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LOW-FREQUENCY AMPLIFIER SYSTEMS

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LOW-FREQUENCY AMPLIFIER SYSTEMS

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PREFACE

The electronic devices and circuitry which accomplish low-frequency amplification contain one or more vacuum tubes or transistors to which are attached appropriate components. The resultant system increases the strength of an electrical signal. Low-frequency amplifiers may consist of a single stage (one vacuum tube or one transistor) or several stages. Many types of electronic apparatus utilize these types of amplifiers to raise the strength of extremely weak input low-frequency signals so that they can be used to accomplish any one of a number of desired end effects. For this reason, it is essential that those concerned with electronics possess a working knowledge of the essential relationships pertaining to this classification of amplifier.

The intent of this book is to extend and further evaluate low-frequency amplifier systems, giving additional attention to the electronic problems of the low-frequency spectrum beyond that offered in a companion volume of this series.* 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 presentation 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 general considerations of low-frequency amplifier systems, low-frequency attenuation, high-frequency attenuation, problems relating to input capacitance, gain

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*Schure, A., *Low-Frequency Amplifiers,* New York: John F. Rider Publisher, Inc., 1959.

in the middle frequencies, transformer-coupled amplifiers, direct-coupled voltage amplifiers, phase-inversion methods, cathode followers, the meaning of inverse feedback, reduction of distortion by inverse feedback, voltage feedback, amplifier-gain inverse feedback, current feedback, low-impedance computations, distortion, bias voltage considerations, volume compression and expansion, pre-emphasis and de-emphasis, decoupling considerations, square-wave analysis, control of frequency response, applications of low-frequency amplifiers, general considerations of vacuum-tube amplifier design, design problems, design procedure for resistance coupled pentode amplifiers, design considerations for transformer-coupled low-frequency amplifiers, higher audio-frequency amplification, lower audio-frequency amplification, mid-range gain reduction, transistor parameters and circuit factors, points of reference, design considerations relating to transistor bias, bias design equations, problems and examples of actual design procedures relating to low-frequency amplifier transistor systems, determination of operating characteristics, class-A transistor amplifier output-stage design, and class-B push-pull transistor output stages. Thus, a foundation is provided upon which further concepts relating to low-frequency amplifier systems 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.

New York, New York
September, 1959

A. S.

CONTENTS

Chapter 1

FUNDAMENTAL COUPLING METHODS IN LOW-FREQUENCY SYSTEMS

1. General Considerations

Low-frequency amplifiers are seldom used singly, particularly in audio systems where the signal voltage is expected to be converted to power for operating a reproducer of some kind. The gain multiplication possible when a signal already amplified by one or more tubes is used as the input for a succeeding amplifier can be an advantage. Aside from certain losses that occur in the coupling systems, the total voltage gain available from such cascaded stages is the product of the individual stage gains.

In establishing any system for coupling the output of one tube to the input of another, it is essential to prevent the direct plate voltage of the first tube from disturbing the grid bias of the second. Yet the coupling system must be able to pass the signal voltage with a minimum of amplitude and phase modification with various frequencies.

Low-frequency coupling systems include direct coupling, iron-core transformer coupling, capacitive-resistive coupling, and capacitor-inductor (*impedance*) coupling. All of these ordinarily produce frequency distortion of the signal to some extent. The degree to which this distortion can be reduced is the criterion of a coupling network's performance. In some applications, it is necessary to modify the signal in some way, as well as to amplify it, thus presenting another type of problem to the circuit designer.

2. R-C Coupled Voltage Amplifiers

In this arrangement a capacitor of suitable size is used to perform the double function of blocking the d-c at the plate circuit of the first tube from the grid of the second and of passing the a-c component of the signal on to the second grid circuit. Figure 1 shows the various components generally found in a two-stage amplifier. The following symbols used to designate the components are conventional ones.

The coupling capacitor C_c is sometimes called a *blocking capacitor* because it offers a very high impedance to dc. The reactance of C_c must be low with respect to R_g for all frequencies to be amplified, since these two components form a voltage divider across

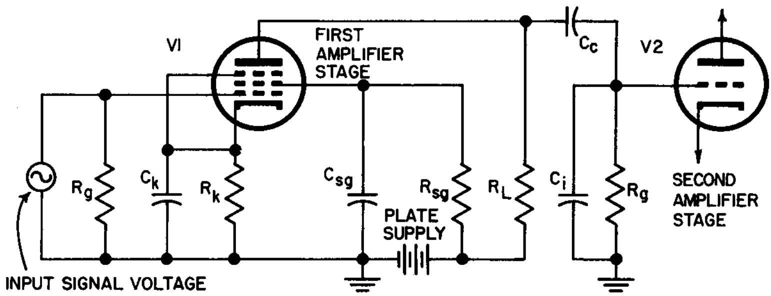


Fig. 1. A standard pentode R-C coupled circuit.

the output of the previous stage. Unless C_c has low impedance, a voltage drop representing a loss will appear across it, and this voltage will not reach the grid of the second tube. Obviously, the signal voltage drop that appears across the second R_g is the voltage applied to the input of the second amplifier.

Assuming that all components have been selected with a view to optimum transfer of signal energy from the output of one stage to the input of the other, our principal concern then becomes one of determining the behavior of this coupling circuit in terms of frequency response. We shall examine one at a time the factors that contribute to the kind of frequency-response curve typical of an R-C coupled circuit as illustrated in Fig. 2.

Ordinarily, without coupling capacitor C_c and the second grid resistor R_g (Fig. 1) in the tube circuit, the amplification of the stage can be calculated for a triode from equation (8) and for

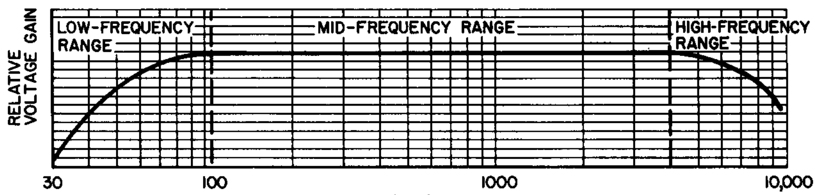


Fig. 2. Frequency response of a typical R-C coupled circuit.

a pentode from equation (20). Observe that frequency as a factor for consideration does not appear in either of these equations. Hence, over a frequency range of moderate width—such as the entire audio spectrum—a single-tube amplifier may be considered essentially flat in response. The drop-off in relative gain at both ends of the audio band, as shown in Fig. 1, must therefore be due to adverse effects of the coupling system comprising C_c, R_g, and the grid-cathode-plate effects in the second tube. Let us examine separately the low- and high-frequency attenuations evident from the curve.

3. Low-Frequency Attenuation

The equivalent circuit of the pentode amplifier presented in Fig. 1 is shown in Fig. 3 as it appears to a low-frequency signal. Note that many of the component parts present in the schematic diagram (Fig. 1) are not illustrated at all in the equivalent low-frequency diagram (Fig. 3). These parts are not required in the analysis of equivalent circuit operation since they play little or no part in the action at low frequencies. C_k and R_k are bias components across which a fixed voltage is developed; C_{sg} and R_{sg} are specifically intended to establish correct d-c screen conditions, and they have a negligible effect on amplifier performance at very

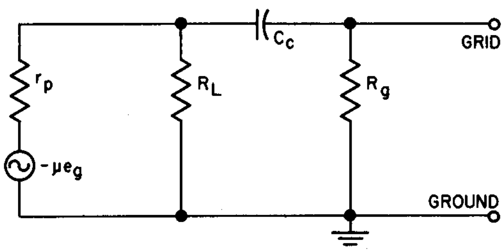


Fig. 3. The low-frequency equivalent circuit of the amplifier shown in Fig. 1.

low frequencies; finally, C_i, illustrated in Fig. 1 as an "unreal" component (it is not an actual capacitor but consists of the combined capacitive actions of other parts of the grid circuit), may be ignored at low frequencies. C_i will be discussed in Section 4. To simplify the discussion, we assume that Z_L is a purely resistive load R_L in the equivalent circuit.

The branch consisting of C_c and R_g is in parallel with the load resistance R_L and will therefore reduce the voltage gain obtainable from this stage. In order to minimize this loss, it is advantageous to make R_g as large as possible. But R_g is in series with the grid, serving as the grid-to-ground return. Because there is always a possibility of small grid currents, the tube manufacturer generally specifies a maximum value for this resistor, varying from 1 or 2 megohms down to 100,000 ohms for some tubes. Once R_g has been selected, C_c must then be considered. The signal-voltage drop across C_c must be made as small as possible so that the drop appearing across R_g will correspond as closely as possible to the drop across R_L. Thus, the capacitive reactance of C_c *at the lowest frequency for which the amplifier will be used* must be quite small as compared with R_g.

It is generally agreed that a 1-to-10 ratio of X_c to R_g is large enough to keep the loss at the chosen frequency inappreciable. Suppose that the amplifier system must meet the specification of good response right down to 20 cps and that the grid resistor of the following stage R_g has been made equal to 500,000 ohms. The capacitor needed to fulfill this requirement may be found as follows. X_c must be no greater than 50,000 ohms at 20 cps; thus:

$$X_c = \frac{1}{2\pi fC}$$

$$50{,}000 = \frac{1}{6.28 \times 20 \times C}$$

$$C = 0.16\ \mu f \tag{1}$$

A capacitor of this large size is seldom used in audio coupling circuits for four reasons: (1) large coupling capacitors encourage *motorboating* and other forms of amplifier instability; (2) a physically large capacitor increases hum and noise pickup; (3) large capacitors are more prone to develop d-c leakage of serious magnitude. Such leakage permits a portion of the d-c voltage from the preceding plate to be impressed on the grid of the coupled tube, thus changing the bias conditions and the operating point of the tube. As we have seen, incorrect bias leads to intolerable distortion. (4) Because of its bulk, it has considerable self-capacitance which adds to the effective value of C_i.

When an amplifier consists of several coupled stages, the reactance of C_c at low frequencies becomes even more detrimental to low-frequency response. The deficiencies of coupled stages are multiplied by cascading. For example, if a stage gain at 20 cps drops 10% due to the reactance of C_c and there are three such stages, then the total loss in response may be found as follows. 10% loss of lows means that 0.9 is the relative low-frequency response of the amplifier compared to its *middle-frequency* response. For three stages, the response then is $(0.9)^3 = 0.73$. Hence, the low-frequency response (at 20 cps) is only 73% of the response for the middle frequencies.

4. High-Frequency Attenuation

As the frequency of the signal in an ordinary R-C coupled amplifier rises, the reactance of C_c takes on a rapidly diminishing significance since X_c is inversely proportional to f. As the middle frequencies are reached, C_c behaves like a virtual short-circuit and may be dropped from consideration entirely. At the same time, another source of loss begins to appear; this is C_i, shown in Fig. 1. C_i is the effective capacitance shunting the input to the second stage and represents a path by which the signal can pass off to ground without appearing as a voltage drop across the grid of of the tube. C_i is the combination of: (1) stray wiring capacitance in the grid circuit, (2) grid-to-plate interelectrode capacitance, (3) grid-to-cathode interelectrode capacitance, and (4) *Miller-effect capacitance.* Miller-effect capacitance is due to the influence of amplification on the input capacitance of a tube. As a result of the Miller effect, the input capacitance of any amplifier tube is not merely the sum of the grid-plate capacitance C_{gp} and grid cathode capacitance C_{gk} but is given by the expression:

$$C_i = C_{gk} + (1 + A)C_{gp} \qquad (2)$$

where A is the voltage amplification of the tube. To appreciate the size that C_i may take on, consider the following example.

Problem 1. What is the input capacitance of a triode, assuming a stray input wiring capacitance of 1.0 micromicrofarad ($\mu\mu$f) and a voltage gain of 20?

Solution. The tube manual provides this information for the tube:

$$C_{gk} = 4.2\ \mu\mu f$$

$$C_{gp} = 3.8\ \mu\mu f$$

$$C_i = 4.2 + (1 + 20) \times 3.8 + 1.0$$

$$= 85\ \mu\mu f$$

The important of Miller-effect capacitance is evident from this example. Of the total C_i of 85 $\mu\mu$f, almost 80 $\mu\mu$f may be traced directly to the influence of amplification on input capacitance. This clearly illustrates why grid-plate capacitance, although ineffective directly across the input from grid to cathode, must be held to a minimum if C_i is to be repressed.

The equivalent circuit of our sample amplifier at high frequencies is shown in Fig. 4. Note how C_i shunts the grid circuit, being increasingly effective as a source of loss at the high frequencies where XC_i becomes smaller and smaller. Since the effective shunting capacitance is a function of the μ of the tube, the greater the amplification factor the more trouble is caused by the Miller

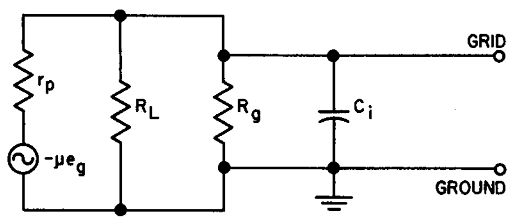


Fig. 4. High-frequency equivalent circuit of amplifier shown in Fig. 1.

effect. Low-μ tubes are not troubled to nearly the same extent, but their amplification is low. Thus, the amplifier designer must compromise between an excellent frequency characteristic and a high gain unless he wishes to use correction and equalization circuits.

Problem 2. What is the input capacitance of a 6CB6 sharp-cutoff pentode, assuming a stray wiring capacitance of 1.0 $\mu\mu$f?

Solution. The interelectrode capacitances for this tube are given as:

$$C_{gk} = 6.3\ \mu\mu f$$

$$C_{gp} = 0.020\ \mu\mu f$$

$$A = 3720\ \text{(approx)}$$

$$C_i = 6.3 + (1 + 3720) \times 0.02 + 1.0$$

$$C_i = 81.7\ \mu\mu f$$

These figures answer a question that often arises in connection with the extent of the Miller effect as applied to pentodes with their very much reduced grid-plate capacitance: the very high gain cancels the low grid-plate capacitance, thereby bringing about an input capacitance of the same order of magnitude for the pentode as for the triode. The reader will appreciate therefore that pentodes could not be used as high-fidelity amplifiers if their grid-plate capacitance were as high as that of triodes. High C_{gp} coupled with a high-amplification factor would make these tubes useless for anything but very low-frequency amplification where even a large C_i would have little effect on frequency response.

5. Gain in the Middle Frequencies

From the previous discussion, it is evident that the gain of any uncompensated amplifier will tend to be greatest in the middle range of frequencies—from 1000 to 4000 cps—where neither C_c nor C_i causes much trouble. When the load is purely resistive, the following equation may be used for finding middle-frequency gain:

$$\text{Voltage gain} = \frac{G_m R_L}{1 + \dfrac{R_L}{R_g} + \dfrac{R_L}{r_p}} \tag{3}$$

This equation is based upon the fact that the voltage drop across the coupling capacitor C_c is negligible and the bypassing effect of C_i is very small at these frequencies. In this case, the voltage gain equals $G_m\, R_{eq}$. R_{eq} is used instead of merely R_L because the grid resistor and the plate resistance are all in parallel with R_L, as in the equivalent circuit of Fig. 5. (The input signal is shown coming

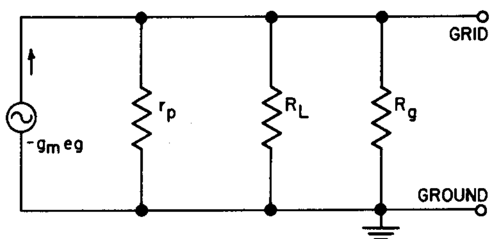


Fig. 5. The equivalent circuit of an amplifier for the middle frequencies.

from a "constant-current generator," in which r_p is indicated in parallel with the generator whose output current equals $-G_m e_g$.)

