$$

in which f_2 is the upper limit, f_1 is the lower limit, f_r is the resonant frequency, and Q is the actual figure of merit of the coil. For double-tuned circuits, the bandwidth may be shown to be:

$$f_2 - f_1 = \frac{\sqrt{2}\, f_r}{Q} \tag{51}$$

for the condition of critical coupling and coils that have identical Q's.

In practical circuits where slightly closer coupling is used, equation (51) is modified to read:

$$f_2 - f_1 = 1.5\, k\, f_r \tag{52}$$

where k is the actual coefficient of coupling and f_r is the resonant frequency.

Problem 7. What coefficient of coupling is required in a bandpass amplifier to provide a bandwidth of 7.5 kc at a frequency of 1.5 mc?

Solution. Equation (52) is solved for k, thus:

$$k = \frac{\text{bandwidth}}{1.5 \; f_r}$$

Substituting:

$$k = \frac{7.5}{1.5 \times 1500}$$

$$= 0.0033$$

Since this is relatively loose coupling, it is apparent that the required bandwidth in this case can be easily obtained. To determine how close to critical coupling this is, we might investigate the order of magnitude of critical coupling for a pair of typical tuned circuits.

If equation (49) is differentiated* and set equal to zero to determine the k required for maximum amplification, it is found that:

$$k_{cc} = 1/ \sqrt{Q_p Q_s} \qquad (53)$$

where k_{cc} is the coefficient of critical coupling (since this coupling provides maximum amplification) and Q_p and Q_s are the Q's of the primary and secondary circuits, respectively. Assuming that typical tuned circuits at a frequency of 1.5 mc have Q's in the order of 100, let us solve equation (53) for the coefficient of critical coupling for these values:

$$k_{cc} = \frac{1}{\sqrt{100 \times 100}}$$

$$= \frac{1}{100}$$

$$= 0.01$$

From this it is seen that the coupling obtained in the solution to Problem 7 is far looser than critical. Similarly, if we now substitute the value for critical coupling in equation (52), we can determine what the bandwidth would be under these circumstances.

$$\text{Bandwidth} = 1.5 \times 0.01 \times 1500 \text{ kc}$$

$$= 22.5 \text{ kc}$$

---

* Derivations requiring the differential or integral calculus are not included. Those interested in such derivations may find them in any standard work on electrical engineering.

Several conclusions may be drawn from these results. The example used above is that of an i-f amplifier one might find in a radio receiver. In such an application, it is important to design for as much gain as possible and for good selectivity; at the same time, the band-pass must be sufficient to provide approximately equal amplification for all frequencies that lie in the modulation sidebands. Assuming that an overall i-f bandwidth of about 5 kc would offer the necessary selectivity and uniform amplification, then we might conclude that:

(a) loose coupling, such as that obtained with k = 0.003, provides the right bandwidth and hence the desired selectivity but that the gain may suffer intolerably. (See the response curve for loose coupling in Fig. 10.)

(b) critical coupling results in maximum gain but produces an excessive bandwidth which might cause the selectivity to degenerate severely.

These conclusions explain why i-f amplifier systems in even the smallest radio receivers utilize at least two sets of double-tuned transformers. Adding cascaded resonant systems results in gain multiplication as well as sharpening the response curve (i.e., reducing the overall bandwidth). Thus, we might start with loose coupling and a bandwidth slightly greater than necessary; with the addition of a second stage, the gain picks up and the bandwidth narrows. By proper design, two or more stages, coupled properly, can yield almost any resonable combination of gain and bandwidth for radio reception in normal areas. Experience indicates that the overall bandwidth of two stages of double-tuned transformer-coupled amplifiers is about 80% of that of a single stage; three stages display a bandwidth about 71% of that of a single stage.

13. Stagger-Tuned Bandpass Amplifiers

The requirements of very-wide-band amplifiers cannot be met by successive stages of r-f amplifiers, all tuned to the same frequency, unless steps are taken to lower the Q of each tuned stage. For example, it is seen from equation (37) that the bandwidth is inversely proportional to Q for single-tuned r-f amplifiers; similarly, equation (51) indicates that the same relation holds for double-tuned circuits. Reduction of Q may be accomplished by connecting resistors of suitable value across both primary and secondary coils in double-

tuned systems. As might be expected, this results in a serious loss of gain which must be compensated for by adding more stages of amplification.

Instead of reducing Q intentionally, wide-band performance may be realized by the process of *stagger-tuning;* although this method also requires additional cascaded stages to bring the gain up to the necessary level, a properly adjusted staggered system is definitely superior to a resistance-loaded arrangement with respect to system gain and bandwidth. For this reason, stagger-tuning is almost uniformly encountered in the i-f amplifier systems of television receivers in which bandwidths of 4.0 mc or more are desired.

When the primary and secondary circuits of a double-tuned r-f transformer are slightly detuned — one higher and the other lower in frequency than the central or resonant point — the change in the frequency-response curve of the secondary appears as though an increase in coupling has occurred. That is, detuning has the same effect as closer coupling. The new (or equivalent) coupling coefficient for the case where the two circuits are identical in all respects except the frequency to which they are tuned is given by the equation:

$$k^1 = \sqrt{k^2 + \left(\frac{f_2 - f_1}{f_c}\right)^2} \qquad (54)$$

in which k is the actual coupling coefficient, f_2 is the higher resonant frequency, f_1 the lower frequency, and f_c the frequency midway between f_2 and f_1.

Let us assume that we have an amplifier which yields a response curve like that of curve A in Fig. 11 when it is double-tuned, with both primary and secondary resonant at the same frequency. Such a curve assumes specific values of Q for each of the windings of the coupling transformer. If the primary of the transformer is now detuned to f_2 and the secondary to f_1, the output voltage of the amplifier will drop but the bandwidth will increase so that the response curve has the shape shown in Fig. 11B. The loss of gain is due to detuning while the increase in bandwidth results from the tighter equivalent coupling coefficient as given in equation (54). Now, however, it is possible to adjust the circuit Q's and the amount of detuning so that the overall gain rises to the same point as that of the double-tuned curve (curve C, Fig. 11) *without an equivalent loss of bandwidth.* The resulting response curve has a bandwidth which
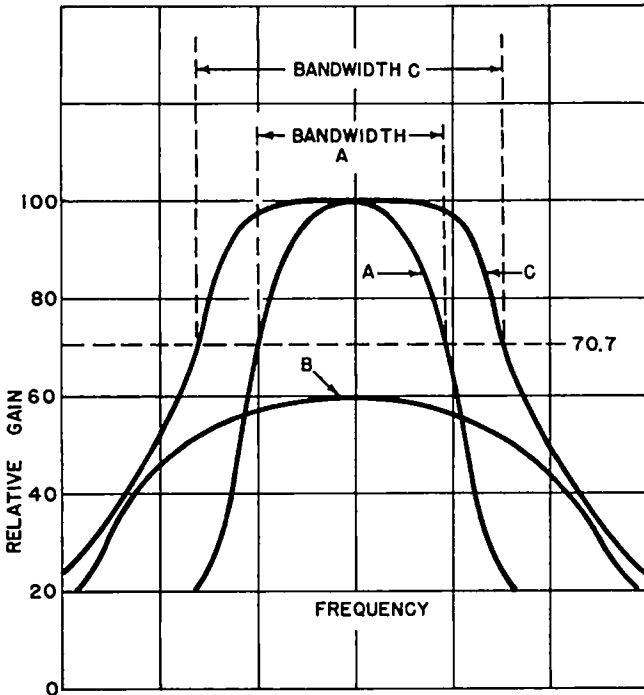
Fig. 11. Comparison of stagger tuning with double-tuned circuit.

is approximately 1.8 times greater than that of the double-tuned arrangement as shown in Fig. 11A. Thus, a staggered pair performs almost as well as two double-tuned stages with respect to voltage gain and, at the same time, can be given greatly increased bandwidth.

14.   Preventing Oscillation in R-F Amplifiers

In a carefully wired triode amplifier circuit, sustained oscillation may still occur due to plate-to-grid positive feedback if the circuit is adjusted for even reasonable gain. Such feedback is an inherent property of triode circuits and cannot be eliminated without incorporating some form of neutralizing device. In general, regeneration in receiver r-f amplifiers is now controlled by employing pentodes rather than triodes thereby making possible far greater voltage gains without instability.

Several other sources of regenerative feedback may exist in r-f voltage amplifiers using pentodes. Many of these are obscure and

difficult to locate so that it is impossible to prescribe definite procedures for eliminating them. Each unstable amplifier is an individual problem in itself. There are some general rules-of-thumb, however, that are often helpful in eliminating such sources of feedback during design and construction of equipment.

(a) Impedances common to two or more stages almost always exist in high-frequency multistage amplifiers. The effects of these may be minimized by avoiding the use of the chassis as a common ground and by returning sensitive grounds to the same point in each case. Carefully designed and installed decoupling filters often prevent common-impedance regeneration from occurring.

(b) Bypass capacitors should always be selected for the lowest possible value of inherent inductance.

(c) Regeneration resulting from the interaction of stray electrostatic and electromagnetic fields may be minimized by careful orientation of the components and by means of adequate shielding.

15.   Review Questions

1. Define the bandwidth of an amplifier. Upon what factors does the bandwidth of an amplifier depend?
2. Why is the bandwidth of an amplifier in accordance with the definition you have given often referred to as "the half-power bandwidth"?
3. Explain why a constant-current generator form of equivalent circuit is better suited for the analysis of pentode action than the constant-voltage type?
4. A tuning coil has an a-c resistance of 3 ohms and an inductance of 40 mh. Predict the approximate bandwidth of this coil when set up in a resonant circuit having a reasonably high Q.
5. What are the advantages of double-tuned amplifier circuits as contrasted with the single-tuned variety?
6. Define critical coupling. What are the principal characteristics of a critically coupled amplifier?
7. What bandwidth may be expected from a double-tuned amplifier circuit containing two identical coils, each having a Q of 120 at a resonant frequency of 3.5 mc? (Assume critical coupling.)
8. A given double-tuned resonant circuit consists of two coils, one with a Q of 100 and the other with a Q of 120. Determine the coefficient of critical coupling for this circuit.
9. Explain why most radio i-f amplifiers use at least two sets of double-tuned transformers.
10. Describe what is meant by stagger-tuning. What are its significant advantages?

Chapter 3

PRINCIPLES OF R-F POWER AMPLIFIERS

16. General Considerations

It was mentioned at the beginning of Chapter 2 that r-f voltage amplifiers are usually operated class-A since the principal characteristic of interest is that of high voltage gain rather than high plate efficiency. In the case of power amplifiers, however, voltage gain is relatively unimportant; good design and sensible economy demand operating conditions that make the most of the d-c power supplied by the source. That is, to permit the use of tubes of reasonable size for a given power, the d-c to r-f conversion efficiency must be high. Thus, normal transmitter power amplifiers are operated class-C except in those cases where the amplifier follows a modulated stage. In this connection, class-B linear operation is called for.

The high efficiency of class-C (60% or better) makes its use almost mandatory where high power output is required. Since high-frequency amplifiers operate over a very narrow range of frequencies relative to the carrier frequency, resonant grid and plate circuits may be used. These act as filters and deliver virtually perfect sinusoidal current and voltage although the plate waveform is severely distorted by class-C operation. The grid-bias for class-C operation varies between 1.5 and 3 times cutoff and in most instances is supplied by a separate rectifier-filter power supply. Bias voltage may also be obtained by a series grid resistor or grid leak. Since the grid is driven positive during part of each input cycle, the flow of grid current can

be used to develop the required bias voltage across the resistor. In this system of bias, however, the tube and its associated components are in constant danger of burnout should the driving r-f voltage disappear since the bias voltage would also vanish and excessive plate current would flow. As a preventive measure, a combination of grid-leak and fixed bias is quite common.

17. Voltage and Current Relations

Reference will be made to Fig. 12 throughout this discussion. The circuit diagram in (A) is not intended to represent a practical amplifier but has been drawn in simplified form to facilitate analysis.
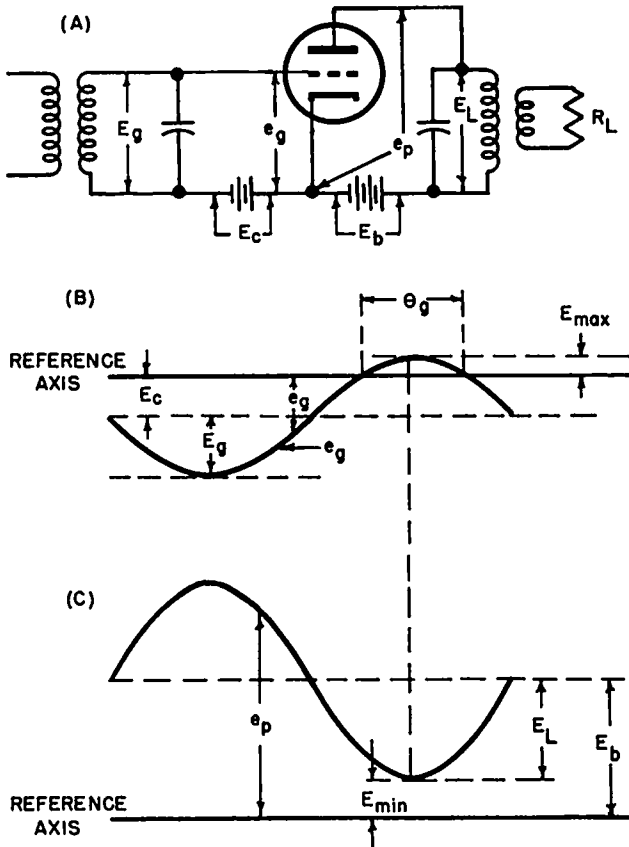
Fig. 12. Electrical relations in a class-C power amplifier.

Grid voltage — The driving voltage applied to the grid of the triode is a combination of the fixed bias voltage E_c and the alternating signal voltage from the preceding stage E_g. The resultant grid voltage is symbolized by e_g. The waveforms are given in Fig. 12B. Observe that E_c, e_g, and E_{max} are measured from the zero or reference
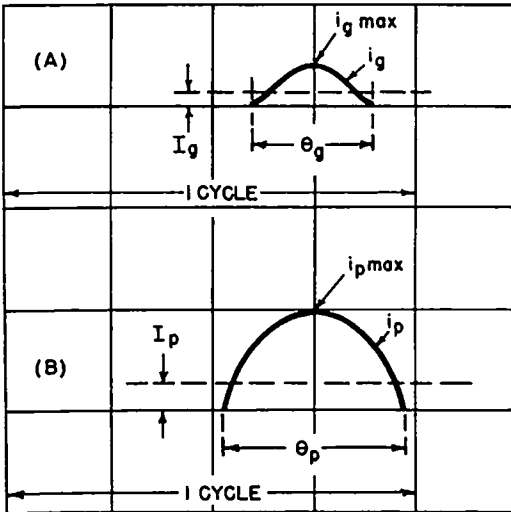
Fig. 13. Plate and grid current waveforms in a class-C power amplifier.

axis (cathode potential in this case), while E_g — the signal voltage — is determined by measuring the amount by which it displaces the bias voltage. Θ_g represents the *angle of grid current flow* (in this case approximately 90°); this, of course, merely means that the grid current flows for approximately one-quarter of the input cycle.

Plate voltage — The pulsating plate voltage is the resultant of the d-c supply voltage and the voltage drop that appears across the plate load. That is, the instantaneous magnitude of e_p is a function of E_b and the instantaneous value of E_L. The waveform relations are shown in Fig. 12C. It is important to note the phase relationship between the grid and plate voltage curves: the maximum instantaneous grid voltage (E_{max}) is in phase with the minimum instantaneous plate voltage (E_{min}).

Grid current — Grid current flow occurs only when the grid is driven positive by the signal voltage. Its angle of flow (or fraction of the input cycle over which flow takes place) is given by Θ_g (Fig. 13B).

Plate current — The fluctuating plate current in a class-C amplifier has an instantaneous value that depends upon *both* the instantaneous grid voltage and the instantaneous plate voltage. In order to determine i_p at any instant, it is therefore necessary to refer to the characteristic curves of the particular tube, locating i_p for given values of e_g and e_p. It will be found from such an analysis that the plate current flows in the form of a short pulse for somewhat less than half a cycle and *that the maximum instantaneous plate current flows at the time when the instantaneous plate voltage is at its lowest point (E_{min})*. The reason for this is clear when it is remembered that the phase of maximum positive grid voltage E_{max} is the same as that of minimum plate voltage E_{min}. Due to the amplification factor of the tube, the grid has much more influence over the plate current than the plate voltage has so that maximum plate current must flow for E_{max} even though the plate voltage at the same instant is at its lowest value. This is a very important consideration with respect to plate efficiency as will be demonstrated later.

