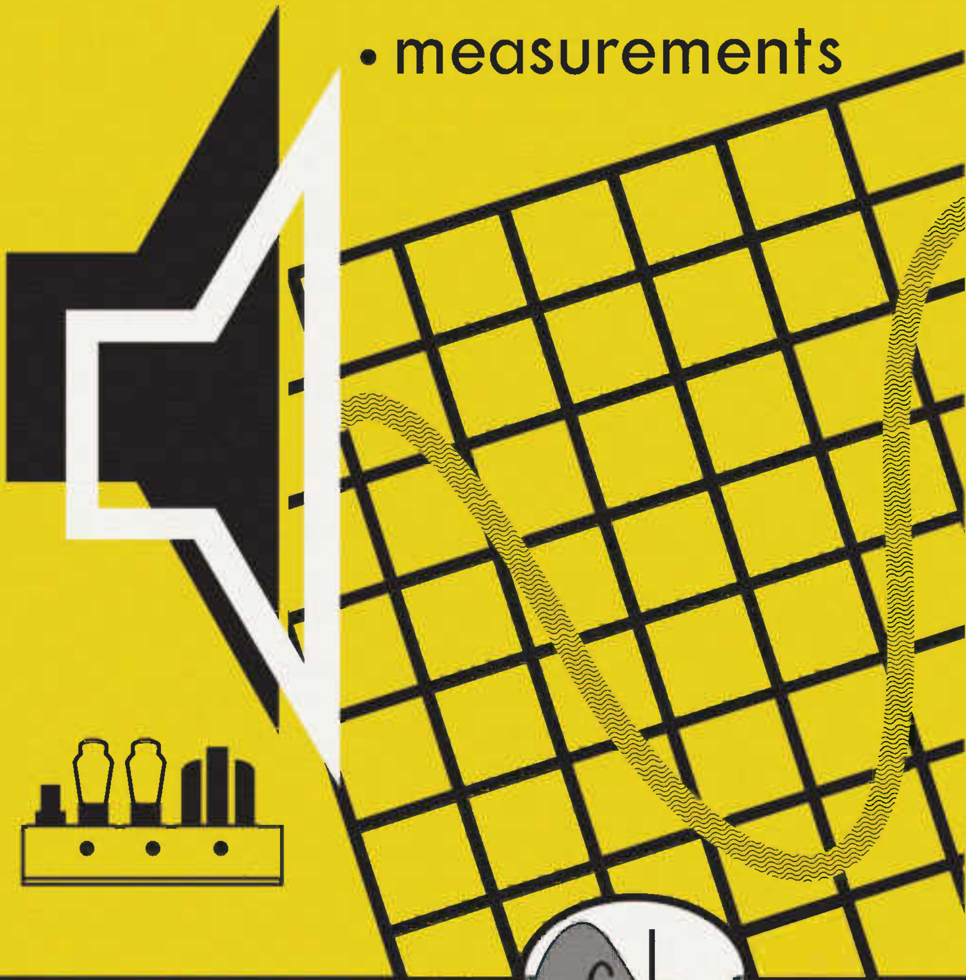
high-fidelity

- design
- construction
- measurements



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HIGH-FIDELITY

- **Design**
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Introduction

AUDIOPHILES are a unique group. They are thoroughly saturated with the tantalizing idea that sound, a complex arrangement of fantastic waveforms, can be stuffed into complex electro-mechanical apparatus and emerge amplified, but so unchanged, that the listener is deluded into thinking it is the unmolested original. To achieve this, they surround themselves with equalizers, crossover networks, amplifiers of assorted sizes, tried and trusted circuits (with and without variations), pursuing their goal with amazing decisiveness of purpose.

Realistic results have been secured. A listening test between audio equipment used but a decade ago and modern audio systems would impress itself even on the untrained ear. Unfortunately, audio quality seems to have increased cost as a running mate, and many of those who want and would appreciate fine audio equipment must let their purse and not their own technical qualifications act as the limiting factor. Also, there are many audio men who prefer to construct, design, and experiment with their own audio systems, rather than work within the confines of a commercially available set-up.

If you belong to these groups, then this book was written for you. Information-packed almost to the point of embarrassment, you can mull and dally a pleasant path through audio design, audio measurements, or audio construction. Be warned, however, that some of this very practical and useful information will cling to you and unless you are wary, will inevitably result in better amplifiers, better speaker arrangements, less hum, improved frequency response—in short, in more faithful, realistic and enjoyable reproduction of both speech and music. This might have the effect of starting you off in search of the perfect amplifier (if you are not already so started) and can only result in your having many happy hours in continuing the design and development.

Originally appearing in RADIO-ELECTRONICS Magazine, the material contained in this book represents the actual experience of many men to whom audio is either a daily business or hobby. The names of these contributors appear in the Table of Contents. We gratefully acknowledge the fact that their significant contributions to the field of audio have made this book possible.

MARTIN CLIFFORD

High-Fidelity

design

measurements

construction

Chapter 1

Audio Feedback Design

BEFORE we consider the amplifier as a whole, it is rather instructive to see what happens to a single tube with feedback. For an example which is reasonably typical, we have in Fig. 101 the characteristics of a

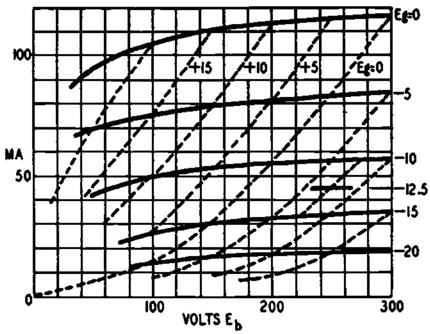


Fig. 101. Negative feedback from plate to grid makes the 6AQ5's characteristic (shown in solid lines) look like those of a triode tube (broken lines).

6AQ5 tube: a miniature which will give you 3-4 watts. It is a pentode, and at the standard working point, $E_b = 250$ volts, $E_g = -12.5$ volts, $I_b = 45$ ma, the transconductance is about 4,000 μmhos and the impedance about 50,000 ohms. This last figure is the real problem: the optimum load is 5,000 ohms, so that if you transform down to use a 15-ohm loudspeaker, the speaker, will be looking back at 150 ohms. It might as well be an open circuit, as far as damping down speaker resonances goes.

Let us see what happens to the tube characteristics if we apply negative feedback. To make the calculations easy, let us apply feedback of

1/10 of any plate voltage change to the grid. Starting at the point $E_g = 0$, $E_b = 300$, let us reduce E_g to –5 volts; then if E_b drops by 50 volts we shall feed back +5 volts to the grid, and we shall be on the same E_g line as the starting point. Similarly, the point $E_g = -10$, $E_b = 300-100$ is on this $E_g = 0$ line. We can follow this same idea through and get a sufficient number of points to give us a smooth curve, as shown by the dashed lines in Fig. 101. The same technique can be employed to produce a family of curves.

The feedback tube curves appear as broken lines. If you will forget that we have a pentode you will see that the new characteristics are those of a triode, with an impedance of 2,100 ohms at the original working point. If you connect this plate-grid feedback around a pentode and put it in a black box, anyone measuring the characteristics will think they have a triode. There is only one difference: you can swing this triode down to a plate voltage of 50 without running into grid current; the corresponding limit when the screen and plate are joined together is about 170 volts.

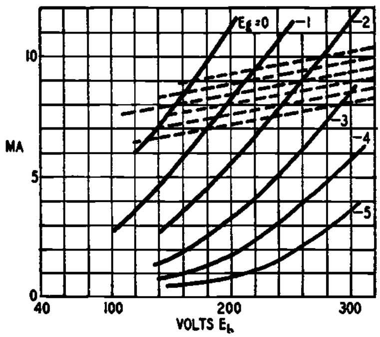


Fig. 102. Negative current feedback will alter the curves of a triode so they will take on the high-impedance characteristic of a pentode (broken lines).

The results of Fig. 101 are well worth a closer study. In the area below $E_g = 0$ there is no grid current, because the feedback circuit will be an a.c. coupling. A pentode with negative feedback of this kind has the same characteristics as a triode, including the triode positive grid region, but you can work over the whole set of characteristics without grid current trouble. This is stressed because there are audio men who do not believe that tetrodes can sound as good as triodes. Cast your eye over those curves and you can see they are better than triode curves!

We can, to some extent, reverse this effect. A typical triode characteristic is shown in Fig. 102. This triode has a mu of about 50, and an impedance of about 20,000 ohms. By putting in a cathode resistor of 2,000 ohms we can make a change of 1 ma in the plate current feed back 2 volts to the grid. The effect of this is shown in part by the broken lines in Fig. 102. These are part of the characteristics of a "black box" tube having an impedance of about 105,000 ohms, which is getting on toward the pentode class. If you need a pentode, and only have a triode handy, this is one way of making the circuit think it sees a pentode.

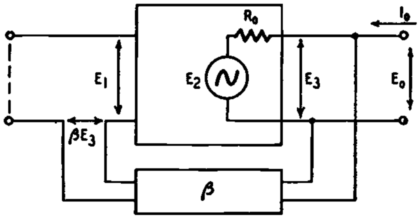


Fig. 103. Voltage feedback network used for calculating the output impedance.

These two examples of modified tube characteristics have been considered because they provide a useful background to the general discussion of amplifier impedances. Ultimately any amplifier can be considered as a 4-terminal network in a box, and if you don't look inside you cannot be certain that it is not just one tube, with a transconductance of 1 amp/volt, perhaps! Now we can turn to the general amplifier circuit. Fig. 103 shows the general voltage feedback ampli-

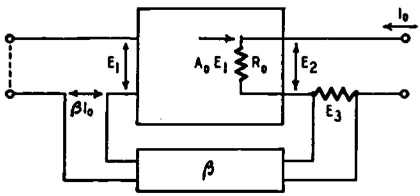


Fig. 104. Network using current feedback.

fier, with input short-circuited and output connected to a generator. This generator produces a voltage of E_0 at the output terminals, and the current flowing into the amplifier is I_0. The amplifier itself has an impedance of R_0 and an *open-circuit* gain E_2/E_1 of K_0. (The gain of the amplifier without feedback is K). Notice that you are using the open-circuit gain which will be quite a lot higher than the usual loaded gain. For a triode the difference is 6 db, but for a pentode or tetrode it may be much more. The feedback network β is assumed to be of such high impedance that it does not affect the impedances.

In the output mesh, we have this equation:

$$E_0 - E_2 = I_0 R_0$$

Now, $E_2 = K_0 E_1$ and since in this particular case the only input is that provided by the feedback network $E_1 = \beta E_3$ or, indeed, $-\beta E_0$ since E_3 and E_0 are the same here.

Arranging our equations in order, we have:

$$E_2 = K_0 E_1$$
$$E_1 = -\beta E_0$$
$$E_2 = K_0 \beta E_0$$
$$-E_2 = K_0 \beta E_0$$

Thus $-E_2 = K_0 \beta E_0$ and we can substitute in the previous equation:

$$E_0 + K_0 \beta E_0 = I_0 R_0$$
$$E_0 \ (1 + K_0 \beta) = I_0 R_0$$
$$\frac{E_0 \ (1 + K_0 \beta)}{I_0 \ (1 + K_0 \beta)} = \frac{I_0 R_0}{I_0 \ (1 + K_0 \beta)}$$
$$\text{or } \frac{E_0}{I_0} = \frac{R_0}{1 + K_0 \beta}$$

Of course E_0/I_0 is the impedance seen by the generator connected to the output, and if the feedback were absent, $\beta = 0$, the impedance would be R_0. With negative feedback the impedance is reduced by the factor $(1 + K_0 \beta)$.

This feedback is voltage feedback. We could use a circuit like that of Fig. 104, in which the feedback voltage depends on the *current* in the output circuit. To keep the formulas very simple, the resistance across which the feedback is picked off will be assumed to be small, just as before we assumed that the β network was of infinite impedance. In calculating this circuit we work in terms of current: the amplifier is assumed to have a transconductance of A, under short-circuit conditions, so that it produces an output current of $A_0 E_1$ for input of E_1. When we apply an additional current of I_0 we have a current of $(I_0 + A_0 E_1)$ through R_0, so that the voltage across the output terminals is:

$$E_2 = R_0 \ (I_0 + A_0 E_1)$$

The feedback network (including the small resistor across which E_3 is produced) delivers a voltage βI_0 to the input terminals, and since the input has no other supply $E_1 = \beta I_0$.

Thus $E_2 = R_0 \ (I_0 + A_0 \beta I_0)$.

The admittance seen at the output is I_0/E_2 and is:

$$\frac{1}{R_0 \ (1 + A_0 \beta)}$$
$$\text{or } \frac{1}{R_0} \times \frac{1}{1 + A_0 \beta}$$

If there were no feedback ($\beta = 0$) this would be just $1/R_0$. With current negative feedback, therefore, the output *admittance* is reduced by the factor $(1 + A_0\beta)$. This means, of course, that the impedance is *increased* by this factor.

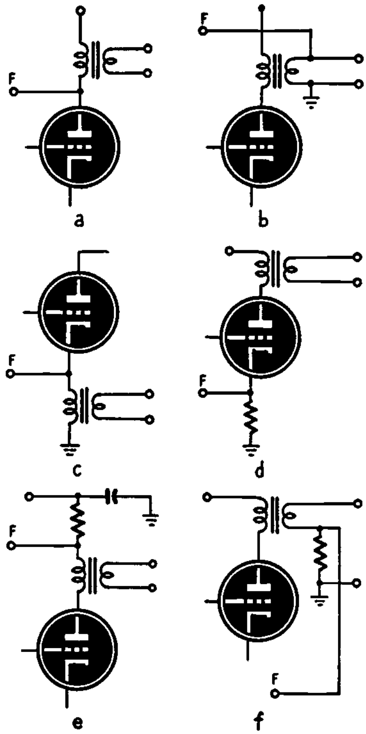


Fig. 105. Six ways to pick off feedback F. a, b, and c, are voltage feedback; others are current feedback.

It must be noted that the factor A_0 is not the same as K_0. This means that we must make a separate calculation when determining the output impedance. When using voltage feedback we must use the equation

$$M = \frac{\mu R}{R + R_0}$$

for the gain of the last stage in determining the gain factor K, but $M_0 = \mu$ for the impedance factor K_0. If, however, the feedback network uses resistors of normal value, we may find it more accurate to write

$$M_0 = \frac{\mu R'}{R' + R_0}$$

where R$'$ is the input impedance of the feedback network. We also ought to bring the output transformer losses into R′. Usually this means that K_0 is about 3 times the usual value of K.

By using positive feedback, the term $(1 + K_0\beta)$ can be made less than unity. Thus, positive voltage feedback increases the output impedance and positive current feedback reduces the output impedance. We can mix positive voltage feedback and negative current feedback to give a very high impedance, for example, without losing too much gain. As examples of what can be done, one small amplifier uses a 12AT7, which has an impedance, at the high side of the output transformer, of something around 5 megohms. This is better than pentode performance. Another amplifier, using pentodes, gives an impedance of

Table for Phasing Feedback Connections				
No. of stages	Circuits of Fig. 105			
	a	c	d	e
1	g−			g−
2	k− g+	g− k+	g− k+	k− g+
3	g− k+	k− g+	k− g+	g− k+
4	k− g+	g− k+	g− k+	k− g+
Circuits b and f depend on transformer sense.				

0.1 ohm when designed to work into a 25-ohm load: this uses positive current and negative voltage feedbacks. There are good reasons for these designs: the pentode is needed to get power at low supply voltage, the 12AT7 to get gain from a single bottle, and the extreme impedances are "musts."

The circuits for giving voltage and current feedback are summarized in Fig. 105. F is the feedback voltage in each case, and a, b, c are voltage feedback, d, e, f are current feedback. The table is a summary showing whether the feedback should go to cathode (k) or grid (g) of an earlier stage to be negative or positive. The plus signs show positive feedback, the minus signs negative. Thus in a 3-stage circuit you must feed back from cathode to cathode (circuit d) or from anode to grid (circuit e) to get negative current feedback.

For the sake of completeness we should examine the bridge feedback circuit. In this the feedback does not affect the impedance. Tube impedances are not so constant that anyone wants to keep them unaffected.

Now let us consider the input impedance. A typical input circuit is shown in Fig. 106, and although you will notice a resistor between grid and cathode, this could be a more complicated impedance. The biasing arrangements are also neglected. In this circuit, the input to the amplifier section K is V_1, and the voltage developed by the feedback circuit is $V_2 = K\beta V_1$. We must have $E_0 = E_1 + E_2 = (1 + K\beta)\ E_1$ and from this we could deduce that the gain would fall by a factor $(1 + K\beta)$ when feedback was added, because we now need an input of E_0 to produce E_1 at the grid.

The current which flows through R is given by

$$I_0 = E_1/R = E_0/R\ (1 + K\beta).$$

So far as any circuit connected to the input can tell, this current is produced by E_0, so that the input impedance is

$$\frac{E_0}{I_0} = R\ (1+\ V_1\beta).$$

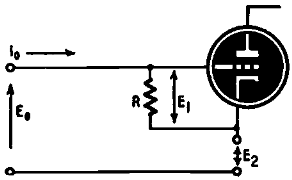


Fig. 106. Amplifier input circuit having series-connected feedback at cathode.

The impedance therefore has been increased by a factor $(1 + K\beta)$. This may be 10 to 100, so that a resistance of 1 megohm, which is all we can normally use because of gas current in the tube, looks like 10-100 megohms to the external circuit. The input capacitance is reduced, too, which is important if you are using a crystal pickup. Notice, however, that positive feedback in this connection will reduce the input impedance, since then $(1 + K\beta)$ is less than unity.

An alternate way of connecting the feedback is shown in Fig. 107. The current which flows into this circuit is

$$I_0 = \frac{E_0 - E_2}{R_1 + R_2}.$$

Now $E_2 = -K\beta E_1$ and $E_1 = E_0 - I_0 R_1$.

We therefore have

$$E_2 = -K\beta\ (E_0 - I_0R_1)$$

$$\text{or } E_2 = -K\beta E_0 + K\beta I_0 R_1$$

$$\text{and } I_0\ (R_1 + R_2) = E_0 + K\beta E_0 - K\beta I_0 R_1,$$

which we rearrange

$$I_0\ (R_1 + R_2 + K\beta R_1) = E_0\ (1 + K\beta),$$

$$\text{giving } \frac{E_0}{I_0} = R_1 + \frac{R_2}{1 + K\beta}.$$

The R_1 term is the impedance of the external generator, so that the input impedance, R_2 without feedback, is reduced by the factor $(1 + K\beta)$ with negative feedback. As before, if the feedback is positive this is an increase in input impedance. It would seem to be quite possible to apply negative feedback to the cathode and positive to the grid to produce an extraordinarily high input impedance. The negative feedback would stabilize the gain so that the term $(1 - K\beta)$ in the positive feedback equation could be made quite small.

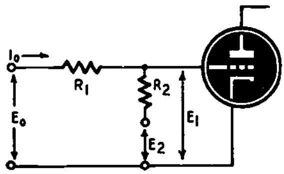


Fig. 107. *A parallel hookup for feedback.*

In many applications it is necessary to provide a controlled value of input impedance, which involves the use of the circuit of Fig. 106 with an additional shunt resistor across the input. Often, too, we need a good 600-ohm output impedance, and then a high impedance shunted by 600 ohms is used.

This has only been a general survey of the problems of impedance control. If an amplifier is needed with some special impedance properties, a closer study may be required, but for almost all jobs the discussion here will be sufficient. Always the effect of feedback is to modify the apparent tube characteristics, and it is probably wise to point out one important thing. If you turn a tetrode into a triode by means of feedback, the optimum load is unchanged: you should not try to match this "triode." You can check this statement very carefully, both with modified characteristics and by actual experiment. This is specially important if you are using a high degree of feedback to get a low impedance for damping a loudspeaker, or to get a high impedance for some other purpose. And don't forget, if you want a low impedance, to allow for the resistance of the output transformer windings. This is also important.

Chapter 2

Equalizer Design

EQUALIZER circuits look deceptively simple: a few resistors and a capacitor or two; but even when prescribed values are used they may fail to deliver the expected results. Many articles and books on the subject safeguard themselves against this possibility with a postscript to the effect that "The foregoing method will provide rough values. *Exact values* for any specific application will have to be determined by test." In other words, "This gives you a general idea; if you have all the necessary test equipment and enough patience, you *may*

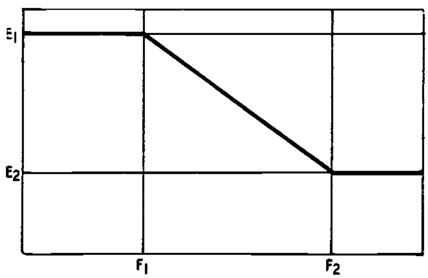


Fig. 201. Ideal equalizer characteristic.

eventually get it right. If you haven't—well, just hope for the best."

Fortunately, this rather pessimistic approach to equalizer design is quite unnecessary if you understand and use the right methods. An equalizer has to provide a definite boost (or drop) in the response of a circuit between two specified frequencies. The *amount* of boost or drop may be expressed in decibels, or as a voltage ratio (db units are generally used in audio work). Since the two limit frequencies—called "turnover points"—are usually fairly far apart (at least in standard recording practice) an ideal equalizer would have the characteristics shown in Fig. 201. The response of the circuit (E_1) is perfectly flat up to the lower turnover point (F_1); then it drops *uniformly* to the desired level (E_2) at the high-end turnover frequency (F_2); then flattens out again.

Fig. 202 shows two basic resistance-capacitance equalizer circuits. The arrangement in Fig. 202-a is for high-frequency boost; Fig. 202-b is for low frequency boost. (Inductors can be used instead of capacitors—with the opposite characteristics, of course—but capacitors are generally used because they are free from hum-pickup troubles, and are usually much less expensive than inductors).

Both circuits are essentially voltage dividers. In Fig 202-a the top resistive element (R1) is short-circuited at the highest frequencies by capacitor C_H. Thus at the higher frequencies the output voltage nearly equals the input voltage, giving the effect of a high-frequency boost. In Fig. 202-b the lower resistor (R2) is open-circuited at low frequencies

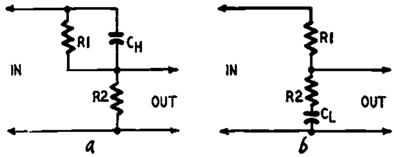


Fig. 202. Basic R-C equalizer circuits. (a) high-end boost. (b) low-end boost.

by the capacitor C_L. As the frequency is decreased, the reactance of C_L rises. This reactance is in series with R2, consequently the signal voltage across this series combination goes up with a lowering of frequency.

Two mistakes are commonly made in dealing with these circuits:

1. It is assumed that they produce ideal steps with a slope of 6 db per octave (an octave is a 2:1 frequency ratio), and that the turnover frequencies are exactly at the opposite ends of the step. The actual slopes are invariably less than 6 db, and the turnover points are never exactly at the ends.

2. The effects of the input and output impedances in the circuit where the equalizer is connected are neglected.

The most important actual characteristics of each circuit can be found by checking its response at three frequencies. Two of these are the turnover frequencies F_1 and F_2; the third (F_3) is the geometric mid-frequency $F_3 = \sqrt{F_1 \times F_2}$. F_3 is useful because the actual response curve has its steepest slope at this point.

The insert in Fig. 203 is a generalized equalization curve showing the important factors. (Although this represents the performance of a high-frequency boost circuit—Fig. 202a) it can be applied just as well to bass-boost equalizers (Fig. 202-b) by simply reversing the curve. Frequencies F_1 and F_2 are the "turnover" points of the curve.

The straight line A in Fig. 203 gives the relation between the over-all height of the step and the ratio $R1/R_T$ or the ratio between the reference frequencies F_1 and F_2. Curve A has a uniform slope of 6 db

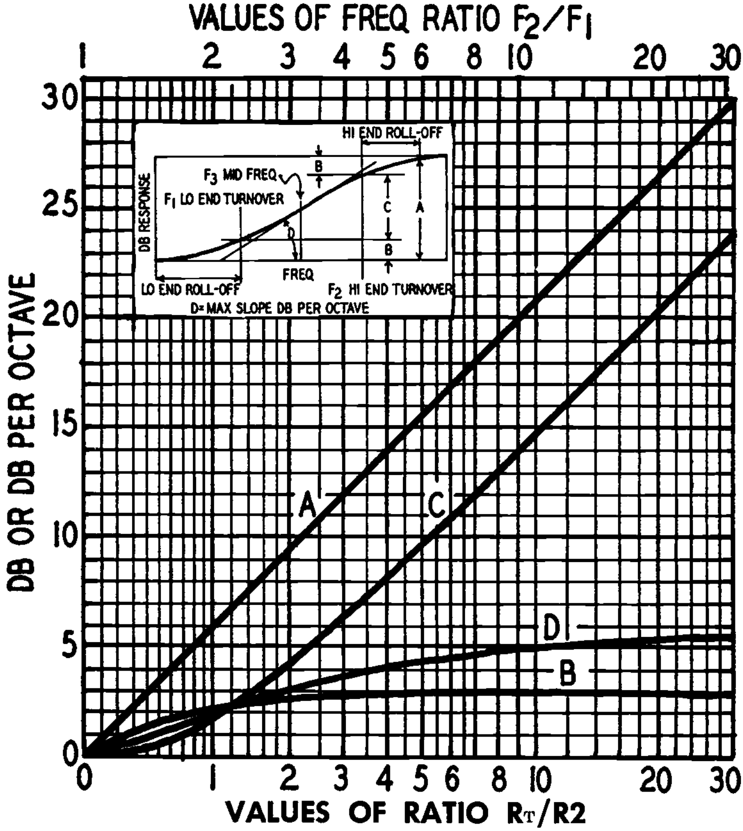


Fig. 203. Equalization curve (insert), and chart for finding characteristics with circuit combinations. "A" is over-all height of the step; "B" is height of roll-off at each end of the step; "C" shows net boost (or drop) between turnover frequencies F_1 and F_2; "D" gives maximum slope at the midfrequency point F_3.

per octave. Curve B show how much the response at the turnover frequencies F_1 and F_2 departs from the nearest level portion of the over-all response curve; for big steps this is very nearly 3 db. Curve C shows the difference in response between the turnover frequencies F_1 and F_2; curve D gives the *maximum slope* in db per octave (at mid-frequency F_3).

Some typical solutions based on the use of this chart are shown in Fig. 204. In Fig. 204-a, for an over-all step height of 3 db, the turnover frequencies have a ratio 1.4/1 (half an octave). The maximum

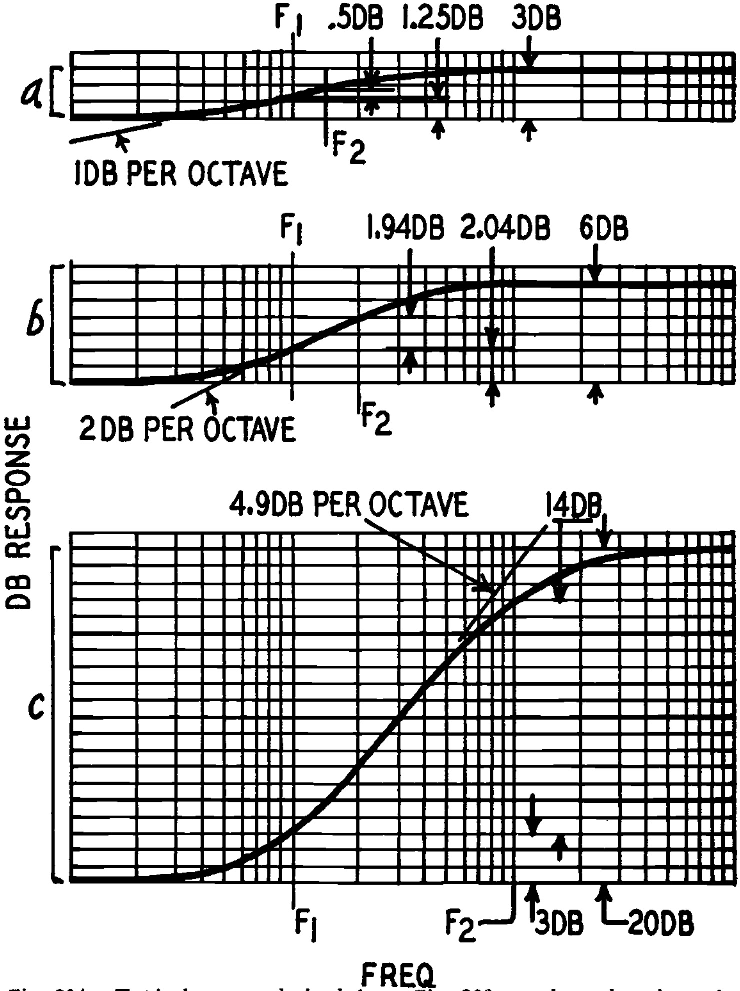


Fig. 204. Typical curves derived from Fig. 203. a, b, and c show the separation between F_1 and F_2. Note that the maximum slope is still less than 6 db per octave.

slope is about 1 db per octave, and there is only about 0.5 db difference in level between F_1 and F_2. There is a rolloff of 1.25 db. between each of the turnover frequencies and the ultimate levels.

In Fig. 204-b, F_1 and F_2 are separated by an octave; the maximum slope is 2 db per octave, and the difference in level between F_1 and F_2 is nearly 2 db. There is a 2-db rolloff at each end.

In Fig. 204-c, F_1 and F_2 are separated by a decade (3.32 octaves or a 10:1 frequency ratio), giving a 20-db over-all step, and a 14-db difference between F_1 and F_2. Even in this case the *maximum slope* is still less than 6 db per octave, and the *average* slope over the 10-to-1 frequency range is only 4.5 db per octave.

Terminal impedances

Now for the second source of error. If an equalizer designed according to the methods in the preceding section is connected between two existing amplifier stages, the results may not conform to expectations. The predicted response would be obtained *if* the equalizer input voltage was the same at all frequencies. Unfortunately, the input impedance changes with frequency, and so does the voltage. The grid-circuit impedance of the following stage also changes with frequency and affects the performance. These plate- and grid-circuit impedances must be included in the design calculations for the equalizer.

High-end boost

Fig. 205 is the basic circuit of Fig. 202-a modified to include these

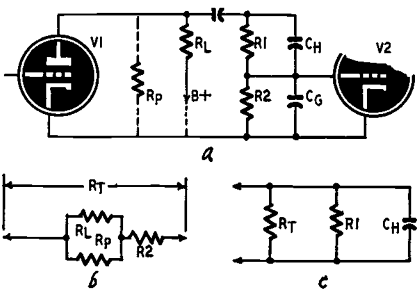


Fig. 205. (a) Complete h.f.-boost equalizer showing elements affecting its characteristics. (b) Equivalent source impedance for finding l.f. turnover point. (c) Equivalent circuit for finding h.f. turnover frequency.

source and output impedances. The low-end turnover frequency is found by making the reactance of C_H equal to R1. But the remaining resistance in the loop (R_T) is now R2 in series with the parallel combination of R_p (the plate resistance of V1) and the load resistance R_L (Fig. 205-b). The high-frequency turnover, F_2, is found by making the reactance of C_H equal to the parallel combination of R1 and R_T (Fig. 205-c). In practice it will only be necessary to find one of these frequencies. F_1 is the easiest to calculate.

As an example in designing an equalizer, suppose that R_p is 1.5 megohms, R_L is 330,000 ohms, and R2 is 470,000 ohms. A high-frequency-boost step of 15 db is required between F_1 and F_2, starting at a low-frequency turnover (F_1) of 2,000 cycles. We can use the equivalent circuit in Fig. 205-b to find the value of R_T. R_p in parallel with R_L gives about 280,000 ohms net plate-circuit resistance; this added to 470,000 ohms (R2) gives 750,000 ohms for R_T. A 15-db boost calls for a ratio $R1/R_T$ of 4.6 (curve A in Fig. 203), so that R1 works out to about 3.5 megohms. A standard value of 3.3 megohms would be close enough. The circuit will show a boost of 2.9 db (curve B in Fig. 203), when C_H has a reactance of 3.3 megohms at 2,000 cycles (25 μμf).

Now the grid-circuit impedance of V2 enters the picture. If the

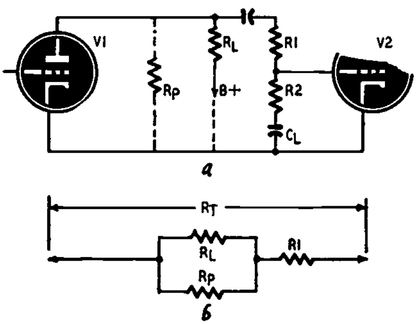


Fig. 206. (a) Low-frequency-boost equalizer. (b) Equivalent input impedance.

input capacitance (C_G), is 25 μμf, the total attenuation will be 6 db more than anticipated, because C_H and C_G form a 2-to-1 capacitive voltage divider at the highest frequencies. Using a smaller value for R2 will reduce the effect of C_G. With R2 equal to 100,000 ohms, R_T becomes 380,000 ohms. R1 then works out to 1.75 megohms (standard value 1.8 megohms). C_H would then be 50 μμf for a 2,000-cycle low-frequency turnover. With C_G equal to 25 μμf, the added loss is only 3.5 db.