6. Transformer-Coupled Amplifiers

From the standpoint of the design engineer, there is an appreciable advantage in the use of transformer coupling between the stages of a low-frequency voltage amplifier. First, there is a negligible plate-supply voltage drop through the windings of a well-designed transformer so that the supply voltage may be more moderate; second, transformers in themselves may have voltage gains which multiply the overall stage gain. Figure 6 illustrates the general circuit conditions in a transformer-coupled two-stage

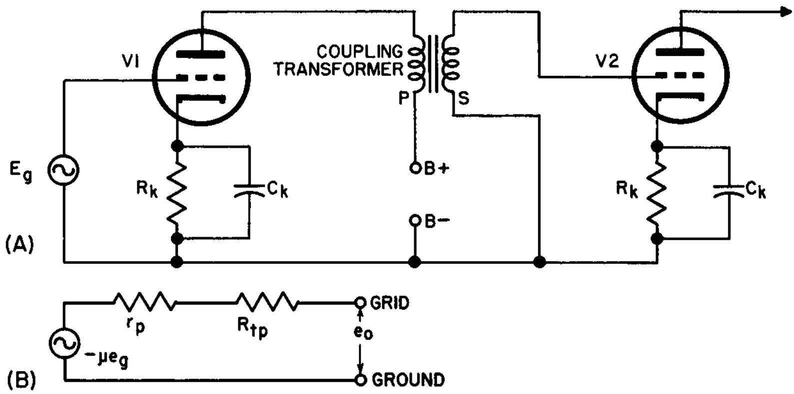


Fig. 6. (A) Circuit of a two-stage transformer-coupled amplifier. (B) Equivalent circuit of transformer-coupled amplifier at middle frequencies.

amplifier, together with the equivalent circuit for medium frequencies between 1000 and 4000 cps.

In the equivalent circuit, the plate resistance of the tube r_p is in series with the d-c resistance of the transformer primary winding R_{tp} and the output voltage of the system e_o is given by:

$$e_o = \mu e_g \frac{n_s}{n_p} \tag{4}$$

in which n_s/n_p is the secondary-to-primary turns ratio of the transformer. In most voltage amplifiers, this ratio is greater than unity so that a voltage stepup is realized. Equation (4) is applicable only if the reactance of the transformer winding is large compared to its d-c resistance and to the plate resistance of the tube. From this, we may conclude that transformer coupling may be quite effective for triodes of relatively low plate resistance but cannot be used successfully when the tubes are pentodes. Another important conclusion is that transformer coupling fails at low frequencies (below 100 cps) because the primary reactance may fall to a very low figure as the frequency is lowered.

At higher audio frequencies, voltage gain again diminishes, this time as a result of the shunting effect of the distributed capacitances of the primary and secondary windings of the transformer. Analysis shows that the drop-off rate at low frequencies due to the falling reactance of the transformer primary is about as severe as the corresponding drop-off in R-C coupled amplifiers. On the other hand, the high-frequency attenuation is much greater in transformer-coupled voltage amplifiers. Taking all these differences

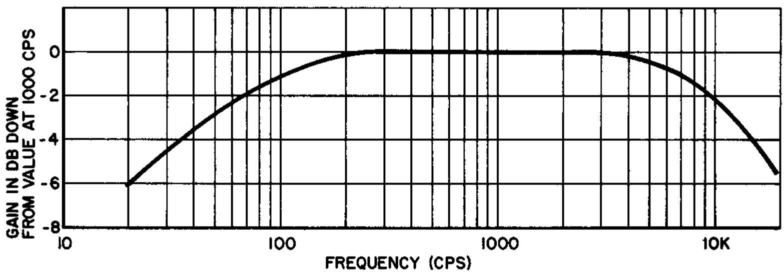


Fig. 7. Attenuation curve of a low-cost replacement audio transformer.

into account, it is understandable why R-C coupling is used whereever possible. Transformer coupling is popular, however, in certain power-amplifier circuits where the control-grid resistance must be kept very low or where impedance matching between two very unequal impedances is required. Figure 7 shows the high- and low-frequency attenuation pattern of a typical "replacement" quality audio transformer.

The difference between a poor transformer and a high-grade one is most evident in its response at high and low frequencies. Good low-frequency response requires large primary inductance; good high-frequency response demands skillful reduction of the distributed capacitances of both windings. Both tend to require more expensive design and construction, and larger size. By their very nature, transformers tend to pick up hum voltages and often require careful orientation with respect to each other and various

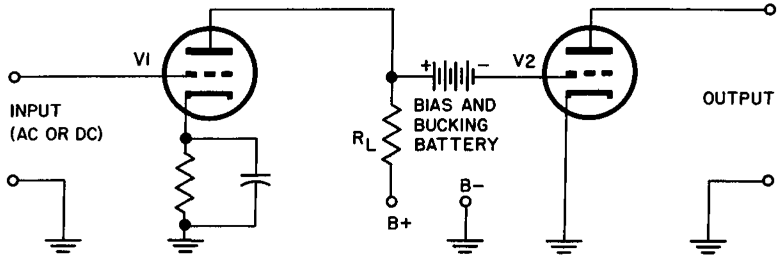


Fig. 8. Bucking battery method of direct coupling.

other components on the amplifier chassis. For this reason, R-C coupling is favored for high-gain, sensitive low-level stages. Transformers find their widest application in output and power stages.

7. Direct-Coupled Voltage Amplifiers

The coupling methods described in the previous paragraphs are equipped to prevent the positive potential on the plate of the first tube from reaching the grid of the second amplifier. In capacitive coupling, the capacitor passes only the a-c component of the signal; thus, the grid of the second tube may be returned to ground through a suitable resistor. Transformer coupling accomplishes the same objective by making use of electromagnetic field transfer of energy from the plate of one tube to the grid of the next without transferring, at the same time, the d-c potentials present.

Both of these coupling systems, however, are troubled by undesirable frequency characteristics, as we have shown. Furthermore, neither of them can be used in d-c amplifiers for operating relays, recording styli in electrical measuring instruments, or other applications where a small direct signal current is to be built up into a larger direct current or voltage.

In the past the circuit illustrated in Fig. 8 was suggested as a possible answer to the direct-coupling problem. The coupling capacitor, as is evident by inspection, is replaced by a bucking battery in the grid circuit. This battery must have sufficient potential to cancel the positive plate voltage and to supply the bias voltage required by the second tube. Let us see what is wrong with this circuit.

The voltage drop across resistance R, which is common to the plate circuit of the first tube and the grid circuit of the second, is not constant with changing input. As the signal changes, the plate current of the first tube varies, causing the voltage drop across R, and hence the bias of the second tube, to change. This moves the operating point of the second amplifier so that distortion is quite likely to set in. In addition, except for laboratory instruments, a battery of the size needed here would be completely impractical.

A substantial improvement in direct-coupled design is shown in Fig. 9. To use this method, it is necessary to provide a separate bias supply of relatively large value as shown in the circuit diagram. A voltage divider is placed between the plate of the first amplifier and the negative end of the auxiliary supply. Any voltage variation will be transmitted from the plate of the first tube to the grid of the second, but the grid swing will not be as large as its plate-variation counterpart. If, for example, the values of the two resistors R1 and R2 are equal, then the grid swing will only be half the magnitude of the plate-voltage change. This is not always a prohibitive disadvantage, however, since the available signal is often larger than need be. Although the voltage of the bias supply is sometimes almost as great as the main supply voltage, the circuit has the important advantage that both power supplies have a common ground. Thus, both potentials may be obtained from the same rectifier-filter system. Also, if more stages are added, the same bias supply will suffice for these. The static voltages shown in Fig. 9 are merely examples to indicate relative potentials. With

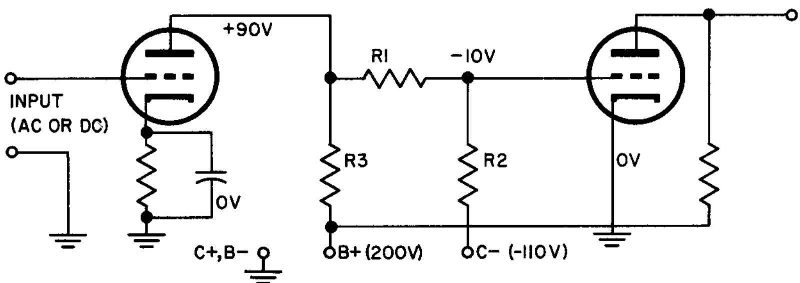


Fig. 9. Auxiliary bias method of direct coupling.

given supply potentials, R1, R2, and R3 must be carefully selected — in conjunction with tubes V1 and V2 — to provide the proper anode and grid potentials, of course.

In the circuit just described, the cathodes of the amplifiers were held at close to the same d-c potential. Another method (Fig. 10) of direct coupling, originated by E. H. Loftin and S. Y. White, approaches the problem from a different point of view. Instead of attempting to bring the variations of the plate potential down to the appropriate level for the next stage, the grid of the second

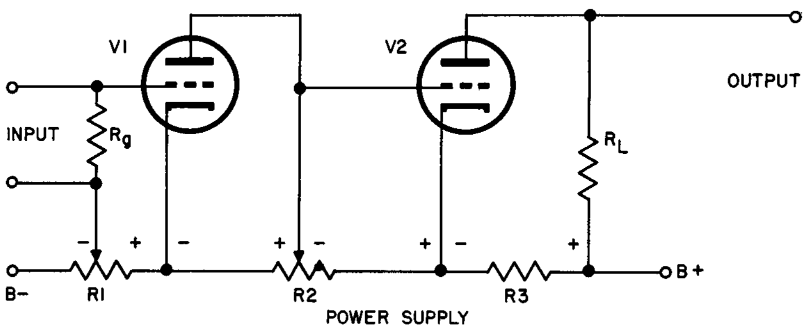


Fig. 10. The Loftin-White direct-coupled basic circuit.

tube is simply connected to the plate of the first tube, but the cathode of the former is raised positively so that the grid will be negative with respect to it. One method of accomplishing this is illustrated in Fig 10. A voltage divider is connected across the main power source. Connections from the various grids, plates, and cathodes are made to points on this divider so that the cathode is always positive with respect to the grid of the same tube by an amount equal to the required d-c bias. Also, the plates are joined to the same voltage divider so that they are sufficiently positive with respect to their cathodes to the extent dictated by the tube characteristics.

One fundamental defect of all direct-coupled circuits is the *drift error.* Since the second stage responds to any d-c variation on its grid, a slight drifting of d-c voltages due to warmup or other causes in the first stage is amplified and appears as a signal to the second tube. In the first place, this limits the number of stages that can be used in cascade since the drift error becomes intolerable soon after a few successive amplifications. Many attempts at compensation and stabilization have been made but the problem is by no means solved. Two-stage direct-coupled amplifiers are now in common use in the input phases of high-fidelity amplifiers where balanced circuits help to cancel out the effects of drift. Except for these, d-c amplifiers of more than a single stage are still confined to the laboratory where they may be adjusted by experienced personnel. For commercial and industrial use, where the amplification of direct currents is necessary, the d-c input is usually *chopped* up into a pulsating d-c input, the a-c component of which is then amplified and used in the output. Although this appears to be a roundabout method, the stability possible in R-C coupling more than makes up for the added complexity.

8. Review Questions

1. Explain what difficulties present themselves when direct coupling between two amplifier stages is attempted.
2. Which portion of the audio-frequency spectrum is affected by an interstage coupling capacitor that is too small in capacitance? Explain.
3. Describe the Miller effect. What part of the audio-frequency spectrum is affected by the Miller effect in a high-gain vacuum-tube amplifier?
4. What is the effect of d-c leakage in a coupling capacitor?
5. The grid resistor of the second amplifier of a cascaded pair is 1 megohm. What is the smallest coupling capacitor that will provide good frequency response at 50 cps?
6. Explain why the careful selection of the coupling capacitor is increasingly important as the number of cascaded voltage amplifier stages is increased.
7. How is transformer coupling superior to R-C coupling? How is it inferior?
8. Explain why transformer-coupled amplifiers suffer from low-frequency attenuation. What is one cause of high-frequency attenuation in transformer coupling?
9. Describe the Loftin-White direct-coupling system, using a diagram to help you.
10. Explain why d-c amplifiers are not yet popular for consumer equipment.

Chapter 2

PHASE INVERSION AND INVERSE FEEDBACK

9. Phase-Inversion Methods

The connection of a single-ended driving stage to push-pull operated tubes requires either an interstage transformer with a centertapped secondary or some equivalent means of providing two signal voltages of equal magnitude 180° out of phase.

Transformerless phase inverters make use of the phase-reversal characteristics of vacuum tubes. Although countless circuit variations are available in vacuum-tube literature, the two arrangements described below are the most common and illustrate the general approach to the problem.

Figure 11 shows a twin-triode type of phase inverter; this circuit may be adjusted to give excellent grid-input balance to the push-pull grids. Inversion action is provided by triode T2. The output voltage of T1 is applied to the grid of T3; a portion of the same output voltage is also applied through resistors R3 and R5 to the grid of T2. The output voltage of T2 is applied to the grid of T4. When the output voltage of T1 swings positively, the plate current of T2 rises. This increases the voltage drop across the load resistors R3 and R5 and causes the plate of T2 to swing in a negative direction. Thus, as the output voltage of T1 becomes positive-going, the output voltage of T2 becomes negative-going and is, therefore, 180° out of phase with the output voltage of T1. To realize equal voltages at both push-pull grids, then $(R_5 + R_3)/R_5$ must equal the voltage gain of T2.

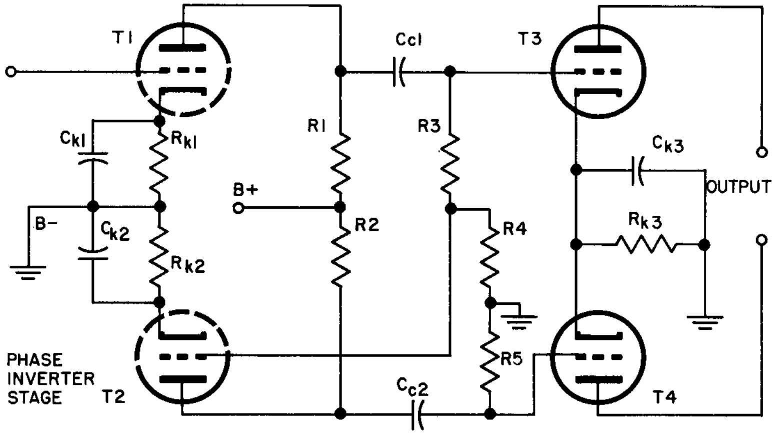


Fig. 11. Twin-triode type of phase-inverter circuit. This circuit may be adjusted to give excellent grid input balance to the push-pull grids.

Since most modern phase inverters take advantage of the convenience of twin triodes for T1 and T2 and since the balance of characteristics is generally quite good in this type of tube, R4 should be made equal to the sum of R3 and R5. The value of R5, for a good first approximation in design, should be made equal to R4 divided by the voltage gain of T2. Also, R3 should be equal to R4 − R5.

Figure 12 shows a single-tube phase-inverter circuit; with only one tube, it does not have the gain of the two-tube circuit (Fig 11).

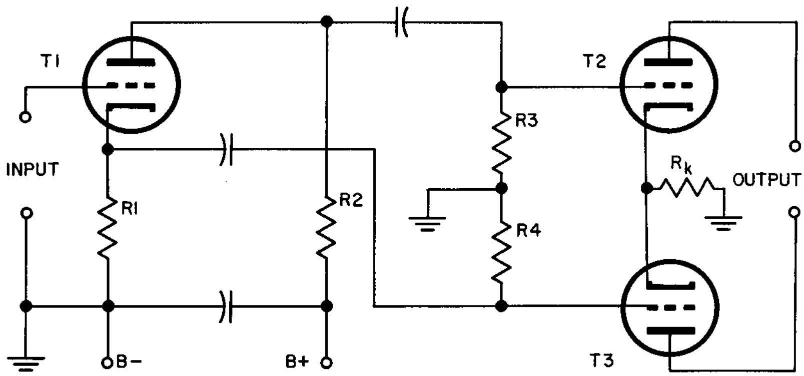


Fig. 12. Cathode resistor type of phase inverter.

Resistors R1 and R2 are relatively large and are equal in value. As the grid-input signal to T1 becomes positive-going, the plate current increases, causing the voltage drop across R2 to increase as well. This applies a negative-going voltage to the grid of T2. At the same time, as the voltage drop across R1 increases for the same reason, the top of this resistor becomes more positive. Since the grid of T3 is connected to this point through the coupling capacitor, it experiences a positive-going voltage. Thus, the grids of T2 and T3 are being driven 180° out of phase to meet the requirements of push-pull operation.

A phase inverter of this type is bound by the physical laws of *cathode followers.* When output is taken from the cathode resistor, it may be shown that this output voltage can approach, but never equal, the input voltage. Instead of obtaining gain, then, a slight voltage loss is experienced in the phase-inversion process. If the stages that precede the phase inverter have built the driving voltage up to, or slightly past, the point where full output from the push-pull stage may be realized, this loss is inconsequential. On the other hand, the circuit of Fig. 11 does provide gain and is more often used in standard amplifiers for this reason.