Average plate and grid currents — The average values of plate and grid current (I_p and I_g) are also indicated in the curves of Fig. 13. These are the currents that would be read by d-c milliammeters connected in series with the plate-supply lead and grid-bias lead, respectively. The average grid current I_g (as well as the instantaneous magnitude i_g) is determined by both the d-c grid bias and the signal driving voltage; likewise, I_p is controlled by several factors including the grid bias, the driving voltage, the plate-supply potential, and the size of the plate-load impedance. With all other circuit parameters fixed, the load impedance may be adjusted so that I_p is the desired figure. This is generally accomplished by varying the coefficient of coupling between the load and the output circuit.

18. Required Driving Power

The instantaneous grid power needed to drive a class-C amplifier is defined as the product of the signal voltage (E_g) and the grid current (i_g), all being instantaneous values. That is:

$$\text{Driving power} = E_g i_g \qquad (55)$$

The instantaneous grid voltage on the driven tube, however, is the algebraic sum of the instantaneous driving voltage and the d-c grid bias, or:

$$e_g = E_g + (-E_c) \qquad (56)$$

since E_c is always negative. Thus:

$$E_g = e_g + E_c \qquad (57)$$

Substituting the value of E_g from equation (57) into equation (55) we have:

$$\text{Driving power} = (e_g + E_c) \; i_g \qquad (58)$$

Expanding:

$$\text{Driving power} = e_g i_g + E_c i_g \qquad (59)$$

The first term on the right side of equation (59) represents the power dissipated by the grid of the tube or the grid power loss; the second term is the power dissipated in the bias source. The significance of this equation thus becomes clear. The driving source must supply sufficient power to supply *both* the grid losses and the losses in the bias component or components. The equation may also be justified by observations in the operation of transmitters: if a battery is used as a bias supply, the battery voltage will slowly rise over long periods of transmitter operation. Although the chemistry of dry batteries does not include rechargeability, power absorption does result in voltage increase. The necessary power must, therefore, come from the driving source.

19. Power and Efficiency

It was emphasized in Section 17 that the maximum instantaneous plate current i_p flows when the instantaneous plate voltage is minimum (E_{min}). Since the instantaneous plate voltage is the difference between the supply voltage E_b and the instantaneous voltage drop across the load (E_L), then:

$$e_p = E_b - E_L \qquad (60)$$

or

$$E_L = E_b - e_p \qquad (61)$$

From equation (61) it is evident that the useful output voltage across the load grows as the voltage drop within the tube diminishes. For an instantaneous plate voltage equal to zero, all the supply voltage will appear as useful output voltage and $E_L = E_b$. This is

a condition of 100% efficiency and cannot, of course, be realized in practice.

This development does demonstrate, however, why the efficiency of a class-C amplifier is high, generally ranging upward from 60% to as high as 85%. *The power source supplies energy to the system largely when most of this energy will be absorbed by the tuned circuit and hence appear as useful power.*

The high efficiency of class-C amplifiers may also be said to be dependent upon the angle of plate-current flow Θ_p. If the plate current flows for a very short interval and if the center of this interval is in the same phase as minimum plate voltage, the power dissipation in the plate-cathode circuit of the tube will be very small. As the flow interval lengthens, more and more power appears as heat in the plate circuit because, now for a larger part of the time, plate current is flowing while the plate voltage is above its minimum value. This indicates that the design of a class-C power amplifier calls for as small an angle Θ_p as can be obtained, i.e., for extremely large values of grid bias. In contrast to this, a small Θ_p means, in turn, low power input and that under these conditions the power output will be correspondingly small even though the efficiency is high. It is found in practice that a good compromise is obtained by arranging for Θ_p of the order of 130° with an efficiency in the vicinity of 70% to 80%.

The foregoing analysis also suggests that conditions be arranged so that E_{min} is as small as possible. A low value of E_{min} encourages high-efficiency operation. Here again a compromise must be drawn. If the minimum plate voltage is permitted to drop below the maximum grid voltage E_{max}, secondary emission effects become dangerous because the grid will tend to attract secondary electrons from the plate. This will result in excessive grid current which reduces operating efficiency and increases the demand for driving power. To a great extent, the magnitude of E_{max} finally selected will depend upon how much driving power is available. If the drive is low, then E_{min} must be made appreciably greater than E_{max} in order to reduce both grid current and the demand for driving power.

20. Class-C Power Amplifier Operating Equations

Although there are literally dozens of equations that may be successfully applied to the solution of class-C power amplifier design

problems, relatively few of these are consistently encountered in practical work. The most important ones are summarized below.

Power output — The power output is defined as the product of the effective (rms) output voltage developed across the plate load and the effective output current. That is:

$$\text{Power output} = E_L(\text{eff}) \times I_p(\text{eff}) \qquad (62)$$

It is often more convenient to compute the power output from the peak values of output voltage and current. Since effective values are found by dividing the peak values by $\sqrt{2}$, then:

$$\text{Power output} = \frac{E_L(\text{peak}) \times I_p(\text{peak})}{2} \qquad (63)$$

Plate efficiency — Plate efficiency is defined as the ratio of a-c power output to the product of the average values of plate voltage and plate current at maximum signal input.

$$\text{Plate efficiency} = \frac{\text{Power output}}{E_p(\text{av}) \times I_p(\text{av})} \qquad (64)$$

*Average plate current** — Average plate current is found from:

$$\text{Average plate current }(I_p) = 0.27 \times i_p(\text{max}) \qquad (65)$$

*Peak output a-c current** — Peak output a-c current flowing in the plate circuit of the tube may be determined from equation (66)

$$\text{Peak output a-c current} = 0.46 \times i_p(\text{max}) \qquad (66)$$

To illustrate the application of these equations, consider the following example:

Problem 8. A tube having the characteristic curves shown in Fig. 14 is to be used as a class-C power amplifier in a radio transmitter. The d-c plate supply voltage is 1250 volts, and the average plate current is to be 0.175 ampere. Find the power output, load resistance, plate efficiency, grid bias, and excitation voltage required for full output. $E_g = E_p$ is a graph of the coordinates for which the plate voltage is equal to the grid voltage.

Solution. The peak plate current is first found from the specified average plate current given in the problem, using equation (65):

$$i_p(\text{max}) = I_p/0.27$$

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* These values are based upon the assumption that the plate-current angle is about 150°, or somewhat less than half a cycle; this is normal for class-C amplifiers.
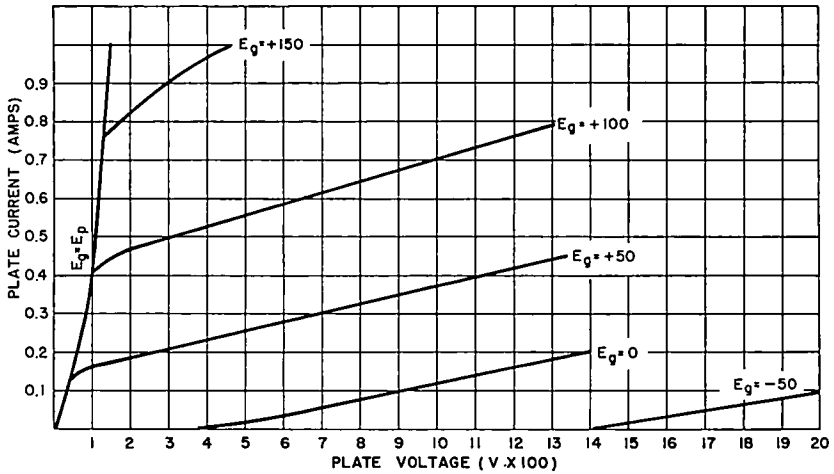
Fig. 14. Characteristic curve of a typical transmitting tube used on a shipboard transmitter. The line labeled $E_g = E_p$ is a graph showing where the plate voltage is equal to the grid voltage.

$$i_p \text{ (max)} = 0.175/0.27 = 0.648 \text{ ampere}$$

Since this plate current is assumed to be obtained at the instant when the plate voltage and the positive grid voltage are equal, we refer to the $E_g = E_p$ line in Fig. 14 and find that this condition occurs for:

$$E_{max} = 120 \text{ volts}$$

$$E_{min} = 120 \text{ volts}$$

The output current is obtained from the peak value of plate current obtained above using equation (66) or:

$$\text{Peak output a-c current} = 0.46 \times i_p \text{ (max)}$$

$$= 0.46 \times 0.648$$

$$= 0.298 \text{ ampere}$$

Since it is customary to give information in terms of rms values rather than peak, this figure may be converted to rms in the usual manner. That is:

$$\text{rms current} = 0.707 \times 0.298$$

$$= 0.211 \text{ ampere}$$

The output voltage is the difference between the d-c plate voltage and the minimum instantaneous plate voltage. Thus

$$\text{Output voltage} = 1250 - 120$$

$$= 1130 \text{ volts peak}$$

or

$$\text{rms output voltage} = 1130 \times 0.707$$

$$= 799 \text{ volts}$$

Power output is determined from equation (63) as follows:

$$\text{Power output} = \frac{E_L \ (\text{peak}) \ \times \ I_p \ (\text{peak})}{2}$$

$$= \frac{1130 \times 0.298}{2}$$

$$= 169.4 \text{ watts (approx. 170 watts)}$$

The plate efficiency is obtained by using equation (64)

$$\text{Plate efficiency} = \frac{\text{Power output}}{E_p \ (\text{av}) \ \times \ I_p \ (\text{av})}$$

$$= \frac{170}{1250 \times 0.175}$$

$$= 77.7\%$$

The necessary load, expressed as a resistance, may be found by dividing the output voltage by the output current, thus:

$$\text{Load (ohms)} = 1130/0.298$$

$$= 3500 \text{ ohms (approx.)}$$

Although the grid bias is not very critical for a class-C amplifier, it should be at least twice the cutoff value for the tube or even slightly greater. Referring to the curves in Fig. 14 and interpolating between the appropriate curves, it is seen that the cutoff bias for a plate potential of 1250 volts is approximately −40 volts. Hence, a bias voltage of from −80 to −100 volts would be suitable. Assuming that a bias voltage of −100 volts is used, the peak excitation voltage from the driver stage may be computed as follows: the peak excitation voltage must exceed the grid bias by the required peak positive-grid voltage. Thus:

$$\text{Peak excitation voltage} = 100 \text{ volts} + 120 \text{ volts}$$

$$= 220 \text{ volts (peak)}$$

21. Biasing Methods

Tubes operated in class A or B require a bias source that is completely independent of the exciting voltage. This is so because the excitation may be continually changing. In the case of class-C amplifiers, however, bias may be obtained directly from the exciting source since this voltage is fixed and presumed to be constant.

Figure 15 illustrates a simple but entirely effective method for obtaining grid bias from the excitation voltage. Since it is expected

that the stage that drives the class-C amplifier will provide a suffici-
ently wide voltage swing to drive the grid positive during part of
the cycle, the grid will draw current. Rectified grid current flowing
through R_g will therefore produce a voltage drop having such a
polarity as to make the grid negative with respect to the cathode (or
filament) of the class-C tube. A radio-frequency choke (rfc) is

Fig. 15. Grid leak bias
applied to a class-C
power amplifier.
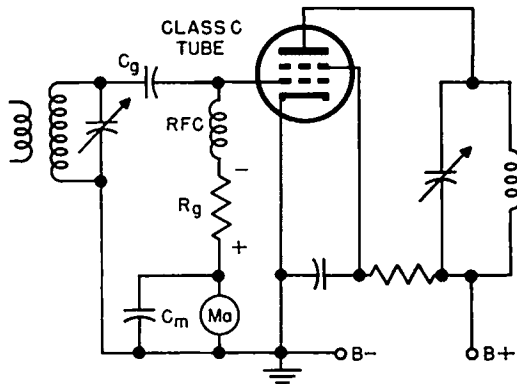
generally placed in series with R_g to prevent the dissipation of ap-
preciable amounts of r-f energy in the resistor. In practical trans-
mitter amplifiers of this type, it is common to find a milliammeter
in series with the rfc-R_g combination. This meter gives an excellent
indication of the drive being supplied to the amplifier and may also
be used as an aid in neutralizing triodes. (Neutralization is dis-
cussed in a later section.) C_m bypasses the milliammeter to prevent
the passage of stray r-f current that may be present through the
meter coil or shunt.

A serious disadvantage of using pure resistive bias in the fashion
shown in Fig. 15 is that, if for any reason the excitation should
vanish, the bias would be lost and the plate current of the tube
might rise to a dangerous figure. To prevent this, combination bias
is often used (Fig. 16). In this system, a battery or separate power
supply provides sufficient bias to keep the plate current at a safe
value should the excitation fail. It has already been mentioned that
the direction of flow of grid current is such as to charge the battery;
that is, the same amount of power is delivered to the bias source
whether that source is a resistor, a battery, or a separate power
supply.
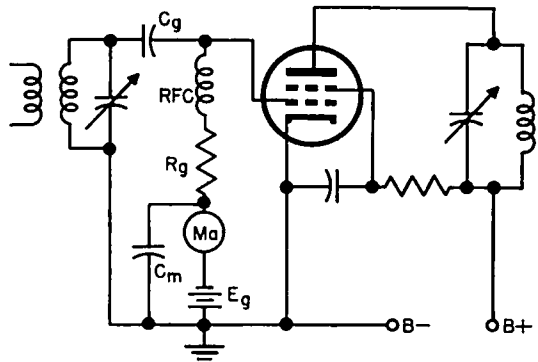
Fig. 16. A class-C amplifier supplied with combination fixed and grid leak bias.

Many lower power class-C amplifiers utilize another form of combination bias commonly termed "safety" bias. In this arrangement, the fixed bias battery or supply is replaced by a cathode resistor just large enough to allow safe plate current to flow should the excitation disappear (Fig. 17). The reason for selecting a minimum-value resistor in this circuit is obvious when it is recognized that the bias voltage drop across R_k represents an equivalent loss of plate-to-cathode voltage. For example, if the bias voltage required for a 1250-volt tube is in the vicinity of −100 volts and if all of it is provided by a cathode resistor, the plate-voltage loss is close to 10%. By obtaining most of the required bias from the grid resistor R_g and only enough for "safety" from R_k, the plate voltage loss may be reduced materially.

22. Grid Excitation of Class-C Amplifiers

Class-A and B amplifiers are normally operated with the restriction that the output voltage vary in direct proportion to the excitation
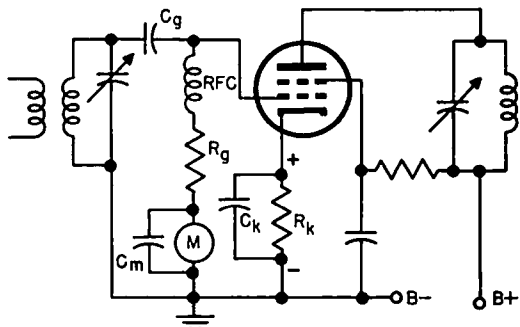
Fig. 17. A class-C amplifier provided with "safety" bias.

voltage. Since this restriction does not exist in class-C operation, it is informative to investigate the effect of increasing the excitation to higher values.

