*SECTION 2*

# ADVANCED
# PRACTICAL
# RADIO ENGINEERING

TECHNICAL ASSIGNMENT

AUDIO FREQUENCY AMPLIFICATION PART I



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*AUDIO FREQUENCY AMPLIFICATION PART I*

## FOREWORD

It is often the case that the most glamorous activity is not the most useful and important. Mention electronics to the layman, and he is apt to think of radar, ultra-high frequencies, television, and the like. Yet one of the earliest applications of the electron tube is still one of the most important, namely, audio-frequency amplification.

It was the telephone company that was one of the first users of the vacuum tube. The existing microphone-receiver type of amplifier had never really been satisfactory; the vacuum tube amplifier was so overwhelmingly superior that its adoption was a forgone conclusion, and today the vast network of telephone trunk lines is equipped with thousands of repeater amplifiers, that compensate for the attenuation experienced in the transmission of speech (audio) signals along these lines.

The audio amplifier is an indispensable component of every standard broadcast, f.m. and television receiver; it is to be found in every broadcast and television studio: all motion picture studios and theaters employ it; hotels, ball parks, auditoriums, and even restaurants use it as public address and sound reinforcing equipment, and the phonographs and juke boxes are more than ever one of its most important applications.

It can therefore be readily appreciated that this assignment is one of the most important and most practical in the entire course. It begins with an exposition of the meaning and use of decibels; a method of measuring power ratios that applies not only to audio amplifiers, but to communication systems in general.

Application of this method is then made to typical amplifier problems to show how it facilitates computation and design procedure; of course the illustrative problems are thoroughly practical and representative of good engineering practice.

Following this, the significance of frequency response curves will be taken up, and then the analysis of various types of voltage amplifier stages, including screen bypass and self-bias circuits. Finally, the various types of audio transformers and their applications will be treated in detail.

The concluding topic will be push-pull audio amplifiers. These are analyzed in detail, together with the distortion products that may arise if the amplifier is improperly designed. Since high-level output stages employ this type of amplifier, the material presented is of paramount importance.

The student, upon concluding this assignment, will find that he has obtained comprehensive and detailed instruction in this important subject, which will be invaluable to him in his daily work. In view of this, he may find it profitable to review this assignment at a later date to help fix its contents in his mind.

E. H. Rietzke,
President

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### AUDIO FREQUENCY AMPLIFICATION PART I

## SCOPE OF ASSIGNMENT

This assignment will deal with audio amplifiers, which are designed to amplify electrical voltages and currents corresponding to sound (audio) waves. Such amplifiers are measured as to their amplification and power-handling capabilities in terms of decibels (db), hence the assignment will first take up this unit of measurement, and how it is applied to amplifier calculations.

The next topic to be discussed will be that of frequency response, what it means, and how it can be used to determine the performance of an audio amplifier with regard to fidelity of reproduction. Following this, specific analysis of transformer-coupled and resistance-coupled stages will be made with regard to frequency response, as well as the effects of screen bypass and cathode bypass capacitors on the low-frequency response of an amplifier.

The concluding topic will be that of push-pull power amplifiers. The analysis will be mainly graphical in nature, and the various modes of operation, such as Class A, Class $AB_1$, Class $AB_2$, and Class B, will be discussed. Practical examples will enable the student to make his own determinations as to power output, grid drive, etc., of tubes he may subsequently have under consideration for a power amplifier stage, so that he can employ plate voltages, bias, etc., different from that specified by the manufacturer in the tube manual.

## DB CALCULATIONS

*POWER RANGE IN COMMUNICATION WORK.*—Ordinarily one speaks of a one horsepower electric motor, or a hundred horsepower gasoline engine, or compares a 100-watt lamp with a 60-watt lamp, and the like. Or he may speak of a 30,000 KW alternator in a power plant, and a 250-watt automobile generator.

Note that as the size gets larger, a change in scale is employed: kilowatts instead of watts. Indeed, in some radar sets the peak pulse output may be expressed in megawatts, and the average power in kilowatts.

But seldom is one confronted in the same line of equipment with a range from one to ten billion, or $10^{10}$, such as from a 1 watt motor to a 10 billion watt motor. In communication work, however, such a range is quite common. The quietest sound that can just be heard may be one ten-billionth that produced by an airplane motor, and even louder sounds are to be encountered in nature.

In everyday practice, a range in sound of 1,000,000 to 1 is common at an orchestra concert. The signal picked up by a radio receiver may vary by perhaps this amount if the set is moved from a point close to the transmitter to a location far away. In an audio amplifier, for example, the ratio of maximum output to noise in the amplifier may be specified as not to be less than 1,000,000 to 1.

All this points to the need for a computer in attempting to handle the huge number that may be encountered in a communications problem. The complications, however, are purely arithmetic in nature, and it would appear that some simpler and cheaper solution than a computer should be available. Such is the case: the use of logarithms furnishes a rather satisfactory means for handling such astronomic ranges.

The reason that the logarithm is so useful where large numbers are involved is quite simple. The logarithm is an *exponent*; it is the exponent to which ten must be raised to obtain the number in question. A number, such as ten, when raised even to a moderately small exponent, gives rise to a fairly large number. Thus $10^7$ is ten million.

This will be made even more clear by the following example. The numbers in the left-hand column of the following Table are each ten times the one above. They therefore form a GEOMETRIC progression. The numbers in the right-hand column are their logarithms. Note that they form an ARITHMETIC progression, whose terms grow more slowly.

| Number | Logarithm |
|---|---|
| 1 | 0 |
| 10 | 1 |
| 100 | 2 |
| 1,000 | 3 |
| 10,000 | 4 |
| 100,000 | 5 |
| 1,000,000 | 6 |
| 10,000,000 | 7 |
| 100,000,000 | 8 |
| 1,000,000,000 | 9 |
| 10,000,000,000 | 10 |

Hence, if instead of speaking of a range of 10,000,000 to 1, one speaks of the logarithms of such a range, the numbers involved are merely 7 to 0. As will be shown very shortly, the logarithm of the number is known as the BEL, and one-tenth of this logarithm is the familiar decibel.

*LOGARITHMIC RESPONSE OF THE EAR.*—There is another reason why the logarithm is a useful measure of audio power. The ear (and also the eye) has to respond to an enormous range of stimulus, namely in the case of the ear to sounds that may vary in power by ten billion to one, or more.

Compare this with an ordinary voltmeter, which can measure ON ANY ONE SCALE a range of perhaps 100 to 1. For example, on the 100-volt range, a voltmeter can indicate down to perhaps 1 volt. If 0.1 volt is to be read, one will have to switch to the 10-volt scale, or even lower. The ear, on the other hand, cannot switch from one scale to the other; it must indicate continuously from the quietest to the loudest sound. It must be sensitive to weak sounds, and yet not overload when loud sounds are present.

Nature has accomplished such a range and yet provided sensitivity at the low end of the power scale by making the ear logarithmic in its response. Thus at low levels the ear can detect small changes in actual power, and at high levels it can detect only large changes in power, so as not to overload. Referring to the Table, the left-hand column can represent the range in EXTERNAL STIMULUS, or sound intensity, and the right-hand column will represent the logarithmic response of the ear, or SENSATION in the brain.

As an example, if the power or stimulus is doubled, the increase in sensation in the brain is as the logarithm of two, or log 2 = 0.3011, or only 30.1% increase in sensation for a 100 per cent increase in stimulus. A thousandfold increase in stimulus produces but a log 1000 = 3-fold increase in sensation; a million-fold increase in stimulus produces but a 6-fold increase in sensation, and so on.

If then the ear reacts in a logarithmic manner, it is natural to rate the sound levels presented to it in this manner and hence to employ the decibel scale. In short, the decibel is a "natural" unit for measuring acoustic power or the electrical counterpart of such power.

*GAIN MEASUREMENTS.*—A third reason for the use of decibels is that it facilitates the computation of overall gain of an amplifier system when the gain of each individual unit is known. To make this clearer, consider the following example. In a broadcast studio, the power output of the microphone is amplified by a pre-amplifier by a factor of say, 1000 times.

This output goes through a so-called mixer system where it is combined or mixed with the outputs of other microphone preamplifiers, but in so doing, the mixer unavoidably cuts the power down to 1/10. Then a studio amplifier amplifies it by a factor of 2000, so-called attenuator "pads" cut the power down by a factor of 1/8, and a line amplifier multiplies the output by a factor of 200.

What is the ratio of the final output to the initial input? Clearly, the above factors must be all multiplied together, so that the overall "gain" is

$$\alpha = 1000 \times 1/10 \times 2000 \times 1/8 \times 200 = 5{,}000{,}000$$

Although the product here is rather simple, in general it can be arithmetically complicated. Multiplication and division are in general more difficult to perform than addition or subtraction.

This indicates that $\alpha$ can in general be more readily calculated if the logarithms of the individual factors are found and added algebraically. Thus, log 1000 = 3, log 1/10 = -1, log 2000 = 3.3, log 1/8 = .9, log 200 = 2.3, so that log $\alpha$ = 3-1 + 3.3 - .9 + 2.3 = 6.7, and only log 6.7 = 5,000,000. Indeed, it is not even necessary to find the antilog; one merely takes log $\alpha$ = 6.7, multiplies it by 10, and obtains a value for the gain of 67 db.

Similar considerations apply to transmission of the signal over say telephone lines. Suppose a mile of line attenuates the power to 1/3. What attenuation will 2 miles of line produce? Clearly, the second mile of line will attenuate the output of the first mile to 1/3, and the output of the first mile is 1/3 of its input. Therefore the output of the two miles of line is 1/3 × 1/3 = 1/9 of the initial input.