10. The Meaning of Inverse Feedback*

A successful amplifier always has a greater signal voltage in its output circuit than in its input circuit. When a portion of this amplified voltage is reintroduced into the input terminals, the voltage is said to be "fed back," and the process is called *feedback.*

Both the amplitude and the phase of the fed-back voltage are important in determining what is to occur next. If the feedback is *positive,* that is, in phase with the incoming signal, the input voltage is strengthened and again amplified. Since the feedback process is continuous, the same percentage of the newly amplified voltage will again be fed back to further strengthen the input. Obviously this is a cumulative process, the input signal growing stronger with every cycle. After a very short interval, the tube will be overloaded and may break into uncontrolled oscillation at a frequency determined by the constants of the circuit (R, L, and C).

*See Schure, A., *Inverse Feedback*, New York: John F. Rider Publisher, Inc.

The voltage fed back from the output to input may just as well be 180° out of phase with the incoming signal. This is termed *negative* feedback, since it bucks the incoming signal, effectively reducing the output. If, however, the output diminishes, the feedback voltage diminishes also. Therefore, unlike positive feedback, negative feedback is not cumulative, is self-limiting, and can contribute to amplifier stability in that any tendency of the output amplitude to increase due to random effects is immediately offset by increased negative feedback that tends to maintain the *status quo*. Negative feedback is often referred to as *inverse feedback*. The term *degeneration* is also categorically applied by some authors to the condition of negative feedback, although it should be reserved for one particular form of feedback, as we shall see.

11. Reduction of Distortion by Inverse Feedback

An audio-amplifier stage, no matter how well designed, always contains spurious frequencies in its load; these are produced within the stage itself. We have seen how incorrect bias gives rise to harmonics that are not present in the input voltage to the audio amplifier; overdriving the amplifier has similar effects. Thus, it may be stated unequivocally that every audio amplifier delivers (a) an amplified version of the grid input, whatever it may be, and (b) new voltages that originate in the amplifier stage, including hum, noise, and harmonic distortion.

A significant improvement in response may be realized by introducing measured amounts of inverse feedback, that is, by feeding back a part of the output voltage to the grid circuit out of phase with the input signal. As an example, consider what occurs when 20% of the total output voltage is used for feedback; we must emphasize that this *total* 20% includes 20% of item (a) and 20% of item (b) above. As both voltages are fed back out of phase, both (a) and (b) will be reduced in magnitude. However, the amplified version of the grid input [item (a)] may be restored to its former level by increasing the *input to the grid of the amplifier* without proportionately increasing the level of the undesired voltage [item (b)]. This latter statement is true for two reasons: (1) if the distortion originates in the amplifier stage, an increment in the input voltage cannot cause it to rise in level; (2) the stage gain is reduced by inverse feedback so that amplification of a voltage of internal origin is not as great. An equation for calculating gain reduction will be given later.

The circuit of Fig. 13 illustrates one form of inverse feedback known as *voltage feedback*. It derives its name from the fact that the amount of feedback, or feedback percentage, is proportional to the output signal voltage. The feedback voltage is obtained from a voltage divider comprising R1 and R2. Although C is introduced to prevent the high plate voltage from changing the

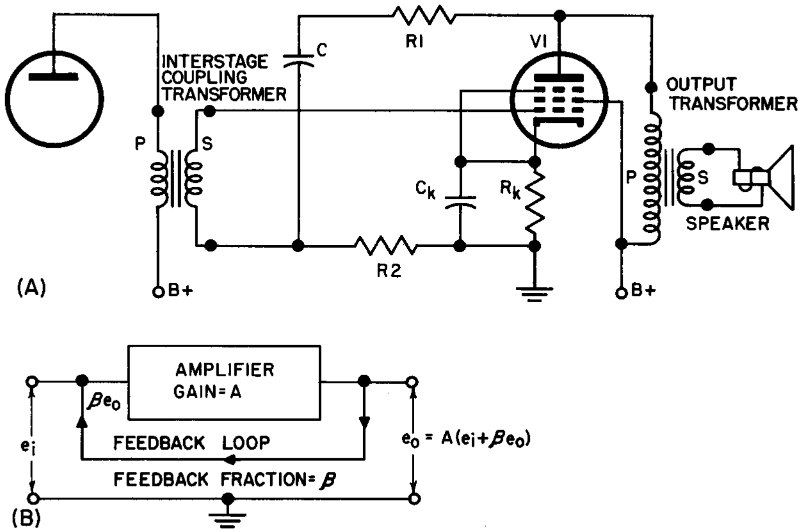


Fig. 13. Voltage feedback. (A) Actual basic circuit. (B) Equivalent circuit.

grid bias, it plays no other significant part in the feedback loop since its capacitance is sufficiently high so that it may be considered a short-circuit for signal voltages.

The actual magnitude of the feedback voltage that appears at the grid is the voltage across R2. From basic voltage-divider considerations, the fraction of the total voltage at the plate of the tube (with reference to ground) which is then used as feedback voltage can be obtained from:

$$\text{Feedback fraction} = -\beta = \frac{\text{R2}}{\text{R1} + \text{R2}} \qquad (5)$$

where β represents feedback (preceded here by a minus sign to indicate 180° phase inversion).

Problem 3. To illustrate what might be considered typical conditions of feedback in an audio amplifier, find the feedback fraction in the circuit of Fig. 13 if R2 = 8000 ohms and R1 + 50,000 ohms:

Solution.

$$\beta = \frac{8000}{50{,}000 + 8000}$$

$$= 0.138 \text{ or } 13.8\%$$

The feedback fraction may, of course, be increased by raising the value of R2 or reducing the size of R1.

12. Amplifier Gain with Inverse Feedback

Referring to Fig. 13B, the reader can compute the gain of a feedback amplifier if the feedback fraction is known. Suppose that a given amplifier stage without feedback has a gain of A. Inverse feedback is then added by introducing a feedback fraction. The actual voltage that this fraction represents is then $\beta \times e_o$, since β is the fraction of e_0 feedback. Thus the net input to the amplifier is $e_i + \beta e_o$. The input voltage – that is, the sum just obtained – when multiplied A times must be equal to the output voltage. Hence:

$$e_o = A\,(e_i + \beta e_o) \tag{6}$$

Since the voltage gain of an amplifier is, by definition, the ratio of output voltage to input voltage, then:

$$\text{Voltage gain} = \frac{e_o}{e_i} = \frac{A\,(e_i + \beta e_o)}{e_i} \tag{7}$$

which reduces to:

$$\text{Voltage gain} = \frac{A}{1 - A\beta} \tag{8}$$

Problem 4. A triode amplifier has a zero-feedback gain of 12. If 20% feedback is used, what is the voltage gain with feedback?

Solution. The feedback fraction is –0.2. Substituting in equation (7):

$$\text{Voltage gain with feedback} = \frac{12}{1 - (12 \times -0.2)}$$

$$= \frac{12}{1 + 2.4}$$

$$= 3.5 \text{ approx.}$$

Note that a minus sign must precede the feedback fraction when used in equation (8) for the reasons given previously. It should also be observed how much the voltage gain is reduced from its zero-feedback figure to the value with feedback. If large percentages of inverse feedback are to be used, the preceding stages must be capable of sufficient gain to make up for the losses introduced by negative feedback.

13. Current Feedback

In the foregoing analysis of inverse feedback, the feedback voltage was taken as proportional to the output voltage, making this a condition of voltage feedback. Negative feedback circuits

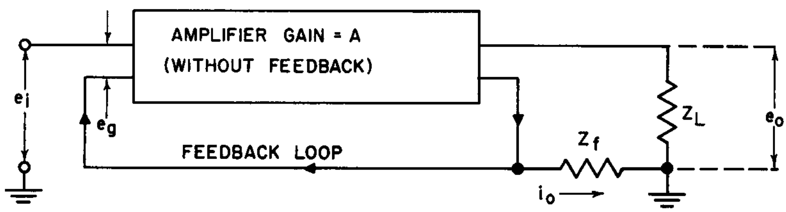


Fig. 14. Amplifier with a current feedback loop.

in which the feedback voltage is proportional to the output *current* are also possible as shown in Fig. 14. Figure 14 should be compared with Fig. 13 to clarify the differences between these two principles.

The voltage gain of this system is again e_o/e_i. The voltage drop across the feedback impedance Z_f is fed back out of phase with the input voltage. Since this voltage drop is the product of output current i_o and Z_f, we can write:

$$\text{Voltage gain} = \frac{e_o}{e_i} = \frac{e_o}{e_g - i_o Z_f} \tag{9}$$

Since the load impedance is Z_L, we can obtain from Ohm's law:

$$i_o = \frac{e_o}{Z_L} \tag{10}$$

Equation (10) may be substituted into equation (9) to yield:

$$\text{Voltage gain} = \frac{e_o}{e_g - e_o\left(\frac{Z_f}{Z_L}\right)} \tag{11}$$

Now the numerator and denominator of the fraction in equation (11) may be divided by e_g:

$$\text{Voltage gain} = \frac{\frac{e_o}{e_g}}{1 - \frac{e_o}{e_g\left(\frac{Z_f}{Z_L}\right)}} \tag{12}$$

But, by definition, the ratio of e_o/e_g is the zero-feedback amplification of the stage. Therefore, substituting A for e_o/e_g yields:

$$\text{Voltage gain} = \frac{A}{1 - A\left(\frac{Z_f}{Z_L}\right)} \tag{13}$$

One common source of distortion is the variation of load impedance with frequency when the load has reactive components. This is true, for example, in the output stage of an audio amplifier or radio receiver which works into an output transformer coupled to a loudspeaker. Since current feedback tends to maintain a constant output *current,* rather than a constant output *voltage,* regardless of changes in amplifier gain or variations in load impedance, this type of frequency distortion is not corrected. For appli-

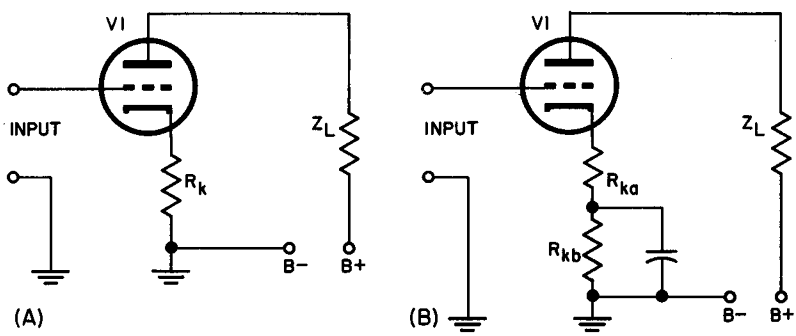


Fig. 15. Common types of current feedback circuits. (A) Current feedback, cathode resistor unbypassed. (B) Cathode resistor partially bypassed.

cations in which this kind of distortion must be minimized, voltage feedback is preferable.

Figure 15 illustrates common types of current feedback circuits. These are generally called *degenerative* arrangements. In Fig. 15A, the cathode resistor R_k is left unbypassed. Since the alternating component of the plate current flows through R_k, a voltage drop is developed across it that is 180° out of phase with the input voltage if Z_L is purely resistive. (When Z_L combines resistance and reactance, the phase angle is not exactly 180°, the phase shift in this case depending upon the relative magnitudes of the resistive and reactive portions of Z_L.) Thus, the feedback voltage is a function of the plate current of the tube and the value of R_k.

The double-duty action of R_k—that is, the production of bias voltage as well as feedback voltage—is often a handicap, particularly in situations where the correct value for R_k biaswise provides too great a feedback fraction. In such eventualities, the circuit of Fig. 15B may prove more satisfactory. Here, bias is produced by the plate current flowing through both R_{ka} and R_{kb}, but only R_{kb} is bypassed. Hence, the feedback voltage is proportional to the drop across R_{ka} and may be adjusted to suit the individual application. The bias voltage and feedback fraction may be determined from the equations:

$$\text{Bias voltage} = I_b\,(R_{ka} + R_{kb}) \tag{14}$$

and

$$\text{Voltage gain} = \frac{A}{1 - A\left(\dfrac{R_{ka}}{Z_L}\right)} \tag{15}$$

Figure 16 shows inverse feedback as typically applied to a two-stage amplifier. Both voltage and current feedback are utilized. Resistor R_{kb} provides current feedback as previously described,. while R_f and C_f supply voltage feedback. Of course, any feedback circuit must be carefully designed to make certain that the feedback voltage is out of phase with the input signal and that there is sufficient voltage amplification preceding the stages controlled by negative feedback to insure adequate overall gain.

When feedback is applied to pentode amplifiers, certain other considerations enter into the problem (Fig. 17). To begin with, both the suppressor grid and the screen bypass capacitor C_{sg} must be returned directly to the cathode rather than to the ground

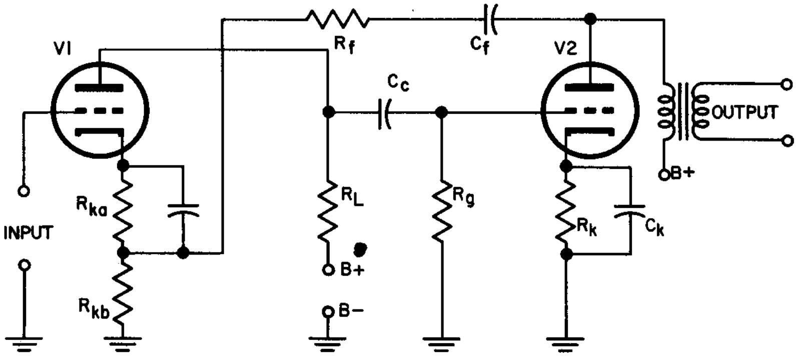


Fig. 16. Inverse feedback applied to a two-stage amplifier.

point of the system. With feedback resistor R_{ka} in the circuit, the cathode is no longer at signal-ground potential so that both the suppressor and screen must be connected directly to it if these elements are to remain at a-c cathode potential despite signal variations. Examining the circuit further shows that the screen capacitor C_{sg} would serve as a shunt across the feedback resistor R_{ka} if the series screen resistor R_{sg} were small in value. The shunting path would be from the top of R_{ka} through R_{sg} and filter capacitor C_f, thence to ground. This feedback short-circuiting action may be prevented by making R_{sg} sufficiently large. Normal pentode voltage amplifiers often employ a large value of screen-dropping resistor, eliminating short-circuit feedback. Voltage feed-

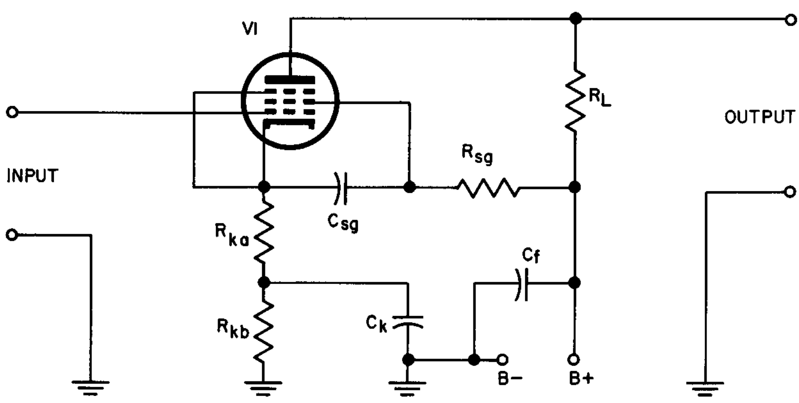


Fig. 17. Feedback applied to a pentode amplifier.

back may, however, be applied to pentodes just as it is to triodes; this includes both voltage and power amplifiers.

14. Review Questions

1. Explain why phase inversion is necessary when a push-pull amplifier stage is to be driven by a single-ended amplifier.
2. Describe two different types of phase-inversion systems and give the relative advantages of each.
3. What are the principal reasons for incorporating inverse feedback into an amplifier system?
4. Discuss how inverse feedback reduces distortion.
5. What is the difference between voltage and current feedback? Under what conditions is voltage feedback preferred? Current feedback?
6. A triode amplifier has a zero-feedback gain of 8. What voltage gain would be expected if 25% feedback were to be incorporated in the circuit?
7. Show that degenerative feedback (unbypassed cathode resistor) constitutes inverse feedback.
8. What secondary considerations are introduced when inverse feedback is applied to a pentode as compared with a triode?
9. Explain how inverse feedback tends to reduce the distortion caused by variations of load impedance with frequency.
10. What equation is used to determine voltage gain with feedback? Show how this equation is derived.

Chapter 3

IMPROVING THE PERFORMANCE OF LOW-FREQUENCY SYSTEMS

15. Volume Compression and Expansion

Low-frequency amplifiers are often called upon to amplify signal voltages which swing over extremely wide ranges, for example, when an orator hugs the microphone too closely and then steps far back to gesticulate at his audience. There are also applications wherein the signal voltage does not vary enough because the voltage sweep has been intentionally reduced by handling.