In Fig. 18, the variation of power output of a transmitting triode tube is plotted against the d-c grid current of the same tube. As the excitation voltage is gradually increased as evidenced by the rising d-c grid current, the power output increments are virtually linear, until point A is reached. This is the point where the minimum in-
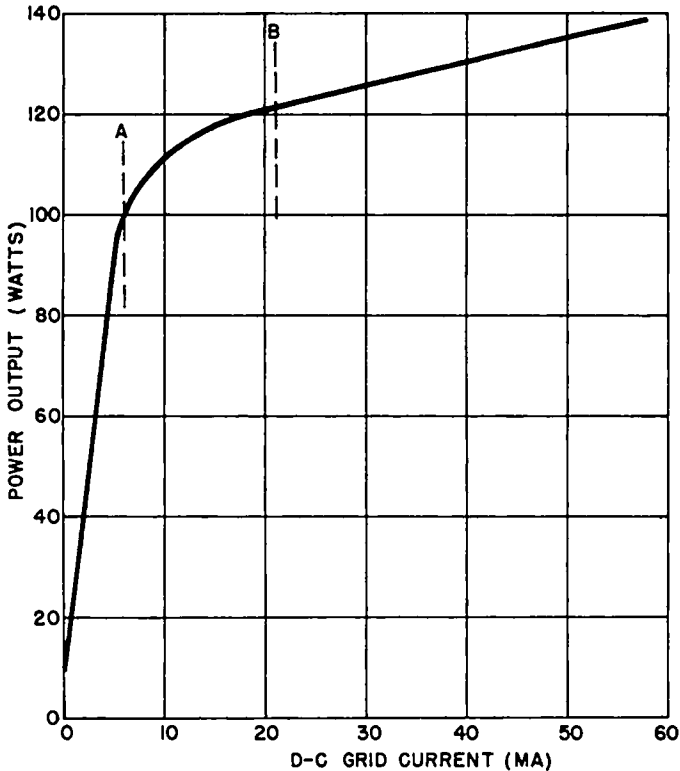
Fig. 18. Generalized curve output power versus d-c grid current for a transmitting triode.

stantaneous plate voltage (E_{min}) is about twice as great as the peak positive grid voltage (E_{max}). The grid current — hence the driving power — is still quite low at point A. Beyond point A, the rise in output power takes place more slowly. Point B on the curve is often called the "saturation" point because of the strong resemblance this

bears for the magnetic saturation curve of iron. At this point, the minimum instantaneous plate voltage is approximately equal to the maximum instantaneous grid voltage, making the grid current and driving power reasonably high as compared with the conditions for point A. This is also the point where the plate efficiency attains its maximum value (about 72% in this case).

With further excitation (i.e., beyond point B), the tube is said to be oversaturated. As the peak output voltage approaches the supply voltage as a limit, the grid current and driving power become very high. Although it might be expected that the plate efficiency would also rise in proportion, this does not occur because, in the over-saturated condition, the plate current waveform is seriously changed preventing further significant increases in efficiency.

A summary of these conditions is shown in Fig. 19. Each section represents a single r-f cycle. For the underdrive condition, the excitation is low, the tube does not reach saturation, and the peak
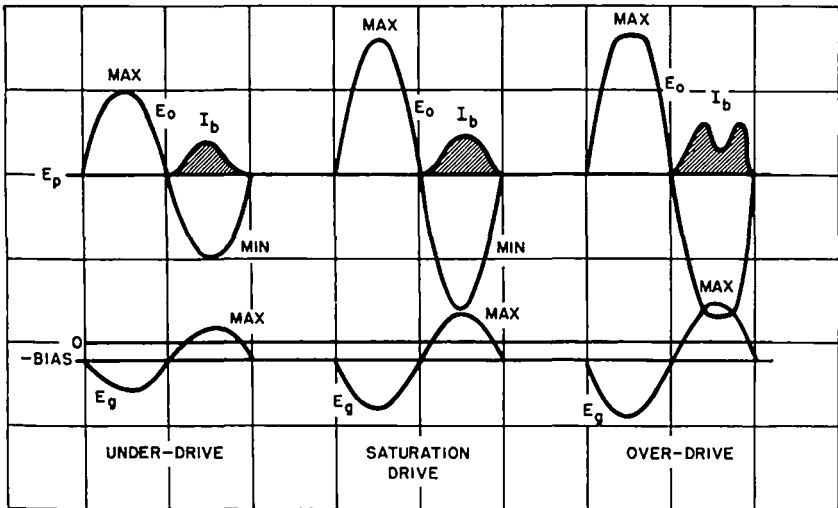
Fig. 19. Voltage and current waveforms in a class-C amplifier for conditions of underdrive, saturation drive, and oversaturation.

output voltage is considerably lower than the plate voltage. Thus, the plate voltage could be raised or diminished appreciably without having significant effect upon the output voltage. In the condition of saturation drive, the maximum instantaneous grid drive voltage attains the same value as the minimum plate voltage, and the peak

output voltage has begun to approach the supply voltage. In this case, a reduction of plate voltage does affect the output because it would then limit the output voltage; the output will not go up appreciably if the plate voltage is raised. When the excitation is finally increased to the overdrive condition, the output voltage does not change materially, but the shape of the plate-current wave is quite different. During the period where the positive excitation voltage exceeds the plate voltage, the plate current takes a sudden dip. This is the effect that limits the output voltage. Should the plate voltage fall below the grid voltage for too long a time, the output current will not be large enough to maintain the output voltage. It should be remembered that the output voltage and current bear a fixed ratio to each other as determined by the effective value of the load impedance. Thus, the output voltage is automatically limited by this effect.

In practice, the proper balance is reached when the peak output voltage is approximately equal to the plate voltage. This is the fundamental reasoning which lies behind the design of plate modulation systems.

23. Review Questions

1. Why does the maximum instantaneous plate current in a class-C amplifier flow at the instant when the instantaneous plate voltage is at its lowest point?
2. Explain why a class-C amplifier operates at a higher efficiency than a class-B or class-A amplifier.
3. When a battery is used as a bias source for a class-C amplifier for a long period of time, its voltage is found to be substantially greater than its original rating. Why does this happen?
4. How is the efficiency of a class-C amplifier related to the angle of flow of the tube's plate current?
5. Assume that you have a class-C amplifier in which the same tube is used as in Problem 8, having the characteristic curves given in Fig. 14. If the plate-voltage supply provides 1000 volts and the resulting plate current is to be 150 ma, determine the bias and excitation voltage required. Find also the power output, load resistance, plate efficiency under full output conditions.
6. What are the advantages and disadvantages of grid leak bias?
7. What are the advantages and disadvantages of battery bias? What significant loss may be attributed to cathode bias?
8. Draw an amplifier circuit in which only battery bias and cathode bias are used. Would such a circuit be likely to be used? Why?
9. Explain what is meant by underdrive, saturation drive, and overdrive.
10. Describe the waveform changes that occur under conditons of overdrive.

OPERATION OF CLASS-C POWER AMPLIFIERS

24. Functions of the Plate-Tank Circuit

The resonant circuit that forms the plate impedance of a class-C amplifier has several tasks to perform besides that of presenting the proper impedance to the plate of the tube for class-C operation. It must be so designed that most of the power delivered to it in the form of r-f current pulses from the tube is transferred to the load; the latter may be a subsequent class-C amplifier, a class-B linear amplifier, or an antenna. Although a certain amount of power loss must be anticipated in the tank circuit, it must be kept as low as possible for reasonable efficiency.

Another important function of the plate-tank circuit is that of discrimination against harmonics. Since the plate-current pulses may be quite distorted and nonsinusoidal, harmonics of varying strengths are always present. The Q of the tank circuit must be made high enough to prevent radiation of harmonics to an excessive extent, or to transfer high-amplitude harmonics to the input to the next stage.

25. Tank-Circuit Efficiency

Tank-circuit efficiency is defined as the ratio of the power delivered to the load to the total power transferred from the tube to the tank circuit. That is:

$$TC \; Eff \; = \; \frac{P_L}{P_t} \; \times \; 100 \qquad (67)$$

in which P_L is power in the load and P_t is power delivered to the tank circuit by the plate of the tube. In terms of circuit Q's, tank-circuit efficiency may be determined from the expression:

$$TC \; Eff \; = \; 1 \; - \; \frac{Q'}{Q} \qquad (68)$$

where Q is the *unloaded* figure of merit of the tank circuit and Q' is the Q of the tank circuit when fully loaded by coupling to a following amplifier stage or to a transmitting antenna.

It follows from an examination of equation (68) that Q' should be as low as possible and Q as high as possible to obtain good efficiency. The extent to which Q' can be reduced is dependent upon two conflicting factors: (1) a low Q' leads to increased efficiency and (2) a low Q' does not discriminate against harmonic transfer as well as a high figure. As always in cases such as this, a compromise is adopted which provides for reasonably good tank-circuit efficiency and, at the same time, assures adequate harmonic suppression. The circuit designer who seeks high efficiency at the expense of having to employ auxiliary harmonic suppression circuits may use a Q' as low as 5 or 6; for good suppression and somewhat lower efficiency, Q' values as high as 20 are often employed.

26. Importance of L-C Ratio in the Tank Circuit

It is possible to establish resonance at a given frequency with almost any reasonable combination of L and C components, but the very nature of the circuits suggests that there is, perhaps, an optimum ratio or range of ratios of L to C. Tank circuits using very high inductance and small capacitance values present to the tube a high-impedance load at both the grid input frequency and the harmonic frequencies. This causes the amplification of the harmonics to increase and the waveform of the output signal to become quite distorted. Although the output power decreases somewhat as a consequence, the input power decreases even more, resulting in a rise of plate-circuit efficiency. A very high L-C ratio has the disadvantage of having high r-f voltages across the tank circuit, necessitating widely spaced tuning capacitor plates and careful insulation; furthermore, the inability of a high-L, low-C tank to suppress harmonics often

requires the incorporation of special circuits for this purpose. When one also considers that a very-high-efficiency stage is not ordinarily suitable for modulation, it can be seen that optimum L-C ratio is not the highest possible L-C ratio.

27.   Factors in Tank-Circuit Design

From the preceding section, it is clear that the required performance of a class-C amplifier may be obtained only by selecting suitable values for L, C, and R_{ac}. (R_{ac} is the a-c resistance of the tank coil as used in previous chapters.)

In beginning the design of a tank circuit, it is assumed that these critical factors are known: the tube to be used and its characteristics, the supply voltage for the plate circuit of the tube, the driving voltage and power, and maximum instantaneous voltage that will develop across the tank circuit during operation. (The last quantity would be obtained from considerations presented in the last chapter.) For these conditions, it may be shown that:

$$2 \pi fL = \frac{E_o{}^2}{PQ'} \tag{69}$$

where f is the resonant frequency, L is the inductance of the tank coil, E_o is the output r-f voltage across the tank circuit, P is the power that is to be delivered to the tank circuit by the tube, and Q' is the loaded Q of the tank circuit, as before. All factors except L are known, provided that a value for Q' is selected as explained previously. Once L has been determined from this equation, the required capacitance may then be computed easily by applying equation (19).

Another relation that is often easier to use and is approximately correct is based upon the fact that the reactance of the coil is materially greater than R_{ac} — which, of course, is almost invariably the case.

$$X_L = \sqrt{R_L R_{ac}} \tag{70}$$

where X_L is the inductive reactance of the tank coil (and the capacitive reactance of the tuning capacitor at resonance), R_L is the load resistance obtained by dividing the output voltage by the output current. (See Problem 8.)

As was shown in Problem 8, the load presented by the sample circuit used in this problem turned out to be approximately 3500
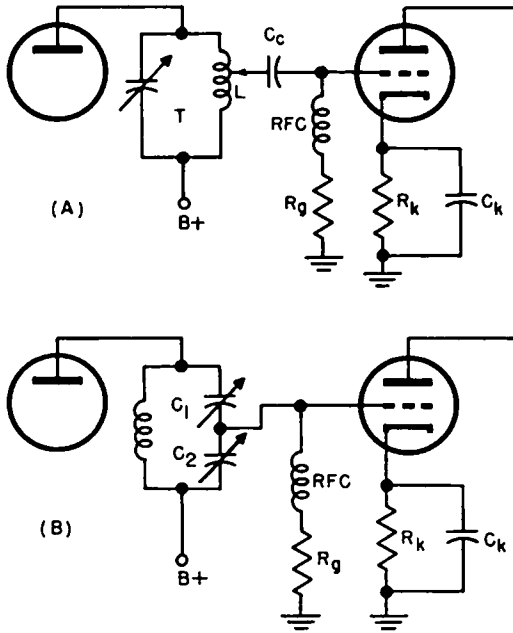
Fig. 20. Capacitive coupling between class-C amplifier tank circuit and load.

ohms. Assuming that the a-c resistance (R_{ac}) of the coil and its connections is 5 ohms, then:

$$X_L = \sqrt{3500 \times 5} = 132 \text{ ohms (approx.)}$$

The inductance L may now be calculated from the working frequency, and C determined from the knowledge that at resonance $X_L = X_c$.

28.   Coupling Tank Circuit to Load

A class-C r-f tank circuit may form part of a power output amplifier which feeds a transmitting antenna or a class-B linear amplifier, or it may be the resonant circuit of an intermediate power amplifier that serves as a buffer between an r-f oscillator and an output amplifier. (In addition to these, class-C amplifiers are also used as frequency multipliers. These are to be discussed later.) In any case, three common coupling categories are used: (1) capacitive coupling (Fig. 20), (2) inductive coupling (Fig. 21), and (3) link coupling (Fig. 22.)

Capacitive coupling — There are two general forms of capacitive coupling in common use. The system illustrated in Fig. 20A, some-

times known as impedance coupling, is suitable only for low-power interstage coupling because the lead containing the coupling capacitor C_c is, in reality, a high-impedance transmission line and cannot be made too long (more than a few inches), particularly at high frequencies without introducing serious radiation losses. By tapping down on the tank coil, an approximate impedance match between the tank circuit and the grid circuit of the next tube can generally be approached; the variable-tap feature also permits the adjustment of the amount of excitation the second amplifier grid receives.

The circuit given in Fig. 20B is a popular form of capacitive coupling. The voltage drop across capacitor C_2 is utilized as excitation for the second stage and, since C_2 is variable, the driving voltage may be adjusted by varying its capacitance. C_1 must be adjusted at the same time, of course, to maintain the total capacitance at the setting required for resonance.

Inductive coupling — A fundamental inductively coupled circuit is illustrated in Fig. 21. In this case, as in capacitive coupling, it is difficult to position the driver and driven stage far enough apart to avoid undesirable interactions, particularly when the power is relatively high. Both resonant combinations represent high-impedance circuits so that all lead lengths must be kept short. Inductive coupling requires two tuned circuits — one in the plate circuit of the driver and the other in the grid circuit of the driven stage — but despite this added complication, it has enjoyed some popularity due to its ability to discriminate against harmonics. To make certain that all the coupling is inductive, a Faraday shield must be inserted between the plate-tank coil and the grid-coupling coil to eliminate capacitive coupling. With proper coupling and an effective shield, harmonic coupling is minimized so that, in this respect at least, inductive coupling is far superior to the capacitive method. In prac-
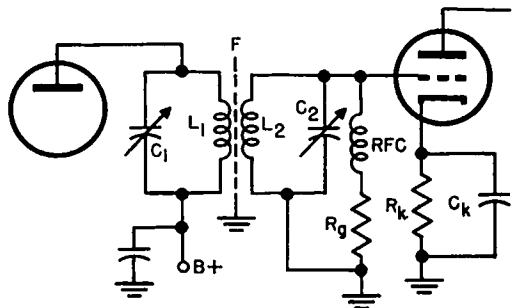
Fig. 21. Inductive coupling between class-C amplifier tank circuit and load.

tice, L_2 and L_1 are placed so that their respective axes are colinear, with L_2 adjacent to the "cold" (a-c ground) end of L_1.

Link coupling — A two-stage, link-coupled class-C amplifier is shown in Fig. 22. In this arrangement, the tuned plate and tuned grid circuits are separated by a distance of one to several feet, thus reducing the possibility of undesirable interaction between the two stages. The transmission line comprising L_2, L_3, and the connecting cable is a low-impedance line; L_2 and L_3 have two or three turns each (depending upon the frequency) and are closely coupled to the
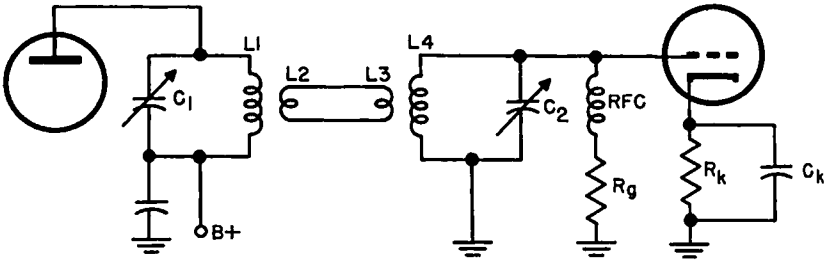
Fig. 22.   Link coupling between class-C amplifier tank and load.

respective resonant circuits. As a result of transformer action from the high-impedance plate coil to the low-impedance line and then back to the high-impedance grid coil, the transmission line carries a very low r-f voltage at a correspondingly heavy current. Its only requirements are, therefore, that it be made of finely stranded wire to reduce skin effect and that it be fabricated of sufficiently heavy conductor to keep its a-c resistance low. Excitation may be varied by changing the coupling at either or both ends of the line.