Increasing the over-all height of the step will offset this loss still further. Using a figure of 20 db, the ratio $R1/R_T$ will be 9, according to curve A in Fig. 203. R1 works out to $9 \times 380{,}000$, or about 3.3 megohms, which again calls for $C_H = 25$ μμf to lift at 2,000 cycles. The loss due to C_G will be 6 db, leaving a net step height of 20 –6, or 14 db.

In any case it pays to reduce the V2 input capacitance to the smallest possible value. One method is to develop a large amount of inverse feedback in V2 by using an unbypassed cathode resistor, or by operating V2 as a cathode follower.

Bass-boost circuit

Now let us turn to the bass-boost equalizer following the basic circuit of Fig. 202-b. Input- and output-impedance factors are shown in Fig. 206. Here R2 is in series with C_L (Fig. 206-a) and the balance of the circuit is the series-parallel combination shown in Fig. 206-b. Fig.

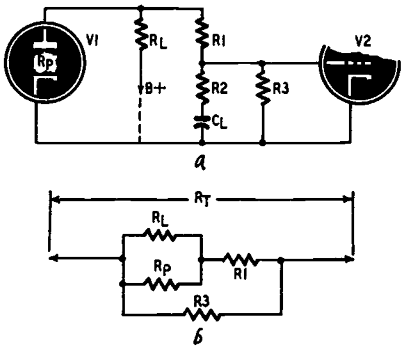


Fig. 207. Workable form of l.f.-boost equalizer, with d.c. return to ground.

206-a would not work in practice because the V2 grid circuit has no d.c. return to ground. Two workable arrangements are shown in Figs. 207 and 208, and the equivalent basic derivation is given under each. In Fig. 207, R3 shunts all the other elements (R2 is effectively open-

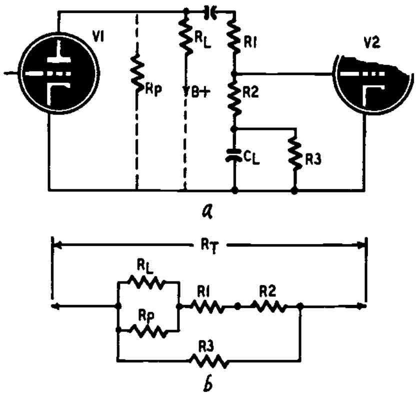


Fig. 208. Another form of low-frequency-boost equalizer. R3 may be a variable resistor for controlling the response.

circuited by C_L). This is obviously the simplest circuit to use. Fig. 208 is better where adjustable boost is required.

Turning back to the question of the errors arising from failure to take circuit impedances into account, take the example shown in Fig. 209. This network is intended to boost both ends of the frequency range. The values of R1 and R2 in the basic circuit of Fig. 209-a should give a lift of about 14 db at each end. If this arrangement is merely inserted between two stages, as in Fig. 209-b (ignoring for the moment the need for a d.c. grid return), the plate-circuit impedance of V1 will reduce the h.f. boost and increase the bass boost, throwing the whole arrangement off balance. With the values shown in Fig. 209-b, the plate-circuit resistance consisting of R_L and R_p in parallel is about 330,000 ohms; using the equivalent circuit of Fig. 205-b we find that R_T is 550,000 ohms, so the ratio $R1/R_T$ is only about 2. Consulting curve A in Fig. 203 we see that this gives less than 10 db top lift (ignoring possible effects due to input capacitance). For bass lift in Fig. 209-b, the

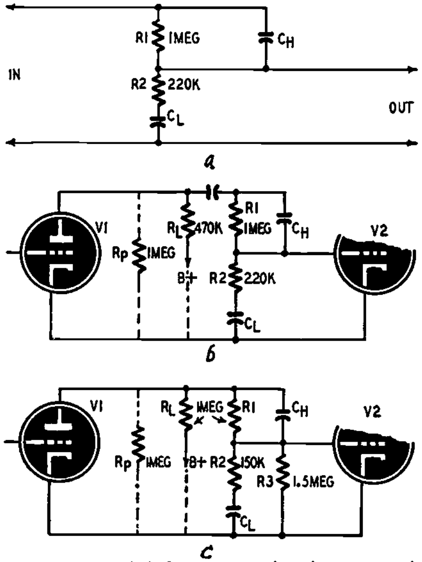


Fig. 209. *High-low-boost circuit. (a) Basic network. (b) Input and output circuits added. (c) Compensation for equal high and low boost by adding R3.*

increased reactance of C_L to over 3 megohms makes $R1/R_T$ about 6, giving nearly 16 db lift.

By careful choice of values in the circuit of Fig. 209-c, using the equivalents shown in Figs. 206 and 208, R3 will not affect the top lift materially, but its value can be adjusted to bring the bass boost to the required equal value. The values shown will give 14 db lift at each end. The reactances of capacitors C_H and C_L must be equal to R1 and R2 respectively at the high and low turnover frequencies. To avoid confusion the effect of grid input capacitance has been ignored; to include it, the procedure for designing the top lift will be as in the earlier example, after which the bass lift can be adjusted by R3 as just stated.

Chapter 3

Crossover Networks

SOME points about the functioning of loudspeaker crossover networks should be clarified. Most classical treatments derive crossover networks from wave filter theory, which in turn is derived from the theory of artificial lines. The resulting designs may not satisfy all the requirements of a crossover network in the best possible way.

The first and most obvious requirement of a crossover network is to deliver low-frequency energy to one speaker and high-frequency energy to another. This is usually taken care of reasonably well, using a roll-off slope to suit the designer's whim.

A second factor (and one that often receives less attention) is the impedance presented to the amplifier by the combined network. Frequently this varies widely over the audio range and includes sizeable reactive components in the vicinity of the crossover frequency. To obtain best performance from the amplifier, a constant, resistive impedance should be presented to it as a load.

The third requirement often receives even less attention. This is the *realism* of the acoustic output from the combination. While this is in some respects a matter of individual conditioning and preference, it depends fundamentally on certain electro-acoustic requirements. One of the most important, and frequently the one least considered, is the phasing, or apparent source of the sound output.

Tests have shown that phase distortion with single speakers is not normally detectable, but when *two* sources are employed, the phase relations between them influence the character of the radiated sound field. The frequency response as registered by a good pressure microphone may be flat, but what about the wave-shapes? The pressure microphone does not answer that, but a pair of human ears can detect such phase peculiarities, and failure to consider this factor has made many dual-unit combinations sound noticeably unreal, even though their frequency response may look perfect, and there may be no measurable distortion.

Source of the sound

You may have checked two speakers for phasing by listening to them while connections to one of them are reversed. Standing some distance in front of them, on the center line (as in Fig. 301), when they are correctly phased the sound seems to come from a point midway between the two units; but when incorrectly phased, two effects can be noticed: There is a deficiency here of low frequencies (due to cancel-

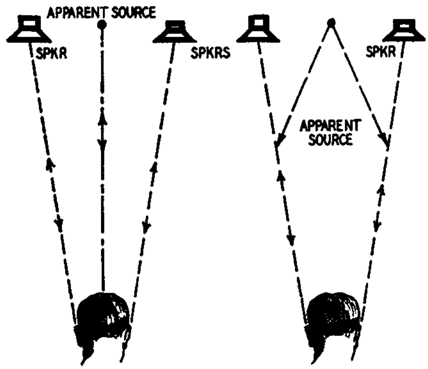


Fig. 301. The apparent source of sound shifts when speakers are not in phase.

lation effects) and the source of sound at higher frequencies does not seem to be associated with the units actually radiating it. This is because the air-particle movement caused by the radiated sound is no longer back and forth along a line from the common source, but approximately at right angles to it. The sound field around the listener's head is perpendicular to what it should be, causing, through our binaural perception, the confused impression which may be called "dissociation effect."

A similar dissociation effect will occur with dual units driven by a crossover system, at any frequency where the two speakers are out of phase. It is best to keep the two units close together. Some favor putting the smaller unit on the axis of the larger one and immediately in front of it. But, however the units are arranged, the dissociation effect can become noticeable if there is an out-of-phase condition at some frequency near the crossover point. The relative phase at crossover can be adjusted by positioning the diaphragms on their common axis so the wave from the low-frequency unit emerges in phase with that from the high-frequency one. When the units are mounted side by side on the same baffle, the sound should emerge in phase at the baffle surface.

Constant-resistance networks

But how does the crossover network affect the relative phase at frequencies near crossover? This question often seems to be overlooked, and neglecting it can cause the trouble just described. Some networks, of the kind employing two or more reactances for each unit, have values adjusted to give an accentuated frequency rolloff. For example, the networks shown in Figs. 302-c and 302-d using values given by the chart

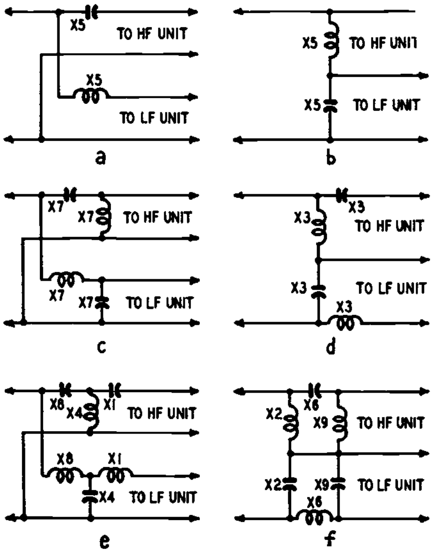


Fig. 302. Examples of crossover networks for two-way loudspeaker systems.

in this chapter, are of the constant-resistance type, giving a rolloff of 12 db per octave; the phase difference between the outputs is always 180 degrees; but using values designed for a sharper rolloff, the phase difference is not the same. At frequencies near crossover, phase difference changes rapidly. Some out-of-phase effect in the vicinity of the crossover frequency is unavoidable unless a constant-resistance type network is used. The chart appears on page 26.

The chart may be used to design any of the six types of crossover network illustrated in Fig. 302. Those in Fig. 302-a and Fig. 302-b give a rolloff of 6 db per octave and a constant phase difference of 90 degrees. For best results the positions of the two diaphragms should be adjusted so the difference in their distances from the face of the baffle is about one-quarter wavelength at the crossover frequency. The phase difference will not be serious within the range where appreciable energy is coming from both units.

The networks of Fig. 302-c and Fig. 302-d give a rolloff of 12 db per octave, and a constant phase difference of 180 degrees, which means that reversing connections to one unit will bring the phase right. The units should be mounted so their diaphragms are in the same plane.

For cases where the frequency response of the units used requires a rolloff steeper than 12 db per octave, the networks of Fig. 302-e and Fig. 302-f are recommended. These give a rolloff of 18 db per octave, and a constant phase difference of 270 degrees. This means that mounting the diaphragms a quarter-wavelength apart at the crossover frequency will give in-phase outputs by appropriate connection.

All these networks are designed to present a constant, resistive impedance to the amplifier over the entire frequency range. Read Chapter 4 for information on how to wind the coils.

The impedance varies

One more point is often overlooked: The networks are designed on the theory that they are feeding resistance loads of the same value as the nominal voice-coil impedance. The voice-coil impedance is *not* pure resistance, so the performance of the networks is altered. The most serious effect is usually due to the inductance of the low-frequency unit's voice coil. By using networks 302-a, -d, -e, each of which feeds the low-frequency unit through a series inductance, this effect can be overcome by subtracting the voice-coil inductance value from the network inductance value derived from the chart. Even if the available data is insufficient to allow this, these networks will minimize the effect, because the inductance of the voice coil will add very little to the effective inductance of the network. In the other networks the shunt capacitor combined with the voice coil inductance will cause a greater variation in input impedance.

Each diagram in Fig. 302 has the inductors and capacitors marked with symbols. These identify the reference line (in the bottom part of the design chart) to be used for finding each component's value. Fig. 303 illustrates the use of the chart to find values for a network of the type of Fig. 302-e, and Fig. 304 shows the actual circuit calculated in this way for a crossover frequency of 500 cycles, at 40 ohms impedance.

The input impedance is the same as each speaker voice-coil impedance. Some prefer to design the crossover network for 500-ohms impedance and use matching transformers at the outputs to feed the individual voice coils. This method has two advantages: The two units need not have the same voice-coil impedance; and smaller capacitors can be used. The range of impedances covered by the chart extends up to 500 ohms (see bottom line of chart) to include such designs.

One modern trend has been to use separate amplifier channels for each unit. In this case the chart can be used for designing an interstage filter to separate the channels, by making the following adjustments: multiply all impedance values by 1,000 (this means the impedance used to terminate each output as a grid shunt); change inductance

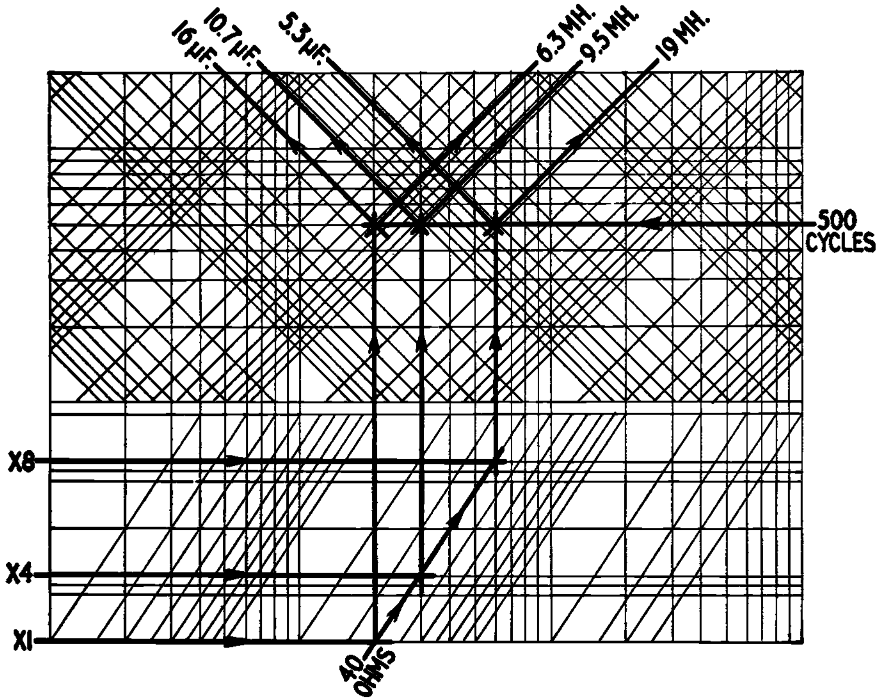
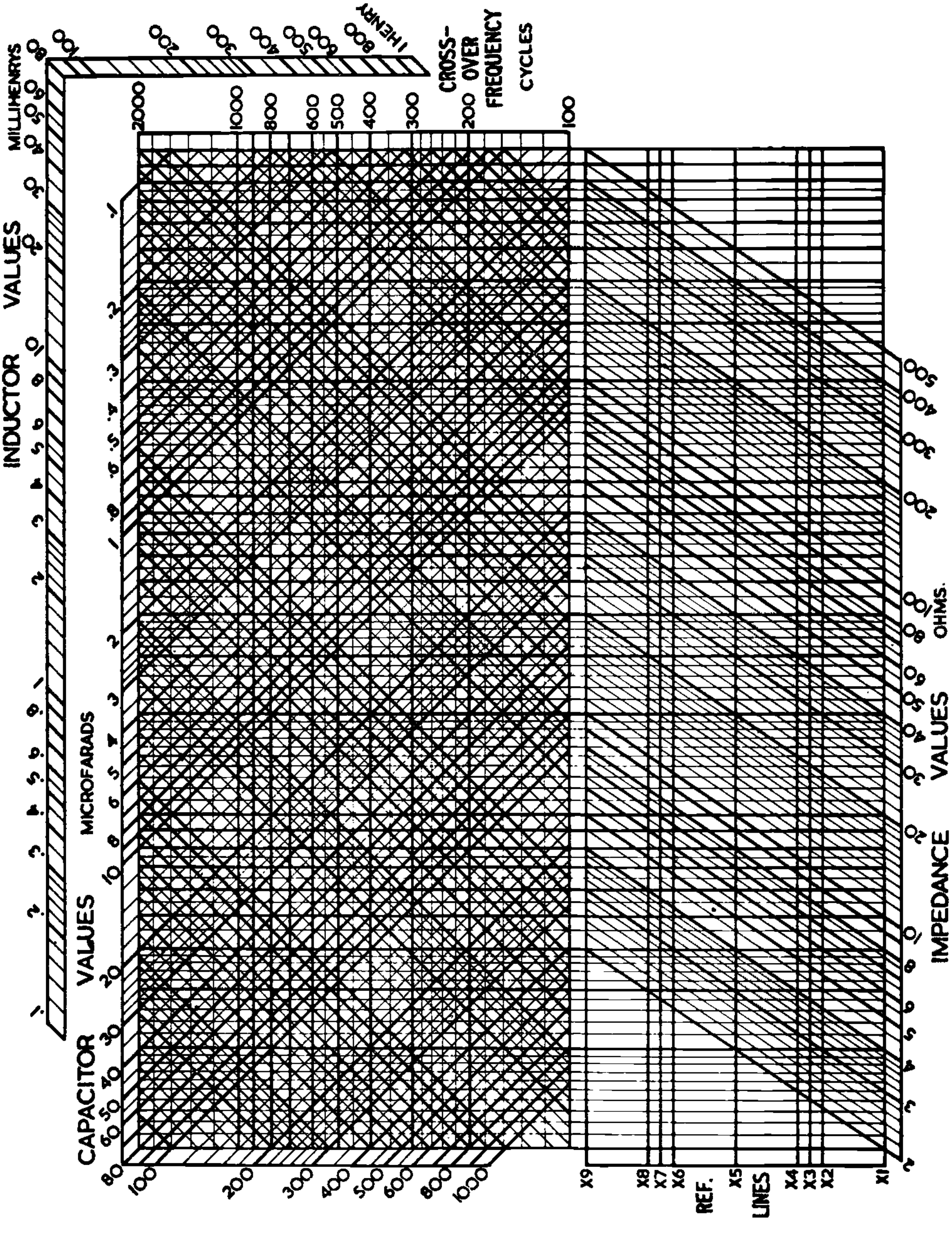


Fig. 303. How the chart on page 26 is used to find inductance and capacitance values for Fig. 302-e. Each speaker has a 40-ohm voice-coil impedance.

values to *henries* instead of millihenries; divide capacitor values by 1,000. Suitable networks for this application are a, c or e, in Fig. 302, since these allow the input to each amplifier circuit to be grounded on one side.

Use of the graph saves a considerable amount of pencil pushing and arithmetic. However, if you have an aversion to charts, Chapter 4 supplies formulas so that you can calculate the values of inductance and capacitance needed for your particular crossover networks.

Crossover Network Design Chart



The characteristics of the final output from any type of network and combination of speakers can only be predicted accurately if the source (amplifier) impedance is known. This can be measured with simple equipment by the methods described in Chapter 8, page 58.

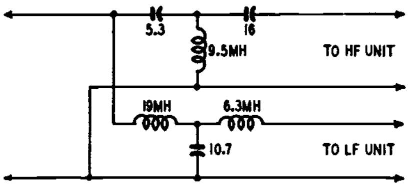


Fig. 304. Crossover-network component values for the circuit of Fig. 302-e derived by the method shown in Fig. 303.

The amplifier impedance should be as low as possible, and practically constant over the entire frequency range to be reproduced. With present-day components, this can be achieved only by using multiple low-impedance output triodes, or by carefully-designed inverse feedback circuits.

Chapter 4

Loudspeaker Network Inductors

THE crossover network design data in Chapter 3 specifies the inductance of the coils in millihenries. Too often, it is hard to find information about how many turns of wire to wind on a form of given dimensions to obtain a required value of inductance.

Fortunately, there are approximate formulas which can be used in designing network inductors. The results obtained will be close enough for practical purposes if care is used in their application.

Many variables are involved in winding inductors of any sort. For instance, the number of turns that any multilayer form will hold depends not only upon the diameter and insulation of the wire used, but also upon the method by which the wire is applied to the winding space provided. Obviously, if the wire is crooked or wound loosely about the form, valuable winding space will be lost. Each turn must lie on top of the preceding turn if the maximum number of turns is to be wound in the space provided.

One good method of winding wire on the coil form is to clamp a hand drill in a vise, then place a bolt or machine screw through the core of the form and fasten it securely with a lockwasher and a nut. For large or rectangular cores, use a wooden form cut to fit the core and drilled for the bolt. The length of the bolt should be such that when the bolt is clamped in the drill chuck, the form will be as close to the chuck as possible. Such an arrangement leaves one hand free to keep the wire taut while the other is occupied turning the drill crank.

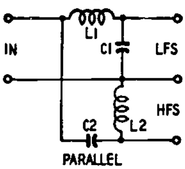
Four circuits are shown in Fig. 401. Those in 401-a and 401-b use the conventional m-derived network; those in 401-c and 401-d have the constant-resistance type of network. Fig. 401-d is the same as Fig. 302-d in Chapter 3. The necessary formulas for calculating values of inductance and capacitance in either case are given.

Since the purpose of this chapter is to supply adequate information for winding the inductors to be used in the networks of Fig. 302 and Fig. 401, we will leave it up to the reader to determine the value of capacitance or inductance required by simply using the formulas shown.

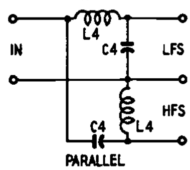
A typical example

Suppose the type of network chosen is the constant-resistance series circuit of Fig. 401-d. Let us also suppose that the forms on which the

a



b



$$L1 = \frac{254 \times R_o}{f_c} \text{ mh} \qquad C1 = \frac{318{,}000}{f_c \times R_o} \ \mu f$$

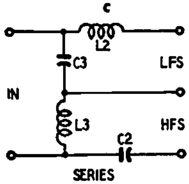
$$L2 = \frac{159 \times R_o}{f_c} \text{ mh} \qquad C2 = \frac{9{,}940}{f_c \times R_o} \ \mu f$$

$$L3 = \frac{99.4 \times R_o}{f_c} \text{ mh} \qquad C3 = \frac{254{,}000}{f_c \times R_o} \ \mu f$$

$$L4 = \frac{224.8 \times R_o}{f_c} \text{ mh} \qquad C5 = \frac{112{,}400}{f_c \times R_o} \ \mu f$$

$$L5 = \frac{112.4 \times R_o}{f_c} \text{ mh} \qquad C4 = \frac{224{,}800}{f_c \times R_o} \ \mu f$$

c



d

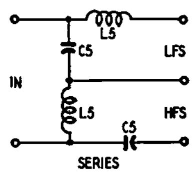


Fig. 401. *Four types of dividing networks and formulas for computing their values.*

inductances are to be made have a measured diameter of 1⅜ inches, and a winding length L of 1 inch is available. The winding depth, C, is equal to half the diameter, D, as shown in Fig. 402-a.

The value of N (number of turns) for a diameter, D, of 1 inch is

TABLE I

Values of N for D = 1", C = D/2, L = 1", R_o = 4 ohms									
Circuits	Type		f_c =	400	500	600	700	800	900	1000
Constant Z	Par.	(L4)		303	270	247	229	214	202	192
	Series	(L5)		214	191	175	162	151	143	136
Conventional m-derived	Par.	(L1)		323	289	264	244	228	215	205
		(L2)		255	228	208	193	181	171	162
	Series	(L3)		202	181	165	153	143	135	128
For R_0 = 8 ohms, multiply all values by $\sqrt{2}$. For R_0 = 16 ohms, multiply all values by 2.										

given in Table I. All values of N in Table I are based on a diameter, D, of 1 inch, a winding depth, C, equal to half the diameter of the form, a winding length (L) of 1 inch, and a voice-coil, R_0, impedance of 4 ohms. Note that in Table I, if R_0 is 8 ohms, you should multiply all values by 1.414 (square root of 2). Similarly, if R_0 is 16 ohms, multiply all values by 2. If, for example, the desired crossover frequency is to be 800 cycles, the value of N (in Table I) is equal to 151 turns. In other words, coils marked L5 in Fig. 401-d should have 151 turns. This value, however, is for a D of 1 inch, and for other diameters a correction factor must be applied to this value of N. These factors

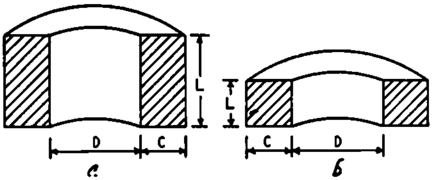


Fig. 402. Coil a is used in Table I. Coil b represents square cross-section coils of Table II.

are given in Table III. The factor for 1⅜-inch-diameter form is 0.79. Hence, all values of N (Table I) should be multiplied by 0.79 if a 1⅜-inch form is used to wind the coils.

In the example: $151 \times 0.79 = 119$ turns.

If the voice-coil impedance of the woofer is 8 ohms instead of 4 ohms, the value of N (119) must be multiplied by $\sqrt{2}$ or $119 \times \sqrt{2} =$ 169 turns.

This value of N is the number of turns needed on the form for a crossover frequency of 800 cycles and an impedance, R_0, of 8 ohms for the loudspeaker voice coil.

TABLE II

Values of N for a square winding area—$D = 1''$, $C = L = D/2$, $R_0 = 4$ ohms

Circuits	Type		f_c =	400	500	600	700	800	900	1000
Constant Z	Par.	(L4)		267	239	225	202	190	178	170
	Series	(L5)		189	169	159	143	134	126	120
Conventional m-derived	Par.	(L1)		284	254	239	214	201	189	180
		(L2)		225	201	184	170	159	150	143
	Series	(L3)		178	159	145	135	126	119	113

For $R_0 = 8$ ohms, multiply all values by $\sqrt{2}$.
For $R_0 = 16$ ohms, multiply all values by 2.

The next step is to determine the size and type of wire to use in winding the inductances. Since, in this example, we are using a coil form having a diameter of 1⅜ inches (1.375 inches), the available winding space is equal to the winding length, L, times one-half the diameter of the form: $L \times 0.5 \times 1.375 = 0.688$ sq. in.

A size of wire which will wind 169 turns in 0.688 sq. in. or 246 turns per square inch must be used ($169/.688 = 246$ turns / in.^2). The number of turns per square inch for various sizes and types of wire is given in Table IV. From this table, No. 16 d.c.c. will wind 271 turns per square inch. However, in the winding space of 0.688 square inch, $271 \times 0.688 = 187$ turns of No. 16 d.c.c. can be wound. This is 18 turns more than the required number and will cause the crossover

TABLE III

Correction factors for various values of D							
Diameter =	$1\frac{1}{4}$	$1\frac{3}{8}$	$1\frac{1}{2}$	$1\frac{5}{8}$	$1\frac{3}{4}$	$1\frac{7}{8}$	2.0
Table I	0.35	0.79	0.74	0.70	0.67	0.63	0.60
Table II	0.90	0.85	0.82	0.79	0.76	0.73	0.71

frequency to be about 140 cycles lower than 800 cycles. Adding one-half of the difference between 169 turns and 187 turns to the required number of turns (169) will raise the crossover frequency to approximately 800 cycles. The foregoing statement will not be true in all cases, but can be used as a means of minimizing the changes in the coils at the final testing of the network.

TABLE IV

Turns per square inch for various sizes and types of wire			
Wire Size	D.C.C.	S.C.C.	Enameled
16	271	321	348
17	329	397	437
18	399	493	548

If the winding area is to be kept square in cross section, the values of N for various crossover frequencies are as listed in Table II. Correction factors for various diameters to be used wih Table II can be found in Table III, and in all cases these factors should be multiplied by the values of N listed in Table I and Table II.

The values of N in Tables I and II were obtained by using Wheeler's and Bunet's formulas for multilayer inductances as given on pages 148 and 149 (third edition) or pages 443 and 444 (fourth edition) of the *Radiotron Designer's Handbook.*

Chapter 5

Multiple Speaker Installation

WHEN only one loudspeaker is used with an amplifier, correct connection is relatively simple. The speaker is connected to the amplifier output tap whose impedance, as marked, most nearly approaches the speaker impedance. Even this simple operation, however, may be beset with problems if the speaker is located far from the amplifier.

When a number of loudspeakers are operated from an amplifier, as in a large sound system, and the power fed to each speaker is different, the fun starts! This chapter is intended to present a few simple rules to help the sound technician do a properly engineered job with preassured results.

The term "transmission line" brings to mind high mathematics, nepers, and surge impedance, especially if associated with television. In audio, things are simpler. An audio transmission line is just a pair of wires that connect one audio device to another, as, for example, an amplifier to its loudspeakers. Surge impedance, propagation constant, etc., don't bother us until the line gets pretty long, say a good fraction of a wavelength. At audio frequencies (15,000 cycles) a 1/4-wavelength line is some 3 *miles* long, and the length increases as the frequency gets lower.

Two things plague the audio man: conductor resistance and shunt capacitance. The former causes power losses (heat) and the latter limits high-frequency performance. Their effects depend on the line impedance.

The impedance of a line is the nominal value (magnitude) of the impedance across the receiving end of that line. The definition is so stated because in high-quality audio amplifiers, the source impedance, looking back into the amplifier, may be but a fraction of the load impedance. Thus the *same* pair of wires can be a 4-ohm line, a 16-ohm line, or a 500-ohm line, depending on the load.

Selecting a speaker line

All conductors have resistance. Accordingly, when current flows in the line, there is a loss of power, the energy being converted into heat. Resistance losses—or copper losses as they are sometimes called—may be minimized by using larger diameter conductors and by making the line as short as possible. Copper losses are most important on low impedance (16 ohms or less) lines since the lower the impedance, the

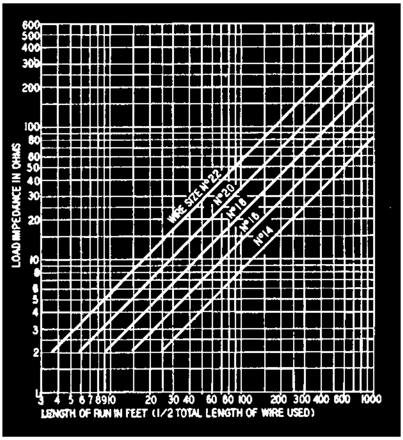


Fig. 501. Chart of maximum line length for given wire size and impedance.

more current flows for a given power, and there is greater I^2R power loss as heat.

Commercial sound engineers usually like to use the smallest possible wire size because the smaller sizes are less expensive and are easier to install and conceal in existing buildings. The smaller the wire size, the greater the resistance per foot and the greater the power loss, so a compromise must be made. Sound engineers allow 0.25-db loss (about 5% power loss) in the wiring to speakers.

To eliminate calculations, Fig. 501 has been prepared to show the maximum length of line which can be used for various line impedances and wire sizes. The chart is simple to use. For example, a single 16-ohm loudspeaker can be connected to an amplifier with No. 18 wire and can be *any* distance up to 75 feet from the amplifier. For No. 14 gauge, the distance may be up to 190 feet.

Three conditions may occur when using the chart:

1. If the point of intersection of the line length and load impedance falls *above* the line labeled No. 22, No. 22 wire or any *larger* size may be used. Smaller wires are not recommended.

2. If the point falls *between* two curves for wire size, the *larger* size should be used.

3. If the point falls *below* the curve for No. 14 wire, the load impedance is unreasonably low for the length of line required and matching transformers to make the line impedance higher must be used. This point will be considered a little later.

Multiple speaker installations

Installing a single loudspeaker involves no problems since its impedance may be used directly in finding the wire size from the chart. When two or more loudspeakers are used, there are two methods of connection: series and parallel. The series connection is often attractive but is not recommended except as an emergency measure. There are two reasons for this: the series connection is unreliable because if one speaker fails the entire series group is silenced. At some frequency a loudspeaker goes through cone resonance and its impedance becomes much greater than its nominal impedance value. Where several speakers are used, especially several makes or several sizes, their resonant frequencies may be different. When one speaker in a series string goes through resonance, there is a change in the sound volume from the other speakers. This effect can be very annoying, and is almost entirely absent in parallel-connected speakers.

When two or more speakers are connected in parallel, their impedances are combined in the same way as parallel resistances. The *equivalent* impedance of the total speaker system is used in the line chart, Fig. 501, in the same way as that of one speaker. The group is connected to the amplifier tap which corresponds to the *equivalent* impedance of the system. If an exact tap is not available, connect to the next *lower* amplifier output tap.