Signal swing reduction is termed *volume compression;* increasing the range of signal-voltage variations is called *volume expansion.* Both actions can be obtained in a single circuit containing a switch that selects either expansion or compression (Fig. 18).

The underlying principle of volume compression is simply that a tube changes its gain automatically as the signal strength changes. For compression the gain must diminish as the signal strength increases. Conversely, when the circuit is arranged so that the gain of the tube rises with increasing amplitude of signal, volume expansion results. Since the gain of a tube is a function of its grid bias, most compression and expansion circuits cause bias to vary with or against signal-amplitude changes to accomplish the required gain modifications.

Figure 18 is a block diagram of one such system. The input signal is first amplified by amplifier A1 in the customary manner. A portion of its output is fed to a special mixer-amplifier tube

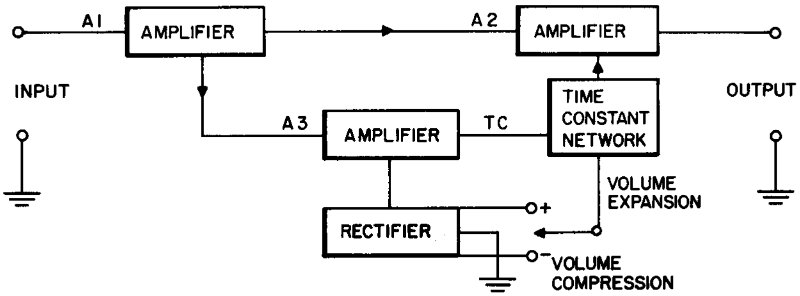


Fig. 18. Block diagram of an automatic compressor-expander, often referred to as a "Compander Circuit."

such as a 6L7, while a second portion is amplified once again by A3. After emerging from A3, the signal is rectified in a circuit that makes available either a positive or negative voltage (with respect to ground), the polarity to be applied to a second grid in A2 being chosen by the switch. TC is a time-constant network that must be carefully selected, as will be explained shortly. Since the magnitude of the d-c voltage obtained from the rectifier—either the positive or negative—is a function of the applied signal, the bias on the second grid of A2 will become more positive or more negative (depending upon the position of the switch) as the signal level increases. Volume expansion takes place in the upper position since the grid of A2 is driven more positive with increasing signal strength, causing its gain to increase. With the switch in its lower position (minus), volume compression is obtained in a similar manner.

It is important that the time-constant network, usually a simple R-C filter after the rectifier, has the proper component values. If the time-constant is too large, the effect will hang over and will continue to hold the gain up (for expansion) or down (for compression) when the need has passed. If it is too small, the action will be so fast that the desired signal may be counteracted and wiped out.

Detailed circuit arrangements for compressor-expanders (often called "companders") may be found in vacuum-tube literature. In many of these circuits, as has been mentioned, the 6L7 is used as amplifier A2 (Fig. 18). This tube is rather unique in that it possesses two signal grids, one of which is a variable-mu type while the other does not have super-control characteristics. The incoming signal, therefore, is applied to the normal grid. This leaves the super-control grid for application of the d-c control voltage from the rectifier. Such an arrangement provides for a relatively wide range of volume control.

16. Other Audio-Frequency Considerations

Pre-emphasis and de-emphasis—In the transmission of modulated radio signals either by amplitude modulation or frequency modulation, the most disturbing noise carried along with the audio lies somewhere between 5000 and 15,000 cps. Since the useful relative strengths of the desired audio signals are generally small in this range and tend to be overriden by the noise, pre-emphasis is provided for these frequencies as compared with the desired frequencies in the lower portion of the audio spectrum. Such pre-emphasis can be accomplished with relatively simple circuit elements (Fig. 19). By properly selecting the resistance and inductance values, the voltage applied to the grid of the audio amplifier may be made much richer in the high-frequency range since the reactance, and hence the voltage drop, in L is larger with higher frequencies. In a radio transmitter, this audio amplifier might be one of the low-level speech-amplification stages immediately following the microphone.

Unless the frequency response of the receiving equipment contains a de-emphasis circuit to compensate for the pre-emphasis at the transmitter, the reproduced audio frequencies will contain too many highs and too few lows and will, therefore, sound shrill and whiny. A suitable de-emphasis arrangement is also given in

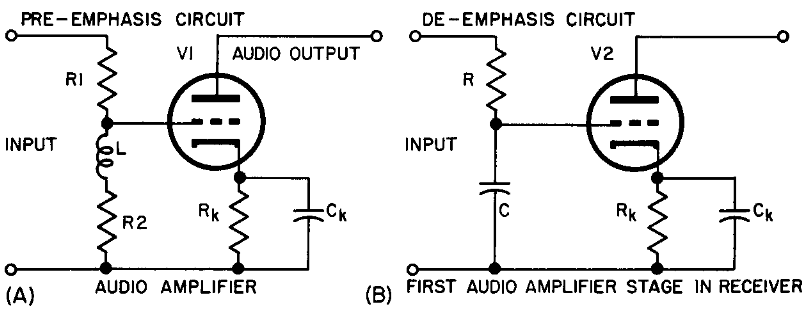


Fig. 19. Pre-emphasis and de-emphasis circuit elements.

Fig. 19. In this circuit, the previously accentuated high frequencies are reduced in amplitude in the grid circuit of the audio amplifier because capacitor C does not develop as large a signal voltage for the high frequencies as it does for the lows.

Pre-emphasis and de-emphasis are time-constant configurations; that is, the time constants must be chosen to obtain the desired

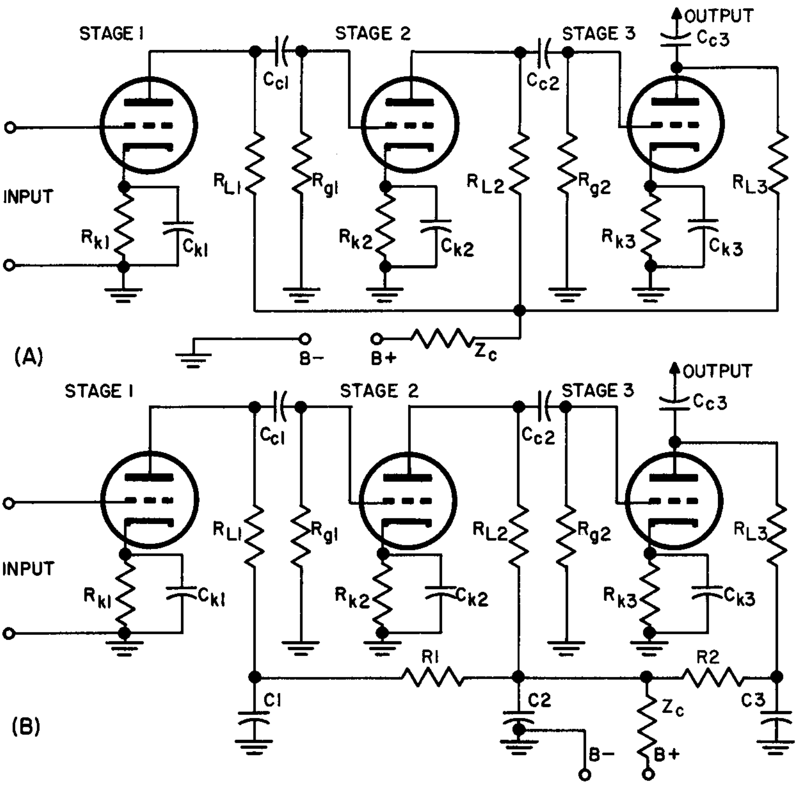


Fig. 20. (A) Common impedance Z_c in d-c supply of three-stage amplifier is often a source of regenerative feedback voltages. (B) Decoupling network added to circuit of (A) to prevent regeneration.

amplitude-frequency characteristic. It is generally considered good engineering practice to use a time constant of approximately 100 μsec in both the pre-emphasis and the de-emphasis circuit.

Decoupling considerations—We have seen that inverse feedback may be intentionally introduced into a low-frequency amplifier to reduce distortion and make the overall frequency response better. Positive feedback, however, is not desirable because it increases distortion and may cause oscillation in one or more stages.

Positive feedback may occur in several ways. If the shielding between input and output wires and components of a high-gain amplifier is inadequate, positive feedback may take place by capacitive or inductive coupling. One of the most common sources of positive feedback or *regeneration* is a common impedance in the power supply (d-c) lead going to two or more stages. In Fig. 20A, Z_c is an impedance in the power supply (e.g. a filter choke or filter resistor) that is common to the plate circuits of stages 1, 2, and 3. Any signal voltage that appears across Z_c due to voltage fluctuations in the plate current of stage 3 will be transferred to the plate circuit of stage 1, thence to the grid of the second tube. Since this transfer is in phase, positive feedback exists with the possibility of instability and oscillation.

Input and output may be decoupled by means of a simple R-C filter as illustrated in Fig. 20B. R1, R2, C1, and C2 acting together provide a high-impedance path for the signal fluctuations which attempt to reach the input circuits of stage 1 from stage 3 output and offer a low-impedance path for these signal voltages back to ground. The network R1, R2, C1, and C2 must have a time constant for the particular application such that voltage variations in the power-supply lead are smoothed out by normal filter action.

Square-wave analysis—By the processes of Fourier analysis, a perfectly square waveform (Fig. 21A) may be shown to be composed of a fundamental sine wave to which all the *odd* harmonics of the fundamental have been added. The Fourier series also shows that the amplitude of each harmonic is inversely proportional to its frequency. Thus the eleventh harmonic will have an amplitude

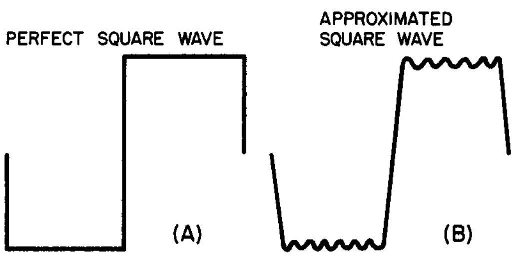


Fig. 21. (A) Perfect square wave is made up of the fundamental and all the odd harmonics of the fundamental. (B) A square wave may be approximated by superimposing up to 10 odd harmonics on the fundamental.

of one-eleventh that of the fundamental. It is usually considered sufficient to include harmonics up to the eleventh for a fair approximation of the square waveform.

Any amplifier circuit that has a flat frequency response and the same time delay for all frequencies it handles will transmit a square-wave input in amplified form to its output without changing its shape. If the frequency response is not uniform or if different frequencies undergo different time delays in passing from input to output, a change in the square waveform must result.

Low-frequency amplifiers may be tested for both frequency response and phase shift by applying a square-wave voltage to the input while the output is observed on the screen of an oscilloscope.

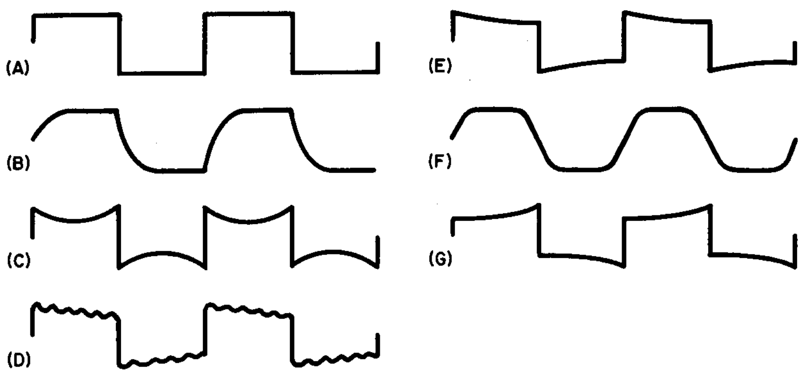


Fig. 22. A group of reproductions of square waves showing important types of attenuations and phase shifts.

The distorted waveforms shown in Fig. 22 are interpreted as detailed below; a complete analysis of the reasons for the particular distortions shown is beyond the scope of this book, but certain outstanding and easily recognized defects can be pointed out as follows:

(A) The input waveform—a more or less perfect square wave.

(B) Some amplitude attenuation of the higher frequencies with concomitant time delay of the higher frequencies, causing a phase shift between the highs and lows.

(C) Attenuation of the lower frequencies without accompanying phase shift.

(D) The result of a peaked response within the frequency range of the amplifier. This effect is sometimes called "ringing" and is usually due to excessive gain in the high-frequency range of the amplifier where the peak occurs.

(E) Some low-frequency attenuation together with an excessive leading angle for the low frequencies.

(F) Attenuation of the high frequencies without accompanying time delay of any of the frequency components.

(G) Slight low-frequency attenuation with excessive low-frequency lagging angle.

17. Control of Frequency Response

It is entirely feasible to design multistage amplifiers that will operate with specific kinds of input and output devices and provide frequency-response characteristics of a desired type. This is especially true when phase distortion is not a consideration. In special applications, it can often prove undesirable for the entire amplifier or even each of its stages to maintain a constant frequency response over its intended range. One or more stages of transformer coupling of the proper design can be utilized to raise the high-frequency gain. This can be done by making the effective Q's of the transformers large and by arranging their construction so that their resonant frequencies fall in the desired range.

Shunting a portion of the plate resistor in a resistance-coupled amplifier with a suitable inductance will cause a decrease in low-frequency gain; the gain at the low-frequency end, on the other hand, may be increased by using one or more stages of double impedance coupling. In this coupling system, inductances of the correct order of magnitude are used in place of the plate-load resistor of the driver stage and the grid-return resistor of the driven stage. Also, capacitors shunted across a portion of the plate-coupling resistor in a normal resistance-coupled amplifier will result in greater high-frequency gain without seriously changing the gain at low frequencies.

When standard coupling methods are used (transformer, resistance-capacitance, etc.), it is not easy to control medium-frequency gain without also affecting the gain at the high and low frequencies. In other words, the medium-frequency gain may be increased or decreased effectively by designing the amplifier for higher or lower gain at both ends of its range. A second method that is sometimes used involves a series circuit of resistance, inductance and capacitance connected from the grid to the cathode of one or more stages. This measure, when properly designed, will yield a change in gain over a small region of the medium frequencies.

There are so many combinations of methods that may be used to give specific desired frequency responses in amplifier systems that it is impossible to deal with all of them in this book. Among these are the employment of high-pass, low-pass band-pass, and band-reject circuits.

18. Applications of Low-Frequency Amplifiers

For the most part, the discussions in the preceding chapters and in those that follow center around audio-frequency amplifier systems which are predominate in low-frequency amplifier applications. Other areas in which low-frequency amplifiers find consistent use, however, should not be ignored. Some of these are:

(a) 60-cycle amplifiers for geophysical work
(b) servo-amplifiers
(c) ultrasonic amplifiers
(d) horizontal-sweep amplifiers for television
(e) amplifiers for low-frequency oscilloscope display

The fundamental design principles of amplifiers used for these purposes are identical with those of audio amplifiers having a reasonable frequency-response range. For example, any well-designed audio voltage and power amplifier will perform in a highly acceptable manner as a 60-cycle amplifier, regardless of its specific purpose. The designer need only provide the necessary voltage gain, power output, and impedance match to obtain the results he requires. As a matter of fact, a 60-cycle amplifier used to drive synchronous motors or to detect faults in underground pipes and cables need not have wide-band characteristics at all; therefore it is generally of much simpler design than a reasonably high-fidelity audio amplifier.