29.   The Need for Neutralization

When an r-f voltage is applied to the grid of a triode, the resulting amplified r-f plate power reacts upon the driving source through the plate-to-grid capacitance of the tube. The effects of this reaction and the methods used to overcome it constitute one of the most important phases of r-f amplifier design. As a matter of fact, its importance is such as to warrant a quantitative discussion on this topic at this point.

Consider first the circuit diagram of the class-C amplifier given in Fig. 23. The capacitor shown between the grid and plate is the inter-

electrode capacitance of the tube C_{pg}. E_{pg} is the r-f voltage across C_{pg}. The grid excitation is 500 volts rms and the output voltage is 5000
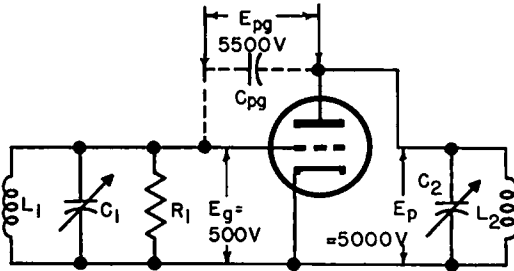
Fig. 23. Fundamental tri-ode class-C amplifier show-ing r-f voltages from grid to ground, plate to ground, and plate to grid.

volts rms; these voltages are exactly 180° out of phase provided that the output circuit is correctly tuned to resonance. Thus, under these conditions, the plate-to-grid voltage is 5500 volts.

Assume now that at the frequency of operation, the plate-to-grid capacitance is such that the reactance is 5000 ohms. (This would be the case, for example, if C_{pg} were 30 μμf at about 1000 kc/sec). For 5500 volts between plate and grid, the current fed back would be of the order of 1.1 amperes. The vectorial representation of these conditions is given in Fig. 24. The applied voltage E_g and the out-put voltage E_p are shown as two quantities 180° out of phase along the horizontal axis, one 10 times greater than the other. The voltage from plate to grid (E_{pg}) is the vector difference between the two

Fig. 24. Vector conditions which obtain in the circuit of Fig. 23.
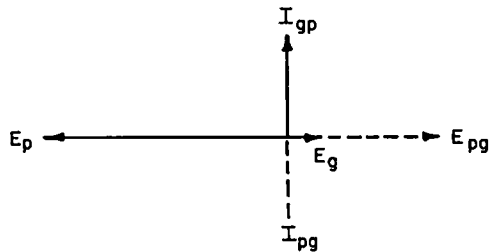
quantities and is represented by the longer broken-line arrow. The current from grid to plate (I_{gp}) due to E_{pg} will lead the latter by 90°.

When the grid voltage is applied and the current flows 90° ahead of the voltage, the impedance of the circuit is capacitive. The mag-nitude of the capacitance must be such as to present an impedance of about 450 ohms obtained from:

$$X_c = \frac{500}{1.1} = 450 \text{ ohms}$$

At 1 mc, the apparent capacitance of the grid input circuit is therefore approximately 330 μμf. On the other hand, if the output of the tube is short-circuited, the input capacitance would be only 30 μμf, as shown in the preceding paragraph. Thus, the presence in the output circuit of a voltage 10 times larger than the grid voltage and in the opposite phase causes the apparent input capacitance of the amplifier to rise to about 11 times its static value.

From the previous descriptive analysis, it is evident that:

$$C_{ai} = C_{pg} (1 + E_p/E_g) \tag{71}$$

in which C_{ai} is the apparent input capacitance, C_{pg} is the plate-to-grid capacitance, and E_p and E_g are the output and input a-c voltages, respectively. This equation must be qualified by the assumption that the circuit is properly tuned so that E_p is 180° out of phase with E_g. The feedback (displacement) current rises with increasing frequency causing the input capacitance to become even more elevated. In most triode amplifiers, particularly as the frequency is raised, plate-to-grid feedback will produce self-oscillation and a serious change in the amplifier operating conditions.

Short of interposing a shield in the form of a screen grid between the plate and control grid, this feedback current can be eliminated only by providing for the flow of an equal and opposite current to make the net current equal to zero. In these circumstances, the grid circuit will remain entirely unaffected by what goes on in the plate circuit.

30. Neutralizing Circuits

Bridge circuits — A class-C amplifier may be neutralized in either its grid circuit or its plate circuit. Grid-neutralization bridge circuits are usually referred to as Rice neutralization, while the plate type is called the Hazeltine method (refer to Fig. 25).

When either the Rice or Hazeltine system is redrawn in conventional bridge form (Fig. 25C), it is easy to see how neutralization is accomplished. If the tap on the plate tank — in this case we are referring to the Hazeltine method — is placed at the electrical center, then the coil may be considered as consisting of two identical halves, L_1 and L_2. When the circuit is constructed symmetrically with re-
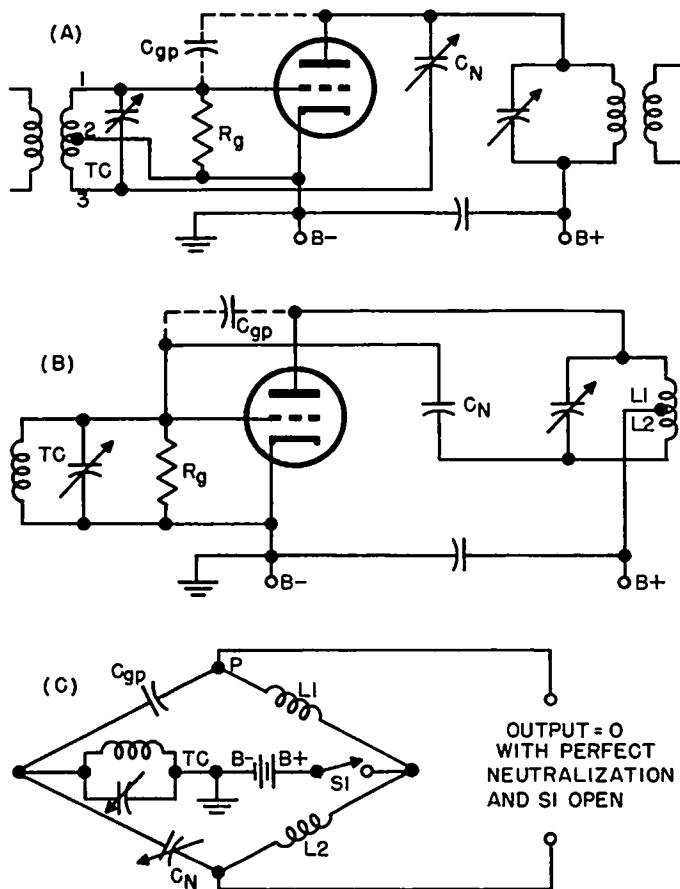
Fig. 25. (A) Grid circuit bridge neutralization (Rice). (B) Plate circuit bridge neutralization (Hazeltine). (C) Bridge equivalent of Hazeltine neutralization circuit.

spect to ground, a setting of the neutralizing capacitor C_n exactly equal to the grid-plate capacitance will result in zero net feedback current since the bridge will then be perfectly balanced and the effective value of C_{gp} with respect to the output circuit will also be zero.

The explanation may also be approached from another point of view, if the reader is not familiar with bridge circuits. Let us refer this time to the Rice arrangement in Fig. 25A. The tuning circuit in the grid (TC) receives r-f energy from the preceding stage. The upper half of the coil feeds the a-c voltage to the grid of the triode, so that the voltage applied across the input of the tube is equal to

that which appears between points 1 and 2 on the coil. An amplified voltage, of course, appears across the plate-tank coil. A portion of this voltage now returns to the grid via the plate-grid capacitance C_{gp}. If nothing else happened at this point, the amplifier would be highly regenerative and unstable. However, at the same time some of the voltage from the upper end of the plate coil is also returned through C_n to point 3 on the grid-tuning coil.

Since the voltage fed back through C_n and the voltage fed back through C_{gp} are in the same phase, the instantaneous voltage across terminals 1 and 3 of the grid coil is always zero. Essentially, this means that the voltage between the grid and plate is also zero (due to feedback effects), since point 1 is connected to the grid and point 3 to the plate. Hence, no displacement current can flow from grid to plate and the capacitance C_{gp} is effectively cancelled.

In practice, the coil tap in either the Rice or Hazeltine circuit is not placed at the exact electrical center since the circuit is not symmetrical with respect to ground; it is usually located about 1/3 of the way up from the "lower" end of the coil. In addition, C_n is always a variable capacitor to make fine adjustments possible.

As implied by the bridge-circuit labels, a criterion for judging the effectiveness of neutralization is based upon the r-f output voltage that appears across the plate-tank coil with excitation from the previous stage but without d-c power applied to the neutralized stage.
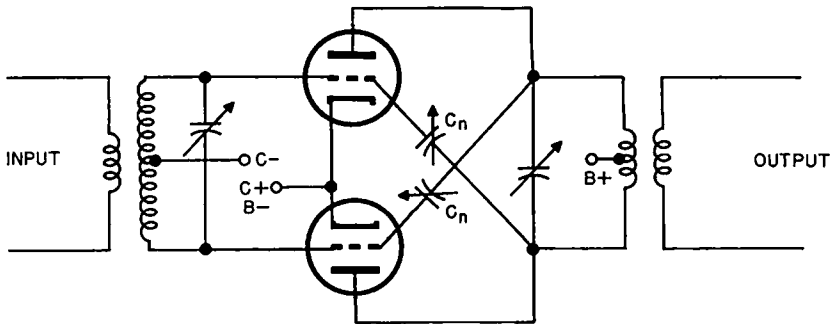
Fig. 26. Neutralization of push-pull class-C amplifiers necessitates two identical neutralizing capacitors in a symmetrical arrangement.

When the bridge is perfectly balanced, absolutely no r-f voltage should appear across the terminal of the plate coil. As a matter of fact, one conventional method of performing the neutralization ad-

justment depends upon reducing the r-f voltage in the output circuit to zero with the help of a sensitive r-f indicator.

Cross-neutralization for push-pull class-C stages — The symmetry and increased power available with push-pull class-C amplifiers have made these circuits extremely popular. Push-pull triodes must, however, be neutralized like single-ended stages, and since two tubes are involved, two neutralizing capacitors are necessary. Figure 26 illustrates the so-called "Ballantine" method of cross-neutralizing push-pull amplifiers.

Study of this circuit shows that it is a combination of Rice and Hazeltine neutralization. If one neutralizing capacitor is removed, both tubes may still be neutralized by adjusting the remaining
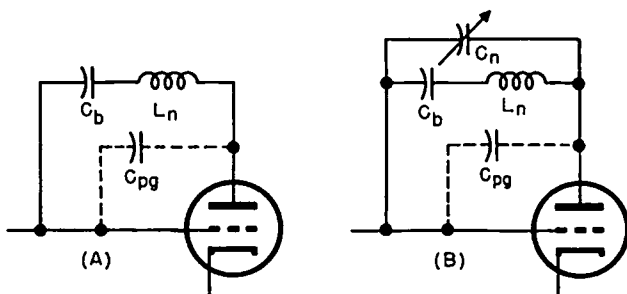
Fig. 27.   (A) Coil neutralization of a triode r-f amplifier. (B) Coil neutralization with compensating capacitor.

capacitor carefully; in this case, one tube will be Rice neutralized and the other Hazeltine neutralized. Such neutralization is quite critical and somewhat unstable, so that the Ballantine combination is much superior. When the circuit is constructed with an eye to perfect symmetry, both neutralizing capacitors may be coupled mechanically and the neutralization performed as a one-control process.

Neutralizing coil circuit — A practical method of neutralizing a class-C triode amplifier, as used in many high-efficiency broadcast transmitters, consists of an inductance of suitable value connected from the plate to the grid of the tube in series with a high-capacitance (low-reactance) capacitor C_b. The latter plays no part in the neutralization process; it merely serves as a blocking capacitor to prevent the coil from short-circuiting the plate and grid (Fig. 27A).

The operation of coil neutralization is fundamentally easy to explain. The current fed back to the grid from the plate by C_{pg} leads

the plate-grid voltage by 90°; that fed back by the coil has a lagging angle of 90° if the inductive reactance of the coil is very much greater than its a-c resistance. The two currents are, therefore, 180° out of phase and cancel each other, provided that the capacitive reactance of C_{pg} is made exactly equal to the inductive reactance of L_n. From a different point of view, L_n and C_{pg} form a parallel resonant circuit with an extremely high impedance at the resonant frequency. Hence, virtually the entire r-f voltage drop appears across the resonant circuit, the circulating r-f current in $L_n - C_{pg}$ is large, but the feedback current is negligible.

Many broadcast transmitters contain neutralizing coils that require no adjustment after installation since the station does not vary its frequency; other transmitters, in which very slight frequency variations are anticipated, contain an additional capacitor (C_n, Fig. 27B) in the neutralizing circuit. L_n is generally made somewhat smaller in reactance than Xc_{pg} and, for a given frequency, C_n is then adjusted to bring the grid-plate capacitive reactance down to the same figure.

31. Neutralization Adjustments

There are two general methods in common use for determining the conditions for correct neutralization during the tuning-up procedure.

Plate circuit r-f indication — The first of these methods requires the presence of either an r-f ammeter in the tank circuit or some means of indicating the presence of r-f voltage across the tank circuit. The latter may be a neon lamp or pilot-light bulb in series with a small loop of wire having a diameter the same as that of the tank coil. With no d-c power applied to the stage being neutralized, the grid is excited by the previous amplifier or oscillator. The neutralizing capacitor is set at maximum capacitance and the plate-tank capacitor rotated until the r-f ammeter reads current or until the neon lamp (or pilot loop) indicates the presence of r-f voltage. The neutralizing capacitor is then slowly unmeshed while the tank capacitor is rocked through resonance. As neutralization is approached, the r-f indication will diminish until, finally, when the tube is fully neutralized, there will be no indication of r-f energy in the plate-tank circuit at all. The use of a neon lamp or flashlight bulb gives only approximately correct neutralization because both of these

indicators affect the small capacitances to a slight degree and cause some detuning.

Grid circuit indication — In the absence of an r-f ammeter, this method is highly recommended especially for low- and medium-power transmitters up to several kilowatts. It does necessitate a grid-circuit milliammeter in the stage to be neutralized; such an instrument is included in every well-designed amplifier, however, since it also serves as a reliable indication of r-f drive. With plate power removed, rotating the tank capacitor through resonance will cause the milliammeter to dip sharply. This occurs because the resonant-tank circuit loads the excitation generator by absorbing large amounts of r-f power from it through the grid-plate capacitance path. The r-f voltage on the grid, as a result, is substantially decreased; this reflects itself in greatly diminished rectified grid current. As C_{gp} approaches neutralization, the dip becomes less and less pronounced until, when neutralization is perfect, the resonant tank is effectively disconnected from the excitation source since the grid-plate path has been cancelled out. In a properly neutralized amplifier, the grid-circuit milliammeter remains completely unaffected as the tank capacitor is rotated fully from maximum to minimum capacitance.

32. Frequency Multipliers

Since any but a pure sinusoidal waveform contains component frequencies that are multiples of the fundamental frequencies, it follows that any distorting amplifier is a harmonic generator or *frequency multiplier.* Since the input to an r-f amplifier is normally a single frequency, second harmonic output is easily obtained by placing a circuit tuned to this frequency in the plate circuit.

The efficiency and power output of a frequency multiplier tend to increase when the grid bias is increased farther beyond cutoff than normal for straight-through class-C amplification. For a typical triode of medium mu, the most favorable conditions for frequency doubling are obtained with a grid bias about 10 times the cutoff value. For this condition, however, greater excitation power is demanded since the driving voltage must be large enough to swing the grid into the positive region against this large bias. At best, frequency-doubled output is about 1/3 that expected from the same tube with the same power input in straight amplification (class-C). To utilize higher harmonics, as in frequency tripling and quad-

rupling, greater bias and correspondingly increased drive is required. Table 2 shows these relationships clearly:

TABLE 2

CLASS-C TRIODE AMPLIFIERS

	Fundamental	*Doubler*	*Tripler*
Grid Bias (× cutoff)	2	10	20
Plate efficiency (approx.)	75%	50%	55%
Plate-current angle	150°	90°	70°
Relative output	100	70	35

Frequency multipliers generally need not be neutralized since the grid and plate circuits are not tuned to the same frequency so that there is little danger of instability due to positive feedback.