It would be desirable to say that the loss of one mile is so much; the loss of two miles is twice as much, and so on. In other words, it would be desirable to rate the loss in such manner that the total effect is the SUM rather than the PRODUCT of the unit-length effects.

This can be done if the logarithm of the loss ratio is used instead of the loss ratio itself, for now we can *add* logarithms instead of *multiplying* ratios. In short, the calculation of loss, as well as that of gain, is facilitated by using db, since this converts a multiplicative process into an additive process.

*GAIN OR LOSS IN DECIBELS.*—The foregoing anticipates to some extent the discussion of the decibel that is now to be presented. It has just been shown that the logarithm is a useful way of measuring sound power ratios: it avoids the use of large numbers, it represents the manner in which the ear hears, and it facilitates gain and loss calculations by substituting addition and subtraction for multiplication and division.

It now remains to show specifically how the decibel is computed and how it is used. The first application is that of DB GAIN. Suppose the input to an amplifier is $P_i$ watts, and the output is $P_o$ watts. Then the power ratio or gain is

$$\alpha = P_o/P_i \qquad (1)$$

The gain in BELS (in honor of Alexander Graham Bell) is:

$$\log \alpha = \log P_o/P_i \qquad (2)$$

Oddly enough, this logarithmic ratio compresses the numbers involved to too small a range, hence the logarithm is further multiplied by ten to furnish the decibel or db gain. Thus, the gain in decibels is

$$10 \log \alpha = 10 \log P_o/P_i \qquad (3)$$

As an example, suppose the input power is 1 milliwatt, and the output power is 20 watts, as indicated in Fig. 1 (A). Then the db gain is

$$10 \log 20/.001 = 10 \log 20000$$

$$10\ (4.3011) = 43.011 \text{ db or } 43 \text{ db.}$$

On the other hand, suppose audio power $P_i$ of 5 watts is applied to an attenuation tee pad (resistive network, as shown in (B), and the output power $P_o$ is .002 watt or 2 milliwatts. What is the db ATTENUATION of this pad?

$$\text{db (att.)} = 10 \log \frac{.002}{5} = ?$$

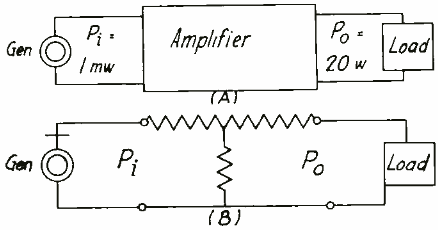



Fig. 1.—Examples of circuits employing amplification or attenuation.

In order to avoid finding the logarithm of a fraction, simply invert the fraction to make it a number greater than one, and put a minus sign in front of the logarithm. This is because

$$10 \log \frac{.002}{5} = 10 \log [1/(5/.002)]$$

$$= 10 \log 1 - 10 \log 5/.002$$

$$= 10 \times 0 - 10 \log 5/.002$$

$$= -10 \log 5/.002.$$

To complete the problem,

$$\text{db (att.)} = 10 \log \frac{.002}{5}$$

$$= -10 \log \frac{5}{.002} = -10 \log 2500$$

$$= -10\ (3.4) = -34 \text{ db.}$$

It is clear from the above results that +db represents gain, and -db represents attenuation or loss.

*DB LEVEL.*—If the power output

of a device can be compared to the power input on a db basis, why cannot ANY value of power be compared to a FIXED or STANDARD value of power? For example, the standard or unit of power may be 1 watt. Or it may be 1 milliwatt, as is currently the value employed, or it may be 6 milliwatts, which was the previous unit of power, or 12.5 milliwatts, as previously employed by the National Broadcasting Company, etc.

When any value of power is compared with the chosen unit of power, the resulting value of db obtained is known as the DB LEVEL, rather than db gain or loss, and the standard is known as the REFERENCE LEVEL. The reason for the use of the word level will be apparent from the discussion to follow.

Consider first the unit power to be chosen. It should be more or less centered in the range of powers that are normally encountered. What is the range? On the high side, powers of 50,000 watts may be encountered; on the low side, an output power of but $10^{-10}$ watt may be obtained from a microphone. An average between these two extremes is the geometric mean, which is

$$\sqrt{10^{-10} \times 50000} = 2.24 \text{ milliwatts.}$$

The actual power unit now employed is one milliwatt.

What is the level for the unit 1 milliwatt? It is simply

$$\text{db level} = 10 \log \frac{1 \text{ mw.}}{1 \text{ mw.}} = 10 \log 1$$

$$= 10 \times 0 = 0 \text{ db.}$$

In short, the unit of power has zero db level; it corresponds somewhat to the zero point on a thermometer scale.

What is the power level of 50000 watts? It is 10 log 50000/.001 = 10 log $5 \times 10^7$ = 10 (7.7) = +77 db. The plus sign is written here merely to emphasize to the student that this is a level HIGHER than 0 db, the unit level; i.e., 50000 watts is higher than 1 milliwatt.

As another example, an amplifier employing a pair of 2A3 tubes in push-pull in the power output stage, has an output of 15 watts. What is the output level? It is 10 log (15/.001) = 10 log 15000 = 10 (4.1761) = 41.8 db.

On the other hand, consider a ribbon microphone. Its output of course varies with the sound power impinging upon its diaphragm; for a moderate-sized orchestra the sound power is that corresponding to 10 BARS, where one bar is a pressure of one dyne per square cm. of the acoustic wave. For the ribbon microphone, the output is approximately $2.4 \times 10^{-3}$ microwatts for 10 bars sound pressure. What is the output level?

Since it is less than the reference level of 1 milliwatt, the ratio will be fractional, the logarithm will therefore be *negative*, so that the db level of the microphone will be negative. In computing the level, it will be simpler to take the reciprocal of the ratio in order to obtain a positive logarithm which is found directly in the log tables (in the manner shown previously). Then a minus sign can be placed in front of it to show it is a negative level.

Thus, the db level of the microphone is

$$10 \log \frac{2.4 \times 10^{-9}}{1 \times 10^{-3}}$$

$$= 10 \log \frac{10^{-3}}{2.4 \times 10^{-9}}$$

$$= -10 \log 4.17 \times 10^5 = -10\ (5.6201)$$

$$= -56.2 \text{ db.}$$

(Note the inversion of the fraction from the first to the second step, with the addition of the minus sign).

*DB LEVEL AND GAIN.*—The reader must suspect by this time that there is some relationship between db level and db gain. This is true, and is very simple to see. In Fig. 2 has been plotted a db scale somewhat similar to the scale on a thermometer. Suppose the ribbon microphone mentioned above is used to

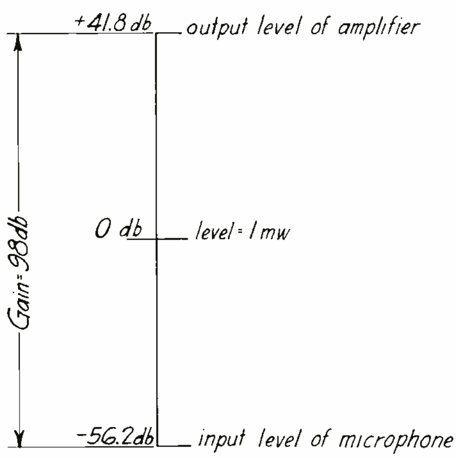



**Fig. 2.—DB scale showing relation between db level and db gain.**

feed the 2A3 amplifier discussed previously. How much gain must the amplifier have to bring the microphone level of -56.2 db up to that of the maximum output of 41.8 db corresponding to 15 watts?

Fig. 2 clearly shows that it must be 41.8 - (-56.2) = 98.0 db. The general rule is simply:

RULE: To find the db gain (or loss) of a device, subtract the input level in db from the output level in db.

*DB LEVEL TO WATTS.*—Often the db level is given, and it is desired to find the actual power. This is a simple process of finding the antilogarithm of the given number. For example, suppose the output level of a phonograph pickup is -24 db. What is its power output in watts (or microwatts)?

$$-24 = 10 \log \frac{P_o}{1 \times 10^{-3}}$$

or

$$\log \frac{P_o}{1 \times 10^{-3}} = \frac{24}{10} = -2.4$$

To avoid negative logarithms and antilogarithms, take the reciprocal $1 \times 10^{-3}/P_o$. Then

$$\log \frac{1 \times 10^{-3}}{P_o} = +2.4$$

$$\frac{1 \times 10^{-3}}{P_o} = \text{anlg } +2.4$$

$$P_o = \frac{1 \times 10^{-3}}{\text{anlg } 2.4} = \frac{1 \times 10^{-3}}{251}$$

$$= 3.99 \times 10^{-6}$$

or approximately 4 μwatts.

(Note that in taking the anlg of 2.4, the quantity 2 shows where the decimal point is to be placed, and 0.4 is looked up in the log table. This then is multiplied by $10^2$ = 100 to get 251).

As another example, suppose that the output level of an amplifier is +31 db (where the plus shows the output is *greater* than 1 mwatt, the reference level). What is the power output in watts?

$$10 \log \frac{P_o}{1 \times 10^{-3}} = +31$$

$$\log \frac{P_o}{1 \times 10^{-3}} = 31/10 = 3.1$$

$$P_o / 1 \times 10^{-3} = \text{anlg } 3.1 = 1260$$

$$P_o = 1260 \times 10^{-3} = 1.26 \text{ watts.}$$

As a further point, if the amplifier is to operate from the phonograph pickup mentioned previously, the gain will have to be

$$\text{Gain} = \frac{1.26}{3.99 \times 10^{-6}} = 0.316 \times 10^{6}$$

or

$$3.16 \times 10^{5}$$

In db it will be 10 log $3.16 \times 10^5$ = 10 (5.4997) = 55.0 db. As a check, subtract the db level of the pickup from that of the amplifier:

31 - (-24) = 55 db gain CHECK.