With regard to ultrasonic amplification between 20,000 cps and 100,000 cps, care must exercised to avoid the loss of high-frequency gain due to the shunting effects of input and output capacitances that always exist in vacuum-tube amplifier circuits, particularly in cascaded systems. In other words, irrespective of their final applications, low-frequency amplifiers are fundamentally the same; if the amplifier is to be used specifically for very low-frequency work

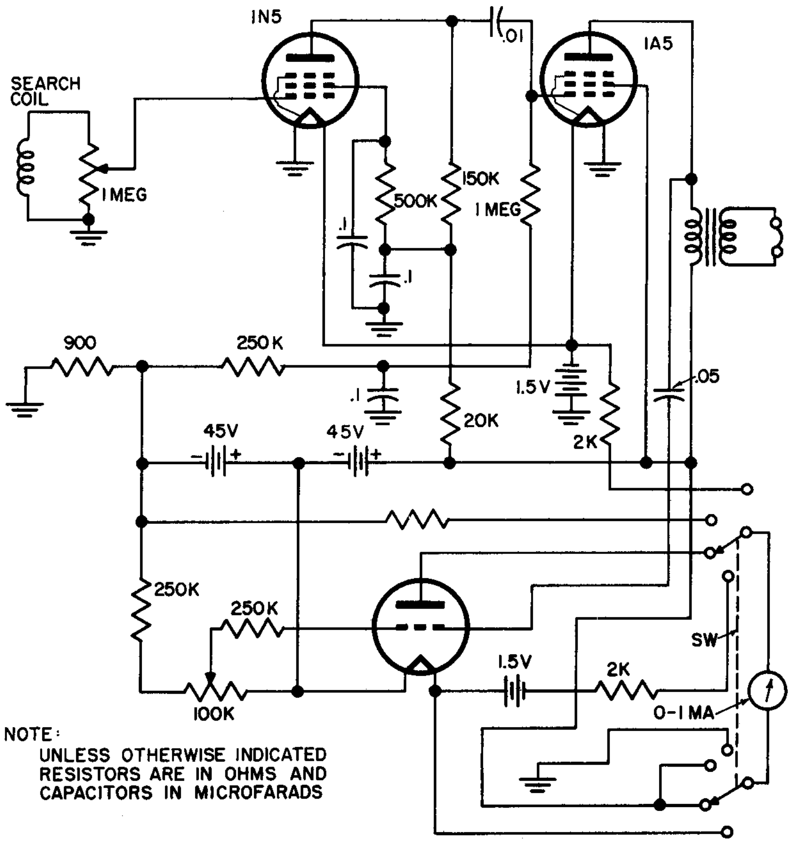


Fig. 23. Circuit diagram of amplifier used in portable geophone.

(as in the case of 30- and 60-cycle amplifiers), then the design must maintain the very low-frequency gain and avoid low-frequency attenuation. On the other hand, ultrasonic applications that employ what may be called "high" low-frequencies, between 20 kc and 100 kc, require design methods which minimize high-frequency attenuation. The following circuits illustrate these points.

The amplifier shown in Fig. 23 is that of a portable geophone[1]

[1]Blankmeyer, "Metal Locators," *Electronics*, vol. 16, December, 1943, p. 112.

used for tracing underground pipes or cables. For example, when a pipe line is being traced, a connection to the 60-cycle a-c line is made between the start of the pipe line and the ground through a large capacitor. A search coil consisting of about 5000 turns of wire wound on a plywood form of about 9 inches in diameter serves as the input device to the amplifier. In use, the exploring coil is moved about until a null is heard in the headphones; this occurs when the axis of the coil is pointing at the center of the pipe carrying the a-c current. In addition, the approximate depth of the pipe can be determined by backing away from the point directly over the pipe (this point is indicated when the axis of the coil is perfectly vertical) until the null occurs at an angle of 45° to the ground. Under these conditions, the depth of the pipe is equal to the distance from the point directly over the pipe to the 45° position of the coil.

The amplifier circuit is completely conventional except for the method of obtaining the bias for the 1A5. This bias is developed

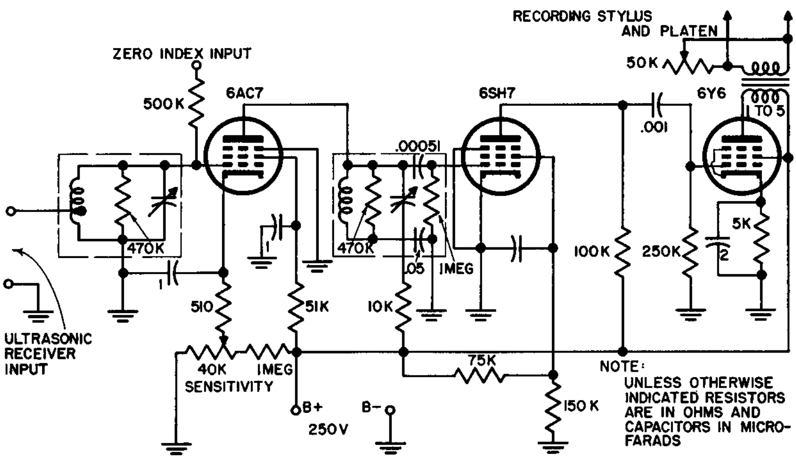


Fig. 24. A three-stage amplifier used for ultrasonic echo depth sounding.

across the 900-ohm resistor in the B− lead to both amplifier tubes. A built-in vacuum-tube voltmeter consisting of the 1G4 tube and associated components may be used to supplement the headphones in determining the precise position for obtaining a null.

The amplifier shown in Fig. 24 is a three-stage arrangement used for ultrasonic echo depth sounding and is designed to operate at a frequency of 21.5 kc. The first two stages are semifixed, tuned to 21.5 kc, and utilize impedance coupling; coupling to the output stage is accomplished through an R-C network of the conventional variety. The input impedance is approximately 15 ohms to match a transmission line from the output of the ultrasonic receiver transducer, in this case a *magnetostriction* type.

The output signal is fed to the recording stylus via a 1-to-5 transformer and consists of 180-volt pulses capable of disintegrating blackbodied carbonized dry recording paper. The 50-K resistor across the secondary of the output transformer is a density control which governs the extent to which disintegration occurs.

The first voltage amplifier (6AC7) is operated in class A with the sensitivity control set in a normal position; the second voltage amplifier (6SH7) is essentially a zero-bias amplifier which obtains some negative grid voltage due to the signal-voltage drop through the 1-megohm resistor in its grid-return circuit. This arrangement provides some limiting action which tends to prevent the output amplifier from being overdriven. The output pentode operates as close to cutoff as cathode biasing will permit. On positive peaks large plate-current pulses flow, each pulse having a steep leading edge and trailing edge. Such fast changes in the magnetic field in the transformer induce relatively high voltages across the secondary; these high-voltage peaks are required for the disintegration-recording method.

At an operating frequency of 21.5 kc, an input voltage of 2μv is necessary to produce a record when the amplifier gain is maximum. For this condition, the bandwidth is 2 kc; that is, the output drops 3 db at a point approximately 1 kc each side of the center frequency.

19. Review Questions

1. Explain why volume compression is used in some forms of recording and broadcasting.
2. Under what circumstances might volume expansion be used in an audio amplifier?
3. What is the value of pre-emphasis?
4. Explain why de-emphasis circuits are widely used in the audio systems of f-m receivers.
5. What de-emphasis time-constant value is normally acceptable? Why?

6. What is the function of a decoupling network?
7. What components are generally found in decoupling networks?
8. Explain how decoupling networks depend upon R-C time-constant considerations.
9. Explain why square waves are particularly suitable for audio-amplifier testing.
10. Which of the factors demonstrated by square-wave testing (i.e., frequency response or phase shift) is more important in audio amplifiers?

Chapter 4

DESIGN OF VACUUM-TUBE AMPLIFIER SYSTEMS

20. General Considerations

Multistage amplifier design involves these general considerations: (a) choice of tubes, (b) choice of coupling methods, and (c) specification of power-supply characteristics. The selection of tubes, coupling and tube components for R-C coupled systems, and power-supply voltages and currents, are entirely in the hands of the design engineer; however, where transformers for coupling are desired, selection is largely a matter of rummaging through transformer catalogs. Since the bulk of the engineering has already been consummated by transformer manufacturers, the design engineer need only choose the units which best fit his needs. This he can do by direct reference to the manufacturers' recommendations.

Although most tube manufacturers also provide data tables for resistance-coupled voltage amplifiers, it is helpful to understand the factors which go into the compilation of such tables. Situations are often encountered for which the tables do not supply the answers; in such cases, it may be necessary to start the design from the very beginning.

In this chapter we shall study the sequential process by which a complete resistance-coupled amplifier may be designed. We shall also discuss important considerations in transformer-coupled amplifier design not covered in preceding chapters.

21. Design Procedure for Resistance-Coupled Amplifier Using Triodes

Although triodes are not used as much as pentodes in resistance-coupled amplifier circuits, they are occasionally encountered in systems demanding large output voltages with less amplitude dis-

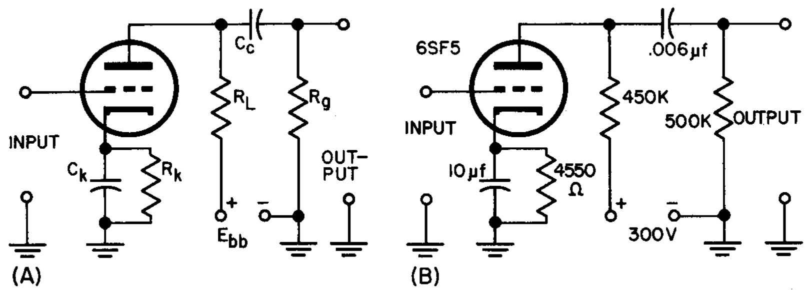


Fig. 25. Pentode voltage amplifier—basic current showing design components.

tortion. Triodes cannot supply as much gain per stage as pentodes, and their performance at the higher audio frequencies is inferior due to their higher input capacitance.

The design procedure for a resistance-coupled triode amplifier is considerably simpler than for pentodes and may be followed with the help of the accompanying text and Fig. 25. The steps in the procedure are given below on the basis of specific problem and tube.

Problem 5. Design a resistance-coupled triode amplifier which will provide the maximum possible voltage gain with a plate-supply potential of 300 volts.

Solution. (1) Since a large voltage gain is the primary requisite, a tube having a large amplification should be selected. A good choice would be a triode such as a 6SF5 with an amplification factor of 100.

(2) After choosing the tube, its plate resistance should be determined from the tube-data book. The plate resistance (R_p) of a 6SF5 is given as 85,000 ohms for a plate potential of 100 volts and 66,000 ohms for a plate potential of 250 volts. Since the voltage drop in R_L (in our final calculations) will be approximately 1/3 to 1/2 of the plate-supply voltage, i.e., half of 300 volts or 150 volts, we might estimate the actual operating plate resistance as 75,000 ohms and use this as a beginning point.

(3) The load resistance R_L is then selected to be from three to six times the magnitude of the plate resistance. Since the greatest possible voltage gain is desired, we choose R_L as six times the plate resistance, or $6 \times 75{,}000 = 450{,}000$ ohms.

(4) The bias must now be adjusted so that the resulting plate current produces a voltage drop in R_L that is from 1/3 to 1/2 the plate potential. Thus:

$$E_b = E_{bb} - I_b R_L$$

or

$$I_b = \frac{E_{bb} - E_b}{R_L}$$

Assuming a drop of 1/2 the supply voltage, this yields:

$$I_b = \frac{150}{450{,}000}$$

$$= 0.33 \text{ ma}$$

(5) The value of the control-grid bias required to produce $I_b = 0.33$ ma is now determined by drawing a load line for $R_L = 450{,}000$ ohms and a supply voltage of 300 volts into the plate characteristics. At a plate potential of 150 volts, the grid bias needed to cause 0.33 ma of plate current to flow is −1.5 volts.

(6) The value of R_k can now be determined from Ohm's law:

$$R_k = \frac{E_c}{I_b} = \frac{1.5}{0.33 \times 10^{-3}}$$

$$= 4550 \text{ ohms}$$

(7) According to general practice, C_k should have a reactance no greater than 1/10 the resistance of R_k at the lowest frequency to be amplified. Assuming that we wish this amplifier to handle frequencies down to 30 cps, then:

$$C = \frac{1}{2\pi f X_c}$$

$$= \frac{1}{6.28 \times 30 \times 455}$$

$$= 8.6 \ \mu f$$

(8) All that remains is the choice of values for R_g and C_c. Assuming that a power amplifier may follow the triode amplifier, the value of R_g is determined by the maximum permissible resistance in the grid circuit of the second tube. For instance, suppose that a 6L6 in class A is driven by the 6SF5; in this case, R_g would be selected as 0.5 megohm to conform with the maximum value given by the manufacturer. It is generally accepted that the time constant of C_c and R_g should not exceed approximately 0.003 sec. Hence C_c may be determined:

$$t_c = RC$$

$$C = \frac{t_c}{R} = \frac{0.003}{5 \times 10^5} = \frac{0.003}{0.5}$$

$$= 0.006 \ \mu f$$

where R is in megohms and C is in microfarads.

Figure 25B provides the complete circuit showing the values computed by the foregoing steps. The voltage gain of this circuit is very close to 65 by actual measurement. Very little change is noticeable in the performance if easily obtainable, standard component values, varying from the computed values by no more than 10%, are substituted. For instance: R_k might well be 4700 ohms rather than 4550 ohms; R_L and R_g might be 0.47 megohm instead of 0.45 megohm and 0.5 megohm respectively; and C_c might be 0.005 μf rather than 0.006 μf – all without significant deviations in voltage gain, frequency-handling ability, etc.

22. Design Procedure for Resistance-Coupled Pentode Amplifier

Since pentodes are used so much more often than triodes in resistance-coupled amplifiers, it is common to begin the design of such a stage with more than just voltage gain in mind. Of paramount importance is frequency response and, unless the range of frequencies covered meets the requirements set forth for the amplifier, no amount of gain will make up for this deficiency. Actual amplification over the range of frequencies is next in importance, while voltage output also requires consideration.

All of these factors are generally specified within the framework of the problem. As we have done before, we shall state the problem and then proceed with the sequence of design steps. Despite the fact that this sequence is concerned with a specific problem, it is generally applicable to the design of any low-frequency resistance-coupled pentode amplifier.

Problem 6. We require a resistance-coupled amplifier having a response range from 30 to 30,000 cps. It is specified that the average amplitude of the end frequencies be no less than 70.7% of the middle range. The amplified output voltage at the mid-range is to be a minimum of 20 volts peak for application to the grid of a 6L6 power amplifier operating in class A (Fig. 26).

Solution. (1) The first step, of course, is to select the tube. Generally, high-gain sharp-cutoff pentodes are used in applications such as this. Let us settle on a 6AU6 (Fig. 27) since this is a miniature tube having the requisite characteristics.

(2) Referring to the characteristics of the stage to be driven by the R-C amplifier (6L6), we find that the largest permissible grid resistance this tube may have as a class-A amplifier with cathode bias is 0.5 megohm. We can show that a good tentative choice for the voltage-amplifier plate-load resistor R_L is approximately 50% of R_g. Thus, we begin with a value for R_L of 250,000 ohms. (Most

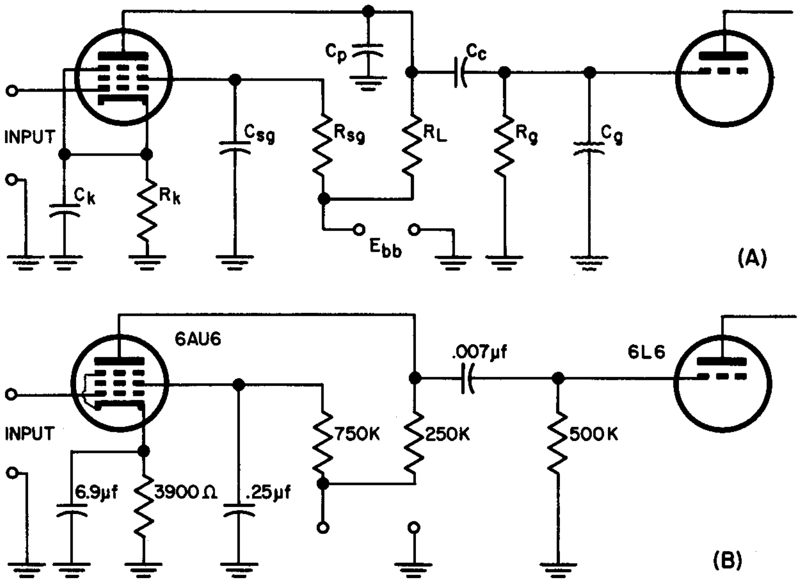


Fig. 26. Pentode voltage amplifier. Basic circuit showing design components.

authorities agree that the range for R_L lies somewhere between 0.2 R_g and 1.0 R_g. Small load resistors tend to yield large output voltages, while the higher values lead to greater voltage gain. In addition, good frequency response is more easily realized with lower values of R_L. In this example, good frequency response is mandatory; reasonably high output voltage is called for, but maximum gain is not; therefore, $R_L = 0.5\ R_g$ appears reasonable.)