Assume that we have three speakers, each with a voice coil impedance of 16 ohms, and that the group is to be 30 feet from the amplifier. The equivalent impedance of the speaker load is 16/3 or 5⅓ ohms. On the chart, Fig. 501, the point of intersection of 5 ohms and 30 feet falls between the curves for No. 18 and No. 16 wire. Thus No. 16 wire is the smallest wire that can be used.

The speaker line should be connected to a 5-ohm tap, but most amplifiers do not have a 5-ohm tap, so the line should be connected to the 4-ohm tap, a common value. The reason for using a lower tap is that an output transformer reflects to its primary an impedance which is the load for the output stage. Connecting a slightly higher load to a tap, as in this case, causes the reflected impedance to be higher in the same proportion, which is better for an amplifier than having a slightly lower load.

When several speakers have their voice coils in parallel, the impedance of each voice coil should be the same. Any parallel system is a *constant-voltage* system and the voltage across each speaker is the same. Each speaker will draw the same power, and, assuming equal efficiency, will deliver the same sound level.

If several values of impedance are used in parallel, as two speakers of 8 ohms and one of 4 ohms, each 8-ohm speaker draws only half the power of the 4-ohm speaker. This unequal power distribution may be useful at times to provide different power levels in speakers. However,

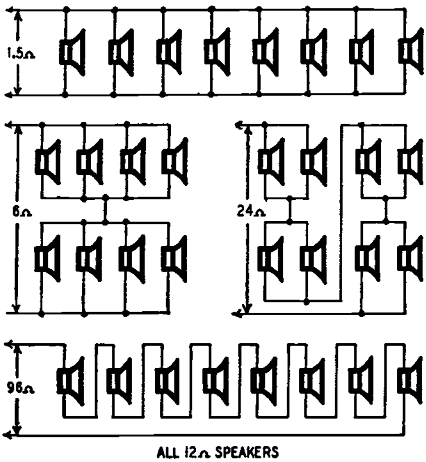


Fig. 502. Four ways of connecting eight 12-ohm speakers to receive equal power.

unequal amounts of power distribution are best distributed with matching transformers and higher impedance lines. Parallel combinations often give nasty combinations of impedances; the line impedance gets below 4 ohms and that makes for short runs or big wire or both. The best method is to use equal impedance and power distribution for these voice coil lines, and keep the line impedance 4 ohms or higher.

Voice-coil impedances range from 2 to about 16 ohms. (Sometimes values outside this range are met, but they are not common). Let us take 12 ohms for some examples because it makes calculating easier. Suppose eight such speakers are connected, all to receive the same power. The arrangements shown in Fig. 502 give over-all impedances of 1.5, 6, 24, and 96 ohms. It is probable that even a multi-ratio output transformer will not provide most of these matchings, so some other arrangement must be figured out, including one or more resistors, as "dummies" to pad out the values. And the number of speakers in an actual installation may not be so convenient for series-parallel connection as our example of eight.

Matching transformers

In discussing the speaker line chart, Fig. 501, it was pointed out that in some cases the length of line is great for the load impedance, and the line loss is greater than 0.25 db unless very heavy conductors are used. For such cases we borrow a trick from the power engineers who deal with power transmissions over long distances. The power boys get around the problem of line loss by stepping up the voltage at the sending end and stepping it down at the receiving end.

Audio engineers use a similar stunt with an output transformer at the amplifier that has a different turns ratio (gives higher voltage) than those ordinarily used with voice coils. Sometimes both voice coil and line impedances are present on the output taps. The usual line impedance is 500 ohms although 250 and 125 ohms are used. Broadcast engineers usually use 600 ohms as a line impedance because telephone lines (used for low levels only) have about 600 ohms characteristic impedance.

Impedances greater than 500 ohms are not used because shunt capacitance in long lines causes rolloff in response at the higher audio frequencies. When using a 500-ohm line, a matching transformer is used at each speaker (or group of speakers) to step down the line voltage to values suitable for voice coils. The speaker matching transformer reflects the voice coil impedance to 500 ohms (or other value depending on the line impedance).

This fundamental axiom is often overlooked, but should be stated for the sake of completeness. The rated power of the speaker or speakers should be at least equal to the rated power output of the amplifier; and the rated power of the amplifier should be equal to the total power desired in the speakers. The first proposition assures that the speakers will not be overloaded, and the second that they will provide adequate sound. Not all amplifiers will put out their rated power even at midrange frequencies, but following this simple rule and using good amplifiers will assure success.

In laying out a 500-ohm distribution system one rule covers all cases of equal or unequal power distribution:

1. Determine the power to be fed to each speaker or group of speakers.

2. Select an amplifier capable of this power, and multiply the amplifier power by the desired line impedance.

3. Divide the number obtained in step 2 by the power desired in the first speaker or group of speakers. This quotient is the primary impedance of the matching transformer for the speaker (s).

4. Repeat step 3 for each speaker or group of speakers.

Some typical cases

The use of the rule is best shown by examples of several types of matching situations.

Fig. 503 shows a single speaker fed by a 15-watt amplifier through

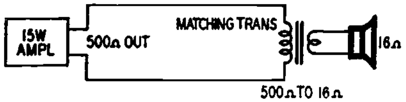


Fig. 503. One speaker on a 500-ohm line.

a 500-ohm line. The power capacity of the speaker is 30 watts and its nominal impedance is 16 ohms. Here the limiting factor is amplifier power and we proceed to step 2: 15 × 500 is 7500. Dividing by 15 (as all the power goes to one speaker) gives us 500 ohms as the correct primary impedance for the matching transformer. The secondary impedance is 16 ohms.

Had there been four speakers to be fed from one transformer, the primary impedance would have been the same, but the secondary would be set for four ohms. In this case the total power for the group is still 15 watts (used in step 3) but the power per speaker is 15/4 or 3.75 watts. The case is shown in Fig. 504.

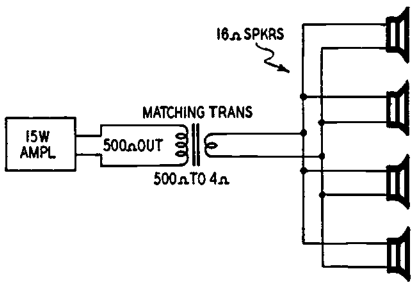


Fig. 504. Power per speaker is 3.75 watts.

Fig. 505 shows four speakers strung along a 500-ohm line. If there is considerable separation between speakers, it is best to use a separate matching transformer at each speaker. Let us say that we have a 20-watt amplifier and the desired power per speaker is 5 watts. The rated power of each speaker being 10 watts, we are safe. In step 2: 20 $\times$ 500 is 10,000. In step 3: we divide 10,000 by 5, giving us 2,000 ohms as the primary impedance for each transformer. Since each speaker has the same power, we need calculate but once. The secondary of each matching transformer is set to the impedance of each voice coil which need not be the same for all speakers.

Let us proceed to a problem shown in Fig. 506 where we have unequal power distribution. A 10-watt amplifier with 500-ohm output is

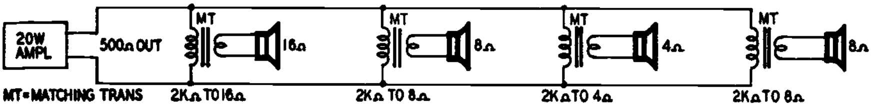


Fig. 505. *Each speaker needs a matching transformer if the lines are long.*

to feed three speakers. The first is to be fed 1 watt, the second 4 watts and the third 5 watts. For this type of circuit each speaker must be equipped with a matching transformer even though they are near each other. In step 2 we multiply 500 $\times$ 10 to get 5,000. Dividing by 1 in step 3 gives us 5,000 ohms for the primary impedance of the matching transformer for the first speaker. The secondary is set for the correct voice-coil impedance. For the second speaker, (step 4), we divide 5,000 by 4 to get 1,250 ohms for the matching transformer primary. Repeating for the third speaker we get 1000 ohms as the impedance for the third matching transformer. To check the calculations, we can combine the three impedances as parallel resistors. We get 500 ohms, which shows that we have a matched system.

The 500-ohm line is not a sacred cow. Lower impedances such as

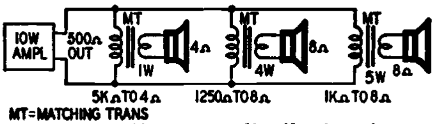


Fig. 506. *If power distribution is unequal, each speaker needs transformer.*

250 ohms will work equally well if larger wires are used for runs over 400 feet. This information is obtained from Fig. 501. A lower impedance sometimes may be required. For high power amplifiers with many speakers the required primary impedance of some of the matching transformers may be greater than 10,000 ohms. In this case the line impedance should be lowered to say 250 ohms, as high quality matching transformers are generally not made with primary impedances higher than 10,000 ohms.

If there are several speakers to be fed from a single matching transformer, the speakers need not be adjacent to each other. For example, one of the speakers of Fig. 504 could be up to 30 feet from the matching transformer, if it is connected with No. 16 gauge wire or larger. Fig. 501 should be consulted to find out if such a scheme (which reduces cost because wire is cheaper than good matching transformers) is practicable.

The maximum recommended length for speaker lines, even at 500 ohms, is 1,000 feet. One reason is that the shunt capacitance in a 1,000-foot line may build up to the point where it is a severe restriction on quality. In high-fidelity work loss of high frequencies is disastrous. In public address work loss of high frequencies causes a reduction in intelligibility and a decrease in the usefulness of the system.

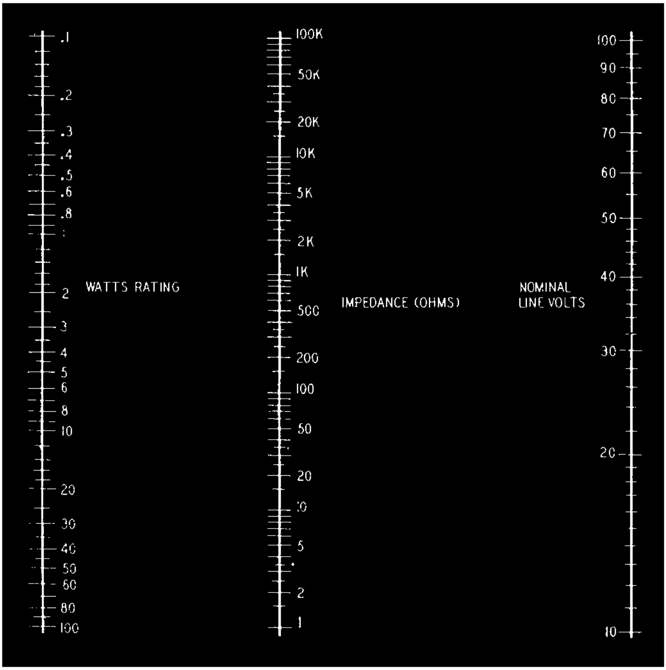


Fig. 507. Constant-voltage chart, showing impedance vs. wattage.

Another reason is power loss. If kept to 0.25 db, that is a 10% power loss. In a 50-watt system it means 5 watts lost. The cost per watt in amplifiers is high, so losses mean money. If a speaker line must be run much over 1,000 feet, it is better to put a booster power amplifier nearer the speakers and feed it on a 500-ohm line at lower power level where a 10% loss means fewer watts. Such a scheme permits equalization to counteract loss of high frequencies. Speaker lines *can* be run over 1,000 feet if the limitations are known and efforts are made to minimize their effects.

The *constant-voltage system* is usually designated by the nominal voltage used. The American standard is a *70-volt line;* the European, a *100-volt line.* The constant-voltage system has one apparent disadvantage—a separate output transformer is required for each speaker. The circuits in Fig. 505 and Fig. 506 are constant-voltage systems. The disadvantage in using multiple output transformers is more apparent than real. True, when an amplifier is working only one speaker, located close to it, there is no point in using more than one transformer for matching the output to the speaker. Doing the matching in two steps adds to cost by one transformer, and also adds slightly to the audio losses. But when the amplifier is feeding a number of speakers, matching problems may well make separate transformers worth while.

The constant-voltage line has another advantage: It is always best to operate dynamic type speakers in parallel, otherwise electrical damping is lost, and peculiar effects due to interaction between speaker impedances are noticed. In constant-voltage operation, all units are always parallel-connected, even when the power delivered to different units is varied. If numbers of speakers are operated in parallel by direct connection, the resulting impedance is so low that much of the output power is lost in connecting lines unless very large cable is used. In the constant-voltage system, impedances can be kept up to a reasonable figure.

A nominal line voltage is chosen, usually 70 or 100 volts. This forms the basis of all the calculations. This does not mean that there is *always* a signal of 70 or 100 volts, because it naturally fluctuates, as audio signal always does. The stated voltage represents a nominal *maximum output level.* Perhaps the easiest way to get the idea is to think in terms of a sine-wave signal, fully loading the amplifier. The amplifier then can be regarded as providing a constant voltage for all the speakers connected to the line, just as an electric line does for all the appliances connected to it. The generator at the power station has a certain maximum load capacity, and consumers' loads may be connected until that capacity is reached, the power taken by each depending on its load impedance and the line voltage. We are quite used to referring to electric lamps and other appliances as, "110-volt, 40-watt," but the same method of rating speakers may seem strange at first.

As we have seen, each speaker is fitted with a transformer to match its voice-coil impedance up to an impedance which accepts the desired wattage when the nominal voltage is applied. Some speakers may be fitted with multi-ratio transformers so their power rating can be adjusted. This makes an installation very versatile, and avoids the loss of power caused when an individual volume control is used on each speaker. You simply vary the number of watts accepted by the speaker. Different voice-coil impedances are also taken care of by the speaker-matching transformer. You can use the technique previously described for multiple speaker connections or the charts in Fig. 507 and Fig. 508.

Fig. 509 is another sample problem. The output transformers for speaker 1 and speaker 2 each have an impedance of 5,000 ohms, so they will accept 2 watts each at 100 volts. (Use the chart of Fig. 507.) Speaker 3 has a voice-coil impedance of 2 ohms, and uses a transformer of ratio 70 to 1. From the chart of Fig. 508 this gives an impedance of almost 10,000 ohms, which from the other chart (Fig. 507) rates at 1 watt for 100 volts. The total wattage load is $2 + 2 + 1 = 5$. Using the

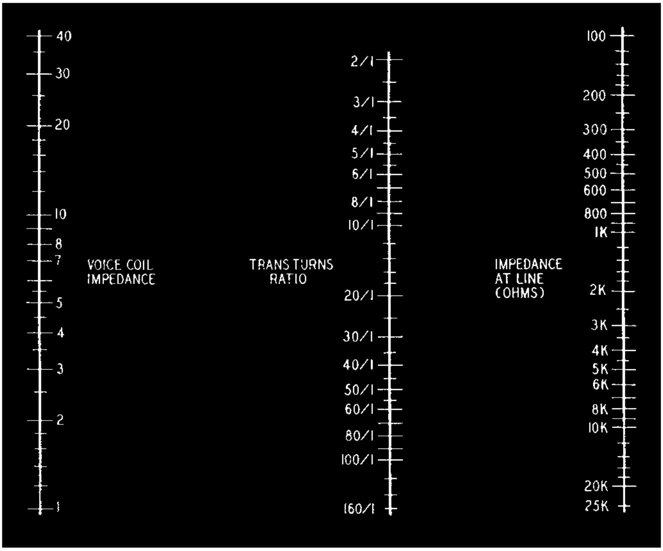


Fig. 508. A chart for obtaining line impedance from turns ratio.

chart in Fig. 507 again, this corresponds to an impedance of 2,000 ohms (still for 100 volts). So an amplifier to supply just this load would need to supply 5 watts matched into 2,000 ohms.

A large amplifier may be used to supply a load smaller than its own output. For example, suppose a 60-watt amplifier is used to feed the foregoing case requiring only 5 watts. The nominal voltage is used to calculate both speaker and amplifier output impedances. A 60-watt output for 100-volt operation should be matched into 170 ohms. The load actually connected is 2,000 ohms. Some amplifiers working into a light load like this will be unstable. To prevent this, a resistance load may be added to absorb the surplus power. In the example suggested, a resistance load to absorb 50 watts would be adequate, and from the chart the value required is 200 ohms. If the amplifier were to be operated continuously at maximum output, this resistor should have a dissipation rating of 50 watts, but in practice a much smaller (10- or 20-watt) resistor could be used.

Sometimes the reverse of the previous problem arises. The nominal load connected *exceeds* the power output of the amplifier. Here matters are adjusted by a different method. Suppose the load is made up of a number of speakers rated at 2 and 5 watts for 100-volt line, adding up to a total load of 80 watts. The load impedance of 80 watts worth of speakers will be 125 ohms. The load for a 20-watt amplifier, 100-volt working, would be 500 ohms. Applying a 125-ohm load to the output of an amplifier designed for 500 ohms would probably cut the output down to about 5 watts, and as well likely cause distortion. So the output must be matched to the actual speaker load of 125 ohms, which, according to the chart of Fig. 507, will give 20 watts at 50 volts, instead of the original basis of calculation, 100 volts. This means the nominal 2-watt speakers, of 5,000 ohms impedance, will get 1/2 watt, and the nominal 5-watt speakers, of 2,000 ohms impedance, will get 1-1/4 watts. Note that this is a reduction of only 6 db, so quite a useful volume will be available, although the amplifier is smaller than one planned for 100 volts. Anyway, if the four-to-one mismatch were used, giving only 5 watts or so, there would be loss of another 6 db, and probably considerable distortion.

Use of the constant-voltage system does not necessarily mean special speaker transformers must be used, so a word about picking suitable transformers from stock lines is needed. Makers of speaker transformers mark them variously in turns ratio or impedance ratio. In the former case the chart of Fig. 508 enables the correct *turns ratio* to be found, but the *actual turns* on each winding must suit the job too. A mike-to-line transformer for a ribbon microphone may have the same ratio of turns as a speaker transformer, but this does not mean that either would do the other's job successfully. A good rule for checking the suitability of speaker transformers with an ordinary ohmmeter is that the winding resistance should be between 2% and 20%, of the impedance for which it is to be used. Less than 2% means its inductance will most likely be inadequate, and 20% or more means the windings will absorb an appreciable portion of the available audio power. If the resistance of the voice-coil winding is too low to register on the ohmmeter scale, the resistance of the high side should be compared with its working impedance. Thus, for example, a winding intended to work at 5,000 ohms should have a resistance that lies somewhere between 100 and 1,000 ohms.

Where transformers are specified by impedance ratio—for example, 7,000 ohms to 3.5 ohms—the maker has stated the *best* impedance at which to work the transformer. Using these impedances, it may be expected to be well over 80% efficient (probably over 90%), and have a good response at low frequencies. But these are not the only impedances at which the transformer can work. The important thing, of

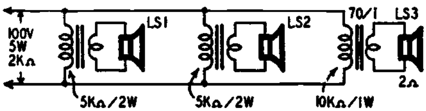


Fig. 509. *A constant-voltage line, with the speakers receiving unequal power.*

course, is that that *ratio* of impedances hold true, so the same transformer could be used for 4000 to 2, 10,000 to 5 ohms, etc. As stated in the previous paragraph, the losses and response must be kept within bounds. If impedances more than two or three times the rated values are used, the transformer's inductance may prove inadequate. If it is used with impedances less than one-third to one-half the rated values it will become quite inefficient.

Chapter 6

Voltage Regulators for Hi-Fi Amplifiers

VOLTAGE-REGULATED power supplies find considerable application for stabilizing high-frequency oscillators in receivers and transmitters. Their purpose is to minimize frequency drift if the line voltage varies. Less well known, but of even greater value, is the use of voltage regulators, especially gas-tube types, as hum and decoupling filters in high-fidelity audio amplifiers. They provide a highly effective and low-cost method of improving the hum level and transient response, and can transform an excellent amplifier into a superlative one.

A voltage regulator is the equivalent of a highly effective low-pass filter consisting of a series impedance A and a shunt impedance B (Fig. 601). We know from applications of Kirchoff's laws that when a current is introduced at a junction of several branches it will divide among the branches in inverse proportion to their resistances or (in the case of a.c.) their impedances. In the low-pass filter, part of the current will go through series element A and part through shunt element B. If the combined resistance or impedance of the series element and the load is 10 times greater than the impedance of the shunt element, 90% of the current will flow through B and only 10% will flow through A and the load.

In a power supply we are dealing with both d.c. and a.c. The d.c., of course, is what we want to apply to the load, while the a.c. is the ripple we want to suppress or bypass. We need a series circuit, which will offer a high impedance to the a.c. and a low resistance to the d.c., and a shunt circuit with high resistance to the d.c. and *a very low* impedance to the a.c. We achieve it by using either resistance or inductance for the series element and capacitance for the shunt element. The capacitor has very high resistance to d.c. and relatively low opposition to a.c., while the inductance has high impedance to a.c. and a low resistance to d.c. Thus we can bypass a good deal of the a.c. without reducing the d.c. too much. Unfortunately, it takes very bulky inductors and capacitors to do the job at hum frequencies (60 and 120 cycles). And when the hum level must be held to 60 db or more below the signal level—which means reducing the ripple to one-ten-thousandth or even one-hundred-thousandth of its value at the rectifier output, four or five filter sections are required.

This is where the VR tube can perform a valuable service. It is an ideal hum filter, and it is very much less costly and bulky than any

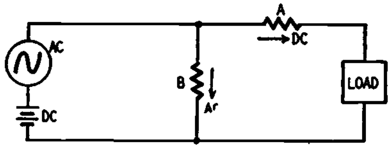


Fig. 601. *Functions of series-element "A" and shunt element "B" in low-pass filter.*

combination of capacitors and inductors or resistors capable of equal filtering action.

The use of VR tubes as hum filters combines perfectly with their use as decoupling filters.

Amplifier "definition"

The ideal amplifier delivers an exact counterpart of the input signal to the load. It neither adds to nor subtracts anything from the *form* of the signal. An amplifier may distort a signal by subtracting some of the tonal values through frequency discrimination, or by adding harmonic distortion and intermodulation. Another addition affects the *definition* of the amplifier.

An amplifier with good definition maintains the distinctness of the individual elements of the signal. Definition in an amplifier can be compared to resolution in a camera. A camera with good resolution will show individual blades of grass and individual hairs in a coiffure; a camera with poor resolution blurs the separate components so that they cannot be distinguished. Similarly, an amplifier with good definition reproduces the individual notes and instruments distinctly; one with poor definition will blur the individual tone elements and instruments until they cannot be distinguished separately.

Definition in an amplifier is very largely a function of transient response. An amplifier with good transient response is nonresonant, nonregenerative and nonoscillating. It is always merely a reproducer, never a generator. Unfortunately, this is much easier to stipulate than to achieve. Most amplifiers are regenerators or oscillators of the triggered type. At some point in the frequency range they are resonant, possess a feedback loop, and will break into momentary oscillation if triggered by a strong-enough impulse. These oscillations are not always audible as such in the output. They are manifested not as constant tones but as "hangover" effects, and consist of a series of echoes of the signal. They may even be pleasing to some because they give a resonant effect which sometimes simulates room resonance. But they reduce the definition of the amplifier, blurring the individual elements and instruments instead of maintaining their distinctness.

Transient oscillations

Fig. 602 shows what happens. The figure at *a* represents the wave-train of a damped (soft pedal) piano trill plotted on amplitude-time co-ordinates. There are definite valleys between the separate tones in the original signal. An oscillating or regenerative amplifier, however, will prolong each pulse with a series of spurious pulses—hangovers or echoes—which not only change the shape of the original pulse, but also fall into and partially fill the valleys between pulses. The result is indicated at *b*. The separation between pulses is no longer distinct

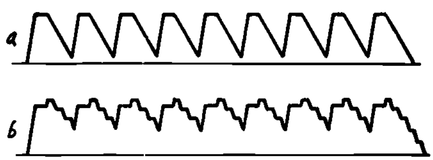


Fig. 602. (a) Original piano trill waveform. (b) Oscillations add "hangover" pulses to each trill cycle.

and sharp, and the total effect is blurred. In fact, the effect is very much like that of playing the same trill undamped—that is, with the loud pedal depressed.

Moreover, if the amplifier is resonant at any frequency, it can be triggered into oscillation at this point by any strong transient regardless of its frequency. In good amplifiers the resonant points will be at the inaudible extremes of the frequency range and the oscillations themselves will not be audible. Nevertheless the oscillations will have two disastrous effects: first, they create intermodulation distortion; second, they are usually of much higher instantaneous amplitude than the signal and may drive one or more stages of the amplifier into the nonlinear regions of their curves and produce violent distortion of every possible form, even at low signal levels.

To be a true reproducer, an amplifier must be as nearly nonresonant, nonregenerative and nonoscillating as possible. To achieve these qualities we make the amplifier frequency response flat away below and away above the audible range (from 10 to 100,000 cycles in good amplifiers). We also try to eliminate or reduce the bad effects of feedback loops in the amplifier.

Feedback loops are inescapable in any amplifier when several stages are fed from the same power supply. Fig. 603 is a block diagram of a

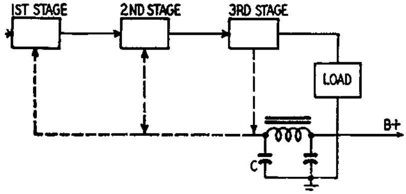


Fig. 603. Possible feedback loops in amplifier with one B supply and no decoupling.

three-stage audio amplifier with a common power supply and no decoupling filters. The dashed lines are the feedback loops created by the common plate-supply line. If capacitor C does not bypass *all* audio voltage on the line to ground, part of the audio voltage in the output stage will be fed back to the first and second stages. Regeneration will occur where the feedback is in phase with the stage signal, and oscillation will start if the amplifier has a resonant point.

With 3 resistance-coupled stages or 2 transformer-coupled stages the feedback will always be in phase over some portion of the frequency range. To minimize feedback we must insert low-pass filters in the feedback loops—filters which will pass the d.c. plate currents but will

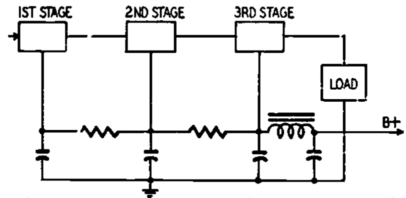


Fig. 604. R-C decoupling networks in three stage amplifier.

bypass most if not all of the signal frequencies. Typical R-C decoupling networks are shown in Fig 604. These will reduce feedback, at normal frequencies; but adequate bypassing below 50 cycles calls for enormous values of shunt capacitance. For this reason cheap amplifiers cut off the low-frequency response very sharply at about 100 cycles.

But as we have already noted, really high-fidelity amplifiers must be designed for a pass-band from 10 to 100,000 cycles. How are we going to decouple effectively at a frequency as low as 10 cycles? One way is to use two or more power supplies. Some expensive amplifiers actually do this. A much simpler and equally effective device, is to use one or more VR tubes at appropriate points in the power-supply loop.

VR decoupling

Any regulator which holds a d.c. voltage absolutely constant, is a *perfect* bypass for a.c. As a shunt unit a VR tube is effective down to subaudible frequencies. The installation of VR tubes for decoupling and hum filtering is simple and inexpensive. Only one VR tube is necessary in most amplifiers. The need for decoupling and hum filtering is progressive; we need much less attenuation of hum and feedback in the final stages of an amplifier than in the input stages. By combining brute-force inductance-capacitance filters with a single VR tube we can usually achieve almost complete hum and feedback suppression in amplifiers of five or even six stages (assuming that the final stage and the drivers are push-pull and that no interstage transformers are used).

The most effective point for a VR tube is the input stage, or the preamplifier for a low-level phono pickup. Fig. 605 shows how to use a VR tube with a G-E-type preamplifier. If a commercial amplifier is used, the 10,000-ohm resistor and VR tube can be inserted at the B plus-input point. If the preamp is home-built, the 16-µf filter capacitors and 33,000-ohm series resistors can be eliminated. In any case, the 150,000-ohm resistor in the frequency-correction network can be removed. This resistor limits the bass compensation at very low frequencies to guard against motorboating. With the much better decoupling provided by the VR tube, this limitation is no longer necessary, and removing the resistor extends the effective range of the bass boost circuit.

Application to any other input circuit is just as simple. The VR tube is simply inserted in the B plus line in place of the usual decoupling network. The stage in which the VR tube is connected should be designed for a plate supply of 150 volts. Most amplifiers designed for a 250-volt plate supply will work all right with a 150-volt supply. In any case, the improved decoupling extends the low-frequency limit and a larger cathode bypass capacitor (at least 25 µf), and a larger grid-coupling capacitor should be installed. In all cases the series resistor R should be adjusted so that a minimum current of 10 ma flows through the VR tube. VR tubes have a range of 6 to 30-40 ma. Adjust R so that the VR's are as close to the center of this range as you can get.

R can then be replaced with a fixed resistor. Two or more stages may be fed by one VR tube provided the total current does not exceed 30 or 40 ma.

The addition of a VR circuit to a well-designed audio amplifier with a very wide, flat frequency range will produce a marked improvement in transient response, especially at very low frequencies. Audio amplifiers using Class AB pentodes and beam power output tubes require a regulated screen voltage for maximum output. This can be supplied by one or more VR tubes operated in series. Use tubes whose total operating voltages equal or approximate the required screen operating voltage. The 0B3 operates at 90 volts, the 0B2 and 0C3 at 105 volts, and the 0A2 and 0D3 operate at 150 volts. The VR-75 (obsolete but often available on the surplus market) supplies 75 volts.

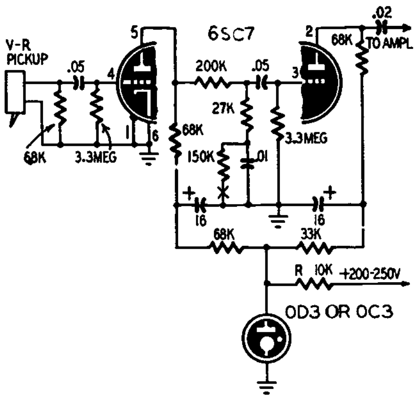


Fig. 605. Adding voltage regulation to a variable-reluctance pickup preamplifier.

You must observe some precautions when using VR tubes. They can oscillate if shunted by a leaky capacitor or an R-C network. The VR tube must light or it will not regulate. VR tubes are light sensitive and will have less voltage across them in a bright light than in the dark. This is not serious.

In a sense, your VR tube is like a bleeder. The amount of current drawn by the VR must be supplied by the power transformer and passed by the rectifier. Take this factor into consideration in your power supply design.

Chapter 7

Speaker Impedance Measurements

IN designing a good sound system, it is necessary to know the *true* voice-coil impedance with the actual load. The nominal impedance (usually that at 400 or 1,000 cycles) is insufficient. The true impedance at some resonant frequency may be as much as ten times the nominal value stated by the manufacturer. If both the impedance of the voice coil and the input to the speaker transformer can be measured over the entire frequency range, the voice coil can be checked for spurious resonances.

The method described in this chapter can also be adapted to measure other impedances, as those of recording cutters and magnetic recorders.

Variation of voice-coil impedance with frequency, together with

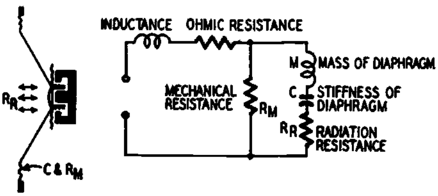


Fig. 701. Equivalent circuit of loudspeaker showing voice-coil components.

the d.c. resistance, gives a good indication of the efficiency of the speaker. A good speaker (including its baffle or flare) has an almost constant impedance. A cheap speaker has pronounced peaks—the impedance at the bass peak may be many times that at mid-frequencies.

Knowing the correct impedance of a speaker is especially important if the speaker is used in combination with one or more other speakers. Multiple speaker connection techniques are described in Chapter 5.

Several speakers in parallel might be used to supply sound to various locations; or several might be used at one location, each covering a different frequency range. In either case correct impedance matching is essential for best results.

What does speaker impedance mean? It is defined as the ratio of the voltage across the speaker voice coil to the current flowing through it (both measured in r.m.s. or both in peak values):

$$Z = \frac{E}{I}$$

What makes up impedance?

As shown in Fig. 701, the voice-coil impedance is made up of the ohmic or d.c. resistance of the wire, radiation resistance due to the dissipation of energy in the form of sound, mechanical resistance because the spider and cone rim are not perfectly flexible, and a number of reactances (which change with frequency). The inductive reactances

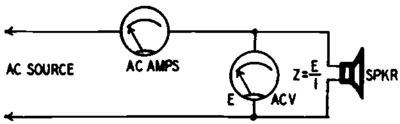


Fig. 702. Voltmeter-ammeter method of impedance checking.

—those which increase with frequency—include that due to the number of turns in the coil and the mass of the diaphragm. The capacitive reactances (inversely proportional to frequency) include those due to the stiffnesses of spider and cone rim.

There is also a reactance due to the elasticity of the air directly in front of the diaphragm.