*VOLTAGE, POWER, AND DB.*—In actual practice it is difficult to measure power, because no ordinary wattmeter instrument is sufficiently sensitive to read down to microwatts, or even milliwatts of power, and moreover it will not read accurately over the entire audio range, and particularly in the r-f region, although more and more specialized electronic instruments are being developed for this purpose.

In the case of audio systems, the frequency range covered is enormous from the viewpoint of the number of octaves involved (about ten) and networks having reactances within them will not have a uniform transmission characteristic over such a range unless they approach a resistance in their characteristics.

Hence it has been customary to specify the performance of a microphone, amplifier, loudspeaker, phonograph pickup, etc., on the basis of its behavior when fed from a resistive source or terminated in a resistive load, as the case may be. For example, in Fig. 3 is shown a microphone feeding an amplifier, which in turn feeds a resistive load simulating a loudspeaker.

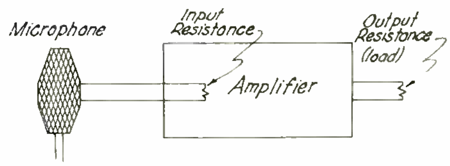



Fig. 3.—Simple audio system showing input and output resistance.

The attitude on the part of the amplifier manufacturer is that his amplifier will perform in accordance with his specifications if the output load is a pure resistance of the specified value, and this value is to be used in testing his amplifier. It is up to the loudspeaker manufacturer to produce a unit which will faithfully reproduce the original sounds when fed from this amplifier whose performance is specified with respect to an output resistance.

Similarly, the microphone's performance is generally specified with respect to a certain value of load resistance. The amplifier connected to it may actually present this load resistance to the microphone, or it may not; it is immaterial in a sense to the microphone manufacturer how his microphone behaves in conjunction with a particular amplifier if he has specified its performance in terms of a certain terminating load impedance.

Actually the situation is not in such a state of anarchy as the above discussion might imply. As will be shown later, the input impedance of an amplifier may either be a pure resistance of the proper value, or it may be such a high reactance as to constitute essentially an open-circuit load to the microphone, as in the case of an unloaded input transformer over most of the audio range. Similarly the performance of the amplifier terminated in a loudspeaker may be reasonably close to its performance with a load resistor over most of the audio range.

Where resistive terminations are employed, the power output $P_o$ can be determined very simply by measuring the output VOLTAGE $E_o$ across the resistor $R_L$. Thus

$$P_o = \frac{E_o^2}{R_L} \quad (4)$$

Voltage measurements are feasible because voltmeters can readily be built which are reasonably accurate over the entire audio frequency range. It will now be of interest to correleate the voltage reading with db level. Let $P_{rs}$ be the power corresponding to the reference standard (now usually chosen as 1 milliwatt). Then

$$\text{db level} = 10 \log \frac{P_o}{P_{rs}}$$

$$= 10 \log \frac{E_o^2}{P_{rs} R_L} \quad (5)$$

Suppose the reference power $P_{rs}$ is developed across a resistance $R_{rs}$, and that the corresponding voltage across $R_{rs}$ is $E_{rs}$. Then

$$P_{rs} = E_{rs}^2/R_{rs} \quad (6)$$

and if this is substituted in Eq. (5) the results

$$\text{db level} = 10 \log \left(\frac{E_o^2}{R_L} \cdot \frac{R_{rs}}{E_{rs}^2}\right)$$

$$= 10 \log \frac{E_o^2}{E_{rs}^2} + 10 \log R_{rs}/R_L$$

$$= 20 \log E_o/E_{rs} + 10 \log R_{rs}/R_L \quad (7)$$

Eq. (7) states that the db level can be measured if the ratio of the given voltage $E_o$ to the reference voltage $E_{rs}$ is known, TOGETHER WITH A CORRECTION FACTOR 10 log $R_{rs}/R_L$ which modifies the result depending upon the ratio of the given impedance $R_L$ to the reference impedance $R_{rs}$.

Specifically, if the reference power $P_o$ is chosen as 1 milliwatt, and if further, $R_{rs}$ is specified as 600 ohms, then $E_{rs}$ is determined in accordance with Eq. (6); its value is 0.775 volt. These are the values used today for the reference level: 0 db = 1 milliwatt = 0.775 volt across 600 ohms.

*THE DB OR OUTPUT METER.*—In Eq. (7), if $R_L = R_{rs}$, the correction factor becomes 10 log 1 = 0 or drops out. Furthermore, if $E_o = E_{rs}$, the first term becomes zero, too, which means that the level is 0 db, as is to be expected. Suppose now that a voltmeter is calibrated so that when

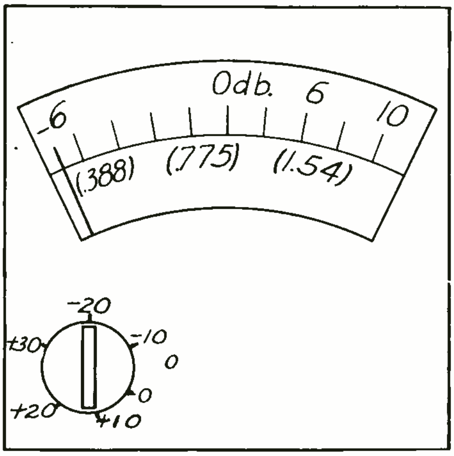



Fig. 4.—A db meter using a voltmeter movement.

0.775 volt is impressed across its terminals, the resulting deflection is marked 0 db on the scale; when, for example, 2 × .775 = 1.540 volts are impressed across its terminals, the resulting deflection is marked (20 log 2 =) 6 db, and so on.

This is illustrated in Fig. 4.

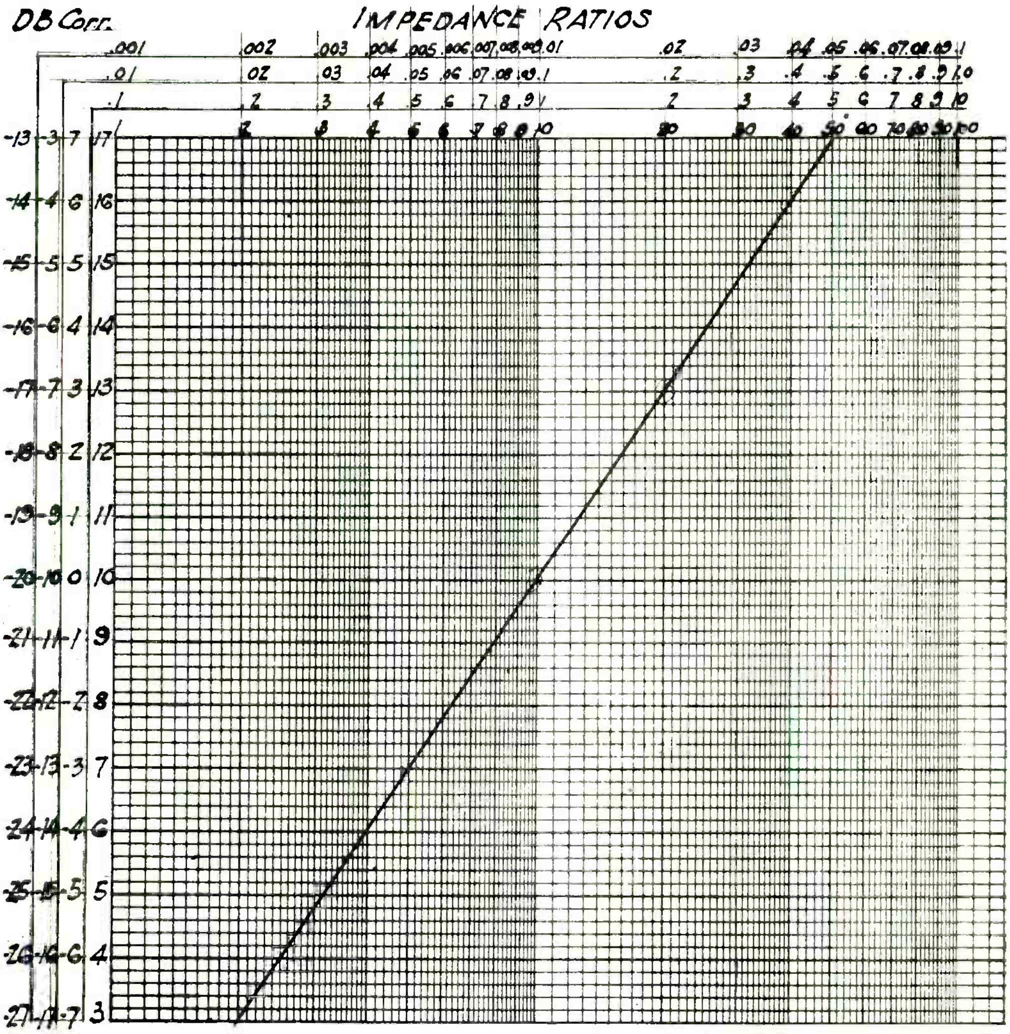



Fig. 5. — Impedance Ratio Correction Factor.