(3) The plate resistance of the tube (R_p), the load resistance (R_L), and the grid resistance (R_g) are effectively in parallel with respect to signal voltages. The equivalent resistance of this group, designated as R_{eq} is given, therefore, by:

$$R_{eq} = \frac{R_L}{1 + \left(\frac{R_L}{R_g}\right) + \left(\frac{R_L}{r_p}\right)} \tag{16}$$

Since the plate resistance of the 6AU6 is very high, particularly at the low values of plate current to be used in this amplifier, the factor R_L/r_p becomes infinitesimal, and we may say that R_{eq} is:

$$R_{eq} = \frac{R_L}{1 + \dfrac{R_L}{R_g}} \tag{17}$$

$$= \frac{250{,}000}{1 + \dfrac{250{,}000}{500{,}000}}$$

$$= 167{,}000 \text{ ohms (to 3 significant figures)}$$

We shall use this value of R_{eq} later for finding the gain in the middle-frequency range.

(4) We must now concern ourselves with the input capacitance to the 6L6. Observe that C_p, C_g, and the stray wiring capacitance are effectively in parallel. As given by the tube-data book, C_p for a 6AU6 is 5 $\mu\mu$f, and C_g for a 6L6 is 10 $\mu\mu$f. Assuming a stray wiring capacitance of 2 $\mu\mu$f, the total shunting capacitance is 17 $\mu\mu$f. The reactance of 17 $\mu\mu$f at the highest frequency to be used (30,000 cps) is:

$$X_c = \frac{1}{2\pi fC}$$

$$= \frac{1}{6.28\,(30 \times 10^3)\,(17 \times 10^{-12})}$$

$$= \frac{1}{2\pi fC}$$

$$= 312{,}000 \text{ ohms}$$

As this reactance is substantially greater than R_{eq}, it may be expected that even the highest frequency for which this amplifier will

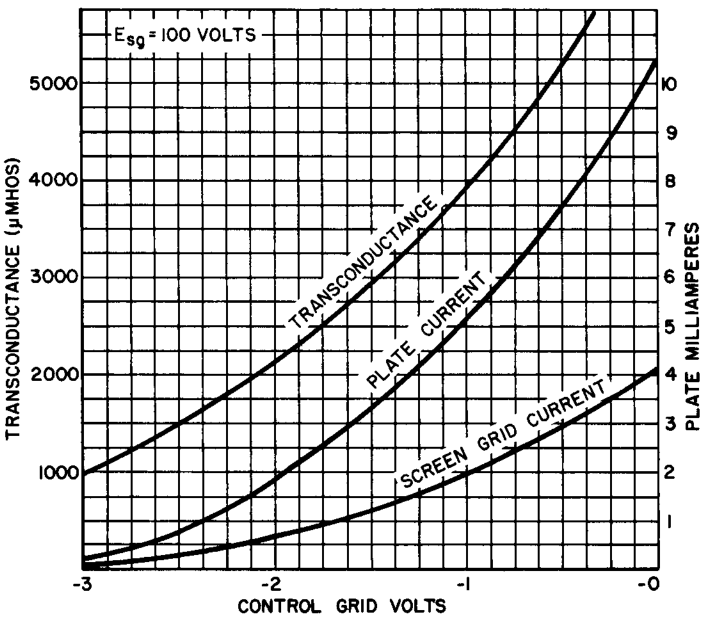


Fig. 27. Characteristics of a sharp cutoff pentode 6AU6.

be used will develop an ample voltage drop across R_{eq} as input to the next stage. We may now consider the original selection of R_L satisfactory for the frequency range in this problem.

(5) The question of the coupling capacitor (C_c) value now arises. For the customary response of no less than 70.7% of the gain in the mid-range for the 30-cps end of the signal, this capacitance must be adequate. It can be demonstrated that for the conditions of (the problem (70.7% response), the coupling capacitor may not have a reactance greater than the sum of the load and grid resistances. That is:

$$R_L + R_g > X_{C_c} \tag{18}$$

Since $R_L + R_g = 0.75$ megohm, the capacitor having this reactance at 30 cps must be determined. Hence:

$$C = \frac{1}{2\pi f X_c}$$

$$= \frac{1}{6.28 \times 30 \times 0.75}$$

$$= 0.0071\ \mu f$$

To insure that the reactance will be small enough, it would be expedient to use a somewhat larger capacitance than this, say, 0.01 μf. (A capacitor of 0.006 or 0.009 μf would also be satisfactory, but these sizes are difficult to procure commercially.)

Summarizing the values obtained up to this point we have:

Tube = 6AU6

$R_g = 0.5$ megohm

$R_L = 0.25$ megohm

$R_{eq} = 167{,}000$ ohms

Shunting capacitor $C = 17.0\ \mu\mu f$

$C_c = 0.01\ \mu f$

(6) Now it is necessary to determine what supply voltage should be selected. E_{bb} is not a particularly critical factor — it must be just large enough to yield a signal-output voltage somewhat greater than the specified value. The problem stipulates that the peak voltage to be applied to the grid of the power amplifier is approximately 20 volts. Since this is relatively low, the power-supply output voltage may be of any value in the normal range provided by commercially available power transformers, say 150 to 300 volts. As a tentative choice, let us determine how a supply potential of 150 volts works out.

(7) Standard design procedure indicates that a plate current (I_b), with voltage drop across the load resistor R_L equal to about half the supply voltage, should now be chosen. (The relationship between voltage drop I_bR_L or E_{bb} depends largely upon the ratio between R_g and R_L. If this ratio is as high as 5 to 1, I_bR_L should be about 0.45 E_{bb}; for smaller ratios down to unity—e.g., $R_g = R_L$

—it is customary to use a plate current which gives $I_b R_L = 0.55\ E_{bb}$.) The ratio between R_g and R_L in this problem is 2 to 1, hence a value of $I_b R_L = 0.5\ E_{bb}$ appears reasonable. Thus:

$$I_b = \frac{E_b}{R_L} = \frac{125}{250{,}000} \tag{19}$$

$$= 0.5 \text{ ma}$$

(8) Referring to Fig. 27, the grid bias required to establish this plate current at 100 volts of screen potential may now be determined. Following the 0.5-ma plate-current line to its intersection with the plate-current curve, it is seen that the required bias is $E_c = -2.7$ volts. For this quiescent condition of operation, the curves also reveal that the screen current will be 0.2 ma and the transconductance will be about 1300 μmhos.

(9) The figures obtained in part (8) of this solution enable us to calculate the following:

(a) Screen-dropping resistor

$$R_{sg} = \frac{250 - 100}{0.2 \times 10^{-3}}$$

$$= 750{,}000 \text{ ohms}$$

(b) Cathode-bias resistor

$$R_k = \frac{E_c}{I_b + I_{sg}}$$

$$= \frac{2.7}{0.7 \times 10^{-3}}$$

$$= 3900 \text{ ohms}$$

(c) Amplification for the middle frequencies

$$\text{Mid-range gain} = g_m R_{eq} \tag{20}$$

$$= 1300 \times 10^{-6} \times 1.67 \times 10^{5}$$

$$= 217$$

(10) The cathode bypass capacitor C_k must be selected with a view to keeping its reactance at the lowest frequency to be amplified (i.e., 30 cps) sufficiently low so that the product of its reactance and the transconductance of the tube (1300 μmhos) will be less than 1.

$$g_m X C_k > 1 \tag{21}$$

Thus, we first solve:

$$XC_k = \frac{1}{g_m}$$

$$= \frac{1}{1300} \times 10^{-6}$$

$$= 770 \text{ ohms}$$

So, since

$$C = \frac{1}{2\pi f X_c}$$

then

$$C = \frac{1}{6.28 \times 30 \times 770}$$
$$= 6.9\,\mu f$$

Since this capacitor would make the product of $g_m X C_K$ equal to unity, we would select a value somewhat larger than this to insure a product substantially lower than 1. Any value from 8 μf to 10 μf would be suitable since both of these are common commercial sizes.
(11) To calculate the value of the screen bypass capacitor, it is necessary to know the dynamic screen-grid resistance of the tube. That this parameter is not given in tube manuals or data books suggests that it is not often used for design purposes. In practice, it is the accepted procedure to use as large a screen bypass capacitor as space and time permit, avoiding electrolytic types if possible. For the conventional sharp-cutoff pentode, a value of 0.25 μf is almost always satisfactory; capacitances larger than this are bulky and unnecessary if the lower limit of the frequency range is in the order of 30 cps. A glance through the resistance-coupled amplifier data for the 6AU6 will corroborate this; as a matter of fact, the largest capacitor listed for any combination of plate voltage and other circuit parameters is only 0.15 μf. Figure 26B provides the complete circuit showing the values computed by the foregoing steps.

23. Design Considerations for Transformer-Coupled Low-Frequency Amplifiers

Mid-range amplification—The design engineer's position with regard to transformer-coupled amplifiers differs substantially from his attitude toward resistance-coupling for the fundamental reason that he can exercise more control over the latter. As we have shown in the preceding section on resistance-coupled amplifiers, every circuit component is subject to design analysis, and the selection of values can be arrived at by relatively straightforward calculation. Although the engineer can specify his needs for a particular transformer, he still must rely upon another engineer—the transformer designer—to meet his specifications. Transformer design and construction is a highly specialized art; nevertheless, anyone who contemplates working with transformer-coupled amplifiers can profit from an understanding of some of the important design factors involved.

A general discussion of transformer-coupled amplifiers was presented in Section 6 and the equivalent circuit of a transformer-coupled amplifier applicable to the middle range of audio frequencies shown in Fig. 5. To grasp the implications of high- and low-frequency attenuation as discussed in this section, we shall now examine the governing factors more fully. The most direct approach to this problem is the development of a general equation which offers, in a single statement, an explanation of all of the attenuation and phase-shift effects encountered in practice. In addition, such an equation in itself can suggest remedies and corrections for undesirable behavior in any specific transformer-coupled amplifier.

Equation (4) states the middle-frequency gain of a transformer-coupled amplifier in this form:

$$e_o = -\mu e_g \frac{n_s}{n_p}$$

in which e_o is signal output voltage, μ is the amplification factor of the tube, n_s is the number of turns in the transformer secondary, and n_p is the number of primary turns. Symbolizing the secondary-to-primary turns ratio simply as n, and solving for gain (*i.e.*, e_o/e_g), we have

$$\text{Mid-range voltage gain} = \frac{e_o}{e_g} = -\mu n \tag{22}$$

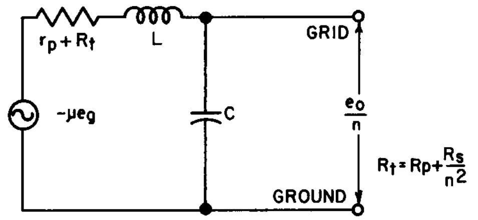
The minus sign before μ was originally introduced to indicate the 180° phase relationship between grid voltage and plate voltage. This may be conveniently written in polar form as:

$$\text{Mid-range voltage gain} = \frac{e_o}{e_g} = \mu n \underline{/180^\circ}$$

High-frequency amplification—Figure 28 shows the equivalent circuit at higher audio frequencies. The equivalent resistance is the sum of the plate resistance of the tube (r_p) and that of the transformer R_t.

$$R_t = R_p + \frac{R_s}{n^2}$$

where R_p is the primary resistance, R_s is the secondary resistance and n is the turns ratio. L is the leakage inductance of the transformer and C is the distributed capacitance of the secondary winding (with certain corrections made for the turns ratio and the capacitance between secondary and primary). It might be mentioned here that there is always the likelihood that the frequency



R_p = TRANSFORMER PRIMARY RESISTANCE

R_s = TRANSFORMER SECONDARY RESISTANCE

n = TURNS RATIO OF TRANSFORMER

Fig. 28. Equivalent circuit of a transformer-coupled amplifier at high audio frequencies.

at which L and C form a series-resonant circuit may fall into the audio-frequency range. In older transformers, the resonance frequency generally could be found between 7000 and 10,000 cps, causing a sharply peaked amplification. In transformers of modern design the effect is still present in many cases but at much higher frequency and greatly reduced in amplitude, so that it may be ignored except in the most critical applications.

To simplify the calculations, the equivalent circuit has been reduced to "unity turns ratio" condition by dividing the output voltage by the actual turns ratio n. Thus, the equivalent circuit becomes a simple series arrangement in which the voltages $-\mu e_g$ and e_o/n are directly proportional to the impedance across which they are developed. Thus:

$$\frac{\frac{e_o}{n}}{-\mu e_g} = \frac{-jX_c}{(r_p + R_t) - j\,(X_c - X_L)} \qquad (23)$$

X_c and X_L are the reactance of C and L, respectively (Fig. 28), r_p is the plate resistanse of the tube, R_t is the resistance of the transformer winding, and $j = \sqrt{-1}$.

Transposing terms to obtain e_o/e_g on the left and dividing through by $-j$, we have:

$$\text{Voltage gain} = e_o/e_g \quad \frac{+\mu n X_C}{-j(r_p + R_t) + (X_C - X_L)} \tag{24}$$

To provide a more easily interpreted form, equation (65) should be converted to polar form. This is easily accomplished by first dividing through by X_C before proceeding with the conversion. The final polar form of equation (24) is given in equation (25):

$$\text{Voltage gain} = \mu n \left(\frac{1}{\dfrac{(X_C - X_L)^2}{X_C^2} + \dfrac{(R_t + r_p)^2}{X_C^2}} \right)$$

$$\underline{/ \; 180^\circ - \tan^{-1}\left(\frac{r_p + R_t}{X_C - X_L} \right)} \tag{25}$$

This equation is interpreted as follows: the quantity μn is the voltage gain of the amplifier in the mid-frequency range [equation (22)]; the factor "180°" is the phase inversion caused by the tube from input to output; the fractional quantity containing the reactive and resistive components under the radical is a factor by which the mid-range voltage gain must be multiplied to determine the gain at high audio frequencies; finally the $\tan^{-1}$ quantity to the right represents the phase shift caused by the reactive and resistive components for high frequencies.

Equation (25) applied specifically to triodes since pentodes are very seldom encountered in transformer-coupled amplifiers. It is extremely difficult to design a transformer whose primary winding impedance can approximate the high plate resistance of pentodes.

Equation (25) reveals several interesting facts. It is possible for the high-frequency multiplier fraction to become greater than 1 if the frequency permits L and C (Fig. 28) to resonate. Should this occur, then $X_C = X_L$, and the first portion under the radical drops out. If X_C is then larger than the sum of R_t and r_p—as it almost always is—then the entire multiplier exceeds unity and the gain rises. Since this would occur only at a relatively high frequency, it causes a resonant peak that destroys the uniform response of the amplifier. It is essential, therefore, to minimize the importance of the $X_C - X_L$ term in the denominator of the fraction; the only practical way to do this is to increase the value of the remaining term under the radical by increasing R_t or r_p or both. But as we will show shortly, it is necessary for r_p to remain as small as possible for good low-frequency response, hence, the only remaining choice is to increase R_t alone.

A second way around the problem is to make the resonant frequency occupy a position that is well above the highest frequency to be handled by the amplifier. Since reduction of L, C, or both will accomplish this, every effort is made to keep the leakage inductance (L) and the distributed capacitance (C) as small as possible. It is evident, of course, that this capacitance can never be made as small in transformer-coupled amplifiers as it can in resistance-coupled types; this explains why the latter are used to the exclusion of the former in all wide-band amplifiers whose response must be uniform over a very large range.

The phase-shift factor in equation (26) indicates that the amount of phase shift caused by the coupling components—and, hence, the overall phase shift—is also a function of frequency due to the presence of X_C and X_L in this term. In low-frequency amplifiers such a variable phase shift has little effect except in equipment using heavy negative feedback. If caution is not exercised in such cases, instability may very well occur as a result of the possible change of negative to positive feedback at certain frequencies.

Low-frequency amplification—Analysis of the low-frequency performance of a transformer-coupled amplifier is based on the

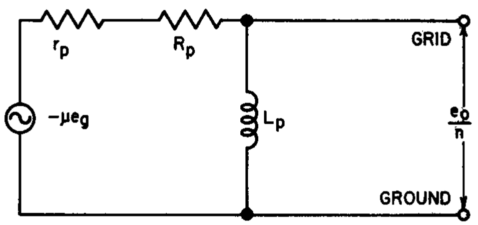


Fig. 29. Equivalent circuit of a transformer-coupled amplifier at low frequencies.

equivalent circuit given in Fig. 29. At the low frequencies, the primary reactance ($2\pi fL_p$) drops and begins to have an appreciable effect upon the stage gain. Since the equivalent output voltage is taken across the transformer, L_p must be shown as part of a voltage divider with r_p as in Fig. 29. As in the high-frequency case, the voltage gain e_o/e_g is proportional to the impedance ratio so that:

$$\text{Voltage gain} = \frac{e_o}{e_g} = \frac{\mu n X_p}{X_p - jr_p} \qquad (26)$$

As before, the polar form is most easily obtained by dividing through by X_p:

$$\text{Voltage gain} = \mu n \frac{1}{\sqrt{1+\dfrac{r_p^2}{X_p^2}}} \quad 180° + \tan^{-1}\frac{r_p}{X_p} \qquad (27)$$

Here again, the product μn is the mid-range gain while the remainder is the low-frequency multiplier. The polar quantity is divided in two parts: 180° is again the tube phase inversion; $\tan^{-1} r_p/X_p$ is the low-frequency phase shift caused by the reactance of the transformer primary winding.