Methods of measuring

Just as there are two main methods for measuring resistance—the Ohm's law method and the Wheatstone bridge—so there are two main methods for measuring impedance! In the Ohm's law method, using the formula $Z = E/I$, where Z = impedance in ohms, E = potential difference in volts, and I = current flowing in amperes, all we need to do is pass an alternating current of suitable frequency through the voice coil and make two measurements (see Fig. 702). In fact if we have an a.c. supply of constant voltage, we could calibrate an a.c. ammeter to read the impedance directly in ohms.

Bridge method

For very precise impedance measurement there is nothing superior to an a.c. bridge—balance being necessary for both the resistive and reactive components (Fig. 703). While the bridge method will yield

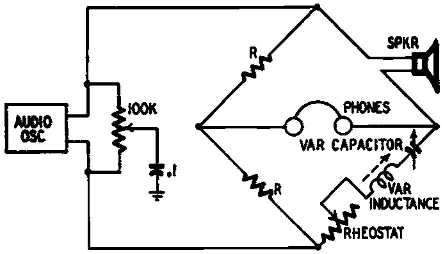


Fig. 703. Bridge for measuring high impedance speakers.

accurate results it is rather tedious, and a faster method is better for the service technician and amplifier designer.

Voltage comparison method

If the same current flows through two components in series, the voltage across those components will be proportional to their impedances. If $Z = E/I$, then Z is proportional to E if I is constant.

To measure the impedance of a loudspeaker (or other nonresistive load) the oscillator, speaker and a 10-watt resistor are connected as shown in Fig. 704. The resistor should have about twice the rated

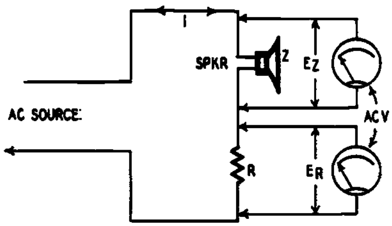


Fig. 704. Measuring speaker impedance with a voltmeter, resistor, and oscillator.

impedance of the speaker under test. The resistor value should be known accurately. The plan is to measure the voltage across the speaker and to compare it with the voltage across the resistor over the audio-frequency range. Since the resistor and speaker are in series the current through both is the same and the impedance of the speaker can be calculated from the following formula:

$$Z_S = R \frac{E_z}{E_R}$$

where Z_s is the speaker impedance, R is the resistor in ohms, E_z is the voltage across the speaker, and E_R is the voltage across the resistor. The oscillator is first set at 1,000 cycles and the amplifier output is adjusted so that conveniently measureable voltages are obtained across both the resistor and the speaker. It is a good point to have the sum of the two voltages less than the maximum of the meter scale in use so that you will not have to change the scale at any possible impedance.

If you have an insensitive voltmeter you may have to have the volume rather loud in order to get readings, but if the neighbors and amplifier can stand it the results will be just as good. Voltage readings are taken across the standard resistor and across the speaker. It is a good plan to start at 1,000 cycles and to sweep continuously down the audio spectrum to 20 or 30 cycles. Since some speakers may have several closely

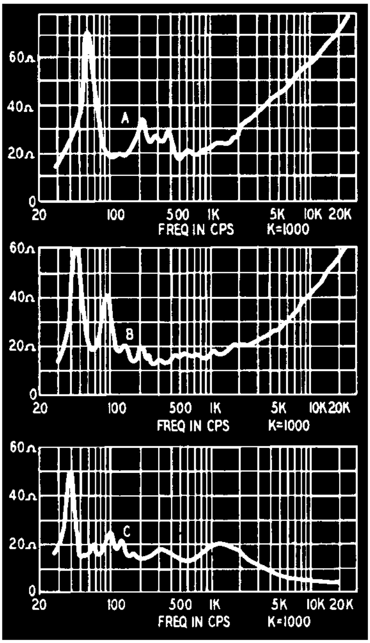


Fig. 705. *Typical speaker impedance curves.*

adjacent peaks, a number of measurements should be made in these ranges to get an accurate picture of the impedance curve. One method is to leave the voltmeter across the speaker after each measurement and to note the significant peaks and valleys as the oscillator is tuned up and down. Readings are made at these peak points along with enough in-between measurements to draw a good curve.

After covering the bass range, the spectrum from 1,000 cycles up should be checked. A smoother curve is usually found in this range, but the technique of continuously sweeping the oscillator up and down the scale will reveal any peaks which exist. The voltage readings are then converted to impedance values.

Fig. 705 presents some typical speaker-impedance curves. Curve A is that of a single-cone 15-inch speaker in an open-back cabinet. Curve B is the same 15-inch speaker in a 7-cubic-foot bass-reflex corner cabinet. Curve C is a two-speaker system with dividing network at 800 cycles. Both speakers are horn-loaded. It is immediately apparent that though all these speakers are rated at 16 ohms, such a rating is only nominal, and much higher impedances are actually present at many frequencies. The high-impedance peaks in the bass range are produced whenever there is a tendency for the voice coil to resonate, either because of resonances in the speaker itself or in combination with the air loading in the cabinet. These peaks in the bass range tell us a good deal about the speaker system. The fairly smooth rise of impedance at the higher frequencies is due to the inductance of the voice coil. This inductance is really too high in the treble range but is needed in the bass. In dual-voice-coil speakers or two-speaker systems this impedance rise can be eliminated by designing each driver for its particular response range.

The high peak in curve A occurs at the resonant frequency of the speaker cone. Mounting the speaker in an open-back cabinet has done little to damp this resonance and seems to have added a couple of new ones at higher frequencies. In the reflex cabinet the air loading raises the frequency at which the cone resonates but reduces the amount of resonance and adds a lower frequency resonance of the reflex cabinet. These effects can be seen in the impedance curves which are very helpful in adjusting reflex baffles. The horn-loaded speakers show a more uniform impedance curve down to the bottom peak. This peak is at the cutoff frequency of the horn and the resonant frequency of the low-frequency driver. The driver has been selected to resonate at this point to hold up the low-frequency response where the horn falls off.

It might well be asked why the speaker impedance rises at these resonant points. It is helpful to look at it this way. The a.c. voltage from the amplifier sends current through the voice coil so that it vibrates in the magnetic field of the speaker. This vibration of the coil in the magnetic field causes it to generate an a.c. voltage of opposite sign to the driving voltage. When a resonant frequency is reached the mass of the voice coil and cone just balances the compliance of the cone and the air chamber, and the coil vibrates back and forth much more vigorously. The voltage generated by the voice coil increases, opposing the driving voltage and reducing the current. Thus the impedance of the unit rises. This is a desirable counterbalance because it is important to reduce the power input at these resonant points to avoid a loudness peak. However, if the amplifier has a high internal impedance, as the speaker impedance rises, the voltage across the speaker will also rise, thus increasing the tendency to resonate, and creating a peak. See Chapter 8 for methods employed in calculating the internal impedance of an amplifier.

As an example of this condition, Fig. 706 shows the frequency response as measured across the voice coil of a 15-inch speaker in a 7-cubic-foot bass reflex corner baffle when driven by amplifiers of different damping factors. These curves were determined by varying the internal impedance of a high-quality amplifier.

In all these cases the amplifier produced a flat frequency curve into a resistor load. The frequency deviations shown are simply the effects of the variations in speaker impedance with the variations in both frequency and the damping factor of the amplifier. Note that with damping factors of 8 or higher the effect on the response of even large changes in speaker impedance is negligible. With a damping factor of

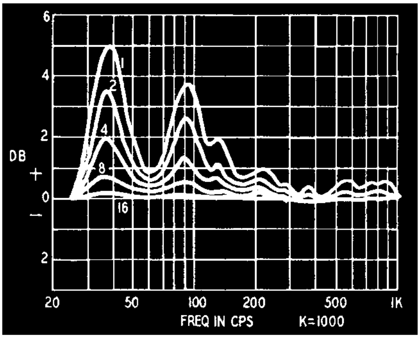


Fig. 706. Effect of amplifier damping factor on speaker response.

4 the rise in frequency response is just noticeable. With factors of 2 or 1 the bass peaks are pronounced and may be noticeably boomy.

Another important angle is the damping effect of the amplifier on the natural speaker system resonances when the system is subjected to sudden bursts of tone or transient impulses. These shock impulses tend to throw the speaker into vibration at its resonant points unless the speaker is *critically* damped. Part of this damping is provided in the construction of the speaker itself; part is supplied by the air loading of the cabinet or horn and by the internal impedance of the amplifier. Damping is not greatly affected when the amplifier internal impedance drops to less than 1/8 or 1/10 of the voice coil resistance.

You will be on safe ground if your amplifier has a damping factor of at least 3 over the entire range of audibility.

Internal impedance measurements are therefore important tests for amplifier constructors. When coupled with speaker impedance measurements the information gained can be used to improve the over-all audio performance considerably.

Using an oscilloscope

A more interesting method of determining voice-coil impedance uses an oscilloscope as a voltmeter and has the advantage that both voltages are indicated simultaneously on the screen. As shown in Fig. 707, both the voice coil and a calibrated variable resistor are connected in series and the voltages are applied (via the amplifiers of the scope) to the vertical and horizontal plates.

Here is the procedure: First replace the voice coil by a known resistance (say 5 ohms), and adjust the variable resistor to the same

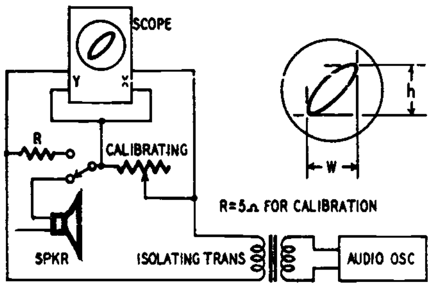


Fig. 707. How to use an oscilloscope to compare voltages and impedances.

value. Now adjust the sensitivity controls of the oscilloscope amplifiers to give a trace consisting of a line at a 45° slope on the screen. This adjustment is important. The horizontal width of the trace must be exactly equal to the vertical height as shown in Fig. 707. After this adjustment, which equalizes the vertical and horizontal scope amplifier sensitivities, the controls must not be moved. The 5-ohm fixed resistor is replaced by the voice coil; and from here on there are two ways to proceed.

Method A: Without touching any oscilloscope control and leaving the variable resistance set at 5 ohms, measure carefully the horizontal width w and the vertical height h of the trace. Both measurements should be in the same units.

Now the impedance is calculated from the formula

$$Z = 5h/w \text{ ohms}$$

This method is handy if a large number of measurements are to be made, and should be used if the trace is *very* different from a straight line. (It will probably be a narrow ellipse.)

Method B: Do not touch the oscilloscope controls. Readjust the variable resistor until the trace is symmetrical about a 45° line (so that the width and height of trace are equal). The voice-coil impedance is then equal to the value of the variable resistor. This method is useful if the trace is very nearly a straight line—the trace *will* be a straight line at one or more frequencies when the voice coil acts as a pure resistance. One of those frequencies is very close to the bass resonant frequency. This method is also better for small oscilloscopes using cathode-ray tubes 2 inches or less in diameter.

Whichever method is used, it is interesting to study the variation in impedance with the type of baffle used. Even if a hand is placed in front of the speaker, a distinct change will occur in the oscilloscope trace as impedance varies.

In all this work a source of audio-frequency voltage is required. Such an oscillator should have an output of at least 5 volts across a load of 10 ohms for low-impedance work. For high-impedance measurements an output of 20 volts across 10,000 ohms is required. These outputs are low and easily satisfied by commercial oscillators.

Chapter 8

Amplifier Impedance Measurements

When an audio amplifier drives a speaker system, there are two impedances which affect the performance. The first (Z_s in Fig. 801) is the impedance of the speaker. Techniques for measurement of

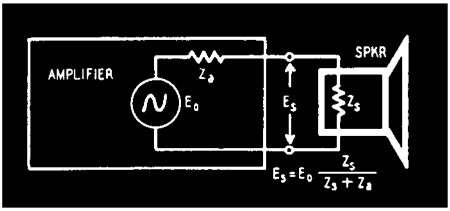


Fig. 801. Basic amplifier impedance circuit.

speaker impedance were described in the preceding chapter. The second impedance we must consider is the internal impedance of the amplifier (Z_a in Fig. 801). The voltage across the speaker is not the full voltage of the amplifier (E_0) because the internal impedance of the amplifier is in series with the speaker. If the amplifier's internal impedance is big enough to be significant and if the speaker impedance varies with the frequency, the voltage across the speaker will not be uniform at all frequencies but will have peaks at the impedance peaks in the speaker. The voltage across the speaker will differ from the fundamental voltage of the amplifier by the ratio indicated in the formula of Fig. 801. This is one reason why an audio amplifier should have a low internal impedance.

The damping factor

The internal impedance of an amplifier may be measured at any one of the amplifier's output taps. A more useful unit is the *damping factor,* which is equal to the rated output impedance divided by actual internal impedance. This makes it unnecessary to state at which output taps the measurements were made. The internal impedance is not a constant value but depends on the frequency at which it is measured. The complete description of an amplifier's internal impedance is a curve showing the damping factor versus frequency.

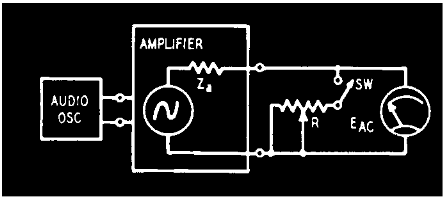


Fig. 802. Circuit for measuring impedances.

Normally an audio amplifier will have an internal impedance of 1/2 to 1/20 its rated output impedance. The damping factor will therefore be in the range of about 2 to 20. This damping factor is determined by the plate resistance of the output tubes, the circuit in which they are used, and the design of the output transformer. The use of feedback around the output stage reduces the effective internal impedance. Amplifiers with triode output tubes such as 2A3's or 6B4-G's without feedback and using traditional transformer design, have damping factors of 2 to 3. Feedback will increase the damping factor. Beam power tubes such as the 6L6 give an amplifier a very high internal impedance. Feedback is essential with these types, and their high gain makes considerable feedback feasible.

There are several ways of measuring electrical impedance. Many of them require bridge circuits or that hard-to-find item, an a.c. milliammeter. The methods to be described have been limited to the simplest equipment possible. An audio oscillator is needed as well as an a.c. voltmeter, a couple of rheostats of about 10 and 25 ohms and a few 10-watt resistors between 10 and 50 ohms. The oscillator should be capable of covering the audio range in which you are interested and should have reasonably low distortion. The voltmeter is preferably of the electronic type, though good results can be obtained with a sensitive volt-ohm-milliammeter with a.c. scales in the 0.5- to 3.0-volt range. Just remember to make all measurements on the same scale if your meter isn't uniformly calibrated on all scales.

One of the simplest methods of measuring the internal impedance of an amplifier is to use the scheme shown in Fig. 802. The plan is to measure the output voltage with no load across the amplifier; then with an adjustable resistor as a load to find at what load resistance the output voltage will be one-half the no-load voltage. The audio oscillator is connected to the amplifier input, and an on-off switch and rheostat are connected in series to the amplifier output terminals. These should be the impedance taps used in your particular installation. The rheostat should be capable of covering a range down to about 1/20 of the nominal output impedance. The voltmeter is placed across the output taps. With the switch open, the amplifier gain or oscillator input is adjusted until a conveniently measurable voltage is obtained, say 2.0 volts. The switch is then closed and the rheostat adjusted until the voltage drops to 1.0. The switch can now be opened and the resistance of the rheostat measured. This resistance equals the internal impedance of the amplifier.

It may sometimes happen with an amplifier of very high damping factor that you cannot get down to one-half the no-load voltage with the rheostats available. In that case the internal impedance can be calculated from the following formula, using voltage readings taken at the lowest resistance setting of the rheostat.

$$Z_a = \frac{RE_0}{E_R} - R$$

Z_a is the internal impedance of the amplifier, E_0 is the no-load output voltage, E_R is the voltage across the load resistance R. For example: if the open-circuit voltage at the 8-ohm tap is 2.0 volts and when the load resistance is lowered to 1.3 ohms (as far as you can go with your rheostat), the output voltage is 1.4 volts, and the internal impedance is

$$Z_a = \frac{1.3 \times 2.0}{1.4} - 1.3 = 1.85 - 1.3 = 0.55 \text{ ohms}$$

These measurements should be made at several different audio frequencies until the entire audio range is covered. A good plan is to check at 20, 30, 50, 100, 1,000, 5,000, 10,000 and 15,000 cycles. If the results are about equal or change smoothly you have a good picture of the amplifier's internal impedance characteristics. If irregular results are obtained, further checking is in order.

The internal impedance curves of two very different amplifiers are shown in Fig. 803. The lower curve (amplifier A) is that of the Brook 12A amplifier. This is a triode output amplifier (push-pull 2A3's), with a high-quality output transformer and with inverse feedback. The upper curve (amplifier B) is that of a home-built beam-power-tube amplifier with a lower quality output transformer and a poorly designed feedback system. However, it is typical of many published designs and of amplifiers used in home music systems.

This type of curve results when the feedback is not uniform over the entire audio range. If the frequency response of the amplifier is not quite flat without feedback, then feedback may help flatten the

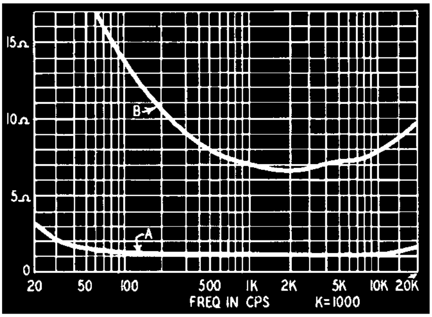


Fig. 803. Amplifier impedance curves.

response, but the internal impedance of the amplifier will rise at each end of the curve. Using the feedback loop around the final stage to adjust frequency response or using it in tone control circuits can cause a lot of trouble in the matter of damping factor. Amplifier B may sound satisfactory with a narrow range speaker system but the superiority of amplifier A with a wide range system is obvious.

Chapter 9

Measurements with the V.T.V.M.

WITH a little imagination and ingenuity mixed with a knowledge of the methods of applying the v.t.v.m, difficult problems in audio measurements can be solved.

The modern v.t.v.m. measures both a.c. or d.c. voltages. It requires essentially no power from the circuit to which it is connected, thus enabling measurement of true r.m.s. voltages in high-impedance circuits and true d.c. voltages in complicated series circuits.

Fig. 901 illustrates a typical phase inverter and output stage as

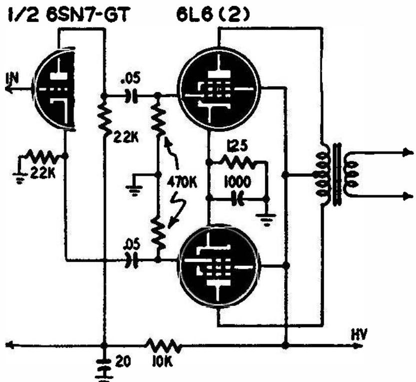


Fig. 901. *Idealized a.f. phase inverter.*

found in many amplifiers. The v.t.v.m. can be used to check the balance of the audio voltages applied to the output-tube grids. In this circuit it is particularly important that the value of the cathode resistor and the plate load resistor in the phase inverter stage be identical. With a constant audio signal applied to the input of the amplifier from an oscillator, the audio (a.c.) voltage appearing on one output-tube control grid should be the same as that on the other output tube's control grid. If the voltages from each grid to chassis differ by more than a few percent, the quality of the amplifier will be seriously impaired.

There are several possible causes of unbalance in the circuit shown. Values of the grid resistors in the output stage should be the same. They should be checked with an ohmmeter and replacements selected which match in value. The value of the coupling capacitor in series with each output grid will affect the balance at low audio frequencies. For best results, these should be the same. The capacitance of the coupling capacitor is chosen to have negligible reactance at the lowest frequency the amplifier is required to reproduce. The .05-μf capacitors in Fig. 901 have a reactance of approximately 31,850 ohms at 100 cycles per second.

At any given frequency of operation the capacitor can be considered a resistor in series with the grid resistance. The reactance of the capacitor is thus indicated as R1 in Fig. 902. The grid resistor is R2. This

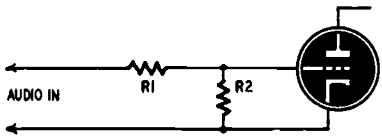


Fig. 902. *R1 indicates coupling reactance.*

simplified diagram shows how the reactance of the capacitor forms a voltage-dividing network, with the grid resistor as the output portion. Since the reactance of a capacitor varies inversely with frequency, it can be seen that the voltage at the grid of the tube would suffer if the capacitor value were too low and its reactance high at the lower frequencies of operation.

Coupling and bypass capacitors

The effectiveness of the coupling capacitor at all frequencies of operation can be measured with the v.t.v.m. Connect an audio oscillator to the input of the amplifier and measure the audio voltage to chassis on both sides, the input and output sides, of the coupling capacitor. Write down the voltages measured on each side, using different audio frequencies from the audio oscillator. As the frequency is made lower and lower, the difference between the voltages measured on the input and output sides of the capacitor will become greater. The low-frequency response of the amplifier can be improved, of course, by using bigger coupling capacitors.

If the coupling capacitor is leaky and permits d.c. to pass, the v.t.v.m. will detect it. Set it up to measure d.c. and connect it to the output side of the coupling capacitor. With the following tube removed, there should be no d.c. present unless the capacitor is defective. Replacement coupling capacitors should be high quality mica, ceramic or paper dielectric, with a voltage rating several times that of the d.c. on the plate of the preceding tube.

The effectiveness of bypass capacitors can be checked simply with the v.t.v.m. and an audio oscillator. With the oscillator connected as shown in Fig. 903, from grid to chassis (adjusted to an output voltage less than the d.c. bias on the grid) and the v.t.v.m. across the cathode bypass capacitor, no audio voltage should be indicated. This test should be conducted at the lowest frequency the amplifier is expected to reproduce. If audio voltage is shown, the stage will be degenerative

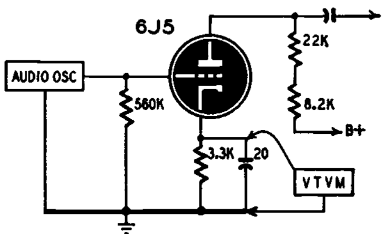


Fig. 903. *V.t.v.m. checks cathode bypass.*

at that frequency and all lower frequencies. To improve the low-frequency response, replace the bypass with a larger one. The reactance of a 20-μf capacitor at 100 cycles is 80 ohms. Since this is small compared with the value of the cathode resistor, 3,300 ohms, the amplifier's response at 100 cycles should be good.

Measuring amplifier gain

The gain of an amplifier is easily measured with the v.t.v.m. The input power and output power as measured by it are converted to *decibels of gain.* Set up the amplifier as shown in Fig. 904 with resistors to simulate the input and output devices. Adjust the volume control for maximum gain. If the resistor in the input is intended to simulate the internal impedance of a dynamic microphone it will be approximately 25,000 ohms. (It is best to get exact data on the internal impedance of a microphone or pickup from the manufacturer.) Adjust the output of the audio oscillator to one volt as indicated by the v.t.v.m. The input power is then found by the formula: $W = E^2/R$.

Since the voltage is 1, E^2 is 1×1 or 1. Dividing by 25,000 gives an input power of .00004 watts. For convenience, this might be called .04 milliwatt.

Supposing the output load resistor is made equal to a speaker impedance of 8 ohms, the output power can be computed in the same way. If the v.t.v.m. reads 13 volts across the 8 ohms, the power output is 13 $\times$ 13 (or 169) divided by 8. This gives 21 watts output.

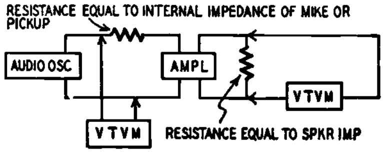


Fig. 904. This block diagram indicates the technique to be followed for measuring the gain of an audio-frequency amplifier. Although two v.t.v.m.'s are shown, a single meter can be used with a d.p.d.t. switch.

The output power divided by the input power is the power ratio and will give the gain of the amplifier in decibels by the formula: db = 10 $log_{10} \times$ power ratio.

A decibel chart

Rather than go to the trouble of finding the *log* and working this out the hard way, the chart of Fig. 905 is supplied to make it simple.

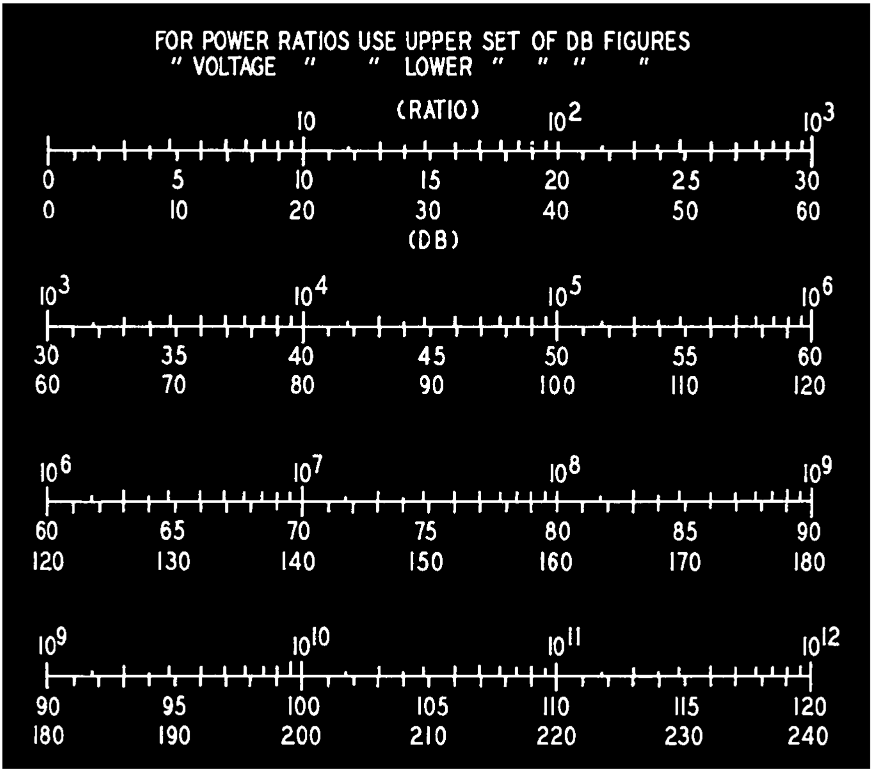


Fig. 905. This chart simplifies conversion of power and voltage ratios to decibels.

Locate the power ratio on the upper scale and read below it the gain in db. For power ratios, read the upper set of db figures. For the example we have chosen, the power ratio is 21 divided by .00004, or 525,000 to 1. Find 525,000 on the upper second scale and read below it approximately 57.3 db. The gain of the amplifier is 57.3 db.

If the amplifier has the same impedance at both input and output, such as the 600 ohms used commonly in line amplifiers, the task of determining gain is considerably simpler. The gain of such an amplifier can be figured from the ratio of the *voltages* appearing across the input and output without converting to power ratios. Using the same setup as shown in Fig. 904, measure the two voltages. Divide the larger voltage by the smaller one. This data can be used to determine gain by the formula: db $= 20\ log_{10} \times$ voltage ratio.

Again, instead of working it out, refer to Fig. 905. Using the lower set of db figures, read the gain opposite the voltage ratio. For example, if the voltage across the input is 1 volt and the voltage across the output is 200, the voltage ratio is 200/1 or 200. Locating this ratio in the chart will show a gain of 46 db opposite the ratio.

The over-all fidelity of the amplifier can be measured by plotting the gain as measured by these methods at a number of frequencies. For example, plot the gain at:

100 cycles	1,000 cycles	7,000 cycles
300	1,500	10,000
500	3,000	15,000
700	5,000	20,000

The variations in gain in db with reference to the gain at 1,000 cycles indicate the over-all fidelity of the amplifier.

Chapter 10

Intermodulation Distortion Tests

THERE are a number of ways of measuring the distortion produced in an audio system. Distortion-factor meters and wave analyzers have been used for many years, but intermodulation analyzers are relatively new. Though no objective measurements have yet been devised that can predict accurately whether or not listeners will like the sound of a particular system (the listeners' ears are, after all, the final criterion), intermodulation tests seem to give better clues to hearer acceptance.

Distortion-factor meters are still most common. A single-frequency sine wave is fed into the amplifier and the output is filtered so that the original frequency disappears. The remainder, which consists of any harmonics generated by the amplifier plus whatever noise is present, is measured. The fundamental output is also measured and the harmonics are then figured as a percentage of the fundamental. For good-quality systems 5% is often satisfactory, but nowadays high-precision professional and high-fidelity equipment has only a fraction of 1%.

The wave analyzer works on the same principle; but instead of measuring all harmonics at once, it filters and measures only one at a time, usually by beating an oscillator with the harmonic to be measured and passing the resultant beat frequency through a very selective filter. The advantage here is that the highly selective filter narrows the bandwidth measured so much that most of the noise is excluded. It also tells just which harmonics are most prominent, giving a clue to the fault, if any, in the system.

The trouble with harmonic measurement is that only distortion in the low-frequency half of the spectrum can be measured; if the fundamental is above the mid-frequency point, all the harmonics fall outside the audio band the equipment covers.

The intermodulation method of measuring distortion is based on the fact that the worst offense to the ears is committed when two or more different frequencies being amplified simultaneously interact due to some nonlinearity in the amplifier. While pure harmonic distortion of a single tone merely creates additional frequencies which are exact multiples of the fundamental, interaction between two or more tones produces sum and difference frequencies which may be totally unrelated harmonically to the originals. The effect is a good deal worse than when some members of a musical ensemble play off pitch or wrong notes, destroying the harmonies or chords. Intermodulation measurements, therefore, usually approximate more closely the subjective reactions of a listener and are generally more useful than simple measurements of spurious harmonics.

The intermodulation method employs two tones: one of low frequency—between about 50 and 200 cycles, and one of considerably higher frequency—anywhere from about 1,000 cycles up. Both tones are fed to an amplifier, and an analyzer connected to the output measures the interaction between them.

The signal generator

Fig. 1001 is a block diagram of a typical intermodulation signal generator. The two oscillators produce sine waves, 60 and 2,000 cycles in this example. (The 60-cycle signal may be provided by the a.c. power line instead of an oscillator.) The two tones are combined in a network carefully designed for minimum distortion.

The resultant wave is shown in Fig. 1001. It is a 2,000-cycle sine wave, with a 60-cycle sine wave as its axis of symmetry. As a more or less standard condition (though no genuine standards have yet been set) the voltage amplitude of the 60-cycle wave is four times that of the 2,000-cycle wave (12 db greater). With this relationship, the intermodulation distortion percentage is usually roughly four times as high as a straight harmonic distortion measurement would be on the same equipment. The exact ratio varies with the cause of distortion.

The output waveform of Fig. 1001 illustrates what happens whenever two or more frequencies are combined—the lower frequencies act as axes for the upper ones, a single composite wave being formed. The shape, amplitude, and position of any portion of the wavetrain about the main a.c. axis depend on the original frequencies, amplitudes, and phase relations. To produce undistorted sound output, the composite wavetrain must reach the loudspeaker in exactly the same condition it assumes at the amplifier input. (The exception is that the ear will tolerate a rather large amount of phase change.)

As the composite generator output goes through the amplifier being tested, the low-frequency signal causes relatively large positive and negative grid-voltage excursions at each stage. If every tube operates on a linear portion of its transfer characteristic up to the maximum grid-voltage excusions in each direction, no distortion takes place. But if—as is always the case, since nothing is perfect—a nonlinear region is encountered during the swing in one or both directions, those alternations of the 2,000-cycle superimposed signal which occur while the

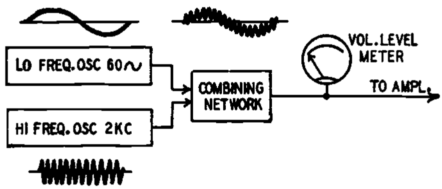


Fig. 1001. *Intermodulation signal generator combines two basic frequencies.*

tube is nonlinear will be either greater or smaller in amplitude than when the tube is operating linearly.

Typical distortion

As an example, suppose one stage in the amplifier is a resistance-coupled 6J5 with a 50,000-ohm plate load resistor, a 50,000-ohm following grid resistor, and 1,000-ohm cathode-bias resistor. Plate supply voltage is 300. According to the resistance-coupled amplifier charts in the RCA tube manual, voltage gain is 13 and maximum a.c. output voltage should be 41. Dividing 41 by 13, we find that maximum signal voltage at the grid should be 3.15.

Assume that the 3.15-volt input maximum is exceeded slightly. On positive input peaks the grid begins to draw current, flattening off the output wave to some extent. Each time the 60-cycle signal reaches a positive peak, therefore, the amplification of the tube effectively decreases somewhat and the 2,000-cycle alternations superimposed on the 90-degree point of the low-frequency wave have smaller amplitude.