The meter readings in db are indicated on the scale from -6 db (corresponding to .775/2 = .388 volt) to + 10 db, (corresponding to .775 × $\sqrt{10}$ = 2.45 volt). The actual quantity read is the voltage; if the impedance across which the voltage is read is 600 ohms, then the reading also corresponds to the db markings on the scale. Often both db and voltage markings are provided, so that the meter can function both as a voltmeter and as a db-meter.

If the resistance across which the reading is taken is other than 600 ohms, the proper correction factor can be added algebraically to the meter reading to give the true db level across the actual load resistance. Fig. 5 provides a curve which can be used to quickly obtain the correction factor.

One curve can serve for all values of the impedance ratio because of the additive nature of the logarithmic function. For convenience four scales are shown; others can be quickly made by dividing the abscissa value, for example, by 10, **and then subtracting 10 db from the corresponding ordinate.**

To see how it is used, suppose the db meter reads 7 db and the load impedance is 6000 ohms. Then $R_{rs}/R_L$ = 600/6000 = 0.1. The next-to-the-top abscissa scale contains this value, the next-to-the-left-hand ordinate scale corresponds to the abscissa scale chosen. From the curve, the corresponding value for $R_{rs}/R_L$ is -10 db. Hence the true reading is 7 + (-10) = -3 db.

Suppose the impedance had been 15 ohms. Then $R_{rs}/R_L$ = 600/15 = 40, whereupon the lowest abscissa and the right-hand ordinate scales apply. For 40, the correction factor is +16 db, so that the true reading is 7 + (+16) = 23 db.

Usually the meter scale covers a limited number of db; in Fig. 4 it ranges from -6 to +10 db. In order to cover the much larger db range encountered in practice, an additional scale-changing switch is employed. A circuit like that shown in Fig. 6 is often employed. Here a ladder-type network is used so as to present as nearly constant (and yet high) a resistance as possible to the circuit under test.

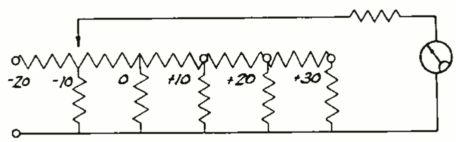



Fig. 6.—Ladder-type resistive network used to change scales on db meter.

As the slider or arm moves to the left in Fig. 6, more and more resistive attenuation is cut out between the source and the meter, and hence the higher the latter reads; or to put it another way, the lower the level that the meter will register up-scale. When, for example it is set on the -10 db scale, if the voltage across the load to be measured corresponds to -10 db, then the pointer will move up to the 0-db mark on the scale. The reading will therefore be -10 + (0) = -10 db. To this, of course, must be added the impedance correction factor, if it is present.

Suppose that the readings are -10 db on the scale switch, and the pointer moves up to +3.5 db, and assume further that the load impedance is 7.5 ohms. The true db level is -10 + (+3.5) + 19 = +12.5 db.

The last factor, 19, is found from Fig. 5. Thus $R_{Ls}/R_L = 600/7.5 = 80$. Although this is off scale in Fig. 5, it is readily found, because 80 = 8 × 10, and the correction factors for 8 and 10 are 9 db and 10 db, respectively, so that the total correction factor is 9 + 10 = 19 db.

*GAIN CALCULATIONS.*—These examples should indicate how the db meter and the correction curve of Fig. 5 can be used to measure db level. Fig. 5 can also be used in the calculation of the db gain of an amplifier. To show this, consider the test setup shown in Fig. 7.

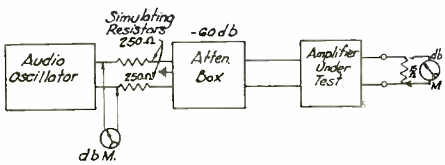



Fig. 7.—Test setup for measuring frequency response of an audio amplifier.

As will be explained in the next section, the gain of an amplifier should not vary appreciably with frequency in the audio range, if faithful reproduction is to be obtained. To test this, an audio oscillator is employed, whose output voltage can be varied in frequency as desired.

Although this oscillator is designed to furnish a nearly constant output voltage over the frequency range, and to have any desired internal impedance by the correct choice of transformer tap, it is usually preferable to place a voltmeter, say of the db type, across the oscillator's terminals, and maintain the voltage constant at this point regardless of the setting of the frequency dial.

The oscillator then appears as a ZERO-RESISTANCE source, by virtue of a rule known as the Compensation Theorem. Simply stated, the oscillator exhibits no regulation or voltage variation, hence it behaves like a source having no internal resistance.

The two 250-ohm so-called simulating resistors following the oscillator then make it look like a 500-ohm balanced-to-ground resistance. This is done because in Fig. 7 it is assumed the amplifier under test has an input of 500 ohms balanced-to-ground, i.e., center-tap grounded. If instead it is unbalanced to ground (one side grounded) then the entire 500 ohms should be placed in the ungrounded side of the system, and a direct connection made on the other side.

The next device shown is an attenuation box. This consists of a number of so-called attenuation pads,* whose function is to attenuate or lower the level of the power furnished by the oscillator. The reason for this is very simple.

Usually the amplifier under test has a fairly high gain, and is designed to amplify a very weak input signal to a relatively high output level, such as that required to actuate a loudspeaker. If a strong signal is impressed on the input of the amplifier, some tube farther on will be driven so hard as to drive

*These will be described in a later assignment.

the grid positive and/or beyond cut-off, whereupon the amplifier will be overloaded and furnish a distorted output.

On the other hand, if the signal is sufficiently low so as not to overload the amplifier, it will be far too weak to measure on the db meter. Hence a strong signal is furnished by the audio oscillator, and is readily measured by the db meter. It is then attenuated by a KNOWN amount by the attenuation box, and then fed to the amplifier.

The attenuation box also is usually designed to be balanced to ground, and to have an impedance of 500 ohms. Fig. 8 shows the resistive configuration normally employed; it is known as an H pad, because each section looks like the letter H lying on its side. Several can be connected in cascade, as shown, to provide more attenuation than one pad alone.

A further characteristic is that if the correct or matched impedance is placed across one pair of terminals, the SAME impedance will be seen looking into the other pair of terminals. The significance of this is that the same power flows OUT of the source whether connected directly to the load or to the pad terminated in the load, but in the latter case a certain *known* fraction of this power is wasted in the pad, and the remainder (also known) gets into the load. As an example, if the attenuation of the pad is 6 db, then only one-quarter of the power flows into the load compared to what would flow into it were it connected directly to the generator.

The attenuation of the pad is adjustable, usually in steps of 1 db. It takes the relatively high level of the output from the simulating resistors and reduces it to an acceptably low level for the amplifier. Then the amplifier raises the level of the signal once more to

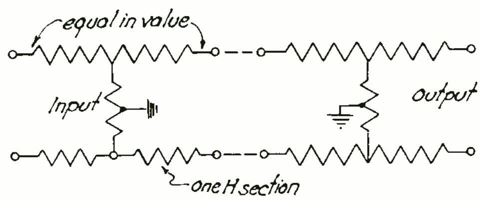



Fig. 8.—Cascade connection of H-pads.

a value at the output sufficient to be read by the db meter. Since the input level, attenuation, and output level are all known, the gain of the amplifier can be readily calculated.

Before doing so it will be well to mention one further point about the attenuator. As stated previously, it furnishes the indicated attenuation only when it is terminated in the proper load. This requires two things: if the load is balanced to ground, the attenuator should be balanced to ground too, as is indicated in Fig. 8; and the load (input of the amplifier) should match the attenuator in impedance.

There are certain standard values of input impedance, such as 500 ohms, 250 ohms, 125 ohms, and perhaps 50 ohms, with 500 ohms (possibly also 600 ohms) as the most common value. Hence the pads in the attenuator are usually designed "to match 500 ohms to 500 ohms;" i.e., a 500-ohm source to a 500-ohm load, and are designed to be either of the balanced or H type or of the unbalanced or Tee type (half of an H type).

If the amplifier has a different input impedance, then it is necessary to interpose between it and the attenuator either a suitable matching transformer, or a matching TAPER pad. (Fig. 9). The transformer has the least attenuation—perhaps only 1/2 db.; the L type taper pad is next and represents the minimum-loss type, and the Tee and

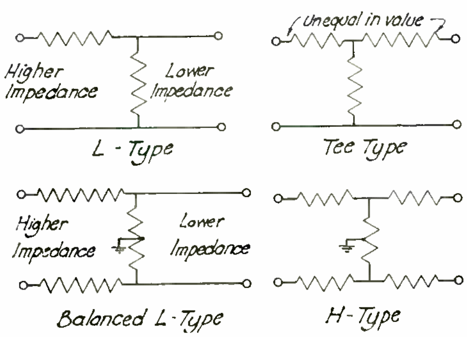



Fig. 9.—Two types of taper pads for matching unequal impedances.

H-type have a greater loss. Normally the L type is used in either the balanced or unbalanced form, depending upon the input of the amplifier. The inherent attenuation of this type of pad is greater, the greater the disparity in the impedances it is designed to match.

Another point is that if the amplifier input is unbalanced to ground, either a suitable so-called "isolation" transformer will be required, as shown in Fig. 10 (A), or else one-half of the attenuator can be used, together with a taper pad, if required, as shown in (B).

In any case, if the attenuation of the attenuator and taper pad is 10 db or greater, then the impedance looking into the generator end a-b is practically a pure resistance, say 250 ohms into one-half of a 500-ohm H pad. In that case the voltage at a-b will be half of that at c-b over the entire frequency range, and therefore there is no real need for the simulating resistance. As a result, it is very often omitted in the test setup, but will be included here in the sample calculations.