A rather unusual and interesting frequency multiplier is illustrated in Fig. 28. In this arrangement, the grids are connected in push-pull and the plates in parallel. The plate-tank circuit, in the setup given here, is tuned to the second harmonic of the fundamental frequency while the grid circuit is adjusted to resonances at the fundamental. In this way, the plate-tank circuit is given a "push" on *each half-cycle* rather than on alternate half-cycles. Thus, there is a full cycle of output produced for each half-cycle of input, and frequency doubling results. Even without a tuned plate circuit, this arrangement provides a recognizable frequency multiplication and is, therefore, sometimes used when the input signal is not a single frequency.

It is sometimes thought that frequency multipliers amplify the harmonics fed to them from the previous stage. Although a *slight* contribution might be made to the multiplied output from this source, almost all of the harmonic content of the final signal is generated right within the frequency multiplier itself, so that it is of little benefit in feeding an impure signal to its grid.

33. Beam and Tetrode Class-C Power Amplifiers

Beam and tetrode transmitting tubes are available up to a plate dissipation rating of about 250 watts. For low- and medium-power class-C amplifiers, such tubes are generally preferred to triodes be-

cause they do not require neutralization except at very high frequencies and may be driven more easily with less excitation power.

This is especially true of beam-power tubes. These tubes employ beam-forming plates and careful alignment of control grid and screen wires so that the electron current from the cathode is focused into a series of beams passing between the grid conductors. The screen-grid current is thus reduced, and suppressor action is obtained by space-charge effects produced between the screen and the plate. The plate characteristic curves of beam-power tubes approach ideal pentode curves in that the knee occurs at quite a low voltage permitting increased plate voltage swing, screen dissipation is reduced, and the efficiency is increased. High-power sensitivity accounts for the ease with which beam-power tubes may be driven to full output.

Circuit and power conditions for pentodes and beam-power tubes are similar in many respects to those encountered in setting up triode class-C power amplifiers, with these exceptions: (1) The d-c screen voltage is an important factor in tetrode and beam-power amplifier design. This voltage should be considerably larger than the maximum grid potential attained with excitation (E_{max}). In this way, a desired value of total cathode current — i.e., current flowing from the cathode to the control grid, screen grid, and plate — may be ob-
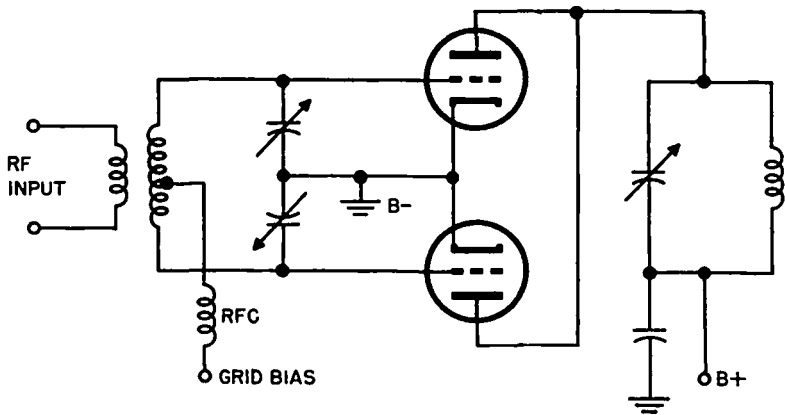
Fig. 28. Push-pull frequency multiplier.

tained without excessive control grid current and with little exciting power. The screen voltage must also be considerably smaller than the plate supply voltage to keep screen dissipation sufficiently small.

(2) The minimum instantaneous plate voltage should not be al-
lowed to drop too low during operation. If this precaution is not
observed, there is danger that a *virtual cathode* condition will occur.
This effect, often observed in beam tubes and pentodes, is the re-
sult of low plate voltage and large space current in the tube; electrons
on their way to the plate slow down or come to a halt on the screen
side of the suppressor (or virtual suppressor in true beam tubes).
These electrons then return to the screen, increasing the screen dissi-
pation unduly. This space-charge effect is valled a *virtual cathode*
because the electrostatic field conditions in the vicinity resemble
those encountered around an actual cathode. (3) A more serious
condition than this occurs if the plate voltage of a tetrode is allowed
to fall too low. If the minimum instantaneous plate voltage drops
below the screen potential, secondary emission electrons will leave
the plate and go to the screen, again causing excessive screen dissipa-
tion and possible tube burnout.

A typical beam-power class-C amplifier is shown in Fig. 29. For
low power applications, screen voltage is obtained from the plate
supply through a series resistor of the proper value, and the screen
is carefully bypassed to common ground through a sufficiently large
capacitor. Since the plate-grid capacitance of a tetrode or beam tube
is very low, danger of instability or self-oscillation does not become
serious except at frequencies above about 25 mc. When necessary,
neutralization is handled in exactly the same manner as in triodes,
except that considerably smaller neutralizing capacitors are called
for.

34. Linear-Tuned Class-B Amplifiers

A class-B amplifier is defined as an amplifier in which the grid
bias is approximately equal to plate-current cutoff so that the plate
current is approximately zero when no exciting grid voltage is
present. The plate current in a class-B stage flows for approximately
one-half of each cycle when an a-c grid voltage is applied. Such an
amplifier yields a pulse of plate current in which both the peak and
average values are proportional to the amplitude of the excitation
voltage. For this reason, such amplifiers are often conventionally
termed *linear amplifiers*.

It is often desirable to amplify an r-f voltage *after* it has been
amplitude modulated rather than before modulation. Although it
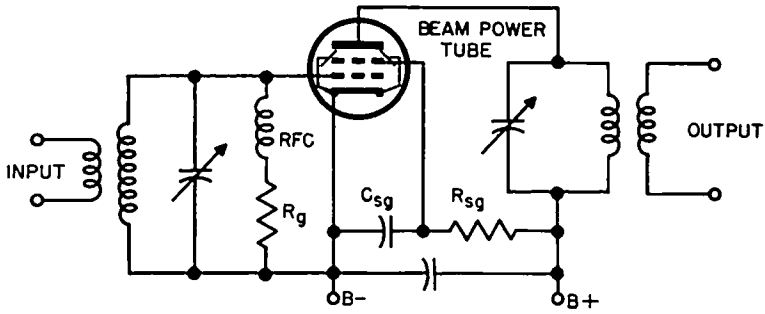
Fig. 29. Beam power class-C amplifier.

is common practice to modulate the final output amplifier in low- and medium-power transmitters, the audio power requirements in high-power systems becomes prohibitive in many cases if this same procedure is attempted. For example, a transmitter having an input
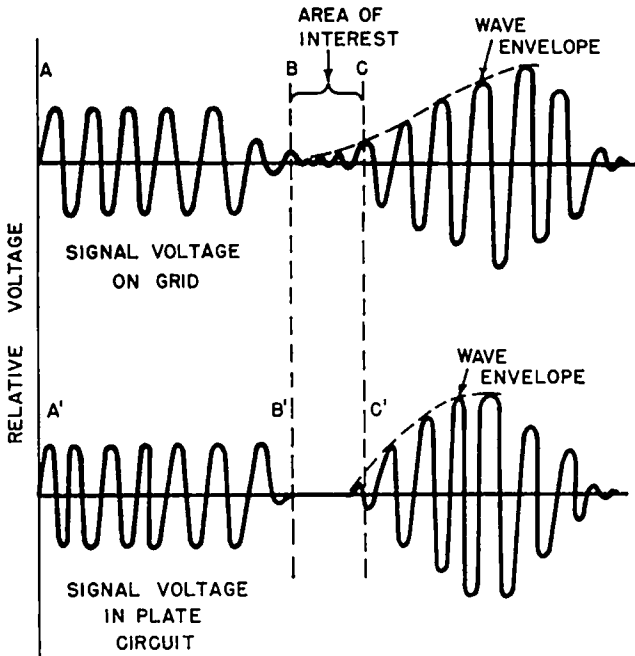
Fig. 30. Grid input waveform (modulated) to a class-C power amplifier and the corresponding plate voltage waveform of the same amplifier showing modulation distortion.

power to its class-C r-f amplifier of 100 watts requires 50 watts of peak audio output power for 100% modulation. This audio power is not difficult to obtain by using small receiving tubes, such as 6L6's, in push-pull parallel. On the other hand, the 500 watts of audio power needed to modulate a 1-kw output stage fully necessitates the utilization of much costlier components; the cost-to-power ratio rises rapidly for r-f input powers in excess of a few kilowatts. Thus, a considerable economy could be effected if it were possible to modulate a 100-watt r-f driver and *then* build up the modulated power output to a higher value.

Unfortunately, however, a class-C power amplifier cannot be used to amplify a modulated voltage without intolerable distortion of the audio component. Figure 30 illustrates why this is true.

Since a class-C amplifier is biased well beyond cutoff, the instantaneous value of the signal grid voltage must rise above a certain minimum figure before plate current begins to flow (Fig. 31). Only the positive peaks of grid voltage large enough to overcome the bias sufficiently to reach or exceed the cutoff value will result in plate-current pulses. Assuming that the input voltage from the previous modulated stage falls below plate-current cutoff (as in the area be-
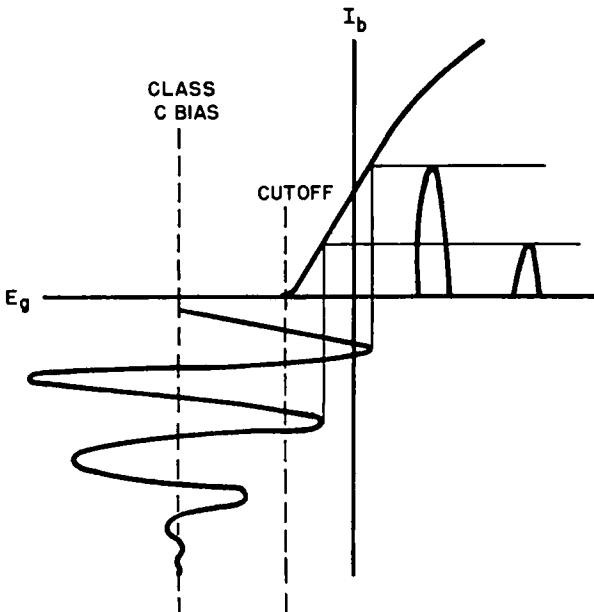
Fig. 31. Plate current pulses due to grid input voltage in class-C amplifier.

tween B and C in Fig. 30) , then for this area, the output in the plate circuit will be zero (B' to C') . For distortionless output, it is necessary that the fluctuations in carrier voltage at an audio rate due to modulation be reproduced exactly so that the envelopes of the waves are identical in shape. Obviously, this cannot be expected to occur in a class-C amplifier for the reasons just given. In Fig. 30, the difference in these envelopes is clearly discernible.

A class-B amplifier, on the other hand, is biased at or near cutoff. For this condition, plate-current pulses flow for all positive values of the impressed signal voltage, producing a plate-circuit voltage that duplicates the input insofar as the upper envelope waveform is concerned. Since the essential difference between class-C and class-B operation is the bias adjustment, the same circuits may be used for either type of amplifier. Class-B amplifiers for amplitude-modulated inputs, however, operate at a substantially lower plate-circuit efficiency than class-C stages. In the latter, the alternating plate voltage is constant so that the design of such an amplifier may be aimed toward maximum efficiency.

In the case of a class-B amplifier handling, as input, a modulated r-f signal maximum efficiency can be realized only at the peak of the modulation cycle. As the instantaneous efficiency at all other times will be directly proportional to the instantaneous plate-voltage amplitude and hence less than peak efficiency, the average efficiency of the class-B stage will be less than that of its class-C counterpart. The maximum efficiency of a practical class-B amplifier may range from 60% to 65% during normal operation. Since the efficiency ranges from zero to maximum for 100% modulation, the average efficiency is in the order of 30% or slightly better. This should be contrasted with the normal operating efficiency of approximately 80% for a class-C amplifier.

The required bias for a class-B amplifier cannot be obtained by the grid leak method, since this method assumes fixed grid drive. As we have seen, a class-B amplifier driven by a modulated r-f signal receives grid voltage of varying magnitude ranging from zero to twice the carrier voltage for 100% modulation. Thus, class-B stages are generally biased by a fixed source such as a battery or a separate power supply.

The peak power output of a linear class-B amplifier does not differ materially from that of an equivalent class-C stage operated under the same conditions of plate voltage and plate current. On the other

hand, the peak power of a 100% modulated r-f wave is four times greater than the unmodulated carrier power. Thus, a tube used for linear amplification of an a-m wave is capable of delivering only about one-fourth as much carrier power as the same tube would provide in class-C operation.

35. Push-Pull Class-C Amplifier Considerations

The conventional method of neutralizing push-pull class-C amplifiers was described in Section 30. Let us investigate other considerations that apply to push-pull operation.

Even harmonics should not appear in the output of a properly balanced push-pull class-C amplifier because harmonic currents flow
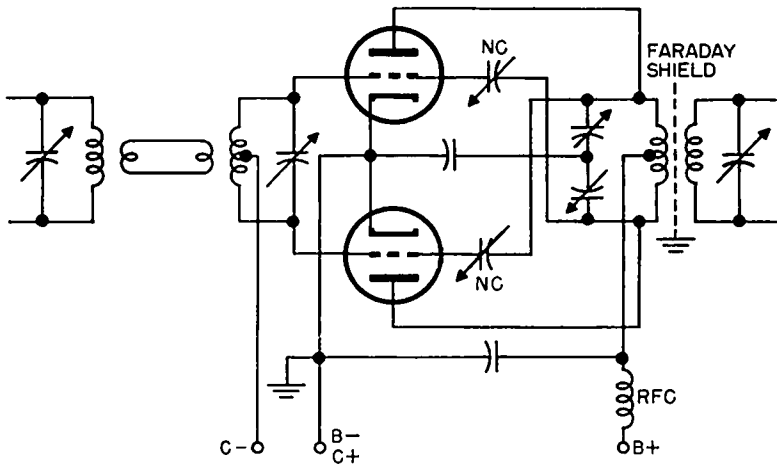
Fig. 32. Use of Faraday shield to reduce harmonic radiation.

in opposite directions through the two halves of the tank coil, causing cancellation of magnetic fields (Fig. 26) . When the load circuit is coupled equally and symmetrically to each half, there should be no even harmonic voltage set up in the load. To insure the necessary symmetry and to provide a low-impedance path to ground for the harmonic currents, push-pull amplifier tuning capacitors are usually of the split-stator type shown in Fig. 32. Another significant advantage of split-stator tuning is that it permits the capacitor rotor to assume ground r-f potential without disturbing the symmetry of the tank circuit.

When strong harmonics are still found in the output of push-pull stages, they may be attributed to stray capacitance between the plate-tank circuit and the coupling circuit to the next stage. With respect to capacitive transfer of harmonics, the two tubes behave as though they were in parallel and feed the next stage (or the antenna) a substantial harmonic amplitude.

To avoid capacitive transfer, push-pull amplifiers should always have an electrostatic shield (Faraday shield) between the tank coil and the antenna coupling circuit (Fig. 32). This will eliminate capacitive transfer almost completely and the residual harmonic will then be a function of the symmetry of balance between the two tubes and their associated components and wiring. In transmitters where the installation of a Faraday shield is difficult or impossible, recourse is sometimes taken to link coupling in which one side of the link is grounded. With sufficient isolation between the output tank coil and the antenna coupling coil, link coupling is capable of producing a large reduction in harmonic radiation.

Although the power output of a push-pull stage is twice that of a single-ended stage operating under the same conditions, the excitation power requirement is also doubled. This is not usually a problem since exciter stages are designed to provide more driving power than necessary for a single-ended stage so that reserve is almost always available.

Another advantage of push-pull class-C amplifiers using neutralized triodes should not be overlooked: in addition to the fact that a stage balanced to ground is easier to neutralize perfectly, this symmetry makes it possible to feed bias and plate voltages to the tubes through *centertapped* coils rather than coils that are tapped some odd fraction of their lengths up from the grounded end. This feature makes band switching far easier in transmitters where frequencies must be changed quickly since the class-C stages need be neutralized on one band only and the neutralization will hold for all the others.