It is likely, too, that negative 60-cycle excursions now take the tube into a nonlinear region, especially with such a low tube-load resistance. At and about the 270-degree point of the low-frequency wave, therefore, given changes in grid voltage produce smaller changes in plate voltage and again amplification decreases. The result is reduced amplitude of the 2,000-cycle alternations superimposed on the 270-degree region of the 60-cycle wave.

Fig. 1002 is a block diagram of an intermodulation analyzer. The amplifier output wave at A is fed first into a high-pass filter which removes the 60-cycle component from the composite signal. This leaves only the 2,000-cycle signal. Since the amplitude of the 2,000-cycle signal is no longer constant (because of the distortion in the amplifier), it appears at the filter output as a modulated wave (shown at B). Its amplitude is normal or maximum at the points which correspond to the 0-, 180-, and 360-degree regions of the filtered-out 60-cycle wave, and less than normal at the 90- and 270-degree points.

The wave at B is exactly like the familiar r.f. modulated wave (except, of course, for the actual frequency, which is 2,000 cycles) and can be detected and measured in the same way. It is first rectified (detected) to yield the d.c. wave at C. Then it is fed through a low-pass filter to remove the 2,000-cycle pulsations. The low-frequency modulation envelope remains at D.

Note one important point. The remaining modulation envelope is *not* the original 60-cycle signal. It is the change in amplitude of the 2,000-cycle signal produced by the 6J5 nonlinearities. Since the amplitude was changed twice during each low-frequency cycle (once by the tube's drawing grid current and once by the negative excursion into a non-linear transfer region), the modulation envelope at D is twice the original low frequency or 120 cycles. If either the negative or positive 6J5 grid excursion alone had caused nonlinearity, that is, a change once per low-frequency cycle, the wave at D would be 60 cycles. Its shape is usually not sine, either, depending on how abruptly the 6J5's characteristic departed from linear.

The wave at D in Fig. 1002 is produced solely by amplifier nonlinearity, which affected the relationship between two frequencies. If the amplifier were linear, the 2,000-cycle signal would have remained constant in amplitude as it was originally, and detection of a wave at B would have resulted in pure d.c. Obviously, then, the amplitude of the wave at D is a direct indication of the amount of distortion present. It is measured by an ordinary rectifier-type volt meter. In some instruments, the circuit is arranged so that the operator can tell whether distortion is greater in the negative or positive direction.

The percentage of intermodulation is equal to the modulation percentage of the "carrier" wave at B. With the analyzer calibrated to present a fixed level to the detector, the meter may be marked directly in percent.

When citing intermodulation figures, the test conditions should be specified. The two frequencies should be given, as well as the amplifier output level. For greatest precision, the amplifier input level and the amplitude relationship of the two frequencies should also be mentioned. Measurements should be made with several sets of frequencies, as the distortion varies somewhat.

An amplifier which has low harmonic distortion ordinarily shows low intermodulation as well. Intermodulation results usually agree more closely with listening tests, however, and give a better indication of performance at high frequencies.

The percentage figure for intermodulation is always higher than that for harmonic distortion, which is why some manufacturers feel it

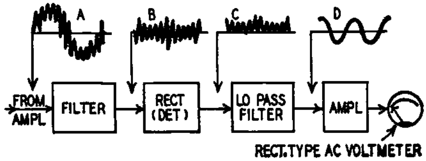


Fig. 1002. The analyzer filters the output signal, detects and measures distortion.

unwise to publish it. A more valid reason is that standards for intermodulation testing have not yet been agreed on, and this may make interpretation difficult.

Standard intermodulation tests

Two standard intermodulation tests have been advocated. The SMPTE test was standardized by the Society of Motion Picture and Television Engineers. The CCIF test is recommended by the International Telephonic Consultative Committee, and is sometimes called the *difference-frequency* intermodulation test.

The SMPTE method requires a l.f. and a h.f. test signal. Usually the first is 100 cycles and has 4 times the amplitude of the second which is 5 kc. Due to nonlinearity in the amplifier, sum and difference frequencies are created. They exist as sidebands at intervals of 100 cycles on both sides of the 5-kc "carrier" (see Fig. 1003-a). X and Y are the l.f. and h.f. signals, respectively.

Distortion due to the nth order sideband is then defined by the fraction

$$\frac{\text{nth harmonic of sidebands}}{\text{h.f. signal amplitude}}$$

The CCIF test uses 2 h.f. test signals of equal amplitude. The frequency difference between them is usually in the range 30-400 cycles. This difference frequency is one of the distortion products due to intermodulation. The distortion it creates is measured by

$$\frac{\text{difference-frequency amplitude}}{\text{sum of h.f. test signals}}$$

Intermodulation also exists between each h.f. signal and the second harmonic of the other (see Fig. 1003-b). This sideband distortion is defined by

$$\frac{\text{sum of both sidebands}}{\text{sum of h.f. signals.}}$$

According to Peterson*, the CCIF test is better than the SMPTE, especially where the audio system has limited h.f. response. Hearing aids, filters, noise suppresors, etc., are systems that fall in this category. A harmonic test on them would be misleading because of the limited h.f. response. The SMPTE test cannot give a true picture because it has a strong l.f. test signal. This means that the h.f. end of the amplifier is not given a real check.

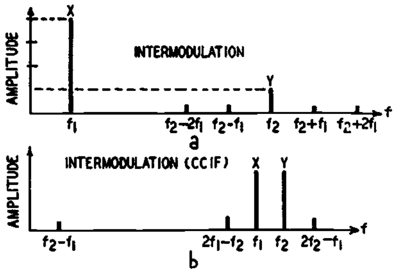


Fig. 1003. Two intermodulation measurement systems. Illustration a is the SMPTE test. Note the amplitudes of X and Y.

A pair of signal generators are required for intermodulation measurements. For the SMPTE test the frequencies may be fixed at 100 and 5,000 c.p.s. as noted previously. For the difference-frequency test the oscillators may be ganged. This gives a constant beat through a wide range. With this setup X and Y may be moved up and down the scale while (f_2—f_1) remains stationary (see Fig. 1003-b).

*Arnold P. G. Peterson, *Technician.* Publication available from General Radio Co., Cambridge, Mass.

Chapter 11

Audio Waveform Analysis

THE frequency response of many amplifiers, provided basic design is good, can often be considerably improved.

Sine wave testing

Distortion, frequency response, overloading, etc., in the audio system can be checked with the following aids: (1) sine waves, (2) square waves, and (3) sweep-frequency records. The first method (illustrated

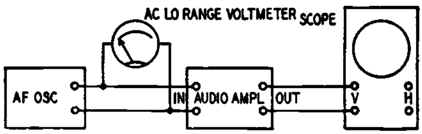


Fig. 1101. *Setup for checking phase shift.*

in the setup in Fig. 1101) consists simply of a variable audio oscillator and a scope. The oscillator feeds a signal into the amplifier, and the scope (across the amplifier output) indicates what has happened to the

Fig. 1102. *Single and double sine wave forms are shown in a and b.*

audio sine wave as it passed through the amplifier. The low-range a.c. or output meter connected across the audio oscillator output should read the same value of signal input for all applied frequencies.

Let us examine some of the waveforms which can be obtained with sine wave testing. We are familiar, it is assumed, with the sine waveforms shown in Figs. 1102-a and 1102-b. Suppose that a sine wave similar to that in Fig. 1103-a is present at the output, *even though no*

Fig. 1103. *Waveform a is distorted sine wave output. Waveform b is a* 60-*cycle since wave modulated with* 400 *cycles.*

signal is being fed into the amplifier input. A logical assumption is that a.c. hum is being picked up at some point in the amplifier. Determine the frequency of the waveform by adjusting the scope frequency control until a single stationary cycle as in Fig.1102-a is obtained; then read the frequency setting on the scope sweep control. If the frequency is 60 cycles, look for cathode-to-heater leakage in one of the tubes, or stray pickup from the filament leads, power transformer, or some external source. If the hum frequency is *120* cycles, an open or low-valued filter capacitor is indicated.

Sometimes a wide, blurred waveform may appear at the output instead of the usual thin, well-defined line. Such indication is caused by oscillation within the amplifier. *Low-frequency* feedback resulting in oscillation may be due to open plate or screen-grid bypass capacitors or to inadequate decoupling between stages. *High-frequency* oscillation may be the result of defective plate or cathode bypass capacitors, or due to regeneration in one or more stages. An open plate bypass

Fig. 1104. *Typical distorted waveforms. Note waveform a is clipped at top while waveform b is clipped top and bottom.*

capacitor in the output stage, and output leads which are too close to the amplifier input circuit can be notorious offenders in this respect. This condition is often caused by improperly shielded grid circuits in high-gain amplifiers, or by plate and grid leads placed too close to each other. Shielding all grid and low-level leads, and properly dressing grid and plate leads almost invariably corrects this latter condition. Be careful to check response after applying these remedies. It may cause appreciable loss of high frequencies.

The waveform in Fig. 1103-b is the result of modulating a 60-cycle sine wave with 400 cycles. Typical distorted waveforms are shown in Figs. 1104 and 1105. Note that the relatively small amount of distortion in Fig. 1104-a results in clipping of the positive peaks only, while the high distortion shown in Fig. 1104-b has flattened both positive and negative peaks. Positive peak flattening is usually due to overloading or insufficient bias. Negative peak flattening can be caused by a shorted cathode bypass capacitor. The waveform in Fig. 1105-a (note the distorted negative peaks) is a typical example of distortion caused by overloading. Fig. 1105-b indicates presence of regeneration *and* overload. The terms positive and negative peaks are applicable literally only when checking the waveform at the plate of a particular stage.

Fig. 1105. *Waveform a is example of overload distortion while waveform b indicates regeneration* and *overload.*

The usual stage-by-stage checks should be employed to determine the source of distortion. First, be sure that the distortion is due to a defect within the amplifier and *not* from excessive input from the signal generator. Connect the scope's vertical leads *successively* across the plate and grid circuits of each stage, beginning with the output stage and working back toward the input until the distorting stage is located. If distortion as indicated by these scope checks suddenly disappears you can be certain that its source has just been passed. Don't forget to check compressor, expander, and phase inverter circuits. In most cases you will find that the distortion is caused by defective tubes, leaky, open or shorted capacitors, improper grid bias, incorrect plate voltage, resistors that have changed in value, power supply hum, etc.

It is important that we positively identify the half-cycle being viewed on the scope. The first step is to check the polarity or phasing of the scope. Connect the negative side of a battery (1.5–4.5 volts) to the ground terminal of the vertical amplifier. Turn up the vertical gain, then touch the positive side of the battery to the hot vertical input terminal. If the trace moves up, the upper half of an alternating input signal will be positive and the lower half will be negative. If this alternating signal is amplified by a single stage before being fed into the scope, phase will be reversed and the positive half-cycle on the grid of the amplifier will be shown as the lower half of the trace on the scope. Thus, the polarity of the observed signal corresponds to that of the scope only when there are an *even number of stages* between the signal source and the scope.

When the signal to the scope is taken from the secondary of a transformer; e.g., the voice-coil winding, check the phasing of the winding. Apply a positive signal of known polarity to the grid of the stage feeding the transformer and note the deflection on the screen.

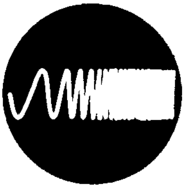


Fig. 1106. Typical response pattern produced by a sweep-frequency record. Use of these records eliminates necessity for audio sweep frequency generator. Sweep-frequency records include marker "pips" at various frequencies for identification of response limits.

Keep the input to the scope low, otherwise the vertical amplifier may overload and distort the trace. When feeding high voltages into the scope, distortion in the vertical amplifier may be avoided by feeding the signal directly to the deflection plates. Before doing this, check the polarity of the plates by using a d.c. source of 20 to 45 volts.

Frequency response

There are several methods of measuring frequency response characteristics. One of the simplest is to use a variable audio oscillator with a db meter (instead of the scope) connected across the output of the audio amplifier. As the audio oscillator is varied slowly from about 30 to 20,000 cycles (with its output kept constant as indicated on the low range a.c. voltmeter), the amplifier response can be read directly as + or – db on the output meter. If no db meter is available, the scope can be used as a response indicator. Simply turn off the horizontal amplifier, so that only a thin vertical line appears on the screen. (This line represents the amplifier output). Adjust the height of the line to cover any desired part of the screen at a certain reference frequency (usually 1,000 cycles). Then vary the audio oscillator as above and note whether the vertical line decreases or increases in length. A 50% change indicates a corresponding voltage drop or gain of 6 db. A drop of this amount would indicate the practical limit of response at that particular end of the band. (First check the linearity of the scope's response over the frequencies to be measured by connecting it directly to the a.f. oscillator output.)

Hum originating in the power supply can be traced with a 600-volt, 1-µf capacitor in series with the scope lead to the vertical circuit and probed to points in the power supply. The amplitude of the 60- or 120-cycle sine wave at these points will indicate the effectiveness and condition of each filter component.

Sweep-frequency testing

Sweep-frequency records (or transcriptions) are available for use without a signal generator. See the typical response pattern in Fig. 1106. Such records include marker "pips" at various frequencies for identification of response limits. An audio sweep-frequency generator may, of course, be used (if the pocket-book can stand the strain); but, in any case, an audio sweep unit is substituted for the oscillator in the test setup.

When the sweep-frequency record is used, you will need a tone arm employing a good magnetic or reluctance pickup, and a small

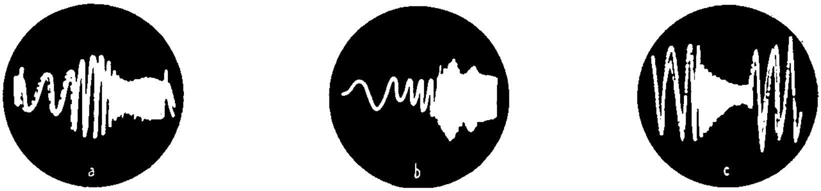


Fig. 1107. Waveforms in a, b, and c give an indication of the frequency response of an amplifier. Such waveforms give a clue as to components to be changed.

amplifier (preferably a twin-triode such as the 12AU7, 6SN7-GT, etc., connected as a two-stage cascade amplifier). Flat response over at least the audio range is necessary for accurate results.

Analysis of the frequency response, distortion, harmonic content, etc., becomes extremely simple with this method. Note the response pattern shown in Fig. 1107-a. This denotes a practically flat response from about 50 to 10,000 cycles (fair for a high-fidelity system). In 1107-b the frequencies from about 200 to 4,000 cycles are markedly emphasized (about 6 db, or twice the height on the screen of the average waveform height), while the response is down at the low-and high-frequency limits. This would indicate that the cathode or coupling capacitors have decreased in value, or that the plate bypass capacitor is too large. It would be wise to check the value of the grid resistors and decrease if necessary. If the amplifier response is generally poor, it can be improved (with a sacrifice in gain) by using negative feedback. Fig. 1107-c indicates a definitely poor response above about 4,000 cycles. If no definite defects are found in the interstage coupling units, try decreasing the value of the plate bypass capacitor (and cathode bypass capacitors). Lowering the value of the coupling capacitor probably would help, but at the risk of impairing the low-frequency response.

Sweep-frequency testing has many advantages when used for audio design or replacement work. The effect of volume controls, tone control and feedback circuits, filters, etc., on frequency response, phase, harmonic distortion, and other factors can be readily observed on the scope screen.

Phase-shift technique

The frequency response can also be determined by measuring the *phase shift* which a signal undergoes in passing through the amplifier.

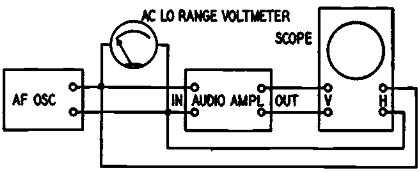


Fig. 1108. *Setup for checking phase shift.*

The testing arrangement is illustrated in Fig. 1108. This method is sensitive because even minute changes in amplitude are accompanied by

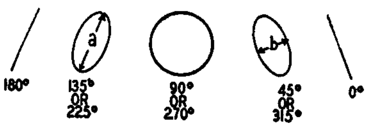


Fig. 1109. *The patterns show phase shift.*

large shifts in phase. The test procedure is the same as that for the amplitude-vs-frequency method except that the input and output signals are compared as to phase by applying them to the vertical and horizontal

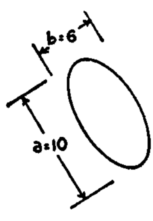


Fig. 1110. *b/a is tangent of angle.*

plates of a cathode-ray oscilloscope. Depending upon the degree of phase shift, the resulting pattern, called a Lissajous figure, will be a straight line, a circle, or an ellipse, as shown in Fig. 1109. The phase shift may be estimated from the shape of the figure, but is best calculated from the formula, $\theta = 2 \tan^{-1}$ (b/a), where θ is the phase-shift angle in degrees, b is the length of the shorter axis, and a is the length of the longer axis. See Fig. 1109 for an illustration of axes a and b.

As an example, in Fig. 1110, the short axis b is 6 units long and the

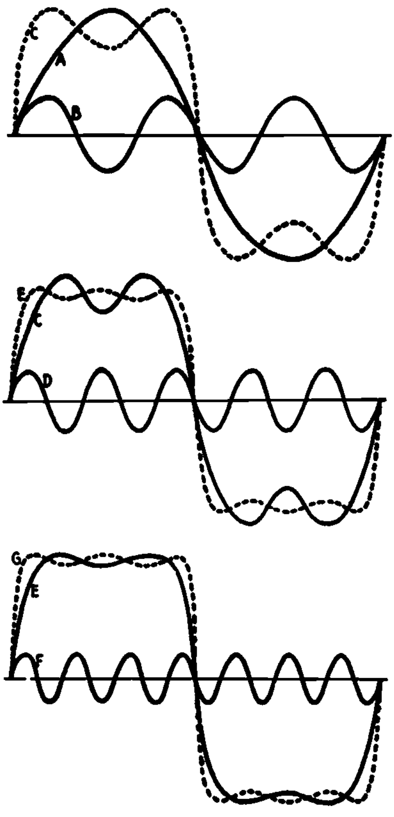


Fig. 1111. *How harmonics add to form a square wave.*

long axis a is 10 units long. Therefore, b/a = 6/10 = 0.6. Looking in a table of tangents to find the angle whose tangent is 0.6000, we find that the closest one is 0.6009, which is the tangent of 31 degrees or 149 degrees. Multiplying these angles by 2 as the formula states, the phase shift is $31 \times 2 = 62$ degrees or $149 \times 2 = 298$ degrees. To determine which of the two is the correct phase angle, note that the Lissajous figure in Fig. 1110 is tipped over to the left. Hence, the phase shift is approximately 62°. Had it been tipped over to the right, the phase angle would be about 298°.

The calculation is made for every frequency fed into the amplifier, from 500 to 15,000 cycles. However, in drawing up the response curve, phase shift, instead of amplitude, is plotted against frequency. If the curve is not sufficiently flat, the faulty stage or stages must be located by point-to-point testing, substitutions made, and the entire procedure repeated as before.

Square-wave testing

The main advantage of square-wave testing is that phase, distortion, and frequency response characteristics of an audio amplifier can be observed at the same time.

A square wave is square because it consists of a fundamental and all its harmonics *in definite phase and amplitude relation to each other.* Strangely enough, square-wave analysis is so simple because square waves are so complex. Theoretically, a square wave is the algebraic sum of a fundamental frequency and an infinite number of its harmonics, all sinusoidal in shape and all having a common time of origin. In practice, however, the 30th harmonic is the highest order of sufficient amplitude to be of consequence. Fig. 1111 suggests how the fundamental and its harmonics combine to produce a square wave

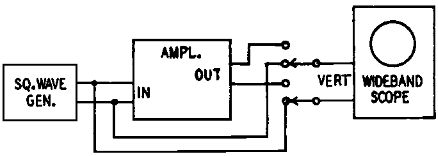


Fig. 1112. Square wave amplifier check.

by flattening the top and steepening the sides of the original sinusoidal waveform.

Any circuit which changes the phase relation or the amplitude of any of the components will distort the square wave. Only when the circuit passes all the frequencies of the square wave without attenuation or relative phase shift can the output be undistorted. This is the principle of square wave analysis.

Square-wave analysis is simplicity itself. The only equipment required is a square-wave generator and an oscilloscope, having a frequency response that will pass the fundamental and all the desired harmonics of the square wave without attenuation or relative phase shift. The square-wave generator is connected to the input of the circuit under test, and the oscilloscope to the output.

Set the fundamental frequency of the square-wave generator at 500 cycles with your equipment as shown in Fig. 1112. The input signal to the amplifier is viewed on the wideband oscilloscope to make sure that it is square. The scope is then switched to the output of the amplifier and the waveform viewed to see if it is square. If it is, the frequency response is satisfactory from 500 to 15,000 cycles. If the output waveform is distorted, the amplifier needs adjustment. The

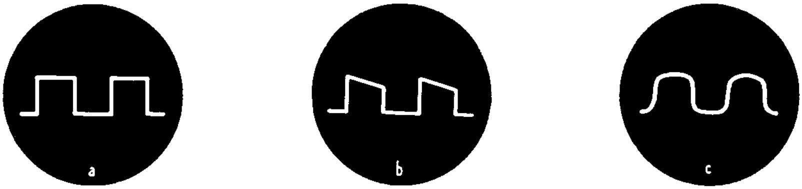


Fig. 1113 a, b, c. *Ideal square wave shown in waveform a. Waveform b reveals drop in low-frequency response; waveform c shows poor low and high response.*

scope is then switched to the output of the next to the last stage, then the stage before that, and so on, until the wave-shape becomes square. This localizes the difficulty in the following stage. Adjustments are made in the amplifier until the waveform assumes a square shape. The amplifier is then passing all frequencies between 500 and 15,000 cycles. There are no multiple-frequency checks, no calculations, and no curves to be drawn!

If no square-wave generator is available, the regular audio oscillator

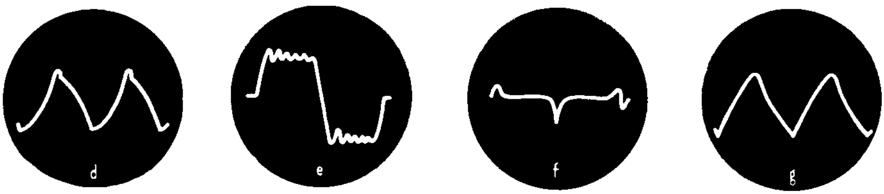


Fig. 1113 d, e, f, g. *Distorted waveforms shown in d, e, f, and g, produced during square wave test of amplifier.*

may be used by adding a simple twin-diode (such as the 6H6) in its output circuit and biasing it to operate as a limiter. Experiment with the bias by feeding a sine wave through the limiter and adjusting the bias until a pattern similar to Fig. 1113-a is seen on the scope.

The square wave shown in Fig. 1113-a is an unattainable ideal. The wave form of Fig. 1113-b indicates a certain amount of distortion, accompanied by a drop in low-frequency response. Fig. 1113-c shows comparatively poor response at both the low and high frequencies. In 1113-d there is considerable distortion and very poor low-frequency response. The sloping sides also indicate excessive phase shift. The drop in low-frequency response points to an open or defective cathode or screen-grid bypass capacitor. The dampened oscillations in 1113-e are usually due to a resonant condition at some particular frequency; check the values of the coupling, cathode, and plate isolating capacitors. Resistance loading of the interstage transformers is a sure cure in some

Square-Wave Component Table
Percentage Width

Harmonic	10% or 90%*	20% or 80%*	30% or 70%*	40% or 60%*	50%
1	.20	.37	.52	.61	.64
2	.19	.30	.30	.19	0
3	.17	.20	.07	–.13	–.21
4	.15	.09	–.09	–.15	0
5	.13	0	–.13	0	.13
6	.10	–.06	–.06	.10	0
7	.07	–.09	.03	.05	–.09
8	.05	–.08	.08	–.05	0
9	.02	–.04	.06	–.07	.07
10	0	0	0	0	0
11	–.02	.03	–.05	.06	–.06
12	–.03	.05	–.05	.03	0
13	–.04	.03	–.02	–.03	.05
14	–.04	.03	.03	–.04	0
15	–.04	0	.04	0	–.04
16	–.04	–.02	.02	.04	0
17	–.03	–.04	–.01	.02	.04
18	–.02	–.03	–.03	–.02	0
19	–.01	–.02	–.03	–.03	–.03
20	0	0	0	0	0
21	.01	.02	.02	.03	.03
22	.02	.03	.03	.02	0
23	.02	.03	.01	–.02	–.03
24	.03	.02	–.02	–.03	0
25	.03	0	–.03	0	.03
26	.02	–.01	–.01	.02	0
27	.02	–.02	.01	.01	–.02
28	.01	–.02	.02	–.01	0
29	.01	–.01	.02	–.02	.02
30	0	0	0	0	0

* Reverse signs given in table for pulse percentages over 50%.

amplifiers. The high-frequency leak or "pip" in 1113-f could be caused by improper shielding or inadequate isolation of the stages. The almost triangular pattern in Fig. 1113-g represents very bad low-frequency response and excessive phase shift.

Not all square waves are alike. They may differ in period or repetition rate, frequency, amplitude, and percentage width. The *period* or *repetition rate* is the reciprocal of the frequency of the fundamental and is defined as the time between the beginning of any two successive pulses. The time duration of the pulse t_1 is called the *pulse width*. These characteristics of the square wave are illustrated in Fig. 1114.

The ratio of the pulse width t_1 to the period t, when multiplied by 100, is known as the *percentage width*: $t_1/t \times 100 =$ percentage width.

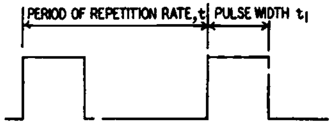


Fig. 1114. Square wave time definitions.

The percentage width depends upon the harmonic content of the particular square wave. It is possible to prove this by a complex mathematical procedure known as a Fourier expansion. This operation is used to analyze complex wave-shapes and to determine the amplitudes and natures of their components. A tabulation of the amplitudes of the fundamental and all its harmonics up to the 30th of square waves of different percentage widths is given in the table of square-wave components. This table lists percentage widths in steps of 10%, from 10% to 90%. The figures in it were obtained by using a Fourier expansion.

Chapter 12

Practical Williamson Amplifier

FOR many years the high-fidelity fan has been looking for the "perfect" amplifier circuit, one that would be flat throughout the audio range, practically distortionless, with adequate power output, and relatively inexpensive. Many circuits have been devised and many have come close to the ideal, but each has left something to be desired. The most popular circuit of recent years is the Williamson amplifier.

Although the original specifications called for the use of a British-

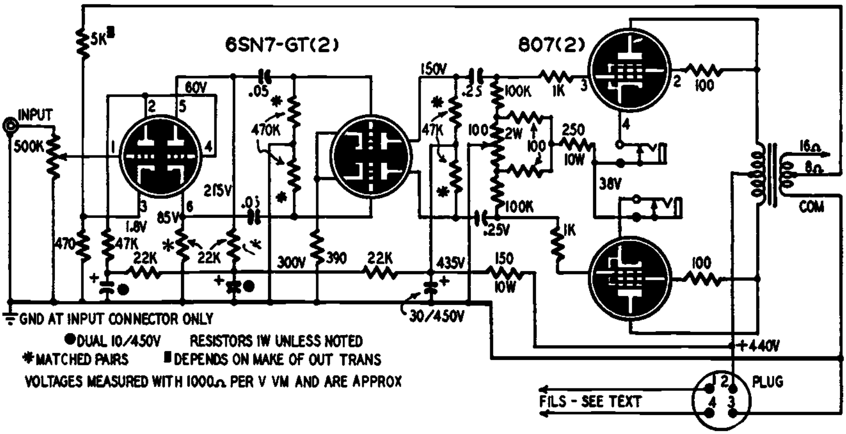


Fig. 1201. *Williamson amplifier circuit (Stancor version). Note use of 807's instead of KT66's and R-C power-supply filtering in first two stages.*

made output transformer (Partridge) and British tube types (KT-66), several American versions have appeared. Complete Williamson amplifier kits, including punched chassis, are now available from several American manufacturers, including Heathkit, Stancor, and UTC.

Other American companies, such as Acrosound, ADC, Peerless, and Triad, are producing special Williamson output transformers. Many distributors are also supplying complete or foundation kits. This amplifier uses Stancor transformers and follows that company's version of the Williamson circuit. (See Figs. 1201 and 1202.)

Construction

The power supply is built on a separate chassis to keep a.c. transformer fields away from the amplifier. Power is brought to the amplifier through a 4-wire cable.

Stancor suggests using two 9 x 7 x 2-inch chassis for the amplifier

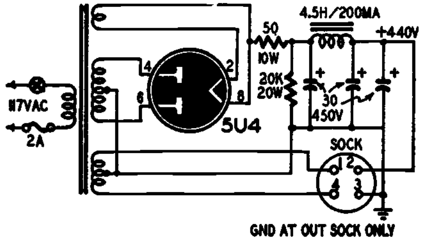


Fig. 1202. Power supply schematic.

and power supply. However, you may find this to be too small for the layout and wiring desired. An 11 x 7 x 2-inch aluminum chassis is adequate.

The power supply itself is conventional and should cause no trouble. The core of the choke should be perpendicular to the core of the power transformer to reduce the effects of stray magnetic fields.

The layout of the amplifier is not critical, because of the isolated power supply. Be sure the 807's have enough ventilation. The heater circuits should be wired first with twisted pair. These leads should run first to the 807's and then to the 6SN7's. The filament center tap is grounded (instead of grounding one side of the filament line) to minimize hum. To prevent ground loops connect all ground returns to a floating bus and ground the bus to the chassis only at the input connector. Be sure to use insulated mountings for metal-shell decoupling capacitors and connect their negative terminals to the ground bus. Use insulated mountings also for the jacks in the cathode circuits of the 807's.

All resistors should have a tolerance of 10% or better. It is very important that the starred resistors be matched pairs. If unbalance occurs in the direct-coupled stage it might bias the tube almost to cutoff. Unbalance in the phase-inverter circuit will cause unequal drive to the 807's. To get matched pairs take an ohmmeter to your parts distributor and measure resistors until you find two that give the same reading within 10% of the required value. Most distributors will let you do this. It is unnecessary to use a Wheatstone bridge to get two perfectly matched resistors; the ohmmeter will suffice.

A few precautions must be taken to prevent attenuation of the high audio frequencies. Do not use any shielded wire; mount all parts as far above the chassis as possible; keep all leads short; avoid cabling wires together; keep your wiring compact. It is amazing how much difference these small details can make. In this amplifier raising parts above the chassis extended the response 3 kc at the high-frequency

Materials for Williamson Amplifier

Resistors: 2—470,000 ohms, (matched), 2—100,000 ohms, 1—47,000 ohms, 2—22,000 ohms, 2—22,000 ohms (matched), 1—5,000 ohms, 2—1,000 ohms, 1—470 ohms, 1—390 ohms, 4—100 ohms, 1 watt, 10%; 2—47,000 ohms (matched), 2 watt, 10%; 1—250 ohms, 1—150 ohms, 1—50 ohms, 10 watt, wire-wound; 1—20,000 ohms, 20 watt, wire-wound; 1—500,000-ohm potentiometer, audio taper; 1—100-ohm potentiometer, 2 watts.

Capacitors: (Paper) 2—0.25 μf, 2—0.05 μf, 600 volts. (Electrolytic) 3—30 μf, 475 volts; 1—30 μf, 2—10 μf, 450 volts.

Transformers: 1 Williamson output transformer (Stancor No. A-8054 or equivalent); 1 power transformer: (Stancor PC-8412 or equivalent); 1 filter choke (Stancor C-1411 or equivalent).

Miscellaneous: Tubes: 2—6SN7-GT; 2—807; 1—5U4-G. 2 chassis, 11 x 7 x 2 inches; 3 octal sockets; 2—5-prong sockets; 2—4-prong sockets; 2—4-prong plugs; 1—input connector; 2 circuit-opening jacks; 2 insulated plate-cap connectors; 1 output terminal strip; 1 fuse holder; 1—2-amp fuse; 1 s.p.s.t. toggle switch; 1 line cord and plug; wire, solder, terminals, hardware.

Materials for Preamplifier

Resistors: 1—10 megohm; 6—100,000 ohms; 3—22,000 ohms; 1—3,300 ohms; 2—2,200 ohms, ½ watt; 1—10,000 ohms, 1 watt; 3—1-megohm potentiometers.

Capacitors: (Paper) 4—.002 μf; 2—.02 μf; 5—.05 μf, 600 volts; (Mica or ceramic) 1—200 $\mu\mu$f; (Electrolytic) 2—20 μf, 350 volts; 1—20 μf, 25 volts.

Miscellaneous: 1—12AX7 or 12AY7 tube; 1—12AU7 tube; 2—9-pin miniature tube sockets; chassis, switches, hardware.

end. Don't use a skimpy power transformer. This amplifier uses a sizeable amount of plate current and the power transformer must be able to deliver. See photo, Fig. 1203, for suggested parts layout of the amplifier and power supply.