Suppose in Fig. 7 (repeated for convenience), the db meter reads +5 db at the terminals of the audio oscillator, and +15 db at the output terminals of the amplifier (15-ohm load resistor). Suppose further that the attenuation in the attenuator box is -53 db. What is the gain of the amplifier?

**A correction factor at the oscillator terminals is required because the impedance there is 2 × 250**

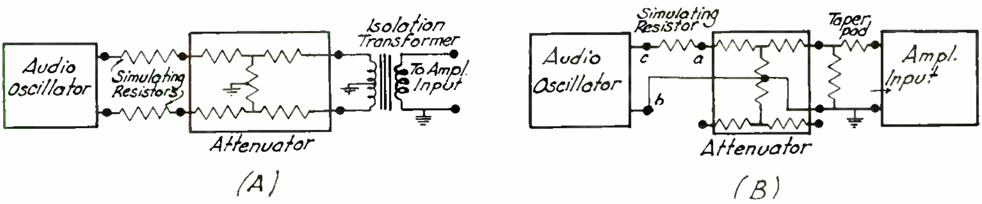



Fig. 10.—Use of an isolation transformer or one-half of an H-pad attenuator.

= 500 ohms for the simulating resistors + 500 ohms in the attenuator, or a total of 1000 ohms. Then $R_{rs}/R_L$ = 600/1000 = 0.6, so that

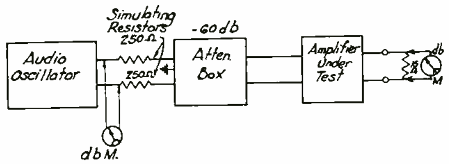



Fig. 7.—Test setup for measuring frequency response of an audio amplifier.

from Fig. 5 the correction factor is -2.2 db. Hence the true reading at the oscillator terminals is +5 + (-2.2) = 2.8 db.

Next there is a 3 db loss in going from the oscillator to the attenuator, owing to the fact that half the power, or 3 db, is lost in the simulating resistors. Hence the level at the attenuator is 2.8 - 3 = -0.2 db. Then, in passing through the attenuator, there is a further loss of 53 db, so that the input to the amplifier is -0.2 - 53 = -53.2 db.

The output level of the amplifier must be corrected because the load impedance is 15 ohms instead of 600 ohms. It was found previously that for 15 ohms the correction factor is +16 db, hence the true output level is 15 + 16 = 31 db. Since the input level is -53.7 db, the gain must clearly be 31 - (-53.2) = 84.2 db. Fig. 11 clearly illustrates the variation in db level from one point to the next in the test setup.

*PRELIMINARY AMPLIFIER DESIGN CONSIDERATIONS.*—It is now possible to examine the method of preliminary design of an amplifier. One has to know the input level and the output level, then one can determine the output stage, and the number and kind of voltage amplifier stages. The method is best presented by means of an example.

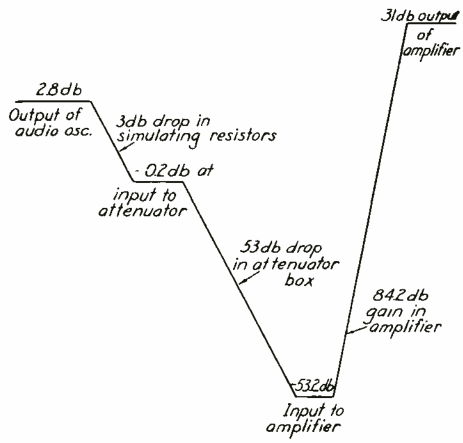



Fig. 11.—DB chart showing how level varies from one point to the next in an amplifier test setup.

Suppose it is desired to design an amplifier which is to be fed by a microphone whose output level at 10 bars is -60 db,-i.e., 60 db below 1 milliwatt. An output power of 15 watts is desired into a loudspeaker system. The output level for 15 watts was previously shown to be 41.8 db.

As a preliminary procedure, one can thumb through the tube manual, where it will be found that a pair of 2A3 tubes operating in push-pull at a plate potential of 300 volts and a grid bias of -60 volts, will deliver 15 watts into a 3000-ohm plate-to-plate load. (The significance of this will be discussed in a following section of this assignment).

The grid swing will be 60 volts peak for each tube, or 120 volts for the two tubes,-i.e., grid-to-grid. Assume a push-pull input transformer is employed, as shown in Fig. 12.

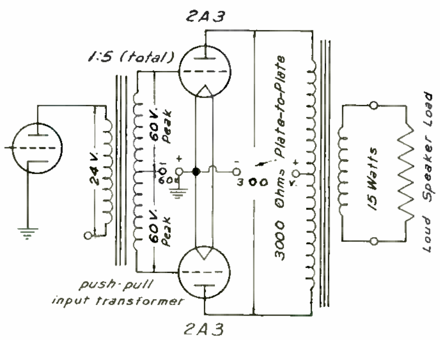



Fig. 12.—**Push-pull 2A3 output stage.**

As can be checked from the manufacturer's catalogue, a reasonable turns ratio for such a transformer is 1:2.5 from the entire primary to either half secondary, or 1:5 from the entire primary to the entire secondary.

Since 120 volts peak signal is required across the entire secondary, 120/5 = 24 volts peak is required to be developed across the primary winding. As will be shown later, the gain of a transformer-coupled stage in the region of flat response (useful range) is closely equal to the mu of the tube feeding the transformer. Moreover, the transformer requires a low-impedance source, around 10,000 ohms or less, so that a low-mu triode is indicated whose $R_p$ is in this range.

In looking through the tube manual, it is found that a 6J5 triode has the characteristics desired. It has a plate resistance of 7700 ohms at -8 volts bias, a mu of 20, and a transconductance of 2600 µmhos. The peak grid swing is equal to the bias voltage, or 8 volts. The tube then amplifies it by a factor of 20, thus providing 8 × 20 = 160 volts across the primary of the push-pull input transformer.

As a factor of safety, suppose the transformer losses (mainly core losses), reduce the amplification by a factor of 0.9, so that the voltage across the primary is 0.9 × 160 = 144 volts. Such a value is more than adequate, for the voltage across the secondary is 144 × 5 = 720 volts, and only 120 volts are required.

This is fortunate, for the tube need be driven but a fraction of the maximum voltage possible (8 volts), and as a result the distortion generated in this tube will be very low. This is a desirable condition for all voltage amplifier tubes, i.e., for the tubes preceding the power stage. The reason is that then the greater fraction of the total distortion permissible is left for the power tubes, which are driven hardest to get the most out of the most expensive tubes in the amplifier, and which therefore tend to generate the most distortion.

The grid drive for the 6J5 tube, for MAXIMUM output, is 120/(5 × .9 × 20) = 1.333 volts peak, instead of the maximum possible of 8 volts. The input voltage from the microphone at 10 bars sound pressure corresponds to -60 db. This does not represent the peak power, which may be 15 to 20 db above this value. Suppose the peak power is 15 db above, then the microphone input level is -60 + 15 = -45 db below 1 milliwatt.

In actual figures the input power is

$$P_1 = (10^{-3}\ \text{watt})\left(\text{anlg}\ \frac{-45}{10}\right)$$

$$= (10^{-3})/\left(\text{anlg}\ \frac{+45}{10}\right)$$

$$= 10^{-3}/\text{anlg}\ 4.5$$

The anlg of 4.5 is found by looking up the anlg of .5 in the log table; the 4 will determine the position of the decimal point, since it represents the factor $10^4$ = 10,000. The anlg of 0.5 is 3.16, hence anlg of 4.5 = 3.16 × $10^4$ = 31600. Therefore

$$P_1 = 10^{-3}/31600$$

$$= 3.17 \times 10^{-8}\ \text{watts}$$

or

0.0317 microwatts.

This is an amazingly small amount of power, and illustrates the marvelous sensitivity of the ordinary vacuum tube amplifier in raising such a small amount of power up to a few watts, or even 50 KW for use to modulate a broadcast transmitter.

Fig. 13 shows the input circuit involved. A step-up input transformer is used so as to step up the signal voltage coming from the microphone to a higher value as applied to the grid. Such a transformer can represent the equivalent of an additional stage of amplification, IF THE IMPEDANCE OF THE SOURCE (MICROPHONE) IS LOW.

For example, if the source impedance is 500 ohms, a step up to 150000 ohms or so is possible. This corresponds to a turns ratio of n = $\sqrt{150000/500}$ = 17.32 times. Thus, a resistor $R_L$ = 150,000 ohms can be placed across the secondary, and it

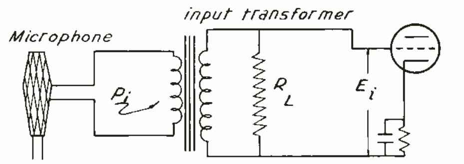



Fig. 13.—The microphone feeds an input transformer, which steps up the voltage to the grid of the first tube.

will match the 500-ohm source when reflected to the primary winding. At the same time, the voltage applied to the grid is 17.32 times that across the primary.

However, in calculating the amplifier gain, the power $P_1$ calculated previously can be regarded as being developed across $R_L$, and the resultant voltage $E_1$ across $R_L$ can therefore be found. Note that for a given source impedance, the greater the step up of the transformer, the higher $R_L$ will be to match the source impedance, and also the greater $E_1$ will be.

There is, however, a limit to the amount of step up feasible; farther on in this assignment it will be shown to depend mainly on the band width desired. A good value for $R_L$ is 150,000 ohms; if the step up is from a 500-ohm source, a turns ratio of 17.32 is involved.