36. Parasitic Oscillations

Parasitic oscillations occur at some frequency other than the fundamental or harmonic frequency. When a transmitter acts in an abnormal fashion without apparent cause, the trouble may often be traced to parasitics.

Each complete r-f transmitter represents a special case in itself; no one can hope to define the specific causes of parasitic oscillations in a specific class-C amplifier, but the general principles are known and several of the more common causes are easily remedied:

(1) If any portion of the plate circuit — this includes connecting wires as well as lumped components — is close to resonance with any part of the grid circuit, parasitic oscillations may occur, since these sections may form a tuned-plate, tuned-grid oscillator if a feedback path is available.

(2) If any element of the amplifier tube tends to display negative resistances and if there is a resonant circuit in series with this element, parasitic oscillations are almost certain to ensue.

Consider the well-designed amplifier circuit given in Fig. 33. This is a grid-neutralized (Rice) circuit with parallel-feed (shunt-feed)
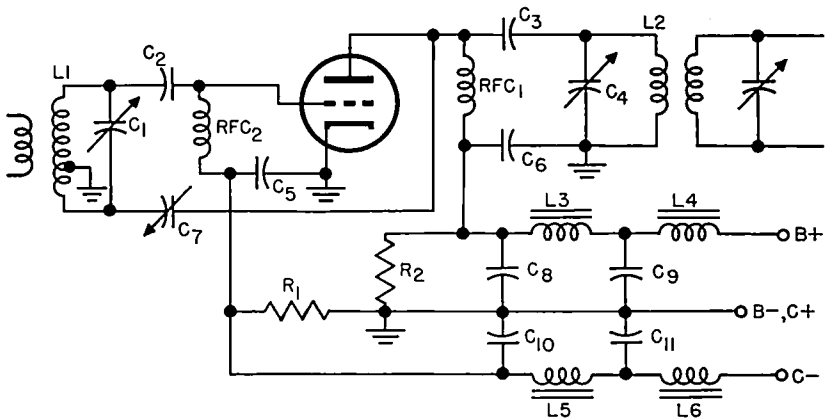
Fig. 33. Class-C amplifier of good design.

of d-c plate potential through an rfc. Shunt feed is a useful method of applying plate voltage in many circumstances because it makes possible a tank coil that is at d-c ground potential on both sides, thus eliminating the shock hazard. Shunt feed is also used for grid bias through the second rfc.

In assembling transmitter components, grid and plate-supply filters may be placed close enough to each other to cause parasitic trouble at frequencies around 25 to 35 cps. Figure 34 represents a portion of the amplifier of Fig. 33 redrawn to show the possible low-frequency parasitic oscillation circuit that may result from magnetic
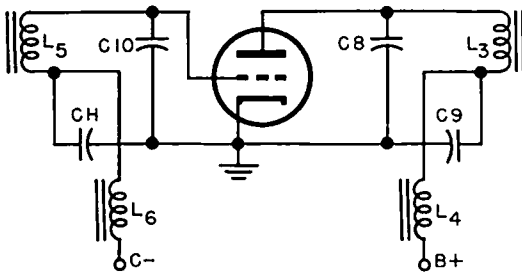
Fig. 34. Filter networks forming a part of the plate and bias power supplies may sometimes cause low-frequency parasitic oscillations.

coupling between filter parts. A design remedy, of course, consists of locating the offending components as remotely as possible from each other and using adequate shielding if required. The values of the capacitances and inductances may also be changed to make the filter sections resonant at different frequencies.

Whenever parallel-feed is used, there is danger of resonance and oscillation due to the presence of the rfc's. This is particularly true at high frequencies where the interelectrode reactance of the tube is small and feedback may easily occur. The relevant sections of Fig. 33 have been redrawn in Fig. 35 to illustrate one of the possibilities for parasitic resonant circuits. Oscillation due to resonant circuits of this sort usually takes place at a frequency that is somewhat lower than the fundamental rf of the system with normal components. The remedy in this case is to select different values of inductance for the rfc's in the grid and plate circuits, always making the plate-circuit rfc higher in inductance.

A frequent cause of high-frequency and uhf oscillations may be traced to the inductance in grid and plate leads of oscillator tubes. Even a short length of wire possesses measurable inductances which may be resonant with the interelectrode capacitances to form uhf tuned-plate, tuned-grid oscillators. Tubes connected in parallel for
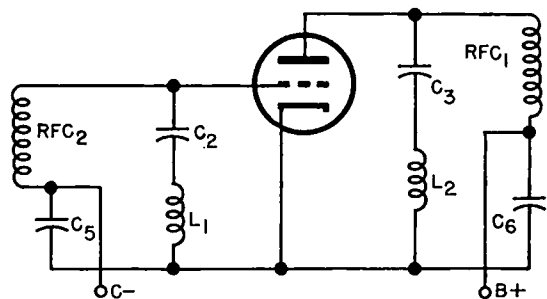
Fig. 35. Circuit that makes possible parasitic oscillations due to resonant combinations formed by rfc's and associated capacitors.

class-C amplification are particularly susceptible to parasitics of this nature. The accepted procedure for eliminating these parasitic oscillations seems to involve the use of a small inductance in series with the plate lead, connected as closely as possible to the plate terminal of the tube. Such an inductance detunes the lead by resonating it at a lower frequency then the grid lead. A low-value carbon resistor is often connected in parallel with the auxiliary inductance to form a "parasitic suppressor." This resistor acts as a "losser" for the parasitic oscillation which is forced to take this path due to the inductive reactance offered by the inductance in the alternate path.

Negative resistance parasitic oscillations are much more common in tetrode and beam-power amplifiers than they are in triodes. If the screen voltage of a tetrode is higher than the minimum instantaneous plate voltage, normal dynatron action may occur, producing spurious oscillation of the parasitic type. The same effect may also occur when the plate voltage of a triode is driven below the grid voltage on negative modulation peaks.

It sometimes happens that when the grid of a triode is driven far into the positive region, secondary emission may begin from the grid wires, causing a diminution of grid current. This effect may go sufficiently far so that an increase of positive voltage on the grid produces a *decrease* in grid current. This is a negative-resistance reaction. Under these conditions, any resonant circuit in series with the grid may go into oscillation. Modern tubes are carefully designed to minimize secondary grid emission, but trouble may still be encountered if tubes are used for the wrong purposes. For example, a tube intended specifically for audio applications should not be used at radio frequencies unless precautions are taken to prevent grid emission.

37. Frequency Limitations

Stray inductances have already been mentioned in connection with parasitic oscillations.

Vacuum tubes, coils, resistors, wiring, transformers all have the common failing that stray capacitances as well as stray inductances make an appearance due to the nature of their structure. Although stray capacitance is sometimes utilized in high-frequency tuned circuits, making it possible to omit a tuning capacitor, even this expedient fails when the frequency rises sufficiently high. This is the

result of the fact that the stray capacitance, small as it may be, is often still too high for tuning purposes. The various significant inductances and capacitances are shown in Fig. 36. The lumped components inserted intentionally for frequency control are L_1, L_2, C_1, and C_2. In the other cases, L_c is the inductance of the leads of the capacitor, L_w is the inductance of connecting wires, L_t is the inductance of the tube leads and elements, C_s is stray capacitance, and C_{gp}, C_{gk}, and C_{pk} are the respective interelectrode capacitances. To improve vhf operation, a rather tedious redesigning procedure is required.

By careful construction of C_k and C_b (bypass capacitors), the respective L_c's can be reduced materially; new shapes and structures for the tuning capacitors C_1 and C_2 have led to great reductions in
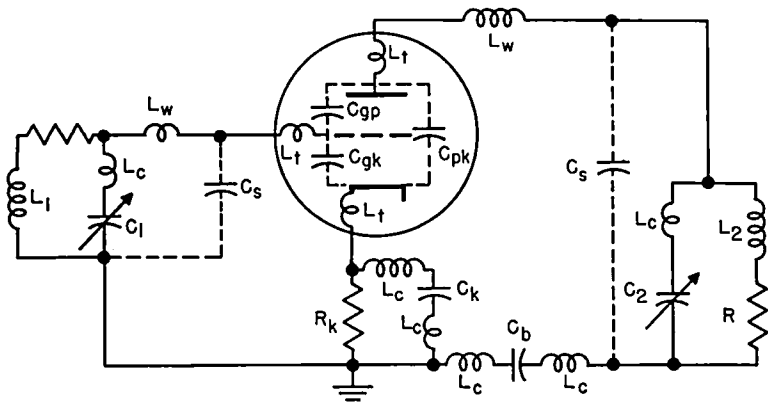
Fig. 36. Inductances and capacitances in a high-frequency amplifier circuit.

the associated inductances L_c. By careful layout, the wiring can be shortened to reduce each L_w, thus approaching the merging of C_{pk}, C_2, and C_s. As this step-by-step procedure is carried on, the frequency of satisfactory operation continues to rise until L_1 and L_2 have become small hairpin loops and C_1 and C_2 have completely vanished leaving only the stray and tube capacitances for tuning. This is the limiting frequency, and, if a further increase in frequency is required, changes in tube structure must be made.

Improvements in tube design have fallen into four categories: (a) wire-lead placement, (b) internal tube capacitance. (c) transit time, and (d) principles of operation that differ from conventional vacuum tubes.

Wire-lead placement — Many new high-frequency tubes are equipped with double leads going to a single grid, plate, or cathode. The twin-lead arrangement often shortens connecting wire lengths appreciably and minimizes the possibility that a single cathode wire will become a common impedance for grid and plate circuit and thus a source of feedback and instability.

Internal tube capacitance — Modern tetrodes and beam-power tubes have greatly reduced internal capacitances and have much lower losses in the glass. This has resulted in a surprising aggregate improvement in high-frequency operation. Constant redesigning of essentially the same tube type has, for example, brought reliable high-frequency operation up gradually from 30 mc to better than 800 mc.

Transit time — At extremely high frequencies the time required for electrons to travel from the cathode to the plate (transit time) becomes an appreciable fraction of the period. At uhf's the alternating grid voltage changes so rapidly that electrons leaving the cathode under the influence of a positive cycle cannot reach the grid before it changes polarity. No output can be obtained under these conditions. For satisfactory operation, the transit time should be less than 0.1 the time for one cycle of the uhf signal. The transit time is fixed by the geometry of the tube and the applied voltages and its effect is to decrease the input impedance of the tube. For example, one tube displays an input impedance of over 1 megohm at 1 mc; at 100 mc, however, the input of the same tube drops to a few thousand ohms. This diminished impedance makes it almost impossible to supply the required grid driving voltage and causes a serious power loss.

By making the tube elements much smaller in size and reducing the spacing between them, transit time has been cut down until reliable operation at reasonable efficiency up to 1500 mc has been made possible if the output is confined to a few watts. Efficiency is poor because close spacing raises both capacitance and grid temperature.

For higher frequencies, conventional vacuum tubes can no longer be used. It has been found necessary to go to tubes such as klystrons, magnetrons, traveling wave tubes and related types to realize satisfactory operation at the "super-high" frequencies.

High-frequency tubes — A detailed discussion of these special-purpose tubes is not within the intended scope of this book. Men-

tion is made of them, however, so that the reader may direct his study in this direction if he wishes. A general class of tubes which make use of the fact that an electron induces a charge of metal electrodes in its path is known as the *velocity- modulated* group and includes the klystron as its major constituent. *Magnetrons* are tubes in which the action is based upon the magnetic effects of electrons in motion. Others such as the *traveling-wave* tube are dependent upon complex wave factors for their operation.

38. Grounded-Grid Amplifiers

In certain applications, particularly in high-frequency f-m and television tuners, triode r-f amplifiers may be used without neutralization if the control grid is maintained at ground potential. When so connected, the control grid behaves as a shield between the input and output circuit as well as a control electrode. A typical basic circuit is given in Fig. 37.

The type of input circuit used to drive a grounded-grid amplifier may take any form, but provision must be made for a d-c return path from the cathode to ground. This function is performed by R_k in Fig. 37. In a circuit of this nature, R_k would have to be selected

Fig. 37. A triode-grounded grid amplifier may be used in r-f applications without neutralization.
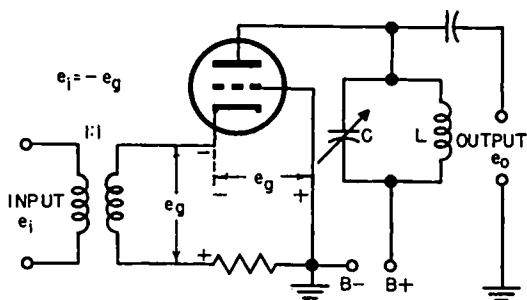
to provide the correct d-c bias for the tube; should a high-impedance input be desired, R_k could be replaced by a resonant circuit — such as the tuned circuit of the preceding stage — and d-c bias established by a battery in series with the cathode ground return.

Since the control grid is at ground potential for both d-c and signal voltages, any feedback current from the plate to the grid will return to the plate circuit directly through ground rather than through the input circuit as in normal grounded-cathode amplifiers.

The grid, therefore, functions as a shield between the input (cathode) circuit and the plate circuit, preventing positive feedback that may encourage instability.

Since a grounded-grid amplifier is rather unconventional in operation, a brief analysis of its gain capabilities will help the reader appreciate its performance and suitability for special uses.

The equivalent circuit of this amplifier is shown in Fig. 38. The input voltage from the previous stage is e_i, the plate voltage of the tube is represented by the equivalent generator $-\mu e_g$, the plate resistor by r_p, and the tuned circuit acting as the plate load by R_L. The algebraic signs on the equivalent generators are based upon the fact that the input transformer is connected so that a positive input voltage e_i makes the cathode negative with respect to ground. On this basis, the grid is positive with respect to the cathode at the same instant. Since the signal is applied between cathode and ground, it may be considered to be in series with the plate circuit and is thus shown in the equivalent circuit.

We can express the Ohm's law relationships in this equivalent circuit by means of the equation:

$$e_i + (-\mu e_g) = i_p (r_p + R_L) \tag{72}$$

From the circuit diagram in Fig. 37, it is evident that the input voltage and the applied grid voltage are equal in magnitude but opposite in sign. Thus:

$$e_g = -e_i \tag{73}$$

The value for e_g in equation (73) may be substituted in equation (72) to yield:

$$e_i + \mu e_i = i_p (r_p + R_L) \tag{74}$$

Solving for e_i, we have:

$$e_i = i_p \frac{r_p + R_L}{1 + \mu} \tag{75}$$

The output voltage e_o is, as usual, the product of the instantaneous plate current i_p times the load resistance R_L, so that the gain may be determined by:

$$\text{Gain} = e_o/e_i = \frac{i_p R_L}{i_p} \times \frac{1 + \mu}{r_p + R_L} \tag{76}$$

or simply:

$$\text{Gain} = \frac{R_L \ (1 + \mu)}{r_p + R_L} \tag{77}$$

The significance of equation (77) can be appreciated by comparing it with the gain equation for a conventional grounded-cathode amplifier:

$$\text{Gain} = \frac{-\mu R_L}{r_p + R_L}$$

The equations differ in two respects: first, the gain of a grounded-grid amplifier is slightly greater than the conventional type in that the number 1 is added to the amplification factor before multiply-

Fig. 38. Equivalent circuit of the basic grounded-grid amplifier.
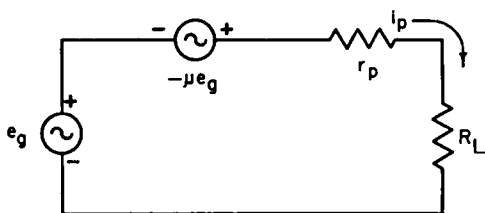
ing by R_L; second, the grounded-grid amplifier produces zero phase inversion in contrast to the 180° inversion found in grounded-cathode amplifiers as is evident from the absence of the minus sign in the numerator of equation (77). Both equations assume resistive loads; this would, of course, be the case when the tuned circuit that forms the plate-load is in resonance with the exciting voltage e_1.

Grounded-grid amplifiers find application in circuits of relatively high frequency where a tuned amplifier without neutralization is to be used and where the noise level of pentodes is too high to be tolerable. Input resistance is relatively low while the output impedance is quite high. At extremely high frequencies, some neutralization may be necessary due to the unavoidable coupling between wires, socket lugs, and so on, but the neutralizing capacitance in such cases is usually very small, making neutralization appreciably easier.