Tests and adjustments

To balance the circuit plug a milliammeter into each of the jacks in turn and adjust the 100-ohm, 2-watt balancing potentiometer until both readings are equal at approximately 50 ma each. If your circuit does not balance—*check the 807's first.*

If the amplifier squeals, reverse the plate leads to the output transformer. This should correct the fault.

The response of the amplifier was checked with a Hewlett-Packard audio oscillator, an RCA audio voltmeter, and a Hewlett-Packard distortion analyzer. The frequency response is superior to broadcast standards and is truly high-fidelity. The unit is free from any inherent circuit noise or microphonics and is very stable. It is well worth the time and trouble put into its construction.

A preamplifier and equalizer should be used with this amplifier to obtain full output and optimum performance when using low-output pickups such as the variable reluctance and some magnetic types. A number of preamplifier-equalizer circuits have been developed by manufacturers and distributors of high-fidelity amplifiers. Some of these include loudness controls and rather elaborate preset equalizer circuits for almost every type of recording characteristic. Others are simpler but equally effective for most applications.

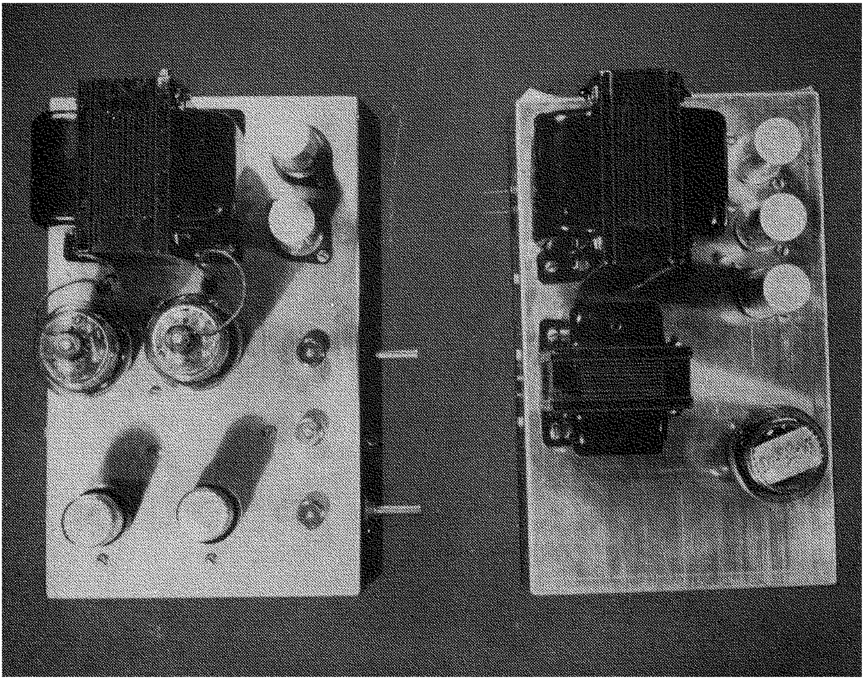


Fig. 1203. Separate amplifier and power-supply chassis reduce hum, allow uncrowded layouts.

Typical of the simpler circuits is the Heathkit WA-P1 preamplifier-equalizer shown in Fig. 1204. A switch selects either of two low-gain inputs for crystal pick-up or tuner, or the high-gain channel for magnetic pick-ups. A two-position turnover switch in the latter channel has positions foı 78-r.p.m. and LP recordings. Separate bass and treble controls provide up to 15 db boost or cut at 20 and 20,000 cycles, respectively. Signal voltage input required to develop 1.2 volts output—the approximate value required to drive the amplifier to full output—is 0.2 volt into the low-gain channels and .004 volt for the high-gain input circuit.

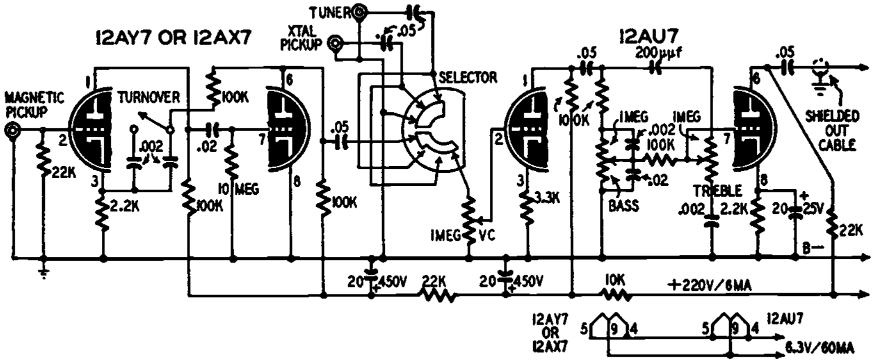


Fig. 1204. *Schematic of the Heathkit model WA-P1 preamplifier-equalizer kit. The turnover switch compensates for the different frequency characteristics used in 78-r.p.m. and LP recordings. Bass and treble controls give flat response in midpositions,* 15 *db boost or cut at opposite ends.* 12*AY7 is a low-microphonic type.*

Other preamplifier circuits can be used with the Williamson amplifier provided the amplifier input requirements are kept in mind. The circuit shown in Fig. 1204 has the advantage of simplicity.

Chapter 13

All-Triode Amplifiers

THESE amplifiers have sufficient flexibility to permit variation of frequency response to suit the individual ear and to compensate for the falling off of high- and low-frequency response which is inherent in some components of the complete sound system.

Deluxe triode amplifier

A number of intangibles go into the construction of an amplifier. Proper balance of highs and lows, correct time intervals for transients and elimination of undesirable transients, adequate power for peaks and good low-frequency response without distortion, and correct coupling to the loudspeaker are a few important points which frequently are neglected. This neglect leads to the difficult to describe but very real sensation known as listener fatigue—a phenomenon not limited to low-priced equipment.

The advantages of having the preamplifier and controls in a separate chassis are many. However, this leads to greater complexity and cost. In this chapter you have two forms of the basic amplifier, a deluxe model on two chassis and a utility model on one chassis. Although both give excellent results, the two-chassis model has a number of additional refinements.

Fig. 1301 and Fig. 1302 are top and bottom views of the two-chassis model, and Fig. 1303 and Fig. 1304 give the schematics of the preamplifier and amplifier. The preamplifier, of course, can be used with any amplifier. Make sure that the power supply can take on the added load represented by the preamplifier.

In addition to having the necessary gain for a magnetic phonograph cartridge, the preamplifier has bass, treble, and loudness controls. Also, a four-position switch, S1, is provided to compensate for the different types of high-frequency pre-emphasis used by the different record manufacturers. Position 1 gives nearly flat response, and will over-emphasize the highs on most records with the possible exception of some 78 r.p.m. foreign discs. Position 2 is best for most 78-r.p.m. records unless scratch requires the use of positions 3 or 4, and also is best for Victor and Capitol LP's. Position 3 provides compensation for Columbia and other LP's. Position 4 is useful for playing worn and noisy recordings. The values of 15,000, 5,000, 3,300 ohms and .04 μf are correct for the G-E cartridge. If the Pickering cartridge is used

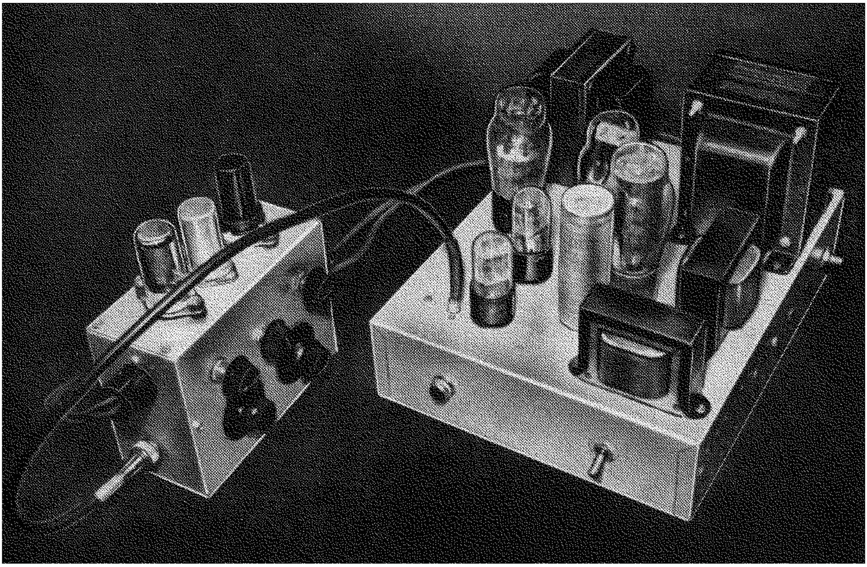


Fig. 1301. View of the two-chassis model of the all-triode amplifier. All the controls on this model are conveniently located on the preamplifier chassis.

these values should be changed to 10,000, 4,700, 2,700 ohms, and .03 μf, respectively. Those who wish still greater flexibility of record compensation can omit S1 and its associated components and substitute a Pickering 132E record compensator in the input circuit.

Switch S2 provides choice of pick-up from a magnetic cartridge or radio. The radio terminal also can be used for input from a crystal cartridge.

The loudness control (or compensated volume control) can be assembled as shown on a 23-position switch. The loudness control is a refinement well worth having. A simple type (such as the IRC, LC1 or the Centralab Compentrol) may also be used. Of course, a

1-megohm volume control can be substituted at some sacrifice of bass perception at low volumes.

The bass and treble controls provide variations as shown in Fig. 1305. These curves have meaning only when considered in connection with the frequency range of the particular type of phonograph pickup used or the radio program listened to, the amplifier, and the speaker. For instance, C1, .0001 μf, in the bass control circuit increases the ratio of high to middle frequencies. If other components give relatively poor high-frequency response, it may be desirable to increase C1 to as much as .001 μf. The position of the bass control ahead of the treble control is no accident. This gives a more realistic bass than if these positions were reversed.

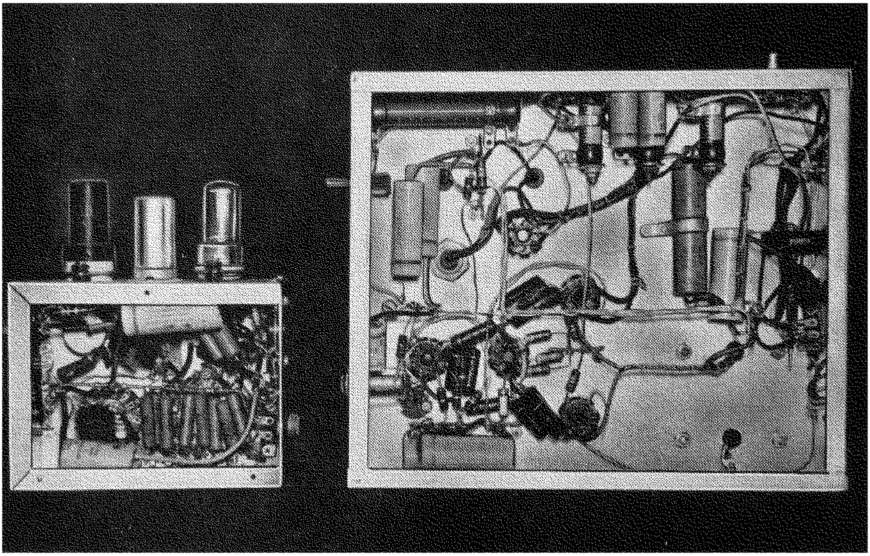


Fig. 1302. Bottom view of the amplifier. A spacious chassis permits a tidy wiring job in the main amplifier.

The 12SC7 and 14AF7 tubes have their filaments connected in series and powered by d.c. from the power supply of the main amplifier. The preamplifier tubes were chosen because of their 150-ma filament current. Except for filament current and voltage, the 12SC7 is identical to the 6SC7, and the 14AF7 is very similar to the 6SN7. Watch polarity when connecting the 100-μfd filament filter capacitor. The plus side of this capacitor goes to the chassis.

Several pairs of resistors and capacitors must be matched for proper balance in the main amplifier. The plate resistor R1 and the cathode load resistor R 2 of the phase inverter must be the same. Plate resistors R3 and R4, grid resistors R5 and R6, grid capacitors C2 and C3, and grid capacitors C4 and C5 should all be matched pairs.

Grid bias of the 6B4-G tubes is a combination of fixed and self bias. The voltage drop across R7 plus the drop across the filaments of the preamplifier tubes gives a fixed bias due to the steady state current through all of the tubes and the bleeder resistor R8. On strong signal peaks when the output tubes draw more plate current, the voltage drop across R7 increases. This increases the bias on the 6B4-G's and prevents overdrive of their grids. Approximately 10 watts undistorted output is available. The additional voltage drop across R9 and R10, although a minor contribution to the 6B4-G's bias, primarily is a balancing device to balance the plate currents of the 6B4-G tubes.

This amplifier has ample volume. However, if more is needed, add

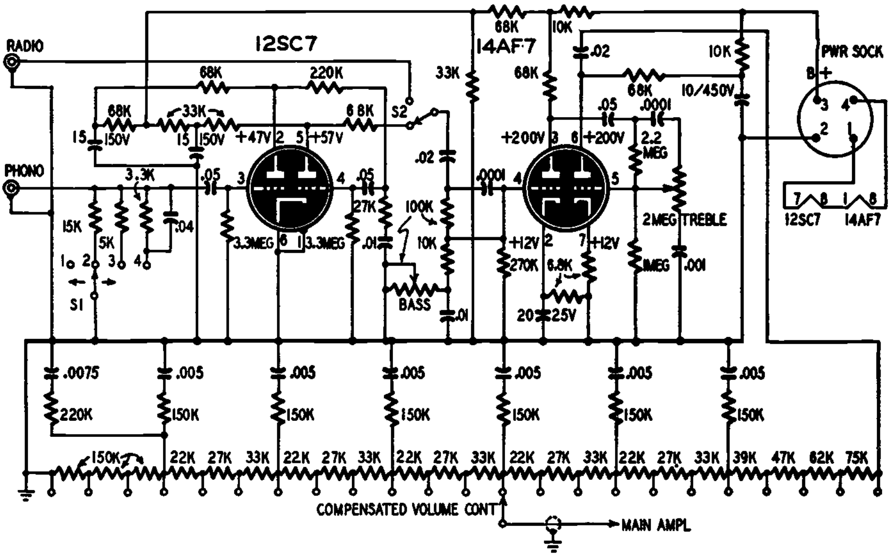


Fig. 1303. Preamplifier circuit. 23-position loudness control permits full-frequency range reproduction regardless of amplifier output level.

the 20-µf, 25-volt cathode-to-ground-capacitor shown in dotted lines on the first section of the first 6SN7.

The power transformer must provide at least 150 ma at 400 to 450 volts to center-tap. Use one with a 200-ma rating. It also must have a 5-volt winding for the rectifier tube and two 6.3-volt windings. These two 6.3-volt sources are required for proper bias and balancing adjustment of the 6B4-G's. One of them can be supplied from a separate filament transformer.

The total plate current supplied to all tubes is about 95 ma.

Bleeder resistor R8 provides additional drain to give a d.c. of 150 ma through the rectifier tube. This 150 ma passes through the filaments of the preamplifier tubes.

One-chassis model

Fig. 1306 and Fig. 1307 show the one-chassis utility model built on the TG-10 keyer chassis and Fig. 1308 gives the schematic. This model requires fewer and less expensive components and only one adjustment,

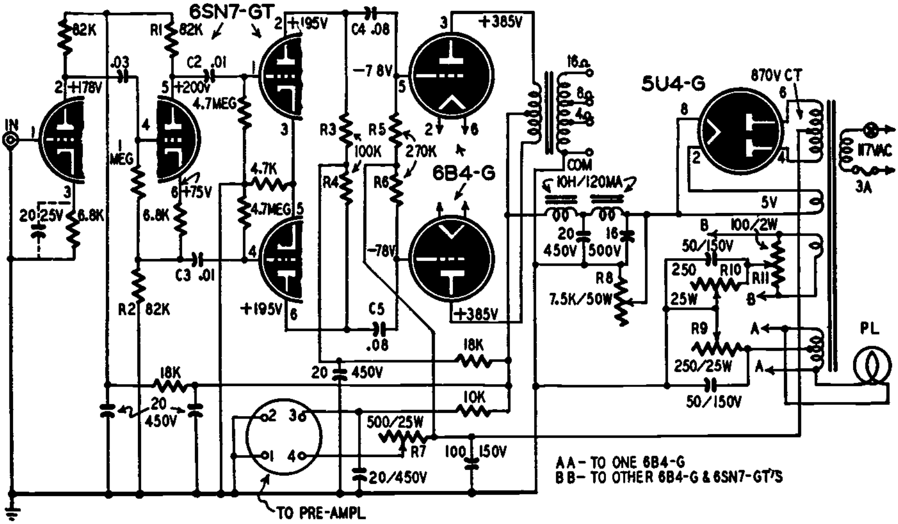


Fig. 1304. Circuit of the amplifier. Full audio output of the unit is 10 watts.

that of R7 to 780 ohms. The 6B4's should be matched for mutual conductance at the time of purchase, if possible. The power transformer should have minimum ratings of 120 ma at 350 volts to center tap. A separate filament transformer can be provided to replace one of the 6.3-volt windings. Do not skimp on the output transformer. Of course, as many of the refinements of the two-chassis model as desired can be incorporated in the one-chassis model.

Construction hints

There is nothing critical about constructing this amplifier. To make sure there is plenty of room in the two-chassis model, a 3 x 5 x 7-inch chassis is advisable for the preamplifier and a 3 x 10 x 12-inch chassis for the amplifier. Do not cable together wires which carry signal currents and do not place components which carry signal voltages on a common terminal strip.

Locate the tube sockets so that coupling capacitors can be placed with short leads from tube terminal to tube terminal. Supply B-plus voltage from a common bus or terminal strip located along one side of the socket layout through the load resistors directly to the socket terminals. Along the other side of the socket layout place a common ground bus to which appropriate connections can be made. Connect this ground bus to the chassis at only one point and locate this point as close to the signal input terminal as possible. In the two-chassis model

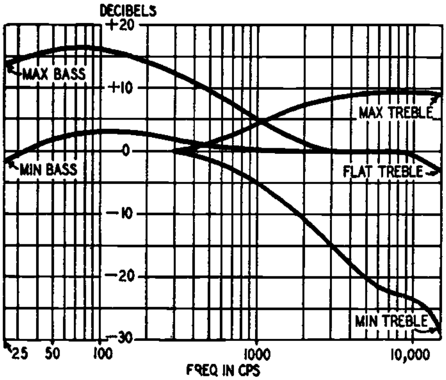


Fig. 1305. The frequency response curves showing the effect of the tone controls.

the one-point grounding procedure should be followed in both preamplifier and amplifier to avoid chassis current effects.

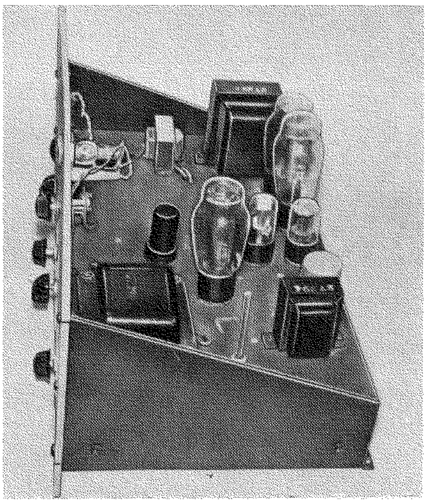


Fig. 1306. Photo of the one-chassis unit.

Resistor R8 in Fig. 1304 gets rather hot. Attach it to the side of the chassis at both ends and bore ventilating holes along the side of the chassis both above and below this resistor.

From the preamplifier to the main amplifier chassis filament ground and signal ground are carried separately and connected together at the main amplifier ground bus.

The 12SC7 and 14AF7 tubes in the preamplifier should be shock-mounted. In the utility model the 6SC7 preamplifier tube should be shock mounted.

In the utility model, a chassis bottom plate helps to cut down direct hum pickup by the compensating network which comprises the input to the 6SC7. It is best to operate the 6SC7 and the 6J5 from a separate filament winding or transformer to reduce hum.

Adjustments

Adjustments must be made in the two-chassis model so that the

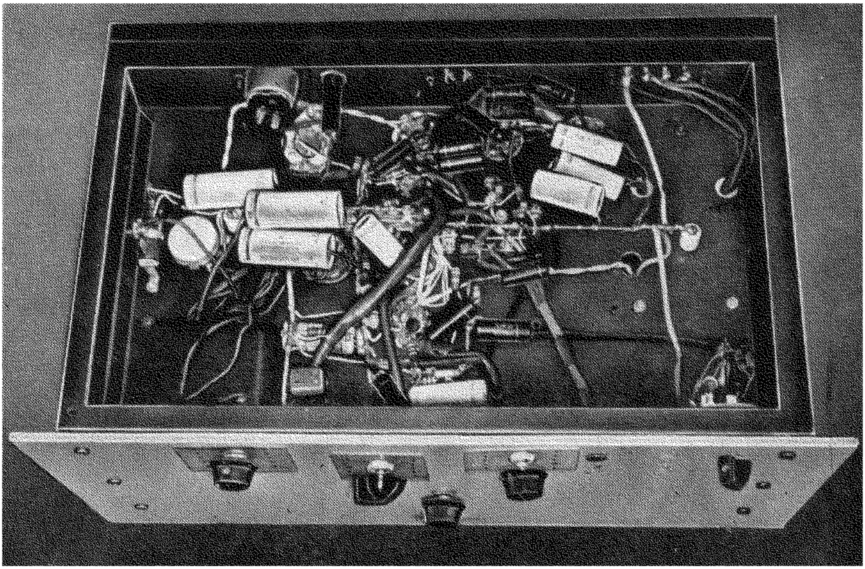


Fig. 1307. Underside of the one-chassis unit. Careful parts placement simplifies wiring, makes neater looking job, simplifies future possible trouble-shooting.

current to each 6B4-G tube is 40 ma and the current to the preamplifier filaments is between 140 and 150 ma.

Proper adjustment of the 7.5 K, 50-watt pot, R8, can be made by inserting a d.c. milliammeter in series with the preamplifier filaments. Adjust R8 for a reading of 150 milliamperes. You will find some interaction between the settings of R7 and R8. Adjust R7 for proper 6B4-G bias and R8 for correct preamplifier filament current. An alternative method is to adjust R7 to 400 ohms and R8 to about 7,000 ohms. Measure the current in the plate circuit of each 6B4-G individually, and by adjusting R9 and R10 balance these two currents at any convenient value between 40 and 50 ma. Then adjust R8 until the voltage between terminals 1 and 4 on the power socket to the preamplifier measures 22 to 24 volts. Next recheck the 6B4 plate current and by means of R9 and R10 adjust to 40 ma each. Again check the filament voltage to the preamplifier and adjust R8 to make this voltage 23 to 24 volts. Repeat the resistor adjustments if necessary until the proper currents and voltage are obtained. Normally R7 is left at 400 ohms.

Materials for Preamplifier

Resistors: 1—3,300, 1—5,000, 2—6,800, 3—10,000, 1—15,000, 5—22,000, 6—27,000, 8—33,000, 1—39,000, 1—47,000, 1—62,000, 6—68,000, 1—75,000, 1—100,000, 9—150,000, 2—220,000, 1—270,000-ohm, 1—1, 1—2.2, 2—3.3-megohm, ½-watt; 1—100,000-ohm, 1—2-megohm potentiometers.

Capacitors: 2—.0001-μf mica; 1—.001, 6—.005, 1—.0075, 2—.01, 2—.02, 1—.04, 3—.05-μf, 400 volt, paper; 1—10-μf, 450-volt, 1—15-μf, 150-volt, 1—20 μf, 25-volt electrolytic.

Miscellaneous: 1—12SC7, 1—14AF7 tubes and sockets; 1—23-position rotary, 1—4 position, single pole, switches; 2—phono jacks; 1—4-prong power socket; chassis, hookup wire, assorted hardware.

Materials for Amplifier

Resistors: 1—4,700, 2—6,800, 1—10,000, 2—18,000, 3—82,000, 2—100,000, 2—270,000-ohms, 1—1, 2—4.7-megohm, ½-watt; 2—250-ohm, 25-watt, 1—500-ohm, 25-watt, 1—7,500-ohm, 50-watt, with adjustable tap; 1—100,000-ohm potentiometer.

Capacitors: 2—.01, 1—.03, 2—.08-μf, 400-volt, paper; 1—16-μf, 500-volt, 5—20-μf, 450-volt, 1—25-μf, 25-volt, 2—50-μf, 150-volt, 1—100-μf, 150-volt, electrolytic.

Miscellaneous: 2—6SN7-GT, 2—6B4-G, 1—5U4-G tubes and sockets; 1—435-0-435-volt, 200 ma power transformer with two 6.3-volt windings and one 5-volt winding; 2—10-h, 120-ma chokes .1—audio output transformer (Acrosound model TO-250 or equivalent, 12.5 watts, 5,000 ohms primary impedance); 1—4-prong power plug; 1—3 amp fuse and fuse holder; 1—phono input jack; 1—s.p.s.t. line switch; chassis, hookup wire, assorted hardware.

Materials for One-Chassis Amplifier

Resistors: 1—3,300, 1—4,700, 1—5,000, 2—6,800, 4—10,000, 1—15,000, 1—27,000, 3—33,000, 2—68,000, 3—82,000, 4—100,000, 1—220,000, 4—270,000-ohm, 1—1-megohm, 1—2.2, 2—3.3, 2—4.7-megohn, ½-watt; 1—1, 1—2-megohm potentiometers; 1—1,000-ohm, 25-watt with adjustable tap.

Capacitors: 2—.0001-μf, mica; 1—.001, 5—.01, 1—.02, 1—.03, 1—.04, 2—.05, 2—.08-μf, 400-volt paper; 1—8, 2—15, 1—50-μf, 150-volt, 5—20-μf, 450-volt, electrolytic.

Miscellaneous: 1—6J5, 2—6SN7-GT, 2—6B4-G, 1—6SC7, 1—5U4-G, tubes and sockets; 1—power transformer, 435-0-435-volt, 200-ma, with two 6.3-volt and one 5-volt winding; 1—output transformer (Acrosound TO-250 or equivalent 12.5 watts, 5,000 ohms primary impedance); 1—10-h, 120-ma choke; 1—s.p.s.t. power switch; 1—s.p.d.t. switch, 1—s.p. 4-position, switch; 1—3-amp fuse and fuse holder; 2-phono jacks; chassis, hookup wire, assorted hardware.

Its value can be changed somewhat if necessary for proper adjustment. As these adjustments must be made with the amplifier on, the secondary of the output transformer must be connected either to a speaker or better to a 10-watt, 4- to 16-ohm resistor during the adjustment to make sure that the transformer does not burn out.

Set the hum-balancing potentiometer R11 for least hum. On the radio input the hum level is more than 75 db below maximum signal.

In the utility model R7 has a different value than in the two-chassis model. Set it at 780 ohms or better, then adjust it to give an average 6B4-G plate current of 40 ma.

In both amplifier circuits, it is important to have equal signal driving voltages on the control grids of the output tubes. Check this by feeding in a steady tone from an audio generator and measuring the signal voltage across the 6B4-G grid return resistors, R5 and R6.

Listening tests

Various versions of this amplifier have been given long-range listening tests using several crystal, General Electric, and Pickering phono-

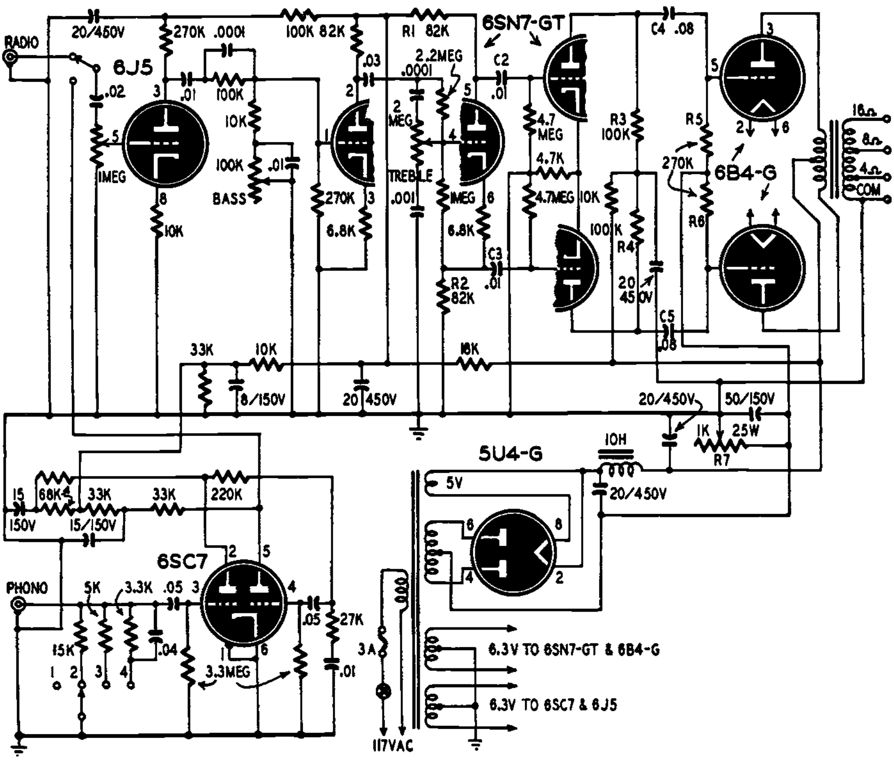


Fig. 1308. Complete circuit of the one-chassis version of the triode amplifier. Tapped output transformer allows choice of different speaker impedances.

graph cartridges, disc recordings from many sources, and AM and FM radio programs. Speakers have included an 8-inch electrodynamic in a conventional console cabinet, a 15-inch Jensen coaxial in both the same cabinet and in a bass reflex enclosure, a single 12-inch electrodynamic in a bass reflex enclosure, and a Klipsch speaker system.

Chapter 14

Miniature-Tube A.F. Amplifier

THE audio amplifier described in this chapter is a miniature, high-fidelity, 4-watt, a.c.-d.c. amplifier that is flat ± 1-1/2 db from 30 cycles to 10 kc, has less than 2% distortion, a gain of 60 db, and a noise level 70 db below 4-watt level. Tubes used are two 50B5's and a 12AU7. The amplifier is constructed on a 4 x 5 x 2-inch aluminum chassis, and over-all measurements are 5-1/2 x 6 x 4 inches. It takes less than half

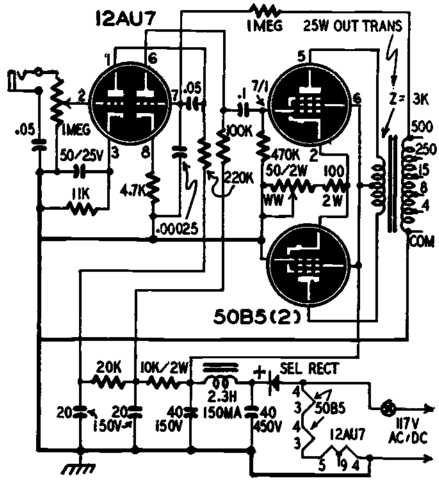


Fig. 1401. *Amplifier schematic. Output power is approximately four watts.*

the space of a conventional amplifier using 50L6's. The push-pull 50B5's are employed in a cathode-driven phase-inversion circuit (see schematic, Fig. 1401), and the two triodes of the 12AU7 are in cascade. This results in 10 times the gain that would be possible if the two triodes were used in the conventional type of inverter circuit.

Although the two triodes in the 12AU7 envelope are identical, the circuit constants are different, being designed for minimum distortion at the levels presented to each.

It is almost impossible to show the actual placement of chassis connections in the circuit schematic, but care in these is most important for low noise level. See Fig. 1402 and Fig. 1403 for layout of components. There are but three chassis connections and they all go to the lugs on the 12AU7 socket. Buses run from here to the metal-can capacitors, the only other chassis connections. (An additional minor one is the 50-ohm bias control for the 50B5's.) The input jack is insulated with fiber washers. One ground point instead of several results

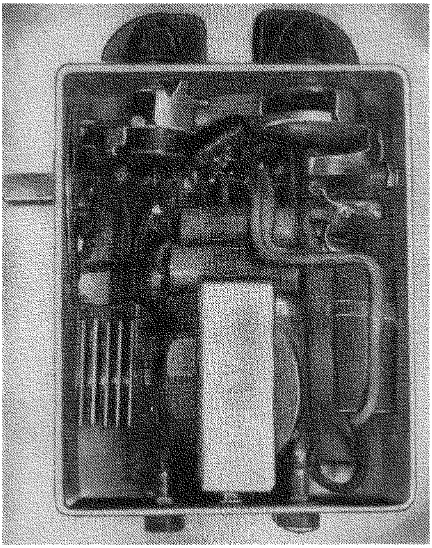


Fig. 1402. Getting components into small space can be done with planning.

in less inductive hum pickup because of elimination of closed low-impedance loops. This feature of the amplifier is a major contribution to its low noise level. The wiring is as compact as possible, and signal leads are close to the chassis and well shielded. Filament leads are twisted tight, the chassis-connected end of the string going to pin 4 of the 12AU7 voltage amplifier.