**Assuming $R_L$ = 150,000 ohms, and $P_1$ = 3.17 × $10^{-8}$ watt, the secondary voltage is <u>readily found to be</u> $E_1$ = $\sqrt{P_1 R_L}$ = $\sqrt{3.17 \times 10^{-8} \times 15 \times 10^4}$ = 6.9 × $10^{-2}$ volt = 69 millivolts. This minute voltage now has to be amplified to 1.333 volts at the grid of the 6J5 tube.**

It will be shown later that for better SIGNAL-TO-NOISE ratio, it is preferable to eliminate $R_L$ since it, as well as the microphone, is a source of thermal noise. The microphone then feeds the unloaded transformer, which appears practically as an open circuit to it (similar to the push-pull input transformer).

If $R_L$ had been included and matched to the microphone, half of the voltage generated in the microphone would have been lost in its own internal resistance, and half would appear across the primary of the input transformer. With $R_L$ eliminated no appreciable voltage is lost in the microphone, so that the voltage across the primary, and hence also across the secondary, doubles, and is therefore $2 \times 6.9 \times 10^{-2} = 13.8 \times 10^{-2}$ volts.

If this value is used, the voltage gain is found to be $1.333/(13.8 \times 10^{-2}) = 9.67$ times. Resistance-coupled stages are usually employed so far as possible for voltage amplification. If one stage is employed, and has a gain of 10 or higher, it will be satisfactory. As excess gain is preferable, although too high a gain is useless and introduces unnecessary problems of instability.

For example, the tube manual shows that a 6J7 operating as a pentode with a load and grid resistance of 100,000 ohms each, has a gain of 41, which is more than sufficient. If desired, the load resistance can be reduced proportionately to bring the gain down closer to 10 as calculated.

The above example indicates how the preliminary design of an audio amplifier would be calculated. It will now be of value to investigate the behavior of the various voltage amplifier stages and push-pull power output stage. The single-ended power output stage has been discussed in a previous assignment.

## FREQUENCY RESPONSE CHARACTERISTICS

*WAVE ANALYSIS.*—The behavior of circuits has been studied by analyzing their behavior to *sinusoidal* voltages and currents. The reason is that computations are quite simple in such cases, particularly if J-operators and/or vectors are employed. However, actual waves encountered in practice, such as the audio signal from a microphone or the video signal from a television camera, are not so simple in wave form nor is their behavior in a circuit as easy to compute.

Fortunately, such wave shapes can first be analyzed and shown to be made up of *sinusoidal components* of different frequencies. Then the behavior of the circuit to each component can be computed, and then the results summed to give the total effect. (This will be illustrated very shortly.)

It was a French mathematician, Fourier, who first discovered this method while analyzing the behavior of a vibrating string initially distorted or plucked at various points along its length. Fourier found that the string vibrated as a whole at some particular frequency; also it could vibrate in halves, with each half of the string vibrating at twice the previously mentioned lowest frequency; it could vibrate in thirds at three times the lowest frequency, and so on. These vibrations could take place ALL AT THE SAME TIME; the motion of the string might be likened to the frenzied motions of a performer of a one-man band.

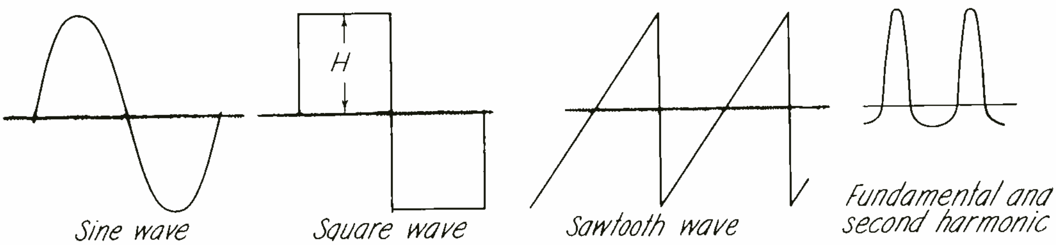



Fig. 14.—Some representative examples of periodic waves.

Fourier called the lowest frequency of vibration the *fundamental* frequency; the vibration in halves, the second-harmonic frequency; the vibration in thirds, the third-harmonic frequency, and so on. From this arose a method of analysis of PERIODIC waves.

A periodic wave is one which goes through its instantaneous values over and over again; i. e., repeats its wave form periodically. Fig. 14 illustrates some simple periodic wave forms, including the simplest of all—the sine wave. It is clear from the figure that all have the common characteristic of repeating their wave form over and over again; theoretically from time immemorial until the present moment of observation.

These waves can be shown to be composed of a number of sinusoids whose amplitudes and phase have special values that act to produce the resultant wave when the instantaneous values are added together. In the case of the sine wave it has but one component, itself; this is a trivial mathematical example. Mathematically one might say that all its harmonics have zero amplitude, but this is merely a play on words; it can just as well be said to have no harmonics.

In the case of the square wave, however, this is not the case. The square wave can be analyzed mathematically and shown to have an infinite number of ODD harmonics. Thus, it has a fundamental component or first harmonic whose frequency or crossings of the time axis are equal to those of the square wave itself.

Then it has a third harmonic, fifth harmonic, seventh harmonic, and so on. The EVEN harmonics all have ZERO amplitude; i. e., they are absent in this wave. Returning to the odd harmonics, we note that the fundamental component has a peak amplitude **$4/\pi$ times the amplitude H of square wave, or $4H/\pi$. The third harmonic has an amplitude one-third of this, or $4H/3\pi$, the fifth harmonic has an amplitude one-fifth of this, or $4H/5\pi$, and so on. This is sum**marized by saying that the amplitudes of the harmonics of a square wave vary inversely as their ORDER.

As to phase, they all pass through zero in a positive direction at the same moment that the square wave itself is passing through zero in a positive direction. This is all depicted in Fig. 15 (A), where it will be observed that the square wave and the first, third, and fifth harmonics all pass through zero simultaneously in a positive direction.

the sawtooth wave shown can also be resolved into a fundamental and infinite number of harmonics. This time even as well as odd harmonics are present and their amplitudes decrease more rapidly than for the square wave, but they too must be of

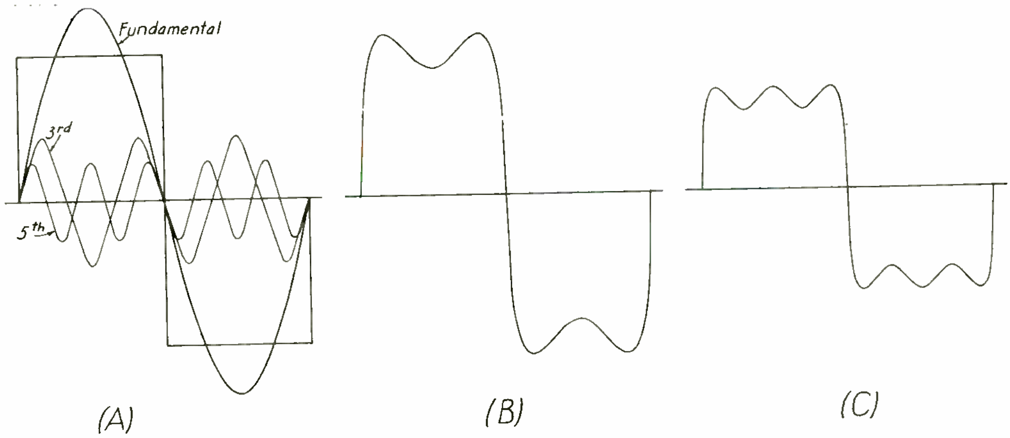



Fig. 15.—Composition of a square wave, showing the first three harmonics, and the approximation to a square wave as the first two and the first three odd-harmonics are combined.

Fig. 15 (B) shows the result of adding the third harmonic to the fundamental; the third harmonic reverses the peak at the center of each half cycle, causing the dip shown and thereby producing a wave that begins to approach a square wave in shape. In (C) is shown the further effect of adding the fi^th harmonic; the sides of the wave bc come steeper, the top becomes flatter with many smaller ripples replacing the few ripples of (B).

As more and more harmonics are added, the wave becomes steeper and steeper on the sides, and flatter and flatter on the top and bottom, with more but finer ripples until in the limit, when an infinite number of harmonics have been included, the wave becomes the square wave shown in (A) and also in Fig. 14.

Referring to Fig. 14 once again,

certain relative amplitudes and phase to produce the sawtooth wave.

The right-hand wave shown in Fig. 14 involves simply two harmonics: a fundamental and a second harmonic. This is the type of wave an overdriven audio amplifier may produce, in that the negative half cycle is more or less clipped. This has been treated previously in the discussion on distortion and its calculation in vacuum-tube am- 'lifiers.

*AMPLIFIER FREQUENCY REQUIRE- MEι.ΊS.*—The question now arises, "What must be the frequency response of an amplifier to pass various types of periodic waves?" In other words, how must the amplifier amplify the various harmonics to preserve the wave shape? Obviously, for best results it must amplify them all equally well to preserve their *relative* magnitudes.

How about phase? An analysis indicates that if the phase shift of each harmonic is in proportion to its frequency, then the output wave will resemble the input wave, and be merely delayed a slight amount in time for the normal phase shifts involved.

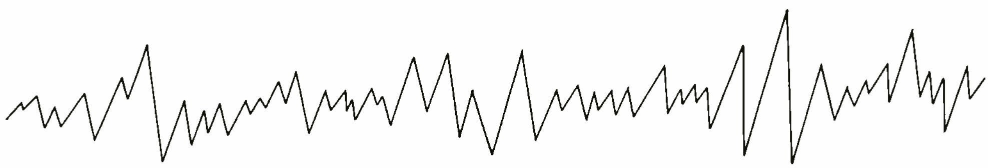


Fig. 16.—An audio wave form, non-periodic in character.