39. Cathode Followers

In a sense, the cathode follower is an inversion of the grounded-grid amplifier. Where the latter presents a low impedance to the driving source and a high output impedance, the cathode follower displays a high input and a low output impedance. Ordinarily used

specifically in audio- and low-frequency circuits, cathode followers may be applied advantageously to certain types of r-f amplifier circuits, particularly in combination with a grounded-grid amplifier as described in the next section. A cathode follower is illustrated in Fig. 39. The cathode follower differs from a conventional triode amplifier in that output is taken as a voltage drop across an unbypassed cathode resistor rather than from a plate load. As an am-

INPUT

OUTPUT

B—   B+
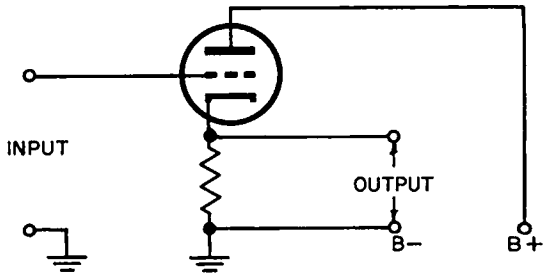
Fig. 39.   Fundamental cathode-follower circuit.

plifier with 100% inverse feedback (degeneration voltage equal to the output voltage), its gain may be found by substituting a feedback factor of 1 into the equation for the gain of an amplifier with inverse feedback. That is:

B—   B+

C_2 L_2

C_c

V1   V2

OUTPUT

INPUT

C_1 L_1

R_k C_k
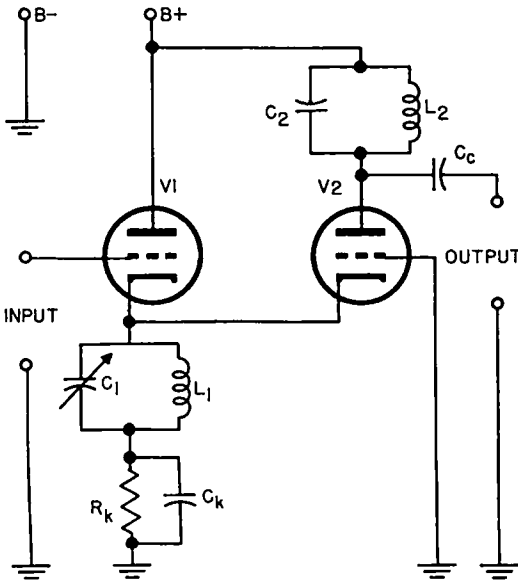
Fig. 40.   A cathode-coupled amplifier as an r-f amplifier.

$$\text{Gain of any feedback amplifier} = \frac{A}{1 - A\beta} \qquad (78)$$

in which β is the feedback factor expressed as a fraction, and A is the gain without feedback.

With $\beta = 1$, then:

$$\text{Gain of cathode follower} = \frac{A}{1 - A} \qquad (79)$$

From equation (79) it is seen that, as A is allowed to grow larger and larger, the gain of the cathode follower approaches unity but can never exceed it. Thus cathode followers are not used as voltage amplifiers, although substantial power amplification is possible.

The real value of a cathode follower lies in its usefulness as an impedance transformer. It may be shown that the cathode-follower amplifier is the equivalent of a conventional amplifier which employs a tube having an apparent amplification factor μ' such that:

$$\mu' = \frac{\mu}{1 + \mu} \qquad (80)$$

and an equivalent plate resistance r_p'

$$r_p' = \frac{r_p}{1 + \mu} \qquad (81)$$

Equation (80) verifies the previous statement that the gain of a cathode follower cannot ever reach unity since μ' must always be less than 1. If a tube having a high μ is used, the equivalent plate resistance becomes very small as implied by equation (81). Thus the output impedance is quite low compared to a conventional amplifier so that the cathode follower finds much application as a high- to low-impedance transformer.

40. Cathode-Coupled R-F Amplifier

One of the most interesting applications of cathode followers to resonant r-f circuitry is that of the cathode-coupled amplifier shown in Fig. 40.

The low input impedance of a grounded-grid amplifier is often undesirable. This objection is nicely overcome by utilizing a cathode follower as a driver stage for the grounded-grid amplifier as in the cathode-coupled amplifier in Fig. 40. Since the cathode follower has

a high input impedance, it is well suited to serve as an amplifier after some input device such as a tuned receiving antenna or a radar antenna; its low output impedance matches the input impedance of the grounded-grid amplifier so that the cathode follower forms a highly satisfactory impedance transformer. When the two triodes are enclosed in one envelope (e.g., a 6SN7, 12AU7, etc.), the resulting two-stage amplifier has both a high input and a high output impedance, gain that compares with that of a pentode, good stability, and low noise level. Both tubes are biased by the common network comprising R_k and C_k. The cathode follower is tuned by L_1 and C_1, power being coupled into the grounded-grid stage directly from the cathode of V_1. The load impedance for the grounded-grid amplifier is the resonant combination L_2-C_2. Power is fed to the final load via coupling capacitor C_c.

41. Review Questions

1. Explain how the tank circuit of an r-f amplifier discriminates against harmonic output.
2. Discuss the relative advantages and disadvantages of high L-C and low L-C ratio.
3. Why is link coupling favored in large transmitters over straight inductive or capacitive coupling?
4. Draw a schematic diagram of a typical triode class-C amplifier. Using this drawing to assist you, explain why neutralization is necessary, tracing the feedback path or paths in the amplifier. Why do pentodes work satisfactorily without being neutralized except at the very high frequencies?
5. Using a diagram to help, explain how bridge neutralization operates in either the grid circuit or the plate circuit of a triode.
6. Describe the process of neutralizing a push-pull stage by the Ballantine method.
7. Explain why radio amateurs prefer bridge neutralization to coil neutralization; explain why broadcast stations favor coil neutralization.
8. With the aid of a diagram, explain how a push-push doubler produces harmonic output.
9. Describe the operation of a linear-tuned class-B amplifier following a modulated class-C stage. Why is linearity necessary in this application?
10. What are parasitic oscillations? What are the usual steps taken to prevent or cure them?
11. Why are push-pull class-C amplifiers popular in small and medium-powered transmitters?
12. How are the frequency limitations imposed on designers by equipment gradually being overcome?
13. Describe fully the operation and usefulness of a cathode-coupled amplifier.

Chapter 5

TRANSISTOR R-F AMPLIFIERS

42. Introduction

Perhaps the most phenomenal aspect of transistor development in recent months is the way in which the high-frequency barrier has been overcome. In the initial stages of transistor research, point-contact types had a considerable advantage over junction transistors as high-frequency amplifiers. By bringing the emitter and collector electrodes very close together, point-contact transistors could be made to oscillate without difficulty up to 100 mc or more while junction transistors had a useful frequency response appreciably below 1 mc. This limitation was largely the result of undeveloped manufacturing processes which have since been perfected.

Frequency restrictions on transistor amplifiers do not appear to be more severe than those encountered in the use of vacuum tubes at the present time. Ingenious fabrication methods and high-quality engineering are advancing the upper frequency limit insistently as research goes on. In this chapter, we shall be concerned particularly with transistor high-frequency amplifiers, recommended circuits for high-frequency applications, and the problems encountered in the design and construction of such amplifiers.

43. Factors Governing Frequency Response

The movement of holes or electrons from the emitter through the base layer to the collector requires a definite time. As short as this

time may be, it must be taken into consideration when thinking in terms of frequency limitations. It has been shown that electrons diffuse through a semiconductor such as germanium at a velocity in the order of 3500 to 4000 cm/sec for each volt of potential difference (pd) per cm. Holes, on the other hand, move much more slowly. A representative figure for the positive carriers is about 1800 cm/sec/volt/cm pd. This signifies, of course, that in either point-contact or junction transistors, signals whose frequencies demand that the carriers traverse the required distances in less time than this cannot be handled. This situation is analogous to *transit time* effects in vacuum-tube amplifiers.

The current carriers do not have clear and unobstructed paths from the emitter to the collector. Thus, the transit time is not the same for all current carriers at any given instant. Since the carriers do not all take the same route, those produced by a specific pulse of signal at the emitter do not all manage to reach the collector at the same time. The differences in arrival times are very small and may be ignored at audio and low radio frequencies. At high frequencies, however, these differences take on added significance because they result in interference between carriers. When the effect becomes sufficiently severe (its severity being a function of frequency), the output signal waveform becomes *dispersed* so that it may no longer be a recognizable reproduction of the input signal. In addition, the
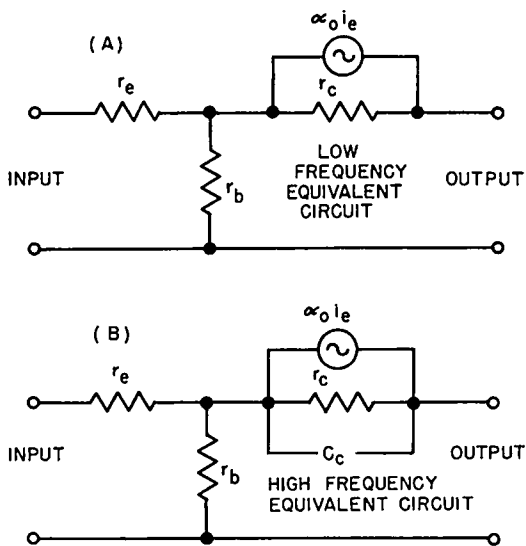
Fig. 41. Low-frequency and high-frequency equivalent circuits of transistor amplifier.

amplitude of the output signal diminishes with consequent loss of gain.

Another important factor to be recognized in discussing transistor frequency limitations is the capacitance of the collector junction. This capacitance may be regarded as a shunt across the collector resistance as shown in Fig. 41. In the equivalent circuit diagrams, r_e is the emitter resistance, r_b is the base resistance, r_c is the collector resistance, $\alpha_o i_e$ is a constant current generator in which the current is the product of the low-frequency current gain (α_o) and the emitter current (i_e), and C_c is the collector junction capacitance. C_c takes on various values depending on the transistor type and method of manufacture. For example, for a grown-junction transistor, C_c may be of the order of 8 μμf while its value may rise to 20 μμf in the case of a fused-junction type.

The effect of C_c is reduced if the source impedance is decreased to as low a value as possible. As may be seen from the equivalent circuits, the base resistance r_b is in series with the source impedance; thus, good performance at high frequencies demands that the base resistance be very low. To reduce transit time effects to a minimum, the base should be made as thin as manufacturing processes permit, but when this is done, the base resistance increases. Since these changes are opposed with respect to improving frequency response, early designs of high-frequency transistors were based upon compromises between them. Today, degradation of frequency response is avoided by incorporating new principles in transistor manufacture. Some of these will be discussed later in this chapter.

44. Effect of Transit Time on Alpha Cutoff Frequency

The time required for the current carriers to move across the base causes large numbers of them to become trapped in the base material. With the minority carriers present in the base for longer intervals, a greater number of recombinations occurs with a consequent loss in current amplification (alpha). An approximate relationship for determining the current gain under actual operating conditions is given by:

$$\alpha = \frac{\alpha_o}{\sqrt{1 + \left(\dfrac{f}{f\alpha}\right)^2}} \tag{82}$$

in which α is the current gain under operating conditions, α_0 is the low-frequency current gain, f is the operating frequency and fα is the frequency at which the current amplification is 0.707 of its low-frequency value. It will be recalled that this is the standard method of determining the significance of a loss of gain and that 0.707 represents a condition of 3 db down from the initial value.

Problem 9. Find the current gain of a junction transistor at 5 mc, if the current gain for low frequencies is 0.95 and the alpha cutoff frequency is 10 mc.

Solution.

$$\alpha = \frac{0.95}{\sqrt{1 \times \left(\dfrac{5}{10}\right)^2}}$$

$$= \frac{0.95}{1.12}$$

$$= 0.85$$

45. High-Frequency Amplifier Circuit Considerations

Selection of the transistor — Modern rate-grown n-p-n transistors such as the 2N168A, surface barrier units such as the SB103, and drift transistors typified by the 2N384 (Fig. 42) are now being utilized in high-frequency transistor amplifiers over a wide range of radio frequencies. The alpha cutoff frequencies of these typical transistors are:

2N168A	8 mc
SB103	75 mc
2N384	100 mc

In selecting a transistor for a given r-f amplifier application, attention should be given to other considerations as well as alpha cutoff frequency, although the latter is definitely the prime requisite for acceptable performance in the r-f range. For example, transistors of the 2N168A group are fabricated with extremely low collector capacitance to eliminate the need for neutralization in many circuits. The collector-to-base capacitance of the 2N168A is only 2.4 μμf as compared with 7 μμf for a 2N335, another transistor recommended for use as an r-f amplifier.

Another important factor is power gain, usually measured at 455 kc for r-f transistors. For the 2N168A the power gain is 39 db (mini-

mum) , while that of a 2N384 is 34 db. On the other hand, the collector capacitance characteristics of the drift transistor are not nearly as good as those of the 2N168A. In common with all design problems, certain compromises must be effected to realize optimum performance in all respects. The decisive factor is, of course, the minimum cutoff frequency permissible for the particular application;
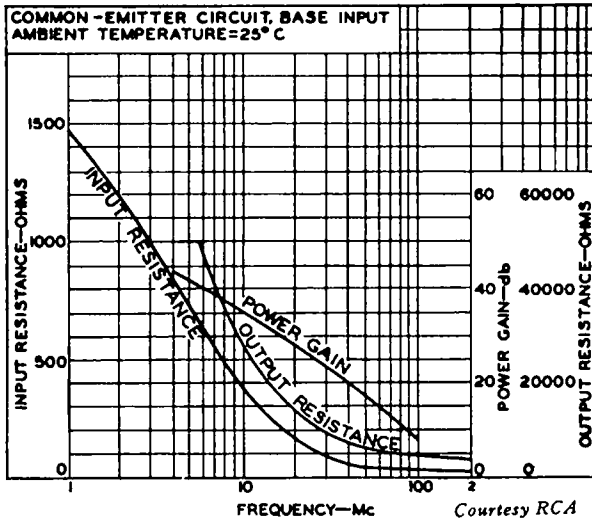
Fig. 42. Start of the junction diffusion process in the manufacture of alloy junction transistors.

from this point, the designer must look for acceptable capacitance characteristics, power gain, supply-voltage characteristics, and cost. A transistor produced by a reputable manufacturer and operated within the ratings given by the maker should, in general, provide as good performance as any other transistor designed for the same application.

Impedance matching — In r-f and i-f amplifier applications, impedance matching problems are encountered in the design of interstage coupling networks and antenna coupling systems. The groundwork for proper matching has already been accomplished by coil and transformer manufacturers so that very little of this task remains for the circuit designer. Unlike vacuum tubes, transistors in r-f amplifiers cannot be coupled through double-tuned transformer systems because their input and output impedances are considerably

lower. Hence, modern transformer design generally is based upon a single-tuned input winding and an impedance-matching output winding as will be shown in the discussions that follow.

Neutralization — This problem has been mentioned in the discussion of collector-to-base capacitance. When instability exists — even in the case of low-capacitance transistors — neutralization to prevent regeneration may be required. This is a relatively simple matter, yet it adds considerable complication in mass production of transistor radios. Often, what is correct neutralization for one transistor may not be correct for another of the same type, making re-neutralization necessary when an i-f or r-f transistor is replaced.

It might be added that the normal precautions regarding lead length and dress, shielding, etc., when working at high frequencies, must be observed with transistor equipment to the same degree as with vacuum-tube circuits.

46. Tuned R-F Amplifiers

Tuned r-f amplifiers in transistor radios can improve the gain, the overall signal-to-noise ratio, and the selectivity. In portable and automobile broadcast band radios, transistors can do as creditable a job as vacuum tubes, with substantially less power usage. A commercial type of tuned r-f amplifier used as a preselector in an automobile radio is illustrated in Fig. 43.

Certain features of this circuit, common to most transistor amplifiers for any frequency, are the base bias resistors R_1 and R_2 set up as a voltage divider in the standard manner; R_3 is the d-c stabilizing resistor which makes the circuit relatively insensitive to ambient temperature changes, and C_2 is the signal bypass capacitor to prevent degenerative effects that might reduce the gain of the stage. R_4 and C_3 together comprise a decoupling network to isolate the collector from the base of the transistor and thus minimize one source of positive feedback that might lead to instability.