The effect of the low-frequency multiplier is easy to discern. For example, if the plate resistance r_p of the tube is relatively high then the ratio r_p^2/X_p^2 may represent a substantial number. When added to one in the denominator, the divisor becomes appreciable and the entire fraction (i.e., the low-frequency multiplier) then

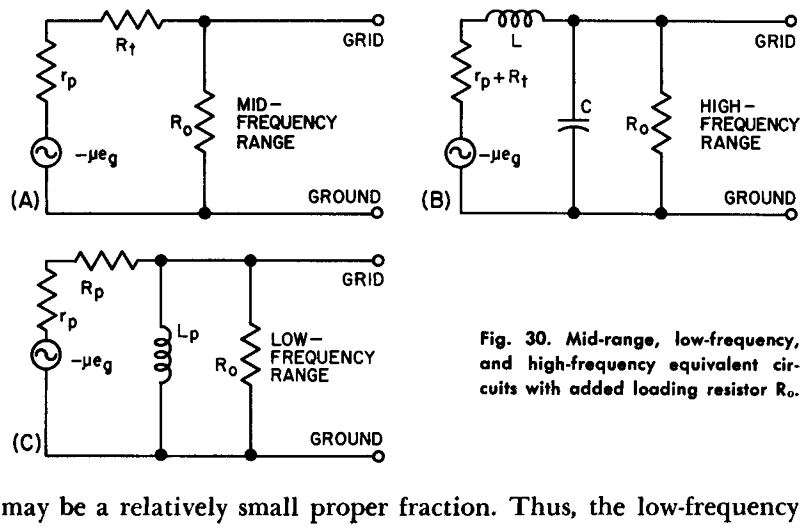


Fig. 30. Mid-range, low-frequency, and high-frequency equivalent circuits with added loading resistor R_o.

may be a relatively small proper fraction. Thus, the low-frequency voltage gain may be decreased seriously, indicating that the response curve would undergo a sharp dip at this end of its range.

Exactly the same effect would occur if X_p were a small fraction of r_p. Evidently, the only way to make the multiplier approach unity—the condition desired for uniform response of the amplifier —is to make r_p small and X_p relatively large.

This at once dictates the avoidance of very high-μ triodes since all such tubes necessarily have large plate resistances. On the other hand, extremely low-μ tubes (which would be expected to provide the best low-frequency response) cannot offer sufficient voltage gain to make the use of transformer-coupling worthwhile. This calls for a compromise. Some of the tube types suitable for such compromise are: the 6S4 which is a miniature type having a μ of 16 and a plate resistance of only 3600 ohms; and the 6C4 with a μ of 17 and a plate resistance of approximately 7700 ohms.

The reactance of the primary winding of the coupling transformer is largely a matter of specification to the transformer manufacturer. X_p is generally made as high as possible by using a core material of modern high-permeability metal, by planning for a large number of primary turns, and by employing an oversized core with plenty of "iron."

Here again, certain compromises must be effected. If the primary turns are increased, the secondary turns must also be proportionally increased to maintain the desired turns ratio, causing the leakage inductance to go up as well. This will almost certainly be detrimental to the high-frequency response. Hence, one meets limitations in all directions—which explains why transformer coupling is not as popular as resistance coupling, particularly in high-fidelity equipment.

Transformers must also be shielded carefully to prevent interaction of their fields with each other, if more than one is used in a given amplifier, and also to avoid induction of hum voltages in their windings due to nearby current-carrying conductors. Very often, orientation of transformers assumes importance in the finished design of an amplifier. It is often found that fields must cross at, or nearly at, right angles to each other to prevent undesirable feedback from one stage to another.

24. Flattening Response of Transformer-Coupled Amplifiers by Resistance Loading

If a small amount of gain can be sacrificed, it is possible to improve the frequency performance of a transformer-coupled amplifier by connecting a resistance of about 100,000 ohms across the secondary winding of the transformer. The effect of adding such a resistor may be obtained from an analysis of each of the equivalent circuits (mid-range, high-frequency, low-frequency) to which loading resistor R_o has been added across the output terminals (Fig. 30).

Effect on middle range—With the addition of R_o to the equivalent circuit of Fig. 6B, we obtain Fig. 30A. An output voltage division now occurs in which the gain μn is reduced by a factor:

$$\text{Mid-range gain reduction} = \frac{R_o}{r_p + R_t + R_o} \tag{28}$$

If R_o is made large compared to r_p and R_t, the fraction given in equation (28) approaches unity and the reduction of gain is seen to be unimportant. For example, with a tube such as the 6S4

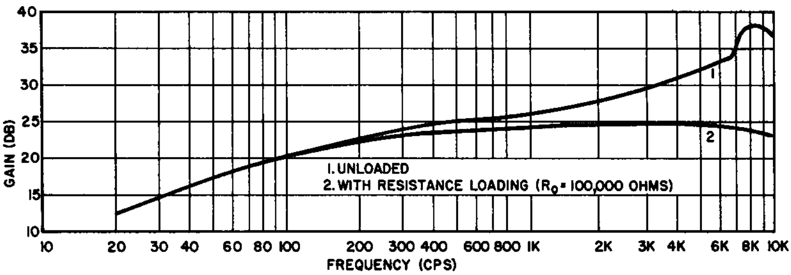


Fig. 31. **Improved performance of medium-cost transformer-coupled amplifier with resistance loading.**

for which the plate resistance is only 3600 ohms, we might use a transformer with an R_t characteristic of perhaps 5000 ohms. Under these conditions, the gain reduction factor given in equation (28) becomes:

$$\text{Mid-range gain reduction} = \frac{100{,}000}{3600 + 5000 + 100{,}000}$$
$$= 0.91$$

which means that the mid-range gain with R_o connected would be 91% of its value without this resistor. If ample reserve gain is available, a reduction of this order of magnitude is inconsequential.

Effect on high-frequency response—By connecting R_o in parallel with C (output capacitance of the amplifier), the Q of the series-resonant circuit comprising L and C will be reduced substantially (Fig. 30B). This would tend to minimize the resonance peak normally found in transformer-coupled amplifiers in the vicinity of 8000 cps; in itself, this is an important improvement in frequency response. If R_o is made too small, however, it is apparent that the high-frequency gain may be reduced to the point where the drop-off may seriously impair the amplifier performance at this extreme of its range. The value of 100,000 ohms is usually about adequate to prevent this.

Effect on low-frequency response—It may be seen from the low-frequency equivalent circuit in Fig. 30C that connecting R_o across the output terminals places R_s and R_o in parallel with L_p. The parallel impedance of this combination will certainly remain more constant than the impedance of L_0 alone as the frequency varies. Thus, at low frequencies an improvement in constancy of output impedance must result in a more uniform response at this end of the range. This improvement occurs in the range between 100 and 800 cps as indicated on the curves in Fig. 31. These curves have been taken for a medium-cost transformer-coupled amplifier of more or less standard commercial design. The overall flattening of the response curve is obvious on inspection.

25. Review Questions

1. What is usually the first consideration in the design of a triode amplifier, aside from frequency-response characteristics?
2. When selecting a load resistance for a triode voltage amplifier, the value is usually made from three to six times the plate resistance of the tube. Justify this rather arbitrary specification.
3. Once the required plate current in a triode-amplifier design problem has been determined, how does one find the bias necessary to produce this condition?
4. The plate current of a triode voltage amplifier is 1.25 ma. Determine the size of the cathode resistor needed to establish a grid bias of —4.5 volts.
5. The cathode resistor of a triode voltage amplifier is 2000 ohms. The amplifier is to have an acceptable response down to 60 cps. What is the minimum value of a cathode bypass capacitor that can be used in this connection?
6. If you wanted the highest possible output voltage from a pentode voltage amplifier of the resistance-coupled variety, would you use a large or a small value for R_L? Justify your answer.
7. Explain exactly what is meant by R_{eq} in equation (27).
8. What is the physical significance of the $\underline{/180°}$ in equation (22)?
9. Explain why transformer coupling is seldom used between pentode-voltage amplifier stages.
10. If for some reason it became important to keep the low-frequency phase-shift in a transformer-coupled amplifier to a minimum, what steps would you suggest taking? [See equation (27).]

Chapter 5

TRANSISTOR AMPLIFIER DESIGN

26. Transistor Parameters and Circuit Factors

If the transistor possessed isolated input and output circuits, the design procedure for transistor amplifier circuits would follow closely the approach used for vacuum tubes. The interdependence of input and output circuits, however, necessitates a somewhat different approach to the problem of amplifier design. For example, input and output impedances of vacuum tubes are quite independent of one another. In a transistor, output impedance depends upon input impedance which, in turn, is governed to some extent by the impedance of the signal source.

To assist us in developing a logical approach to transistor amplifier design, let us first review the major transistor parameters. We note at this point that parameter symbolization has not yet been standardized; thus, the symbols that appear below may sometimes be given in other forms by individual manufacturers. Some confusion also exists relative to applied voltages and their points of measurement. To clarify this, let us first define what is meant by a *point of reference*: for transistors, the reference point is the transistor electrode that is common to both input and output circuits. In the common base configuration, therefore, all other electrode voltages would be measured with reference to the base potential; in the common emitter circuit, the emitter potential becomes the reference point. The parameters may then be listed as follows:

V_b: *Base voltage.* The base voltage is the potential between the base terminal and the reference point.

V_c: *Collector voltage.* The collector voltage is the potential between the collector electrode and the reference point.

V_e: *Emitter voltage.* The emitter voltage is the potential between the emitter terminal and the reference point.

E: Supply voltage. This is the voltage furnished to an individual electrode from an external supply. Subscripts indicate the electrode being supplied, as E_b is the base supply voltage, E_c is the collector supply voltage, and E_e the emitter supply voltage. (Supply voltage is not a transistor parameter, of course, but is listed here because of its close association with the parameters.)

r_b, r_c, r_e: *Electrode resistances.* These are internal resistances of each respective electrode measured to the reference point.

i_b, i_c, i_e: *Electrode currents.* These are the currents flowing in the individual electrode circuits as measured by an appropriate current meter immediately adjacent to the electrode.

The *circuit factors* are the elements contained in the circuit external to the transistor itself. These are:

R_b: *Base resistance.* The resistance external to the transistor through which the base current flows.

R_c: *Collector resistance.* The resistance external to the transistor through which the collector current flows.

R_e: *Emitter resistance.* The resistance external to the transistor through which the emitter current flows.

27. Design Considerations — Transistor Bias

Although common-base current gain (alpha, α) and common-emitter current gain (beta, β) are sometimes listed as transistor parameters, they do not belong in this category. In a sense, they are similar to tube constants such as amplification factor and transconductance in that they cannot be varied *at will* merely by changing circuit conditions. Alpha and beta are extremely important, however, in design procedures. They are defined as follows:

Alpha, or common-base current gain, is the ratio of a small change in collector current to the change in emitter current produced thereby. That is:

$$\alpha = \frac{\Delta I_c}{\Delta I_e} \tag{29}$$

Beta, or common-emitter current gain, is the ratio of a small change in collector current to the change in base current produced thereby.

$$\beta = \frac{\Delta I_c}{\Delta I_b} \tag{30}$$

The simplest biasing arrangement for transistor amplifiers is shown in Fig. 32.

In this circuit, a base current of the proper size is fed into the transistor through R1. This sets up the desired collector-to-emitter voltage, and the collector current in the circuit. Unfortunately,

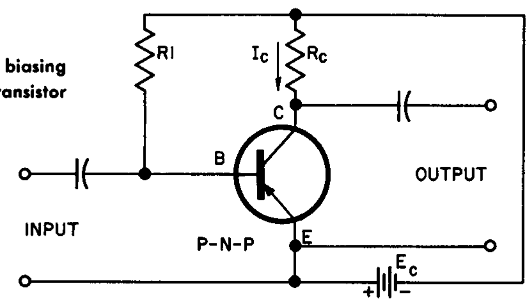


Fig. 32. Fundamental biasing arrangement for a transistor amplifier.

these biasing conditions must not vary despite possible changes in ambient temperature, otherwise the bias becomes incorrect, leading to distortion and instability. In the circuit of Fig. 32, the collector current that flows is given by:

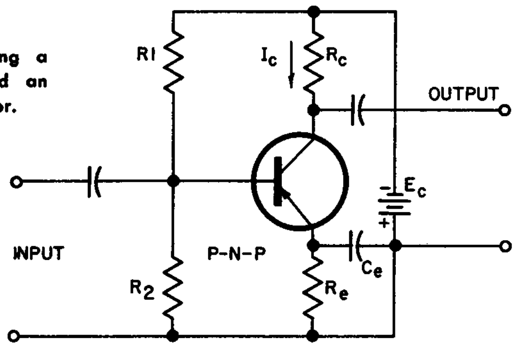
$$I_c^* = \beta \frac{E_c}{R1} \tag{31}$$

Since beta is temperature sensitive, the bias depends on temperature as well. Except in cases where the bias resistor R1 can be adjusted as required, this type of biasing is not satisfactory. Individual transistor differences with regard to beta and leakage currents are sufficient to make this circuit unsatisfactory even if the temperature could be maintained constant.

*The use of upper case "I" indicates a d-c condition rather than a signal current for which lower case is used.

One of the most satisfactory ways to overcome the problem of instability consists of adding a feedback circuit which will make the biasing conditions independent of current gain and the remaining transistor parameters. In addition, rather than supplying the

Fig. 33. Bias circuit using a base voltage divider and an emitter feedback resistor.



base current through a single dropping resistor, it is advisable to connect a voltage divider (R1 and R2 in Fig. 33) in the base circuit for additional stability. Feedback is provided by the emitter resistor (R_e, Fig. 33), thus stabilizing the operating point of the transistor.

For this stabilized circuit, the current in the collector circuit is given by:

$$I_c = \frac{E_c \times R2}{R_e\,(R1 + R2)} \qquad (32)$$

It should be observed that the introduction of feedback has removed beta from the I_c equation altogether, and the collector current becomes a function of the supply voltage E_c, the emitter resistor R_e, and the voltage divider comprising R1 and R2. Capacitor C_e is included in the circuit to prevent excessive degeneration of the signal and the possible loss of gain that might occur as a result of degenerative action.

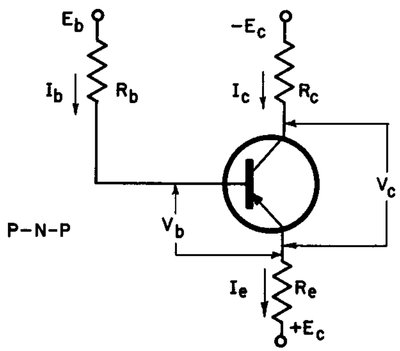
Since equation (32) indicates the independence of collector current as referred to current gain and other parameters, this biasing circuit, when properly designed, can be made to provide optimum performance despite variations in temperature, leakage current (I_{co}), and intrinsic current gain. We require, therefore, a set of design equations that will provide information on circuit values necessary to achieve a specified end result.

28. Bias Design Equations

Figure 34 illustrates a generalized circuit for which the desired equations can be derived.

The significance of the various symbols as applied to this circuit are: $-E_c$ is the collector supply voltage. The minus sign here indicates that a p-n-p transistor is being used. $+E_c$ is the positive terminal of the collector supply. E_b is base supply voltage, obtained in the final circuit by means of the voltage divider consisting of

Fig. 34. Generalized circuit used to derive design equations.



R1 and R2. V_b is the potential between base and emitter, the latter being the reference point as previously defined. V_c is the potential between collector and emitter. I_b, I_c, and I_e are electrode currents. (Note that the direction of each current indicated is that of *electron current.*)

Using a standard engineering approach, the value of the base supply voltage can be determined from the equation:

$$E_b = I_e\,[\,(1 - a)\,R_b + R_e] + V_b - R_b\,(I_{co}) \qquad (33)$$

V_b is a function of temperature and transistor material. Thus, it can be given for germanium and silicon transistors as a constant factor for a given temperature. Taking two extremes, V_b has the values;

	0°C	90°C
Germanium	0.2	−0.1
Silicon	0.7	0.5

We shall designate the figures in the left column as $V_{b\,max}$ and those in the right column as $V_{b\,min}$ for use in the next equation.