Many other points of mechanical construction in the amplifier help stability and keep noise down. For example, the lugs of the phono jack are moved close together. A shielded lead covered with spaghetti runs from this jack to the 12AU7 socket, and the inside lead is drawn out through an appropriate hole to make connections at the volume control. Center socket shields are grounded. Whenever possible, socket lugs are soldered together without utilizing wires. The a.c. plug and switch are kept close together and on the opposite side of the chassis from the input jack. Output jacks are mounted on the other end of the small chassis.

It is necessary to mount the filter choke directly underneath the output transformer, but no trouble has been experienced with hum. The cathode bypass capacitor in the first stage does not raise gain, as it would appear, but eliminates hum. Incidentally, a 450-volt capacitor is used instead of a 150-volt unit on the input of the B-plus filter to minimize maintenance, as it is subject to surges. For the same reason a selenium rectifier instead of a vacuum tube is used. The filter capacitors are exposed to a considerable amount of heat, but the Mallory FP units used are designed for this and have proven that they can take it without going bad.

The cathode phase-inversion circuit of the 50B5's permits economical use of tubes. It is also probably the best for maintaining balance

Materials for Miniature-Tube Amplifier

Resistors: 1—100 ohm, 2 watt, 1—4,700 ohm, 1—10,000 ohm, 2 watt, 1—20,000 ohm, 1—100,000 ohm, 1—220,000 ohm, 1—1 megohm; 1—50-ohm, 2-watt potentiometer, 1—1 megohm potentiometer.

Capacitors: 1—.00025 μf, 3—.05 μf, 1—.1 μf, 1—50 μf, 25 volt, 2—20 μf, 150 volt, 1—40 μf, 150 volt, 1—40 μf, 450 volt.

Miscellaneous: 1—selenium rectifier, 150 ma, 1 filter choke, 150 ma, 1 jack, 1 output transformer, 3—miniature tube sockets, 1—12AU7, 2—50B5, 1—chassis, 2—knobs, assorted hardware.

through wide frequency bands. It is not unstable, tricky, or difficult to adjust. The reason for the wideband performance is that no capacitors are involved. The phase-inverting cathode resistor is made equal to the reciprocal of the tubes' transconductance and is adjustable. (The transconductance of 6B5's is in the order of .0075 mho; thus the resistor is adjustable around 133 ohms.) A cathode-coupled phase-inverter has the gain of a single tube, the signal voltage being effectively split and half applied to each side of the push-pull circuit.

This is how it works. The signal from the 12AU7 is coupled to the grid of the upper 50B5 in the usual way. The cathode of the 50B5 is unbypassed. The signal plate current, passing through the cathode resistor to ground, creates a voltage drop between cathode and ground, which changes at an audio rate. The grid of the lower 50B5 is grounded and its cathode is connected to that of the upper tube. Any voltage appearing across the cathode resistor as the result of signal through the upper tube, also appears between cathode and grid of the lower tube.

There are several ways of adjusting the bias of the tubes for balanced inversion. Probably the best and easiest is to note with a pair of headphones (connected to the 500-ohm output) the point at which the hum is balanced out. The adjustment is brought all the way up until a loud hum is heard in the phones and then gradually worked down to the position where the hum just fades out.

An ideal feedback loop should go from output to input. In this amplifier the volume control is in the first stage, and feedback introduced here would change with the setting of the volume control. To

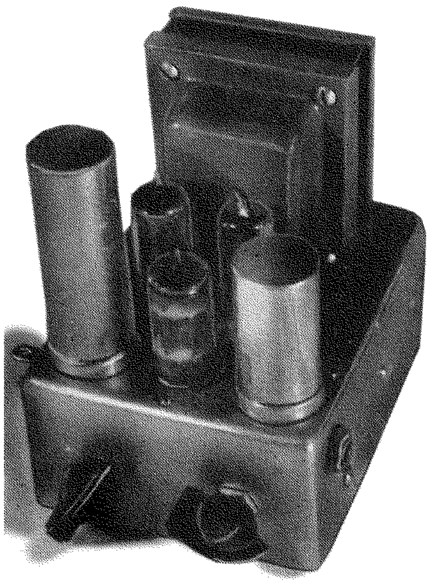


Fig. 1403. *Complete miniature amplifier. Chassis measures only* 4 *x* 5 *x* 2 *inches.*

overcome this difficulty the feedback is applied to the second stage. As the first stage works at a very low level, it contributes negligible distortion. The output transformer and power tubes are the greatest contributors to distortion.

The use of inverse feedback to flatten response calls for a feedback loop with no frequency discrimination. Since no capacitors are used, the feedback loop in our amplifier meets this requirement. The grid resistor of the second 12AU7 triode returns to ground through the secondary of the output transformer.

When the output winding of the transformer is in the feedback loop, there is a frequency at which the phase of feedback is reversed due to the stray inductance of the windings. At this frequency there is oscillation, a.f., ultrasonic, or r.f. While the oscillation may be inaudible, it may overload the amplifier and distort the audible signal. The 1-megohm grid resistor and the 250-μμf capacitor isolate the inductance of the transformer and provide a low-impedance path to ground for r.f. and ultrasonics.

The polarity of the primary and secondary of the output transformer must be correct so that the feedback is negative, not positive. If it is not right on the first try, interchange the connections.

Frequency runs at various levels have been plotted, and there is much improvement in performance when the output transformer is run well under saturation. This is the reason for a 25-watt transformer in a 4-watt amplifier; the larger the transformer, the better the bass response and the lower the distortion.

(Note this warning carefully: the chassis and the capacitor cans in this amplifier are directly connected to one side of the line—and it may be the hot side!) To make the amplifier safe for use, several methods are possible, though none are entirely safe and precaution is still required.

The entire unit may be enclosed in a nonmetallic cabinet so that no metal part can be touched. If the input is "isolated" from the line with a .05-μf capacitor, a phonograph pickup connected to it may be touched with little danger.

The best procedure is probably to bring all points shown connected to chassis to a common negative point, from which the chassis can be entirely isolated. Two-terminal filter capacitors with insulated cans would, of course, be necessary. This method is used in underwriter-approved receivers, most of which also use an .05-μf capacitor between chassis and negative B.

Another method would be to orient the power plug in the wall socket so that the chassis side of the line is the grounded side. The socket and plug can then be so marked that the orientation will be the same in the future. While electrically preferable, this method is psychologically unsound—even a brilliant radioman will forget some day and get burned. If, however, the amplifier is to be connected to a tuner or any other powered device, the polarized-plug trick is the only way to be safe—and it may also be the only way to keep out hum.

Chapter 15

Three-Channel Amplifier

A FLEXIBLE tone-control system is an important part of any amplifier. Most high-quality amplifiers use more or less complex networks of L, C and R. They usually also require one or more supplementary tubes to make up for the loss of gain in the compensating network. This can be avoided by splitting the input stage into three independently controlled channels. In this amplifier, two channels cover only a part of the audio frequency spectrum (bass and treble). The third channel covers the whole spectrum.

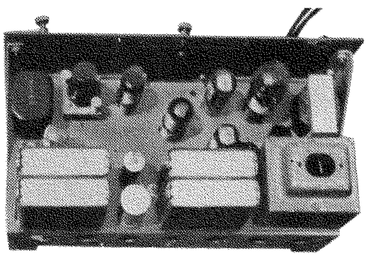


Fig. 1501. Top view of the three-channel amplifier.

There are six separate tone controls. (The photo, Fig. 1501, shows only two, P4 and P6, bass and treble controls. The other four, S1, S2, P2, and P3, were added after the photo was taken.) The circuit is shown in Fig. 1502.

As the amplifier was designed for living room use, 8 watts was judged sufficient output. A pair of 6V6's in class AB1 deliver this at the plate voltage chosen. To lower the output impedance, negative feedback is used from the voice-coil winding of the output transformer over three stages to the 6J5 cathode circuit. The amount of feedback can be changed by varying the 1.2K resistor connected from the undergrounded side of the voice coil to the cathode circuit of the 6J5. A 2K, wire-wound potentiometer can be used as a feedback control. If the feedback causes oscillation, transpose the voice coil leads.

The input of the amplifier is designed for a crystal pickup. R1 is selected to limit signal voltage at point X to 0.5 volt on peaks. From here the signal goes through volume control P1 and then divides into

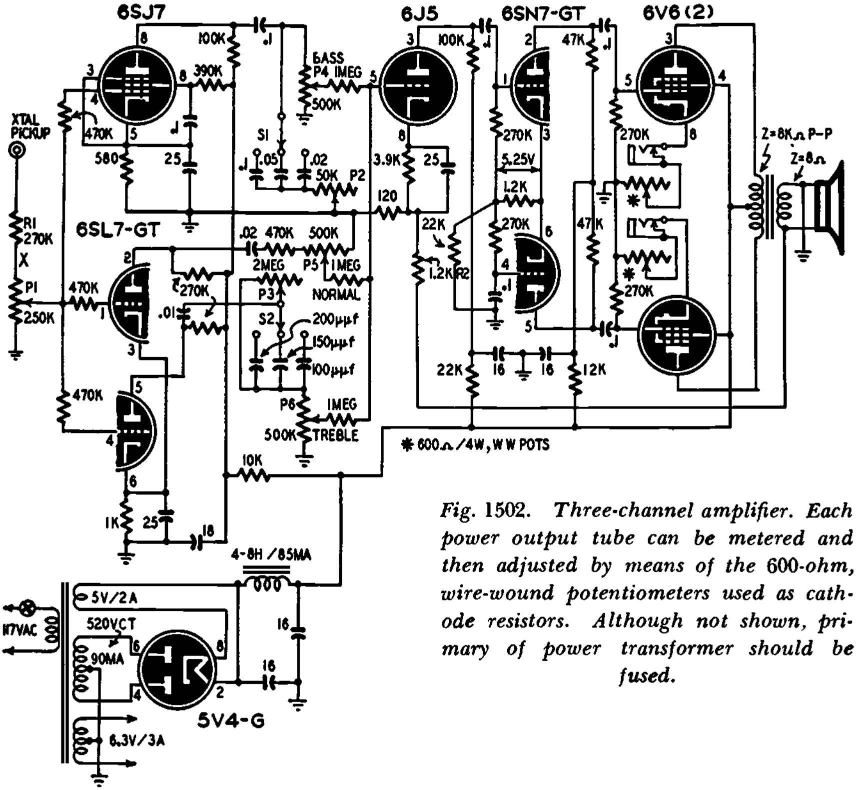


Fig. 1502. *Three-channel amplifier. Each power output tube can be metered and then adjusted by means of the* 600-*ohm, wire-wound potentiometers used as cathode resistors. Although not shown, primary of power transformer should be fused.*

three parts, going to a 6SJ7 and each half of a 6SL7. The output of the 6SJ7 has a low-pass filter in its plate circuit. The crossover frequency of this filter can be adjusted by S1. The amount of bass boost is regulated by P2. The over-all gain of the bass boosting channel is controlled by the potentiometer P4 in the grid circuit of the 6J5.

One-half of the 6SL7 functions as the uncompensated stage which amplifies all frequencies. P5 controls the gain of this channel.

The other half of the 6SL7 is used as the treble channel. This channel has a variable high-pass filter in its plate circuit. S2 selects the frequency at which the channel's response begins to fall. Potentiometer P3 regulates the amount of fall. The over-all gain of the treble channel is controlled by potentiometer P6.

The output of the three separate channels is combined at the grid of the 6J5 amplifier. The 6SN7 which follows the 6J5 functions as the phase inverter using a cathode-coupled circuit. In this circuit, only one of the triode grids is driven. The other triode's grid is at ground potential as far as a.c. is concerned. The signal is fed through the cathode which is connected to the cathode of the first half of the 6SN7. Since both cathodes are well above ground, they vary at an audio rate when a signal is fed to the first triode grid. Because of the

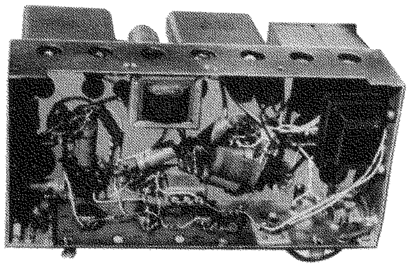


Fig. 1503. The underchassis wiring. The output transformer is at the right.

grounded grid, the second triode's plate has the same phase as the cathode, and differs by 180° from the plate of the first triode.

The 6V6 push-pull output stage is connected in conventional fashion. However, each tube has a separate cathode bias resistor (a 600 ohm, 4-watt potentiometer). A closed-circuit jack in each cathode permits metering cathode current of each 6V6. Adjust the bias to give a cathode current of 35 ma for each tube.

To get the maximum quality from this amplifier, use a high-grade output transformer. The capacitors and resistors on each side of the inverter and push-pull circuits should be accurately matched for best results. Grounds must be grouped by stages, and all groups connected to the chassis at one point. The filament winding on the power transformer was center-tapped and all filament leads were shielded. An underchassis view is shown in Fig. 1503.

Chapter 16

High-Power Amplifier

CHARACTERIZED by reliability, simplicity, flexibility, and ample power, this amplifier has a maximum undistorted (0.5%) output of 30 watts. The input for maximum output is 2.5 millivolts.

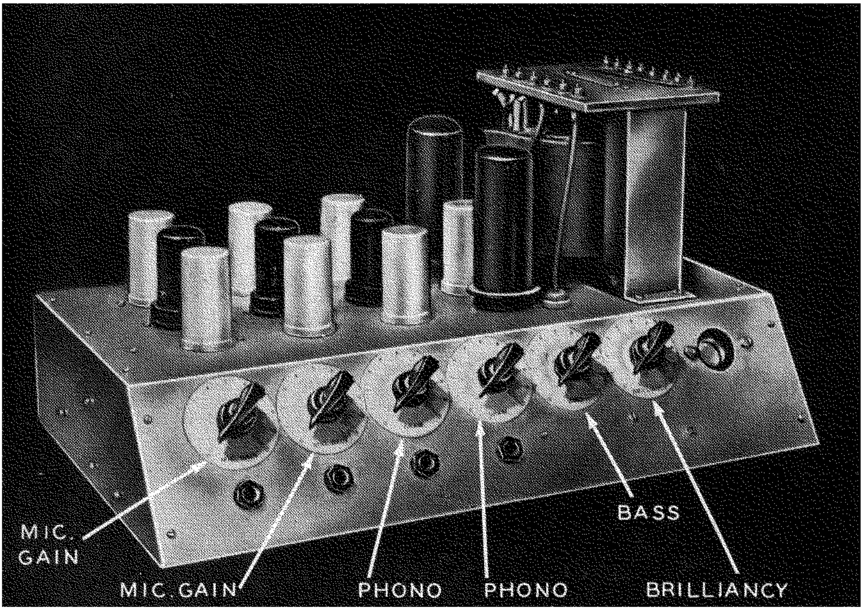


Fig. 1601. *Front view of the high-fidelity amplifier. One of the two phono inputs was later changed to a microphone input. All controls are on the sloping panel.*

Two novel features in the circuit are the tone control stage and the high-gain phase splitter. The input circuit is designed for three dynamic microphones and one crystal pickup, and not two of each as shown in the photograph, Fig. 1601. Simplicity is the keynote and only one microphone transformer is used, a 1:50 Mumetal-shielded type. Mixing is smooth and silent.

Tone control circuits

The outputs of the microphone transformer and pickup are applied in parallel between grid and ground of the tone control tube, a 6SJ7 (see Fig. 1602). The gain of the low-impedance mikes is controlled

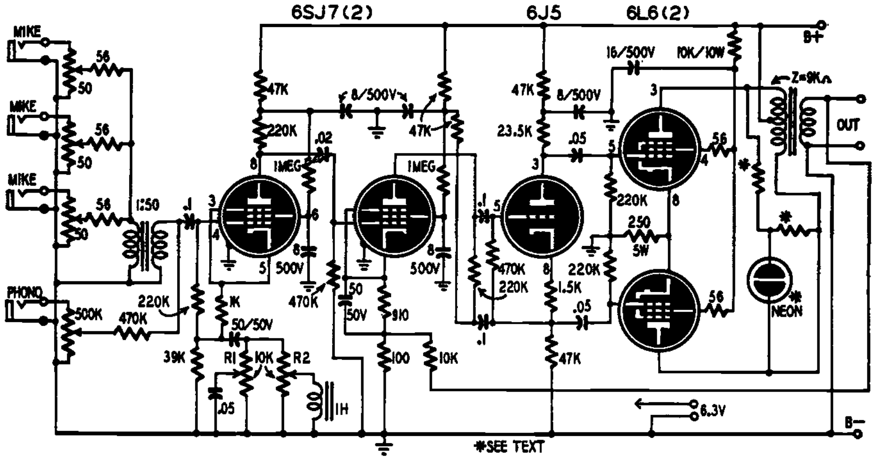


Fig. 1602. Schematic of the amplifier. The second 6SJ7 uses the cathode follower's high input impedance as a.c. plate load.

by 50-ohm pots. A 500,000-ohm pot is needed for the high-impedance crystal pickup. Variable negative current feedback is applied to this tube by the cathode resistors and associated networks.

The correct value of grid bias is obtained by returning the 220,000-ohm grid resistor to a tap in the cathode circuit.

When the moving contacts of the two tone control potentiometers, R1 and R2, are grounded, the impedance between cathode and ground is about 5,400 ohms and is independent of frequency; therefore the negative feedback is also independent of frequency and the gain is constant.

When the moving contact of potentiometer R1 is moved to the other end of its track, the network has an impedance which decreases with rise of frequency—3,500 ohms at 1,000 cycles, and 1,300 ohms at 10,000 cycles (see Fig. 1603). The corresponding decrease in the negative feedback with increasing frequency causes the gain to rise and the control to act as a treble boost. Fig. 1603 also shows the cathode-ground impedance variations with the frequency, with potentiometer R1 at the half-resistance and maximum setting. R1 at minimum indicates a constant impedance throughout the frequency range.

By similar reasoning, potentiometer R2 is a bass boost. The 1-henry choke gives the network an impedance of 3,550 ohms at 500 cycles and 1,310 ohms at 50 cycles. Fig. 1604 shows the cathode-ground impedance variations with frequency, with both half-resistance and maximum

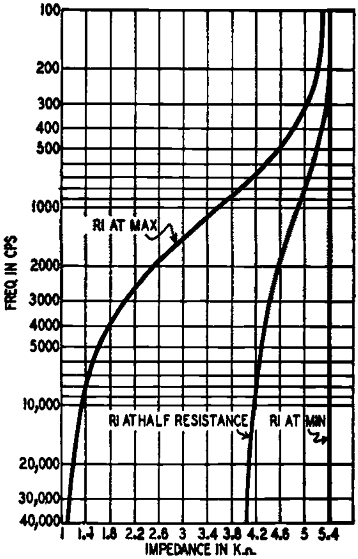


Fig. 1603. Graph at the left shows impedance variation of treble circuit with bass control at minimum.

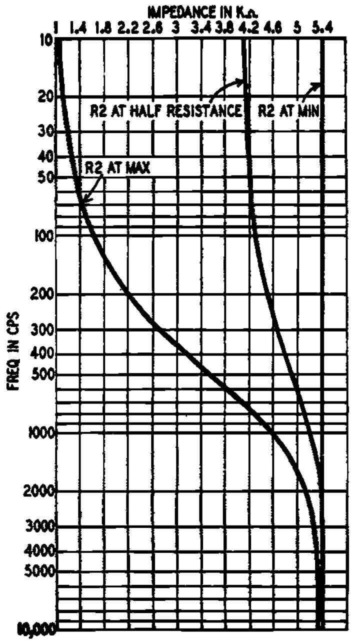


Fig. 1604. The graph at the right shows the impedance variation of the bass circuit with the treble control set at a minimum.

settings of potentiometer R2. With R2 at minimum, the impedance is constant.

The resonant frequency of the choke and capacitor is 723 cycles; but there is no peak in the response curve at this frequency, even with both controls at maximum, because the tuned circuit is very heavily damped by the parallel resistances, R1 and R2. This tone control circuit, although simple, is extremely satisfactory. The table shows how it increases bass and treble response.

A novel phase splitter

The next two stages are considered together. The first is a 6SJ7 operated so the stage gain approaches the amplification factor of the tube. This is achieved by making the plate load of the tube the extremely high input impedance of a cathode-follower phase splitter.

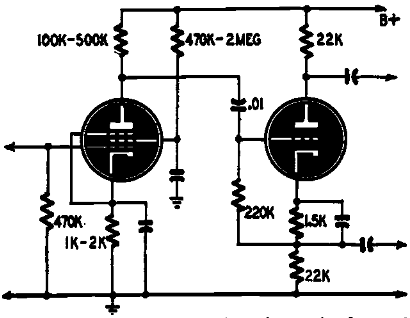


Fig. 1605. Conventional cathode follower phase splitter preceded by a pentode.

The operation is best understood by developing the circuit from a conventional cathode-follower phase splitter preceded by a pentode amplifier whose gain is determined by the values of the plate load resistance and the B-supply.

The input impedance of such a cathode-follower is approximately 10 times the impedance between grid and cathode. In the circuit of

RESPONSE TABLE

Frequency (cycles)	Min. Treble Min. Bass (response) (db)	Max. Treble Min. Bass (db)	Min. Treble Max. Bass (db)
40	−0.5	0	+10.2
100	0	+0.1	+8.6
200	0	+0.3	+6.6
400	0	+0.8	+4.2
1,000	0	+3.2	+1.3
2,000	0	+5.7	+0.4
4,000	0	+8.0	+0.1
10,000	0	+9.9	0
15,000	+1.0	+11.3	+1.0
20,000	0	+10.6	0
40,000	−0.7	+10.3	−0.7

Fig. 1605 this is approximately 2.5 megohms, so that the input impedance of the phase-splitter does not affect the gain of the pentode. However, the maximum value of the plate resistance consistent with a reasonable plate voltage is about 500,000 ohms. This gives a maximum gain of 250 with a 6SJ7 and a plate supply of 300 volts. The gain of the phase splitter being about 0.9, the overall gain is 225.

The phase splitter of Fig. 1605 may be redrawn as in Fig. 1606, where C1 and C2 have negligible reactance at the lowest working frequency. The grid-cathode impedance is now 150.000 ohms (R1 being in parallel with the grid resistor). so that the input impedance of this

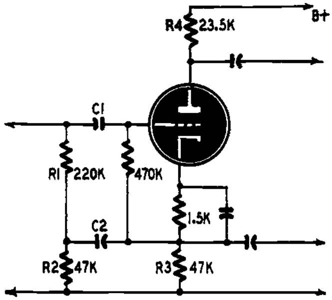


Fig. 1606. *Re-arrangement of the phase splitter of Fig.* 1605 *for higher gain.*

cathode-follower is approximately 10 times 150,000 ohms or 1.5 megohms. If the grid end of R1 is connected to the anode of the preceding pentode and the ground end of R2 is connected by the B-supply, the a.c. conditions of the phase splitter are unchanged and the pentode sees the input impedance of 1.5 megohms as its plate load. The over-all

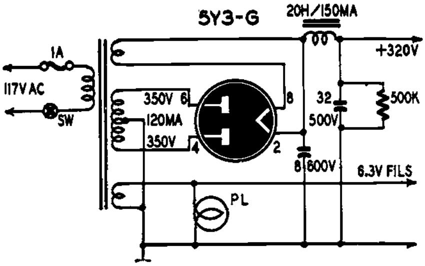


Fig. 1607. *Power supply for the amplifier.*

gain is thus increased to about 1,000. The inherent unbalance is negligible if R2 = R3 = 2 × R4. Comparison of Figs. 1602 and 1606 shows that this is the arrangement used. The constructor may use several methods of obtaining the correct resistance. Possibly the easiest is to use two 47,000-ohm resistors in parallel for R4.

Output stage and B-supply

The remainder of the circuit is conventional. Two small resistors (56 ohms) are included in the screen feeds of the 6L6 output tubes for parasitic suppression and to limit screen dissipation. Considerable negative feedback voltage (about 20 db) is introduced into the cathode of the second 6SJ7 from the secondary of the output transformer.

Extensive decoupling is used throughout to prevent positive feedback. Because of its extremely high gain, the amplifier is very sensitive to noise and microphonics in the first tube. The latter noise is eliminated by rubber mounting the tube socket.

The power pack is conventional as seen from the circuit in Fig. 1607. A photo of the power supply is shown in Fig. 1608. The power transformer supplies 350-0-350 volts to a 5Y3-G full-wave rectifier, and a 500,000-ohm bleeder is connected across the B-supply to discharge the electrolytic capacitor after switching off. A 1-watt resistor is adequate.

As shown in the circuit, Fig. 1602, an output indicator is included

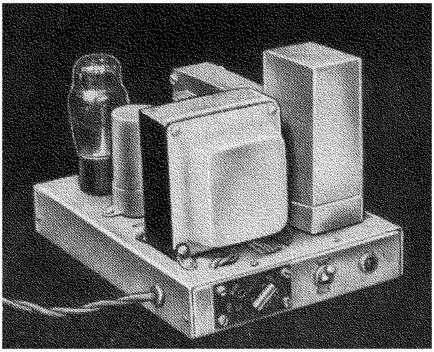


Fig. 1608. The power supply is a separate unit.

in the amplifier. This consists of a neon lamp, with limiting resistors, connected across the primary of the output transformer. The 0.5-watt lamp, a common type of indicator for British standard 230-volt lines, was uncapped and fitted into the octal base of an old burnt-out tube with Plastic Wood. The values of the limiting resistors were adjusted by trial and error until the indicator is fully lit at 30 watts output.

Construction hints

For those readers who contemplate building a similar amplifier, the following constructional notes may be of interest. The chassis of both units are of .064-inch aluminum, and the two chassis measure 15 x 7 x 3 inches and 8 x 6-1/2 x 1-1/2 inches, respectively. The amplifier control panel is set at an angle and the six controls are grouped in a horizontal row, the four input jacks being placed below their respective mixer potentiometers. This, together with a symmetrical layout of the tubes and electrolytic capacitors, gives a neat appearance to the job. An underchassis view is shown in Fig. 1609. Power is carried to the amplifier by a heavy-duty four-wire cable terminated in a female four-point connector. The speaker output is taken from two insulated binding posts at the rear of the chassis.

While the general layout is not very critical, some precautions must be taken to keep the hum at the lowest possible level because of the amplifier's high gain. One good way to keep hum down is to make all the common ground connections to a single bus bar, then ground the

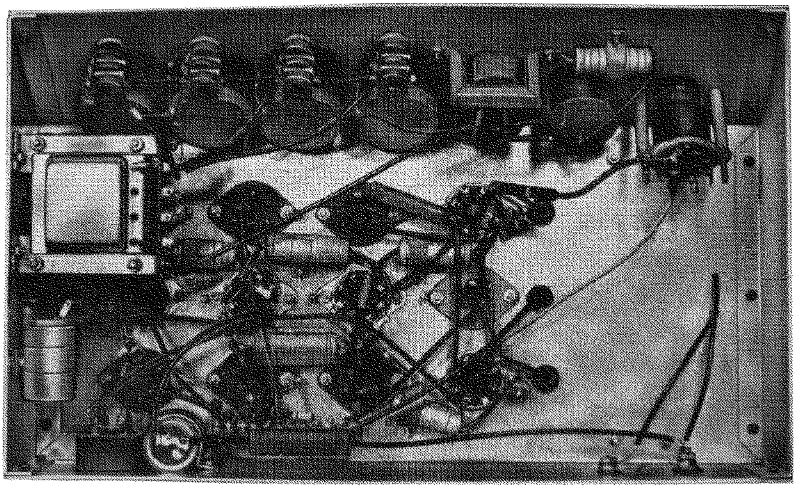


Fig. 1609. *The symmetrical layout under the chassis gives the job a very neat appearance.*

bus bar to the chassis at one point only. This point should be at the input stages or where the signal level is lowest. The heater circuits should be wired with a pair of twisted wires. Do not ground one side of the heaters in the amplifier chassis. The power supply schematic shows one side of the 6.3-volt winding grounded. It is better to ground the centertap of this winding if there is one.

The photographs show the placement of parts, which is not very critical. All resistors and capacitors have 20% tolerance. The 250-ohm, 5-watt common bias resistor of the 6L6 output tubes largely compensates for any slight mismatch of the resistors in the phase splitter circuit. The 23,500-ohm resistor may be 22,000- and 1,500-ohm units in series. The two 50-µf electrolytic capacitors are mounted with their cans isolated from ground. All coupling, decoupling, and smoothing capacitors are rated at least 500 volts, as the B-supply reaches this value before the output tubes are fully conducting.

Chapter 17

Cathode-Coupled Amplifier

ALTHOUGH this amplifier departs from the conventional in many respects, it has no fancy hard-to-adjust features. A frequency run with a Hewlett-Packard oscillator and a G-E VU panel shows it to be flat within 1-1/2 db, 20 to 20,000 cycles. The output transformer governs the frequency response, and that depends on what the pocketbook will stand. Fig. 1701 is a top-side photo and Fig. 1702 is the schematic of the amplifier.

Except for the output tubes, all stages use miniatures. The phase inverter stage—a 6J6 duo-triode—is cathode-coupled; so you don't have to juggle resistors. The only transformer in the amplifier is in the output stage. The final tubes are driven by a pair of 6C4's used as cathode followers for low driving impedance and as a good way to adjust the balance of the final stages. The 6AG5's are triode-connected to make the whole amplifier triodes from input to output! That should satisfy even the most rabid "triodes-are-best" boys.

The first 6J6 is a preamplifier to drive the phase inverter and also an electronic bass-boost stage. The mathematics of the bass-boost stage get rather complicated, but the values shown make it perform well.

The second 6J6 dual triode is the cathode-coupled phase inverter. The grid of one section is fed the audio signal, and the other grid is grounded. The cathodes, being tied together, work in unison. Both plates also vary as the input signal varies, but the signal outputs are 180° out of phase. It's a neat scheme that works nicely. Since the 6J6 is a high-mu triode, good stability and low hum level are important. The phase inverter stage must be shielded with a miniature spring-loaded tube shield to minimize microphonics which 6J6's often develop. The a.c. circuits should be carefully laid out to keep hum pickup low. With these precautions the 6J6 performs well. This type of phase inverter shows no aging effects, as is often the case with other inverter circuits.

The next stage uses two 6AG5's, triode-connected, to drive the 6C4 cathode followers. The fixed bias for the final tubes is fed through the cathode loads of 6C4's; and, by varying a portion of this voltage on the 6C4 grids, the bias can be adjusted so the signal in the final output tubes is balanced.

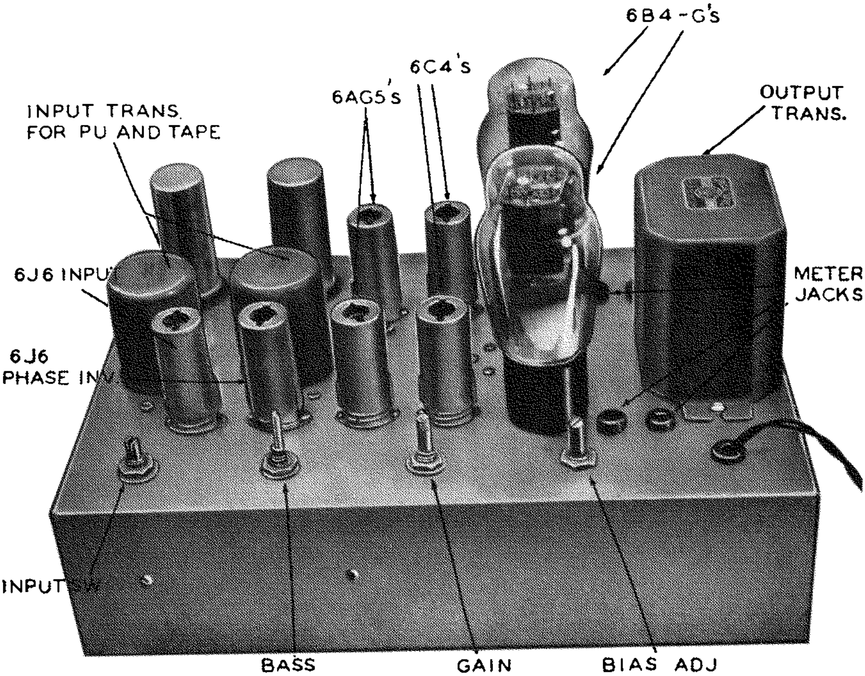


Fig. 1701. Photo of the amplifier. The two input transformers on the left are for a tape recorder and low-impedance pickup. They do not appear in the schematic diagram.