The coupling between the tuned circuits and the corresponding transistor electrodes deserves special study. Mention has been made previously of the need for impedance matching. Here we see one way in which it is accomplished. By tapping down on the secondary winding of the input tuned circuit, the signal source impedance — in this case, the lower section of the tuned coil, L_2 — is made relatively small (about 500 ohms) to match the input impedance of the

transistor. The tap on L_3 is adjusted so that the lower section of this coil presents an impedance of approximately 25,000 ohms to match the collector output impedance. The coil winding data for this circuit is:

L_2 .... 145 turns, tapped 15th turn from lower end.
L_3 .... 145 turns, tapped 80th turn from lower end.

Note how many more turns are included in the load coil section (80 turns to match 25,000 ohms) than in the source coil section (15 turns to match 500 ohms).

Either a 2N94A or a 2N169A n-p-n transistor may be used in this circuit. When tuned from 500 to 1700 kc, a power gain of approximately 20 db may be anticipated. Input coil (L_1) is wound to have an impedance of approximately 500 ohms in this band of
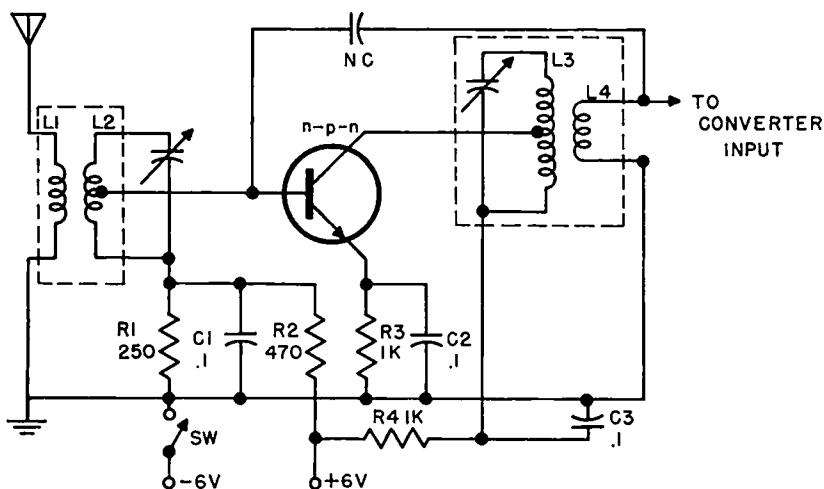
Fig. 43.   An r-f amplifier used in an automobile radio. The transistor must be a high-frequency type having a high alpha cutoff frequency.

frequencies; the same is true of the output impedance at the terminals of L_4. At about 1 mc, the bandwidth of the amplifier exceeds 20 kc at 3 db down from the maximum amplitude at the center frequency. The current drain of the tuned r-f stage is approximately 12 ma from the 6-volt battery source.

If a 2N169A is used, the neutralizing capacitor (NC) could probably be eliminated; if needed at all its value would be in the vicinity of 3.0 μμf. Note that the capacitive feedback through NC must be

out of phase with the collector-to-base feedback which tends to cause oscillation. This makes it very important to connect NC to the proper lead of L_4. In all instances where manufactured coils are utilized, the winding terminals will be found to be color-coded so that errors may be prevented.

47. Converters

Transistor radios normally make use of a converter system in which the functions of local oscillator, mixer, and i-f amplifier are combined in a single stage. Engineering practice at present tends toward a rather standard converter design as shown in Fig. 44.

The mixer (or r-f amplifier) terminates at the input end with the tuned circuit consisting of L_1 and C_1. L_1 is tapped near its "lower"
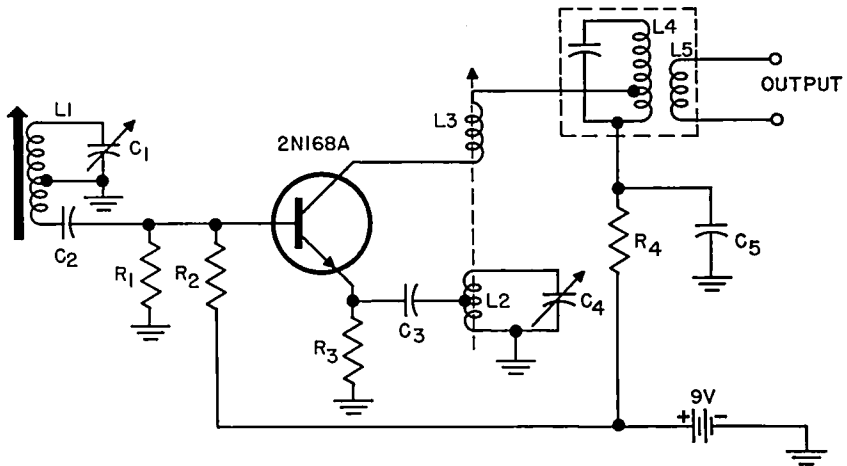
Fig. 44. Standard transistor superheterodyne converter stage. The converter combines the functions of oscillator, mixer (r-f amplifier), and i-f amplifier.

end at a point that will permit matching to the base of the transistor, a 2N168A in this case. R_1 and R_2 comprise the base-bias voltage divider with the values so chosen that the transistor is biased in a relatively low current region. By selecting proper values for R_1 and R_2 (approximately 25K for R_1 and 10K for R_2), the input characteristics of the transistor become nonlinear, thus establishing the correct conditions for mixing.

The mixer section of the 2N168A operates in the common-emitter configuration. R_3 provides the required d-c stabilization as in previous circuits studied. L_2 and L_3 taken together form the oscillatory circuit, with L_2 serving as a "tickler" winding; here again, we find a tap on L_2, placed correctly to match the impedance of the coil to the emitter circuit of the transistor.

As in all converter circuits, the output section of the stage — in this case, the collector circuit — is made resonant to the difference frequency (i-f) of 455 kc. A single-tuned i-f transformer (L_4 and L_5) is employed for reasons outlined in the previous section; the ubiquitous matching tap again appears, this time on L_4, so that the collector impedance may be properly matched to the tuned i-f circuit.

C_2 effectively brings the base of the 2N168A to d-c ground potential so that the oscillatory portion of the circuit is set up in what amounts to the grounded-base configuration. Similarly, C_3 bypasses the stabilizing resistor (R_3) to ground making the mixer section a common-emitter amplifier. R_4 and C_5 compromise the familiar decoupling network intended to prevent undesirable feedback from later stages from reaching the mixer collector circuit through the common battery impedance. For sustained oscillation, L_3 and L_2 must be properly phased to produce positive feedback.

For those interested in transformer selection or design, the input and output impedances of the transistor used become important factors. For the 2N168A used as a converter, the design center characteristics are:

Input impedance ($I_e = 1$ ma, $V_c = 5$ v, $f = 455$ kc)

$$= 400 \text{ ohms}$$

Output impedance (same constant as above) $= 12{,}000$ ohms.

48. I-F Amplifiers

Transistor i-f amplifiers differ but little in major respects from corresponding vacuum-tube amplifiers, as the circuit of Fig. 45 discloses.

The collector current is adjusted by the proper selection of R_1, R_2, and R_3. I-f amplifiers are generally operated at or very close to the class-A condition, so that these values will be different for an i-f amplifier as contrasted with the mixer described in Section 47. The

n-p-n transistor in this circuit, a 2N293, is specifically designed for i-f amplifier applications and requires resistors of the following values for class-A bias:

$$R_1 \ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{10K ohms}$$
$$R_2 \ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{81K ohms}$$
$$R_3 \ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{470 ohms}$$

The collector capacitance for this transistor is sufficiently small (2.4 $\mu\mu$f) and controlled in maufacture to such low variation tolerances that neutralization is unnecessary to obtain an i-f amplifier with relatively high gain and excellent stability. In a typical i-f test
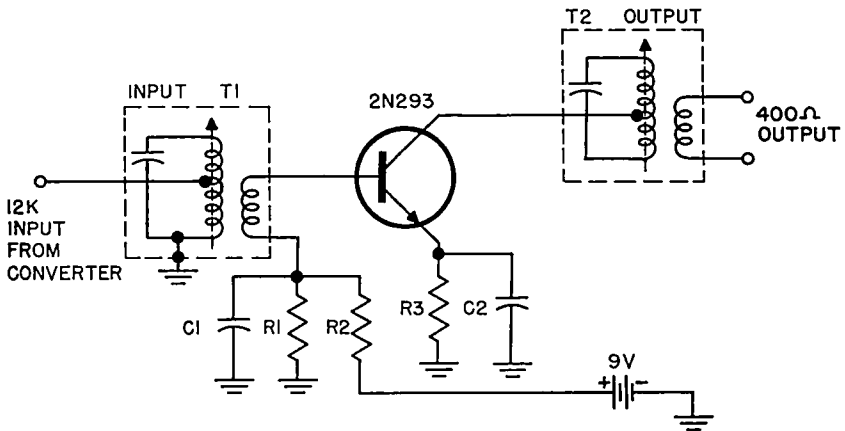
Fig. 45. A representative i-f amplifier for a broadcast band superheterodyne receiver.

circuit, the minimum power gain was found to be 28 db for several production units tested. The average power gain as given by the manufacturer is 39 db, while the power-gain range of variation in representative i-f circuits is stated as 3 db.

Except for broadcast radios used in the vicinity of transmitting stations, as in large cities, the i-f amplifier just described does not generally provide sufficient gain for satisfactory operation. Power gains up to 60 db are easily realized by employing two transistors in cascade in the i-f amplifier section. Each of these stages may be designed according to the circuit of Fig. 45; no circuit value changes are required when two stages are cascaded, provided that the i-f transformers permit proper impedance matching at both input and

output. Neutralization is necessary only when transistors having relatively high collector-to-base capacitances are used. As previously stated, rate-grown n-p-n units with low capacitance are now commercially available. On the other hand, p-n-p transistors have larger inherent electrode capacitances and do require neutralization as a general rule.

49. Automatic Gain Control of R-F and I-F Amplifiers

Transistor radios are subject to the same changes of received signal strength as are vacuum-tube receivers. To compensate for carriers of varying power and the objectionable effects of fading, transistor receivers must be equipped with some form of automatic gain control.

Several different circuit configurations are possible. In each type, a portion of the detector output is converted into a steady d-c voltage proportional to the strength of the received carrier. This voltage is then applied to one or more i-f (or r-f) amplifier stages to control
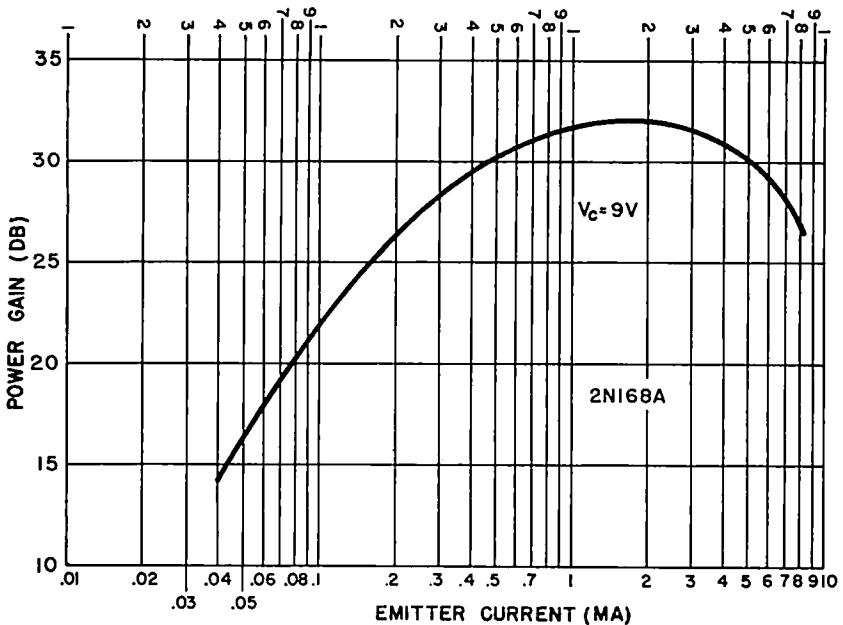
Fig. 46. Curve showing power gain as a function of emitter current. Using only this system for agc, best control is obtained between 1 ma and 0.1 ma of emitter current.

their gain. In vacuum-tube amplifiers, the control voltage is applied as a negative potential to the tube grids, thus varying their trans-conductances and hence their voltage amplification. A transistor r-f or i-f stage must be treated somewhat differently.

As the emitter current of a common-emitter stage is reduced, the power gain diminishes. This characteristic is widely used in commercial receivers as the basis for the design of agc systems. Referring to Fig. 46, it may be seen that the power gain of a typical converter/i-f transistor drops sharply when the emitter current is reduced from 1 ma down to values in the vicinity of 0.1 ma. By feeding back a negative potential from the output of the detector, a fair amount of agc is realized.

A typical circuit in which emitter-current agc is used to control the performance of a broadcast receiver is shown in Fig. 47. For the moment, ignore the auxiliary diode (D) and its connecting leads. A negative voltage from the detector is filtered and smoothed
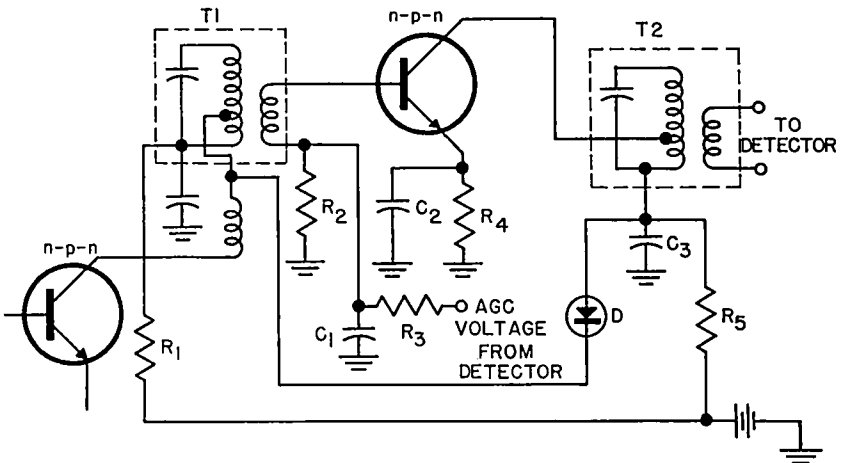
Fig. 47. Emitter-current agc system. The diode, labeled D, is used to ensure good agc action when the signal level is high.

by R_3 and C_1. This potential is then applied to the base of the i-f transistor via the secondary winding of T_1. The larger this potential becomes, the greater is the diminution of gain as a result of the reduction of emitter current with increasing negative base voltage.

Unfortunately, this form of agc works well only with medium to weak signals but fails completely when the signal level is high. This

effect is the result of the flow of collector leakage current (I_{co}), making it impossible to cut a transistor off completely. Hence, when the signal level exceeds a definite value, agc is no longer effective and overload distortion is apt to set in.

Diode D is connected into the circuit to prevent this undesirable limitation from having a serious effect. Initially D is reverse biased by the voltage drops across resistors R_5 and R_1. For weak and medium signals, D does not conduct and so represents an open circuit across the primary of T_1. Under these conditions, D has no effect on circuit performance. As the signal level rises, however, the emitter-current agc system reduces the current, thereby causing a reduced voltage drop across R_5. When the signal has risen sufficiently, D begins to conduct and shunts a portion of the signal to ground; the net effect is that the input signal to the i-f transistor is reduced and further agc action occurs. The use of combined emitter-current and diode agc provides good control over a very wide range of signal strengths.

53. Review Questions

1. Why is transit time an important consideration in the design of transistor circuits for high-frequency amplification?
2. Explain why the capacitance of the collector junction should be made as small as possible for high-frequency transistor applications.
3. Determine the current gain of a transistor at 10 mc if it has the following low-frequency characteristics: $\alpha = 0.92$, $f\alpha = 12.5$ mc.
4. Explain why the alloy-junction process is capable of producing very thin layers of base material.
5. Why is the surface-barrier transistor capable of good performance at high frequencies?
6. What specific constructional feature of a drift transistor enables it to perform well at very high frequencies?
7. Why is it common to find coupling transformer coils in transistor r-f and i-f amplifiers tapped?
8. Explain why neutralization is required in some r-f and i-f amplifiers and not in others.
9. What is the basis for most agc systems in transistor radios?
10. What is the limitation imposed on the effectiveness of agc by I_{co}? How is this limitation overcome?

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