In the design of a bias system, certain maximum and minimum figures must be provided by the manufacturer. In this case, we must know the minimum emitter current for acceptable performance and the maximum emitter current that will permit comparable performance. Similarly, the minimum and maximum values of alpha must be stated, as well as the maximum anticipated leakage current I_{co}. With these figures available, it is then possible to derive an expression that yields the value of the base resistance R_b, as follows:

$$R_b = \frac{(I_{e\,max} - I_{e\,min})\,R_e + V_{b\,min} - V_{b\,max}}{I_{co\,max} - [1 - \alpha_{max}]\,I_{e\,max} + [1 - \alpha_{min}]\,I_{e\,min}} \qquad (34)$$

Finally, the values of R1 and R2 needed to construct a voltage divider that will yield the correct base supply voltage (E_b) as determined from equation (33) may be found from:

$$R1 = \frac{R_b \times E_c}{E_b} \qquad (35)$$

and

$$R2 = \frac{R1 \times E_b}{E_c - E_b} \qquad (36)$$

29. Design Procedure Using Equations (33) through (36)

The basic design equations are used as described in these steps:

Step 1: Using equation (34), R_b is determined for several values of assumed emitter resistance. Since R_e is usually in the range between 500 and 5000 ohms, three values of this order are assumed and, together with other data supplied by the manufacturer's literature, are substituted in equation (34) to find corresponding values of R_b.

Step 2: Any of the determined values of R_b is then substituted in equation (33) to find the voltage needed as base supply potential (E_b). If this voltage is too high or low for the overall power supply plan, another value of R_b as previously determined is tried.

Step 3: Using equations (35) and (36), R1 and R2 are then determined. In using these equations, E_c must have been predetermined on the basis of available or desired battery or batteries.

30. Example of Actual Design Procedure

A 2N190 p-n-p transistor is to be used as a single-stage audio amplifier in the circuit of Fig. 33. The typical characteristics as given by the manufacturer are:

$E_c = 12$ volts	$\alpha_{max} = 0.985$	$I_{e\,min} = 0.80$ ma
$R_c = 5$ K ohms	$\alpha_{min} = 0.970$	$V_{b\,max}$ and $V_{b\,min}$
$I_{co\,max} = 100\ \mu$amp	$I_{e\,max} = 1.25$ ma	for germanium

We might illustrate the design process by selecting a 1000-ohm resistor for R_e. Substituting this, and the values of the other factors given for the 2N190 transistor in equation (34), we have:

$$R_b = \frac{(1.25 \times 10^{-3} - 0.80 \times 10^{-3})(10^3) + (-0.1 - 0.2)}{100 \times 10^{-6} - (1 - 0.985)(1.25 \times 10^{-3}) + (1 - 0.970)(0.80 \times 10^{-3})}$$

Simplifying and solving for R_b, we find that:

$$R_b = 1200 \text{ ohms}$$

As this appears a reasonable value for R_b, we proceed to substitute $R_b = 1200$ ohms and $R_e = 1000$ ohms into equation (33), using design center values for I_e, alpha, V_b, and I_{co}. Design center values are also provided by the transistor manufacturer, and for the 2N190 are as follows:

$$I_e = 1 \text{ ma}$$

$$\alpha = 0.973 \text{ volt}$$

$$V_b = 0.2 \text{ volts}$$

$$I_{co} = 10 \mu a$$

Thus:

$$E_b = 10^{-3}\,[\,(1 - 0.973)(1200) + 10^3] + 0.2 - (1200)(10^{-5})$$

which solves to

$$E_b = 1.22 \text{ volts}$$

To determine the values of the voltage divider components, we first find R1 from equation (35) and then use this value of R1 to determine R2 from equation (36).

Thus:

$$R1 = \frac{1200 \times 12}{1.22}$$

$$R1 = 12{,}000 \text{ ohms (approx.)}$$

And:

$$R2 = \frac{12{,}000 \times 1.22}{12 - 1.22}$$

$$= 1350 \text{ ohms}$$

The completely designed circuit appears in Fig. 35.

31. Determination of Operating Characteristics

A further design analysis of the circuit in Fig. 35 is now possible.

Quiescent collector current—This current is found by substituting quantities already determined in the design procedure in equation (32).

$$I_c = \frac{12 \times 1350}{1000\,(12{,}000 + 1350)}$$

$$= 1.2 \text{ ma}$$

Input resistance—If C_e is made sufficiently large to prevent significant degenerative effects for the frequencies being amplified, then the input resistance of this circuit is given by:

$$R_{in} = (1 + \beta)r_b \qquad (37)$$

In this case, r_b refers to the base input resistance. This figure for the 2N190 is given by the manufacturer as 29 ohms. The beta of this transistor is listed as 36. Hence,

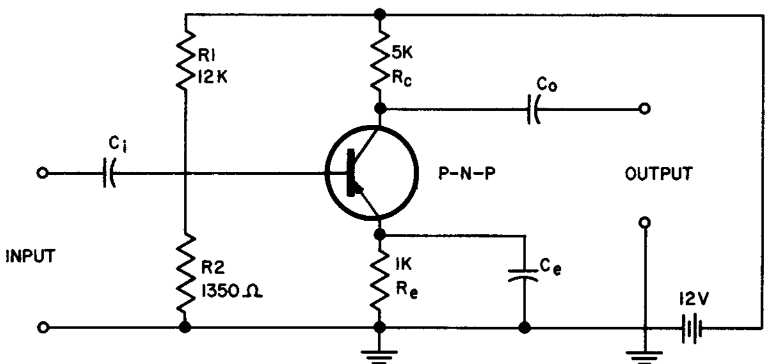


Fig. 35. Specially designed transistor amplifier circuit.

$$R_{in} = (1 + 36) \times 29$$
$$= 37 \times 29 = 1073 \text{ ohms}$$

Signal voltage gain—This is defined as the ratio of output voltage to input voltage and is very closely expressed by:

$$\text{Voltage gain} = \frac{R_c}{r_b} \tag{38}$$

For the circuit of Fig. 35, the voltage gain would be:

$$\text{Voltage gain} = \frac{5000}{29}$$
$$= 172$$

Low-frequency cutoff—The frequency at which the voltage gain is down 3db from its value at design frequency (usually 1 kc) is a function of the generator resistance (r_g), beta, and C_e. Since C_e is usually of the order of 50 μf, then assuming a generator impedance of 1000 ohms to match the input resistance obtained above, we can determine the cutoff frequency by:

$$f_{co} = \frac{1 + \beta}{2\pi (r_g \times C_e)} \tag{39}$$

so for this circuit

$$f_{co} = \frac{1 + 36}{6.28 \times 1000 \times 50 \times 10^{-6}}$$
$$= 120 \text{ cps}$$

32. Class-A Amplifier Output Stage Design

A class-A output stage can be designed most readily by making use of the collector characteristics as provided by the manufacturer. As an example, consider the collector characteristics given in Fig. 36. These represent the variations of collector current as a function of collector voltage with base current as a parameter. The broken-line curve outlines the limit of power dissipation, in this case 200 mw, in the collector circuit. Operation must be confined to the region to the left of the broken line to avoid exceeding the collector dissipation rating.

The design procedure is begun by selecting the value of the collector supply voltage, taking into consideration the absolute maximum rating of the transistor. Assume that the transistor in question has an absolute maximum rating for $E_c = -25$ volts. A good start would be to try approximately half of this, or -12 volts.

A load line is then drawn into the family of curves by laying a straight edge on the graph paper so that the line drawn along this edge will intersect the voltage axis at -12 volts, and proceed

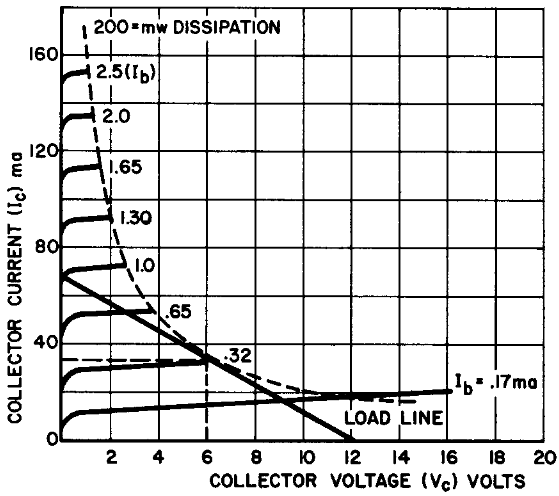


Fig. 36. Collector characteristics for a typical medium-power audio output transistor.

upward and to the left, just skimming the lower curvature of the power dissipation line, until it reaches the current axis. (See load line in Fig. 36.) A load line drawn in this manner insures operation within power dissipation limits at all times.

To permit the output signal to swing equally in the positive and negative direction, biasing conditions are set up so that V_c is exactly half of E_c, or -6 volts in this case. As may be seen from the curves, the base current (I_b) required to establish this bias is approximately 0.32 ma. The bias supply (voltage divider or battery) must then be arranged to supply this base current.

For a base current of 0.32 ma and a collector voltage of -6 volts, the collector current is evidently 34 ma. This provides us with sufficient data to compute the important operating characteristics.

Power output—Power output is obtained from the expression:

$$P_o = \frac{V_c I_c}{2} \tag{40}$$

$$= \frac{6 \times 0.034}{2}$$

$$= 102 \text{ mw}$$

or

$$R_L = \frac{V_c}{I_c} \tag{41}$$

$$= \frac{6}{0.034} = 175 \text{ ohms (approx.)}$$

Power gain—For a power output of this magnitude, the load resistance is negligible compared to the transistor output impedance. For this condition, little error may be anticipated by con-

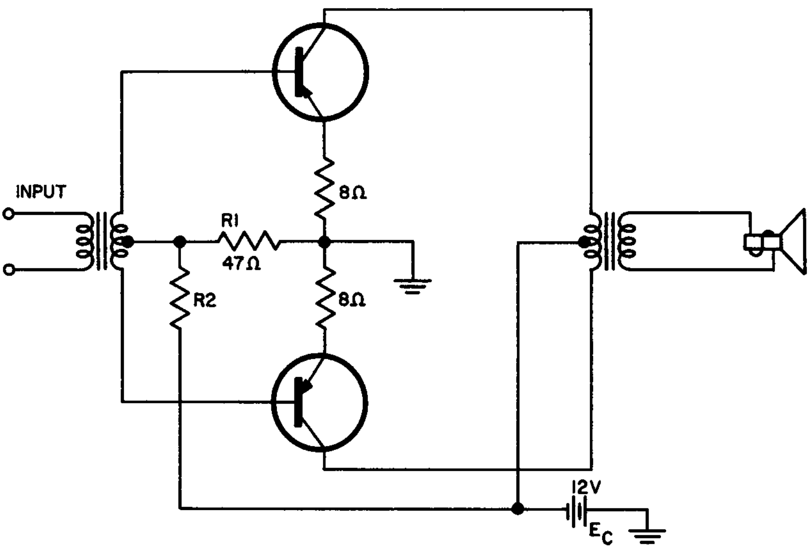


Fig. 37. Typical class-B audio output stage.

sidering the current gain of the transistor essentially the same as the short-circuit current gain beta. Thus, the power gain may be expressed as:

$$\text{Power gain} = \frac{\beta^2 R_L}{R_{in}} \tag{42}$$

in which β is given as 60, and input impedance (R_{in}) is given as 1200 ohms. Thus, the power gain is:

$$\text{Power gain} = \frac{(60)^2 \times 175}{1200}$$
$$= 525$$

33. Class-B Push-Pull Output Stages

A typical class-B push-pull output stage appears in Fig. 37.

The 8-ohm resistors are present to prevent thermal runaway if the junction temperature rises above 60°C. A small forward bias of about 0.1 volt is necessary to prevent crossover modulation distortion.

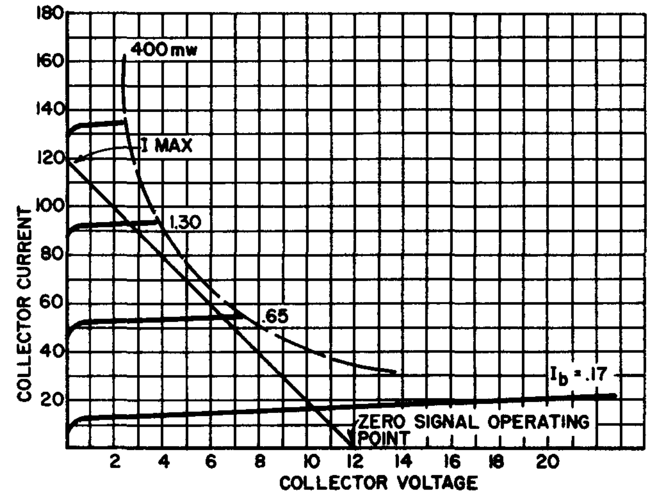


Fig. 38. Two typical P-N-P's in push-pull Class B.

tortion. In general, the value of R1 is recommended by engineering data from the transistor manufacturer, but R2 must be determined. In this circuit, with −12 volts of collector supply potential, R2 would be about 4700 ohms to obtain the required 0.1 volt. In some cases, it might be necessary to make small adjustments on R2 to minimize crossover distortion.

Power output—With a small forward bias applied as explained in the previous paragraph, the no-signal collector current (total for both transistors) is approximately 1.5 ma. For a current of this small magnitude, the collector voltage (V_c) is virtually the same as the source voltage (E_c). The power output for class-B conditions as described may be shown to be given by:

$$P_o = \frac{I_{max} \times E_c}{2} \tag{43}$$

so that for the load resistance given in Fig. 38, the power output is:

$$P_o = \frac{0.120 \times 12}{2}$$

$$= 720 \text{ mw}$$

It will be observed that the power dissipation limit in this case is double that of the single transistor used in class-A in the preceding paragraph. This permits a higher maximum collector current (I_{max}) and, consequently, a higher power output.

The load resistance is different in the class-B circuit, too. This is given by the equation:

$$R_L = \frac{E_c}{I_{max}} \tag{44}$$

$$= \frac{12}{0.120}$$

$$= 100 \text{ ohms}$$

Power gain—Fundamentally, the power gain is:

$$\text{Power gain} = \frac{P_o}{P_i} \tag{45}$$

where P_i is power input, thus:

$$\text{Power gain} = \frac{I_c^2 R_c}{I_b^2 R_i} \tag{46}$$

where R_i is input resistance.

For small values of load resistance, however, the ratio I_c/I_b is the current gain beta [see equation (30)]. Thus:

$$\text{Power gain} = \beta^2 \frac{R_{cc}}{R_{bb}} \tag{47}$$

where R_{cc} is the collector-to-collector resistance and R_{bb} is the base-to-base resistance in the push-pull circuit. The figure given in the manufacturers' literature for R_{bb} is 4000 ohms; we have already determined that R_{cc} is 100 ohms, since R_{cc} is the same as R_L, the load impedance. Hence, the power gain in this circuit is:

$$\text{Power gain} = (60)^2 \frac{100}{4000}$$
$$= 90$$

Note that the power gain in class-B is appreciably less than it is for a single transistor in class-A. This is quite normal and indicates that a class-A driver should precede the output stage if full power output is to be obtained.

34. Review Questions

1. Explain why alpha and beta are not truly transistor parameters.
2. Explain why the biasing system shown in Fig. 32 would not be satisfactory for use in a mass-produced transistor radio.
3. Why does the feedback circuit of Fig. 33 reduce instability in a transistor low-frequency amplifier?
4. Calculate the current in the collector circuit (Fig 33) if the collector supply voltage is -20 volts, the emitter resistance is 1000 ohms, and R1 and R2 are 100,000 and 10,000, respectively.
5. Define I_{co}. Which of the following values would most closely approximate the I_{co} of a good audio transistor? ($1\mu a$, $50\mu a$, 1 ma, 100 ma)
6. Design a transistor amplifier stage, using a 2N190 transistor for a supply potential of 20 volts and a collector resistor of 7500 ohms. All other circuit values and parameters may be taken to be the same as those given in Section 30.
7. Determine all of the operating characteristics of the amplifier stage designed in (6).
8. A class-A amplifier stage for audio output is to be designed around the transistor and its associated characteristics described in Section 32. Assume that you have been told to limit the maximum collector current to 50 ma. Redesign the circuit, taking into account this additional limitation.
9. Determine the operating characteristics of the newly designed stage.
10. Explain why a class-B push-pull output stage requires more driving energy than a class-A stage using the same transistor.

INDEX