In the final stage either 6B4-G's or 6A5-G's can be used. The 6A5-G is preferable because the cathode-type construction further reduces hum.

Cathode bias can be used, but the fixed bias gives increased output and keeps the bias voltage constant under all plate conditions. This bias voltage is stabilized by an 0D3 regulator tube.

For an output transformer you can take your choice. If you want the ultimate in performance, use one of the best quality transformers, so long as it loads the tubes with 3,000 ohms.

The physical layout of the amplifier can take almost any form, provided the usual wiring precautions are observed. Run the a.c. wiring close to the chassis and so that no ground loops are formed. A common-ground bus is better than the usual chassis ground.

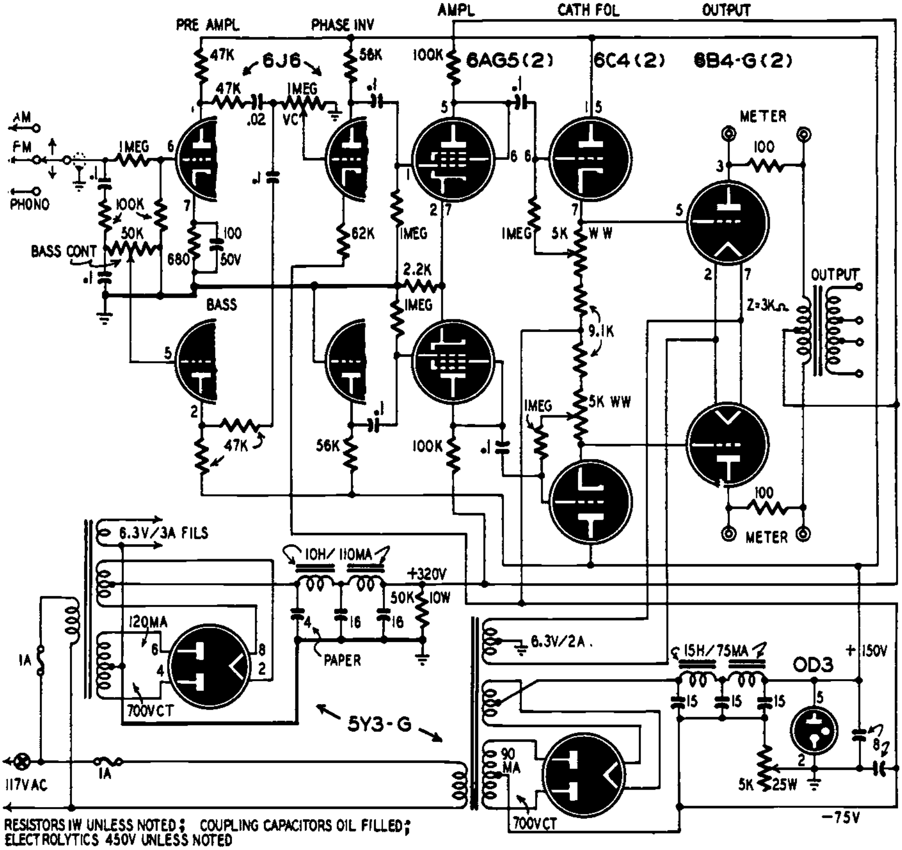


Fig. 1702. *The schematic of the all-triode amplifier. Input operates at low level and has high gain.*

While separate chassis were used for the amplifier and for the two power supplies, both could be on the same chassis if the power transformers are mounted away from the input stages.

Materials for Amplifier

Resistors: 2—100, 1—680, 1—2,200, 2—9,100, 4—47,000, 2—56,000, 1—62,000, 4—100,000 ohm, 5—1 megohm, 1 watt; 2—5,000, 1—50,000 ohm, 1—1 megohm potentiometers; 1—5,000 ohm, 25 watt; 1—50,000 ohm, 50 watt.

Capacitors: 2—0.1 μf, 400 volt, paper; 1—.02, 5—0.1 μf oil filled; 1—4 μf, 600 volt, paper; 2—8, 5—16 μf, 450 volt, 1—100 μf, 50 volt, electrolytic.

Miscellaneous: 1—350-0-350 volt, 90 ma, 1—350-0-350 volt, 120-ma power transformers with filament windings; 2—10 h, 110 ma, 2—15 h, 75 ma chokes; 1—output transformer, 3,000 ohm primary; 2—6J6, 2—6AG5, 2—6B4-G or 6A5-G, 2—5Y3-GT, 1—OD3 tubes with sockets; 2—1-amp fuses and holders; chassis, hookup wire, switches, assorted hardware.

Chapter 18

Distortion Meters

ONE of the most important considerations in appraising amplifier performance is how faithfully the system transmits audio impulses. Only distortion measurements can establish quantitatively the degree of fidelity.

Distortion tests enable the designer to check the effects of circuit changes on reproduction. They are invaluable also to the service technician in determining the effectiveness of repairs and adjustments. In routine maintenance work, the sudden appearance of small amounts of distortion, ordinarily not detectable by ear, usually points to the start of trouble in some component.

In the broadcast field distortion meters are a *must* by FCC rule. Serious audio enthusiasts and technicians who build and service audio equipment are turning to the distortion meter for specific and accurate checks on amplifier performance.

Testing methods

The first distortion meter was a milliammeter in the plate circuit of an amplifier tube. Theoretically, there should be no fluctuation in the *average* plate current of a class A amplifier. At times, however, the meter needle will kick up or down from its steady-state value. Kicks of more than 10 percent of the steady-state reading indicate unsymmetrical amplification, with resulting distortion. The plate-current distortion meter indicates only that distortion is present; the remedy is still a problem.

The principle of modern distortion-measuring technique is simple. A signal from an audio oscillator is applied to the amplifier under test. The output of the amplifier is fed to the distortion meter. This output contains the fundamental frequency of the oscillator, plus any harmonic distortion (multiple or submultiple frequencies) introduced by the amplifier. The output voltage is adjusted to a reference mark on the v.t.v.m. associated with the distortion-measuring equipment. This is called the "calibrate" step. Next, the fundamental frequency is eliminated, leaving only the harmonics. The combined amplitude of these harmonics is then read directly on the v.t.v.m. as a percentage of the total output.

In measurements of this type, perfect impedance matching between all components is imperative to prevent reflections that affect the accuracy.

The heart of the distortion meter is the means used to eliminate the fundamental frequency and leave all the harmonics for measurement. One method uses a bridge circuit. The bridge is highly effective, but requires costly precision components and very careful balancing adjustments. Another method is by means of a T or bridged-T

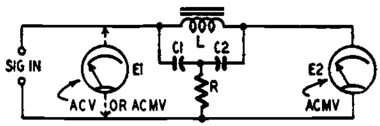


Fig. 1801. Basic distortion meter circuit. It is the familiar bridged-T.

filter network. This also requires balancing, and a switching arrangement for each frequency band, since a given set of high-Q filter components are effective only over very narrow frequency limits.

Of the several methods of measuring audio distortion, the fastest, as well as the one most favored in service and maintenance, is that of checking *total distortion*. The technician usually is not as much interested in the strength of separate distortion components (for example, second harmonic, third harmonic, fourth harmonic, etc.) as he is in the answer to his question, "How *much* distortion is present?" Furthermore, there seldom is time to check individual harmonics and to calculate the square root of the sum of their squares when distortion must be checked after each of many amplifier adjustments.

A widely-used basic distortion-measuring circuit is shown in Fig. 1801. This is the bridged-T network which is the foundation of one of the distortion meters described in this chapter. The network components (L, C1, C2, and R) are chosen to provide a sharp null (zero transmission) at the test frequency. R is made variable and is preset closely for sharp null. When the Q of the circuit is kept reasonably high, the fundamental test frequency is eliminated completely and only the total harmonic voltage (E2) appears across the a.c. millivoltmeter. In use, the amplifier or audio network to be tested is supplied with a signal from an audio oscillator or signal generator having very low distortion. The output of the amplifier then is connected to the input terminals of the network. The meter is switched first across the network input terminals, as shown by the dotted lines, to read the amplifier output voltage (E1). This voltage contains fundamental and harmonics. The meter finally is switched to the output of the bridged-T network, and the small voltage (E2) due to harmonics is read. The

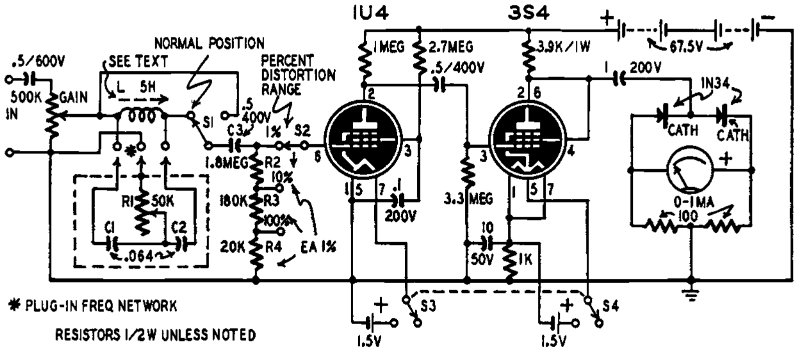


Fig. 1802. Distortion meter which combines sensitivity, low hum, complete suppression of fundamental and ability to operate on several audio frequencies.

distortion then is the ratio of E2 and E1 and is expressed in percentage as 100 (E2/E1). If the input voltage, E1, always can be set to a predetermined level for reference, the millivoltmeter can be made direct reading in distortion percentage, and no calculations will be required.

Complete instrument

Many amateurs and professionals have employed this method of distortion measurement, using homemade equipment. However, there are several difficulties common to almost all setups: (1) the a.c. meter must be capable of checking very small voltages at the output of the bridged-T network. These often are millivoltages which cannot be read on the scales of ordinary v.t. voltmeters. For example, 1 volt may be obtained from the amplifier under test and applied to the input terminals of the distortion-checking circuit. In order to measure 1% distortion, the meter then must show 10 millovolts (0.01 volt) when connected to the bridged-T output. (2) Coil L (Fig. 1801) must have a higher Q than customarily is obtainable with the power-supply filter chokes often used in the circuit by experimenters. If the Q of the bridged-T circuit is not high, harmonics will be attenuated and the meter will not give a true indication of distortion. (3) Usually, no provision is made for easily changing the operating frequency to a new value. (4) Considerable inaccuracy can occur from hum generated by the power supply and picked up by the bridged-T choke.

The distortion meter shown in Fig. 1802 removes these shortcomings: (1) A sensitive electronic millivoltmeter circuit has been provided. This circuit requires no zero adjustment. The indicating

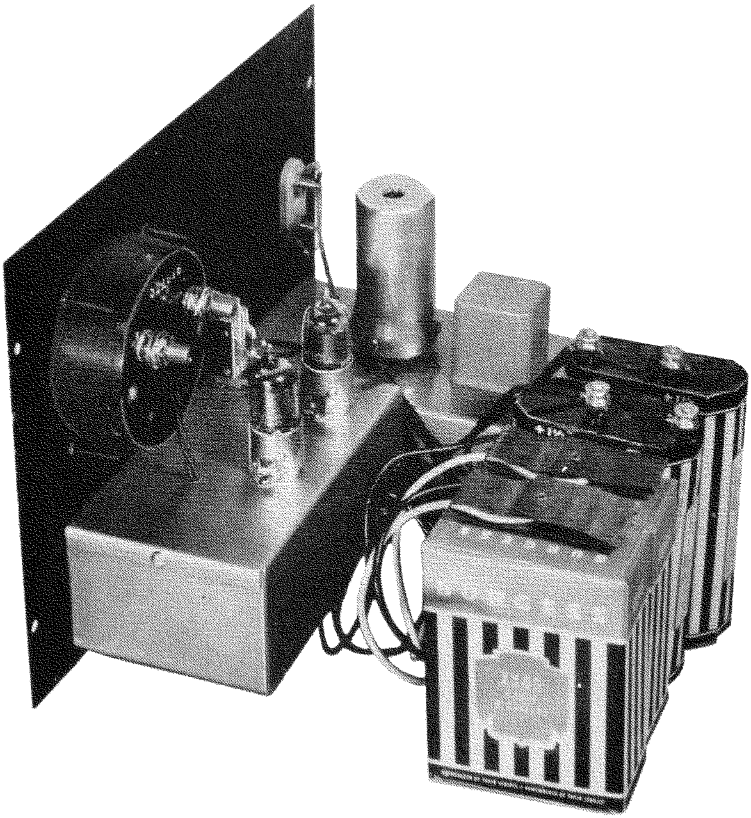


Fig. 1803. *Distortion meter. Audio inductor and plug-in tuning unit are seen at rear.*

meter is a comparatively inexpensive 0-1 d.c. milliammeter (reading linearly direct in distortion percentage—1, 10, and 100% full scale), and only 10 millivolts at the output of the bridged-T network is required for full-scale deflection when the range switch (S2 in Fig. 1802) is in its 1% position. (2) Hum has been eliminated by powering the voltmeter circuit with small, self-contained A- and B-batteries. Since the drain is low, long battery life may be expected. Battery operation also provides complete isolation. (3) Coil L is a special audio inductor. (U.T.C. type VI-C15) which has good Q and adjustable inductance. (4) The test frequency may be changed at will with plug-in frequency units C1, C2, R1. The values given for C1, C2, and R1 in Fig. 1802 are for a test frequency of 400 cycles, since this frequency is supplied by most AM signal generators. Table I gives capacitance and resistance

Table I

Plug-In Bridged-T Network Components

Test Frequency (cycles)	Capacitors C1, C2 (μf each)	Potentiometer R1 (ohms)
50	4	10,000
100	1	10,000
200	0.250	50,000
300	0.11	50,000
400	0.064	50,000
500	0.04	100,000
600	0.028	100,000
700	0.020	100,000
800	0.016	100,000
900	0.0125	100,000
1000	0.010	500,000
2000	0.0025	500,000
3000	0.0011	500.000
4000	0.00064	2 meg.
5000	0.00040	2 meg.

values for other common test frequencies between 50 and 5,000 cycles. There is little point in checking beyond 5,000 cycles, since the harmonics of higher frequencies lie out of the range of most hearing. Usually, three test frequencies, selected in the low, middle, and high portions of the audio spectrum (e.g. 50, 1,000, and 5,000 cycles), will give a satisfactory practical picture of amplifier behavior. It is convenient, for compactness and simplicity, to use plug-in frequency units. However, you can incorporate a switching "tuner" and enclose all components for 10 or more frequencies within the instrument case.

The input gain control allows the meter to be set to a reference level (such as 100%) when the meter is switched to read input voltage. Switch S1 is a spring-return switch resting normally in the position shown, to connect the meter across the output of the bridged-T network. For initial adjustment, this switch is thrown to the right to connect the meter to the input.

The range switch, S2, gives the indicating meter three ranges—0–100%, 0–10%, and 0–1%. At the 100% setting of the switch, the meter has a full-scale deflection of 1 volt, at the 10% setting 0.1 volt, and at the 1% setting 0.01 volt.

The millivoltmeter circuit is simple, employing a high-gain 1U4 pentode resistance coupled to a triode-connected 3S4 driver amplifier for the meter circuit. The full-wave meter circuit consists of a crystal bridge employing two 1N34 germanium diodes and two 100-ohm resistors. If a pair of matched 1N34's aren't available, use a 1N35 duo-diode. The 100-ohm resistors must be matched within at least 1%.

Since the 3S4 must be cathode-biased, the A-battery must be operated above ground. That is the reason for the separate A-battery for

Fig. 1804. A front view of the distortion meter.

the output stage. There is no B-battery drain when the tube filaments are not lighted, so no switching is needed in the high-voltage circuit.

Construction details

The photographs, Fig. 1803 and Fig. 1804, show construction details of the distortion meter. The instrument is housed completely in a 10 x 7 x 8 inch standard metal cabinet. Tuning and millivolt-amplifier units are mounted on separate small box-type chassis. The chassis holding the amplifier measures 5-1/4 x 3 x 2 inches. The tuning unit chassis is 5-1/2 x 3 x 1-1/4 inches. Both of these chassis are enclosed on all sides and accordingly provide good shielding.

The small, shielded inductor L may be seen on the rear of the farther chassis. Directly in front of the inductor is the plug-in can containing C1, C2, and potentiometer R1. The latter is provided with a slotted shaft for adjustment through the hole seen in the top of the plug-in can. The plug-in foundation is a Millen 74001 assembly which has an octal base. The coil form is removed from this unit and the two capacitors and potentiometer are installed in its place. In order to fit into the small can, Mallory *Midgetrol* potentiometers (15/16 inch diameter) were used. If you are compelled to use larger potentiometers, the larger Millen 74400 plug-in cans may be employed. The latter are rectangular cans and, like the type 74001, are provided with octal bases for plugging into a standard 8-pin tube socket.

The A- and B-batteries fit snugly into the cabinet just to the rear of the amplifier chassis in approximately the position they are shown in Fig. 1803. They stand vertically.

The leads from the input binding-post terminals are shielded with braid.

Keep all leads in the meter amplifier as short as possible and use point-to-point wiring. Resistors R2, R3, and R4 must be of close tolerance, varying not more than 1% from specified values. Aside from these points, no special precautions are necessary in construction of the amplifier. In the plug-in units, capacitors C1 and C2 must be matched carefully and should be close to specified values (see Table I) as can be obtained. A glance at the list shows that a number of the capacitances are not stock values, but must be made up with suitable units connected in parallel (e.g., 0.028 μf required for the test frequency of 600 cycles would be made by paralleling 0.02 and 0.008). These capacitors must be high-grade, to insure high Q in the bridged-T network. By using metallized tubulars for the high capacitances, smallest physical size may be secured.

Initial adjustment

If the instrument has been wired correctly and good components used throughout, the voltmeter section will require no adjustment whatever. However, if desired, this part of the instrument circuit may be checked for voltage calibration and linearity. Remove temporarily the lead between C3 and S1 and feed a series of accurately known calibration voltages between C3 and ground, checking the corresponding meter readings. A 1,000-cycle source, such as an audio oscillator, is recommended. In some instances, higher sensitivity might be obtained —full-scale deflection with less than 10 millivolts input with switch S2 at 1% position. However, the absolute voltage level is unimportant in this application. The important thing is that the voltage, whatever its level, be divided *exactly* by 10 and 100 by successive settings of switch S2.

Next, the frequency units must be adjusted to the corresponding operating frequencies. Here is one example, that of the 400-cycle unit. (1) Switch on the distortion meter and allow about 5 minutes warmup time. (2) Set an audio oscillator to 400 cycles and connect its output to the distortion-meter input terminals. (3) Plug the 400-cycle frequency unit into the distortion meter. (4) Set switch S2 to its 100% position and advance the 500,000-ohm potentiometer gain control until a healthy meter deflection is obtained. (5) Adjust potentiometer R_1 in the frequency unit for lowest obtainable meter reading (complete null). (6) With an 8–32 Allen wrench, adjust the inductance set-screw in inductor L for further improvement of this null. (7) Do not touch the setting of R1 at any time afterward unless a routine recalibration is made. Also, do not retouch the setting of the inductance screw in the inductor L. (8) Successively plug in each frequency unit and adjust it to its particular frequency by adjustment of its potentiometer R1 only.

Operating the meter

Checking oscillator distortion: After warming up both the oscillator and distortion meter for at least 5 minutes, connect the oscillator to the distortion meter and set the oscillator output to the desired level. (1) Set switch S2 to its 100% position. (2) Plug-in a frequency unit for the desired test frequency and set the oscillator dial to that frequency. (3) Set switch S1 to its right-hand (input) position and adjust the gain control for full-scale meter deflection. (4) Return switch S1 to its normal (output) position. (5) Set switch S2 successively to lower ranges until an accurately readable meter deflection is obtained. This deflection, together with the setting of switch S2, will indicate the oscillator distortion percentage directly. (6) Rock the oscillator dial back and forth slightly for an improvement in the meter dip. (7) Repeat the procedure at several settings of the oscillator output control, since oscillator distortion often varies with output.

Checking amplifier distortion: Measuring the distortion of an amplifier is similar to the procedure just given for checking an oscillator. There is an important preliminary point, however, and that is to check carefully the distortion of the oscillator which is to be used to supply a test signal to the amplifier. The oscillator distortion figure then must be subtracted from any distortion figure obtained for the amplifier. When checking a complete amplifier system, connect the distortion meter across the loudspeaker voice coil (if you can stand the noise), since the speaker is the normal load of the amplifier. If quietness is a necessity, however, the voice coil may be replaced temporarily with a resistor having the same ohmic value and rated at twice the power output of the amplifier.

To check the amplifier: (1) Connect a low-distortion audio oscillator (whose distortion has been checked and recorded at each intended test frequency) to the amplifier input. (2) Connect the amplifier output to the distortion meter. (3) Allow the oscillator, amplifier, and distortion meter to warm up. (4) Plug-in a distortion meter tuning unit for the first test frequency. (5) Set the oscillator to the same frequency. (6) Set switch S2 to its 100% position. (7) Set switch S1 to its right-hand (input) position, set the amplifier gain control to the desired test point, and advance the gain control of the distortion meter for full-scale meter deflection. (8) Return S1 to its normal (output) position and change the setting of switch S2 for a readable meter deflection. (9) Rock the oscillator dial to deepen the null. (10) Read the distortion percentage from the meter deflection and the setting of range switch S2. (11) Subtract from this figure the distortion of the audio oscillator, previously determined. (12) Repeat the procedure at as many test frequencies and at as many settings of the amplifier gain control as desired.

Special note regarding low test voltages: When the amplifier or oscillator under test delivers an output voltage of 1 or higher, the

Materials for Meter (400 Cycles)

Resistors: 2—100 ohms, ½ watt (matched within 1%); 1—1,000 ohms, ½ watt; 1—3,900 ohms, 1 watt; 1—1, 1—2.7, 1—3.3 megohms, ½ watt; 1—20,000, 1—180,000 ohms, 1%, ½ watt; 1—1.8 megohms, 1%, ½ watt. (Potentiometers) 1—500,000 ohms; 1—50,000 ohms (Mallory U33).

Capacitors: (Electrolytic) 1—10 μf, 50 volts (Miniature metalized tubular); 2—.06, 2—0.5 μf, 400 volts; 2—1.0 μf, 200 volts; 1—0.5 μf, 600 volts. (Mica) 2—.004 μf.

Miscellaneous: 1—Adjustable inductor (UTC VI-C15). **Batteries:** 2—67.5 volts, Burgess XX45 or equivalent; 2—1.5 volts, Burgess 2FBP or equivalent. **Switches:** 1—s.p.d.t., spring-return rotary; 1—single-circuit, 3-position, non-shorting. **Meter:** 1—0-1-ma d.c., Triplett 327-T or equivalent. Chassis; cabinet; dial knobs; input terminals; hookup wire; tubes; sockets, etc.

distortion meter can be set initially to 100%. Under this condition, 1% distortion then can be read at full scale when the range switch is at its 1% setting. On the same range, the first major division of the meter scale (0.1 milliampere) indicates 0.1% distortion and the first small division 0.02% distortion. If the test voltage is lower than 1 volt but sufficient to allow the meter to be set to full scale with switch S2 in its 10% position, then 10% distortion is indicated at full scale when S2 is switched to its 10% position. When the test voltage is too low to permit setting the meter to full scale, simply divide the final indicated distortion figure, indicated by the meter, by the distortion indicated in the initial setting. Example: With switch S2 at its 100%

setting, the meter can be set initially only to the 50% point. In the distortion measurement, with S2 subsequently set to its 1% position, the meter indicates 0.5% distortion. The true indicated distortion then is 0.5 divided by 50, or 1%.

A new approach

With the basic principles of commonly-used distortion meters in mind, here is the theory and construction of a different type of instrument. This new distortion meter was developed in the electronics laboratory of the University of Wyoming. The new meter is inexpensive, easy to operate, and accurate within 2 percent. It can be used with inexpensive audio oscillators with no sacrifice in accuracy.

The new circuit eliminates the fundamental by *phase cancellation.*

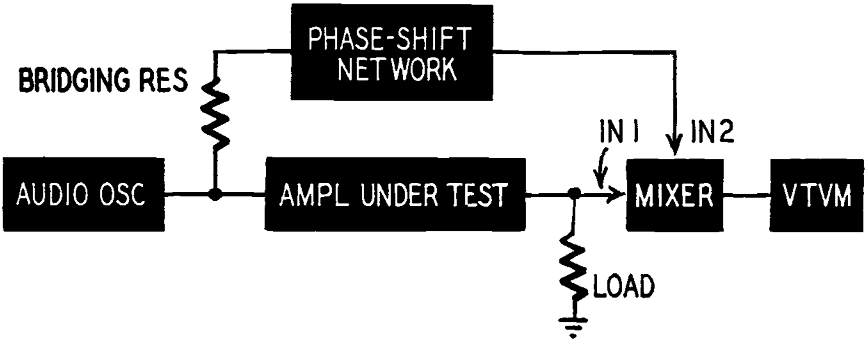


Fig. 1805. *Block diagram of the simplified method of measuring harmonic distortion.*

If we impress two sine waves of equal frequency on the same circuit, a number of results may be obtained. If the two waves are *in phase* (in the same position in their respective cycles at any given time), they will add, and the measured voltage will be the sum of the two individual amplitudes. If the waves are exactly opposite in polarity (*180° out of phase*), the larger will cancel the smaller, and if they are of equal amplitude, they will cancel each other completely.

No doubt you have jumped ahead in your thinking, and see the basic principle involved: *If the output of an amplifier containing a fundamental and harmonics is mixed with the correct amount of fundamental shifted exactly 180°, the resultant will contain only the harmonics.*

Fig. 1805 shows the principle in block-diagram form. The oscillator output is applied to the amplifier. A phase-shift network is "bridged" across the oscillator output. (A *bridging connection* should not be confused with a *bridge circuit.* "Bridging" is a term used in telephone and broadcast work to describe a high-impedance connection across a low-impedance circuit. Program monitors, volume indicators, and other testing devices are bridged across standard 600-ohm lines through high series resistances to reduce loading effects and provide a high degree of isolation. The v.t.v.m. is the most common example of a bridging device.)

The impedance match between oscillator and amplifier is not affected by the bridging connection, but the amplitude of the signal going to the phase-shift network is small. The amplifier output is fed to a suitable resistance load, and the load voltage is applied to one input of the mixer. The output of the phase-shift network is fed to the other mixer input.

Amplifier phase relations

In an *ideal* vacuum-tube voltage amplifier stage *with a resistive load,* the grid and plate voltages are exactly 180° out of phase. This relationship is not true in practical circuits. Tube and stray capacitances,

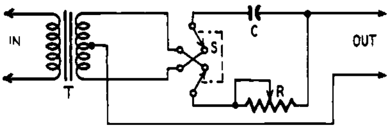


Fig. 1806. *Fundamental phase-shift circuit.*

coupling transformers, capacitors, and reactors upset the perfect resistive condition. In a multistage amplifier, we may have different phase shifts in each stage. The shift-per-stage will also change as the frequency changes. In any case the shift-per-stage cannot vary more than 90° from the 180° ideal. Thus an amplifier with an even number of stages would have an ideal phase shift of 360°, a minimum shift of 180°, and a maximum shift of 540°. An odd number of stages would have an ideal shift of 180°, a minimum shift of 90°, and maximum shift of 270°.

The phase-shift network in this instrument has two functions: (1) It provides a fixed shift of 180° to compensate for an odd or even number of amplifier stages; (2) it can be also varied over a range of approximately ±90° to make up for practical amplifier conditions.

The distortion-meter circuit

The basic phase-shift circuit of the instrument is shown in Fig. 1806. This circuit has been borrowed from industrial electronic equipment. (The Westinghouse *Industrial Electronics Reference Book* explains the operation of this circuit in complete detail. It is available at most large public libraries.) Reversing-switch S gives a fixed shift of 180°. Smaller shifts are obtained by varying R.

Fig. 1807 is the complete circuit. T1 is a *line-to-push-pull-grids* or *plate-to-push-pull-grids* type, with the highest possible step-up ratio. It need not be a high-fidelity type, but the secondary center-tap must be accurately balanced. Resistor R should match the input impedance of T1. R2 and R3 are equal bridging resistors. Their value is not critical and may be from 10,000 to 100,000 ohms. Higher values improve isolation but reduce the voltage to the transformer.

R1 should be about 50,000 ohms. C1 may be from 0.01 µf to 0.1 µf. Its value depends on the characteristics of T1, and can be found with the help of an audio oscillator and a scope. Set the oscillator to 1,000 cycles and connect it to the network input terminals, and to one of the scope inputs (either vertical or horizontal). Connect the other scope

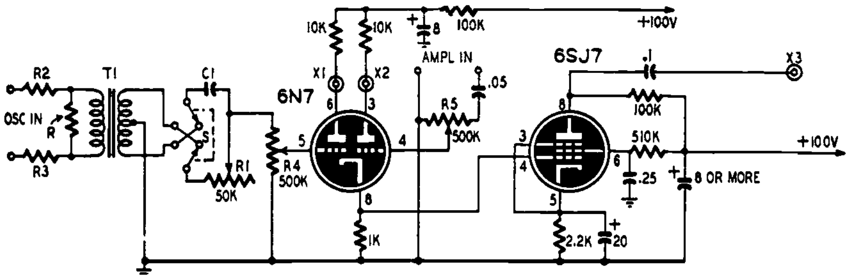


Fig. 1807. Schematic of the complete distortion measuring instrument.

input to X1, and shut off the internal sweep. Select C1 so that by varying R1 the scope pattern changes from a diagonal line to an ellipse, and then to a circle.

A 6N7 mixer is shown in the diagram, but any twin-triode type (6SL7, 6SN7, 12AT7, or 12AU7) may be used. R4 and R5 may be 500,000-ohm volume controls, or a T pad may be substituted for R5 if a calibrated attenuator is preferred.

The mixer is operated as a cathode follower. There is no voltage gain, but the stage has no measurable distortion. X1 and X2 are test jacks for oscilloscope checks. The 6SJ7 is a conventional resistance-coupled amplifier, except that the cathode resistor is double the usual value because of the direct-coupled input. X3 is the terminal for the v.t.v.m.

Testing and operation

After a complete continuity and voltage check, connect a v.t.v.m. to

X3, and short out both mixer inputs. Any reading on the meter indicates hum or noise in the instrument, which must be tracked down and eliminated.

When the unit is in perfect working order, connect an audio oscillator to the OSCILLATOR INPUT terminals and check the controlling effect of R4 on the output. At least 1 volt output should be obtained through the phase-shift circuit. Next, check the amplifier input section of the mixer in the same way.

Actual harmonic distortion measurements are carried out in the following way:

1. Connect the audio oscillator to the input of the amplifier under test, and bridge the phase-shift circuit across the input as shown in Fig. 1805. Adjust R4 for full-scale reading on the highest possible range of the v.t.v.m. plugged into X3.

2. Disconnect the bridging circuit from the input to the amplifier and short out the bridging input terminals. Connect the amplifier load resistor to the AMPLIFIER INPUT of the mixer, and adjust R5 for the same full-scale reading as in step 1.

3. Restore the bridging connection across the input to the amplifier and adjust R1 for *minimum* output. If the v.t.v.m. pointer goes off scale, reverse S and adjust R1 again for minimum output voltage.

4. Repeat steps 1, 2, and 3 for absolute minimum reading. The final reading is the harmonic distortion in *volts.* For maximum accuracy, the v.t.v.m. may be switched to a lower range *only after steps 1, 2, and 3 have been completed.*

The final voltage measurement may be converted to a percentage by the formula:

$$\%\ \text{Distortion} = \frac{100\ E_{min}}{E_{max}}$$

Distortion in the audio oscillator itself will not affect the accuracy of the measurements, since the original signal is completely canceled in the mixer if the final balancing adjustments are made with care.

Materials for Measuring Set

Resistors: 1—510,000 ohms, 2—100,000 ohms, 2—10,000 ohms, 1—2,200 ohms, 1—1,000 ohms, R, R2, R3 (see text) ½ watt; 2—500,000-ohm potentiometers.

Capacitors: (Paper) 1—0.25 μf, 1—0.1 μf, 1—0.05 μf, 400 volts; C1 (see text). (Electrolytic) 1—20 μf, 25 volts, 1—8 μf, 150 volts.

Miscellaneous: 1—6N7 tube; 1—6SJ7 tube; 1 push-pull input transformer (see text); 2 octal sockets; chassis, connectors, terminals, solder, hardware; power supply—100 volts d.c. at approximately 20 ma; 6.3 volts a.c. or d.c. at 1.1 amp; 1 d.p.d.t. switch.