For example, if the fundamental is shifted in phase from the grid or input circuit to the plate or output circuit of a tube by 10°, the second harmonic by $2 \times 10 = 20°$; the third harmonic by $3 \times 10 = 30°$, and so on, then no distortion of the wave will result (provided also that the amplitude relations are maintained), for the various components will line up in the output just as they line up in the input.

However, so far as AUDIO waves are concerned, it is not necessary to obtain this phase relation in an amplifier, because the ear does not seem to care whether the phase shift is in proportion to frequency or not.* As a result, audio amplifier design is simplified considerably over video amplifier design, where phase considerations are at least as important as amplitude considerations.

So far the discussion has centered on periodic wave forms. However, in speech, as well as in television, the input signal may be of a transient nature, and differ from one moment to the next. Fig. 16 illustrates a possible audio wave form. It is non-periodic in nature because it inherently must vary as the words and syllables change in the sentence.

The question arises as to whether such a wave can be resolved into sinusoidal components. The answer is, "Yes, they can." They vary *continuously* from zero frequency or d.c. up to theoretically an infinite frequency; there are no gaps or "holes" in the spectrum as in the case of a periodic wave.

Hence an amplifier that can handle periodic waves should be able to handle transient waves, too. But it is apparent to the student that in actual practise no amplifier can amplify ALL frequencies equally well. The best that can be done is to amplify equally over a certain band or range of frequencies.

*There is some evidence that this is not true in the case of transient (non-periodic) waves, such as the sounds in the tap-dancing, but ordinarily the phase relations are not of importance in audio work.

Fortunately, this is sufficient for all practical purposes. Tests indicate that if an audio amplifier can amplify equally well from say 16 to 20,000 c.p.s., it can reproduce all the sounds that the human ear can hear. This is because the human ear is similarly limited in its frequency range. Similar considerations hold for video amplifiers although here the range is from about 30 c.p.s. to 5 mc., an enormously greater band width.

The problem of frequency response in an audio amplifier is therefore the following: If a source of sinusoidal voltage is used whose frequency can be varied to any value within the audio range, and this voltage is applied at *fixed amplitude* to the input of the amplifier, then the output voltage of the amplifier should be also of constant amplitude over the given range.

The output voltage or the gain $\alpha$ (ratio of output to input voltage) can be plotted as shown in Fig. 17; it is called the frequency response of the amplifier. This is purposely shown as having a certain amount of curvature at the ends of the bandwidth, denoting that the gain falls off at these extremes. This is because actual amplifiers do not have an absolutely flat response over the desired frequency range, and then fall off abruptly, but instead fall off more gradually, as shown.

The reason why audio amplifiers do not have a flat response over an indefinite bandwidth is due primarily to the external impedances such as the load impedance, but also including interelectrode capacitances rather than to the electronic action in the tube itself. The latter action is practically instantaneous until frequencies high in the megacycles are encountered, whereupon transit-time effects come into play.

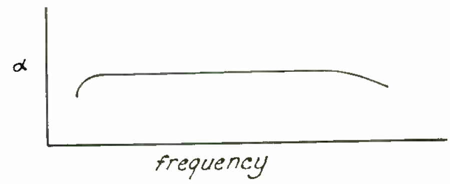



Fig. 17.—Typical frequency response of an audio amplifier.

These will be treated at the appropriate point in the course. For audio amplifiers, however, it is merely necessary to study the behavior of the external impedances with frequency, in order to understand how the resulting frequency response curve comes about.

*THE RESISTANCE-COUPLED AMPLIFIER.*—The first type of amplifier stage to be studied is the resistance-coupled amplifier. It was discussed previously from the graphical viewpoint; this had to do more with the maximum signal output and distortion than with the frequency response.

Fig. 18 (A) shows the circuit, including in dotted lines the interelectrode capacitances which affect mainly the high-frequency response. However, the grid-to-plate capacitance $C_{gp}$ affects mainly the input capacitance of the tube, and can therefore be combined with the grid-to-cathode capacitance $C_{gk}$ plus the stray wiring capacitance.

In the case of a pentode tube, $C_{gp}$ is very small owing to the shielding action of the screen grid, but then the capacity of the control grid to the adjacent screen grid adds to that of the cathode to produce a noticeable increase in the input capacity, and the capacity of the plate to the adjacent screen grid produces an appreciable output capacity.

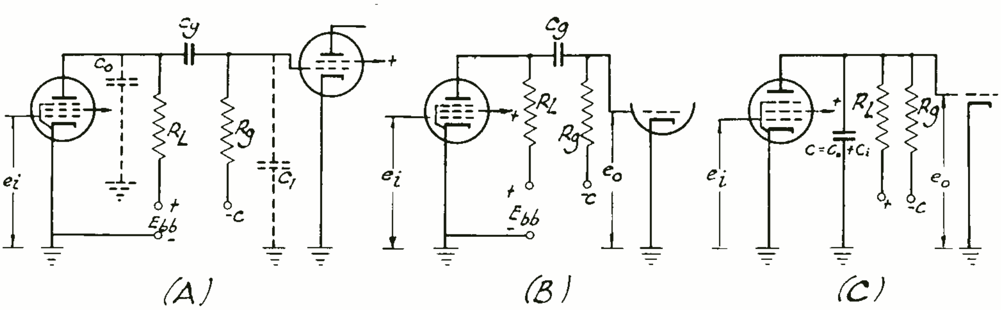



Fig. 18.—Typical resistance-coupled stage, showing the various circuit elements involved.

Hence, let the output capacity be denoted by $C_o$, and the input capacity by $C_1$; these include all the various capacitive components including that of the wiring to ground. At low frequencies their reactance is so high as to constitute a negligible shunt across $R_L$, so that they may be omitted, and the circuit represented as in (B).

At high frequencies, the series reactance of $C_g$ is negligibly small, so that $C_o$ and $C_1$ are essentially in parallel, and constitute a single capacitance $C = C_o + C_1$. Representative values of C are from 20 to 100 μμf; the reactance at 10,000 c.p.s. and higher may constitute an appreciable shunt to $R_L$, particularly if the latter resistance is on the order of 100,000 ohms or higher.

It is therefore convenient to study the behavior of the resistance-coupled amplifier in three steps: its low-frequency response; intermediate-frequency response; and high-frequency response. The low-frequency response will be studied first.

At the low audio frequencies, the reactance of the coupling capacitor $C_g$ prevents the output voltage across the grid resistor $R_g$ from being as large as at the higher frequencies. In other words, the low frequencies are attenuated until at zero frequency, the gain is zero.

This is shown in Fig. 19; the gain drops from its maximum value in the intermediate-frequency range along the curve shown. The action can best be understood by the application of Thevenin's theorem to the equivalent circuit.

In Fig. 20(A) the equivalent circuit is shown: the input grid voltage $e_1$ acts like an equivalent voltage $\mu e_1$ injected in the plate circuit in series with $R_p$. The load is $R_L$ paralleled by $C_g$ in series with $R_g$, and the output voltage is $e_o$ across $R_g$.

Thevenin's theorem states that any linear circuit can be broken into two parts: one containing the generator whose effects are of interest; and the other part containing the terminals across which it is desired to find the voltage or through which it is desired to find the current flow, or both. The first part is considered to be the generator; the second part, the load. The apparent internal resistance of this generator is the impedance that would be measured looking into the terminal at which the break was made; the apparent generated voltage of this generator is that appearing at the same terminals when the second part is disconnected.

This will be clearer when studied in the light of the problem at hand. Referring to Fig. 20, let the terminals, at which the assumed break is to be made, be 1-1. Then all to the left is the apparent generator; all to the right, is the load. The points at which the break is to be made are arbitrary; one chooses those points which will facilitate the solution of the problem.

The apparent generator there-

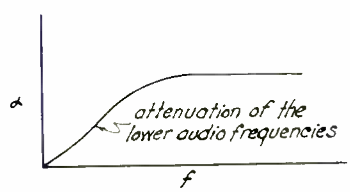



Fig. 19.—The reactance of the coupling capacitor attenuates the lower audio frequencies.

fore contains $\mu e_1$, $R_p$, and $R_L$; the load involves $C_g$ and $R_g$ in series. The internal resistance of the apparent generator or source is that seen looking in to the left from 1-1. The actual generated voltage is ignored in this calculation; there are therefore simply two resistive parallel paths between 1-1; namely, that through $R_L$ and that through $R_p$. Hence the apparent source impedance is $R_L$ and $R_p$ in parallel, or $R_pR_L/(R_p + R_L)$.

The apparent generated voltage is that appearing across $R_L$ when $C_g$ and $R_g$ are not connected. This voltage is simply $\mu e_1$ multiplied by the ratio of $R_L$ to $(R_L + R_p)$, or $\mu e_1(R_L/R_L + R_p)$. This is shown in (B); this is the generator that can be assumed feeding $C_g$ and $R_g$ and developing the output voltage $e_o$ across $R_g$.

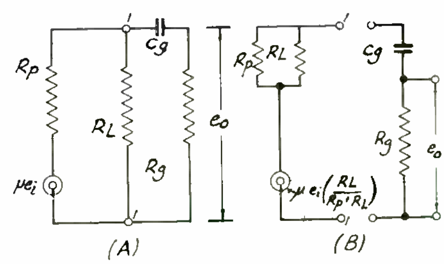


Fig. 20.—Equivalent circuit and application of Thevenin's theorem to it.

Thevenin's theorem has therefore reduced the more complicated circuit of (A) to the simpler series circuit of (B), and has indicated that a simple R-C time constant will be involved. It is quite simple to solve for $e_o$ in terms of $e_1$, and also their ratio $e_o/e_1$ equal to the gain $\alpha$. The result is:

$$\alpha_l = \frac{G_m R_G' \beta}{\sqrt{1 + \beta^2}} \tag{8}$$

and the phase shift is

$$\theta = \tan^{-1}(1/\beta) = \cot^{-1}\beta \text{ (leading).} \tag{9}$$

where $\alpha_l$ represents the gain at low frequencies, $R_G'$ represents $R_p$, $R_L$, and $R_g$ of Fig. 20 in parallel, and $\beta$ is a variable compounded of $2\pi f$ or $\omega$

and the time constant $T_l$ involving $C_g$ and all the resistances connected around it. Thus

$$T_l = C_g\left(R_g + \frac{R_p\ R_L}{R_p + R_L}\right) \qquad (10)$$

and

$$\beta = \omega T_l \qquad (11)$$

At sufficiently high frequencies $\beta^2$ is much greater than unity, so that $\sqrt{1+\beta^2}$ is essentially equal to $\sqrt{\beta^2} = \beta$, whereupon Eq. (8) reduces to

$$\alpha_i = G_m R_G' \qquad (12)$$

where $\alpha_i$ is the gain at intermediate audio frequencies*. This is the region of highest gain; it will be shown subsequently that the shunt capacities begin to pull down the gain once more at the higher audio frequencies.

Hence the gain at either the low or the high end of the band can be expressed in terms of the gain $\alpha_i$ at intermediate frequencies. Specifically, the ratio of low-frequency to intermediate-frequency gain is

$$\alpha_l/\alpha_i = r_l = \beta/\sqrt{1+\beta^2} \qquad (13)$$

This expression can be readily plotted; i.e., $r_l$ can be plotted against $\beta$. As $\beta$ is decreased (which means that for a given value of $T_l$, $\omega$ is decreased), $r_l$ decreases, and the plot is simply the low-frequency response of the resistance-coupled stage in terms of the generalized or normalized variable $\beta$ rather than $\omega$ or f.

*This refers to frequencies from perhaps 500 to 3000 c.p.s. or so, and not to the i-f amplifier of a superheterodyne receiver.

The curve can be used for design purposes, and for any permissible drop in response $r_l$ at any given low frequency $\omega_l/2\pi$, $T_l$ can be computed. However, it is generally preferred to express the drop in response in db rather than numerically as $r_l$. The db attenuation is simply

$$A(\text{db}) = 20 \log r_l \qquad (14)$$

Hence a curve will be plotted between A (db) and $\beta$. This is shown in Fig. 21. As expected, it has the same general shape as the specific frequency-response curve for the stage. This graph is very simple to use, as the following example will show.

Suppose a drop of 2 db is permitted in a resistance-coupled amplifier stage at 40 c.p.s. What time constant is permitted for the stage, and how is this then translated into circuit constants? From Fig. 21, for a 2 db attenuation, $\beta$ is found to be equal to 1.3. Since $\omega_l = 2\pi\,40 = 251$ radians/sec., from Eq. (11), $T_l$ is found to be

$$T_l = \beta/\omega = 1.3/251 = .00518 \text{ sec.}$$

Now suppose a 6SJ7 pentode tube is employed, Fig. 22. Its $R_p$ is over one megohm, (assume, for simplicity, it is one megohm), its $G_m$ = 1650 µmhos, and a value of $R_L$ = 0.5 megohm is chosen. Suppose the tube following this one is also a 6SJ7 type, and a grid resistance $R_g$ of 1 megohm can be used. Then from Eq. (10),

$$C_g = T_l/\left(R_g + \frac{R_p\ R_L}{R_p + R_L}\right)$$

$$= .00518/(10^6 + \frac{1.0 \times 0.5}{1.0 + 0.5} \times 10^6)$$

$$= 0.00389 \ \mu\text{f. or } 0.004\ \mu\text{f.}$$

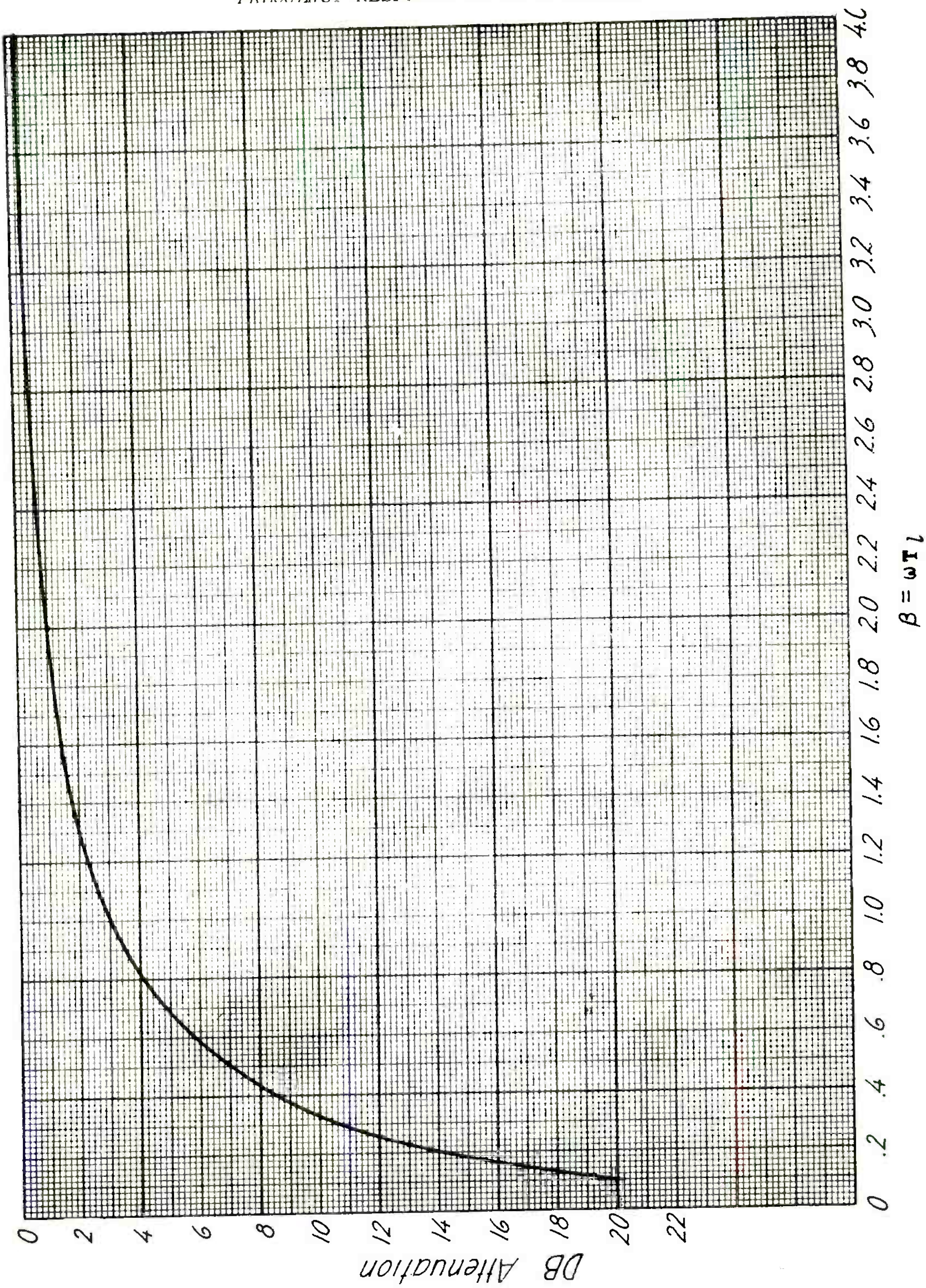



Fig. 21.—Generalized graph of relation between db attenuation at the lower frequencies and $\beta$ for a resistance-coupled amplifier stage.

Such a value of $C_g$ will permit this stage to hold up its response at 40 c.p.s. so that it does not drop by more than 2 db. This, however, depends upon the values of $R_{SG}$ and $C_{SG}$ in the screen circuit, and $R_k$ and $C_k$ in the cathode circuit. Unless these R-C pairs have a sufficiently high time constant, additional low-frequency attenuation will result from these circuits. This will be taken up farther on.

One further point remains, namely the calculation of the phase shift at the specified frequency of 40 c.p.s. (Of course calculations of attenuation and gain can be made at any frequency in exactly the same way as illustrated here.) From Eq. (9), the phase shift is

$$\theta = \tan^{-1}(1/1.3) = \tan^{-1} 0.77$$

$$= 37°35' \text{ leading.}$$

In passing, it is to be noted that the gain at intermediate audio frequencies is, by Eq. (12)

$$\alpha_i = (1650 \times 10^{-6}) \times \left( \frac{1}{\frac{1}{10^6} + \frac{1}{10^6} + \frac{1}{0.5 \times 10^6}} \right)$$

$$= 413$$

The actual measured gain is 238; the difference is due to the fact that the tube has a lower $G_m$ in the actual region of operation on its characteristic curves than the value of 1650 μmhos given.

The actual gain can be either measured experimentally, or determined graphically in a manner described in a previous assignment. Such determination will not be made here because all that it is desired to calculate here is the DROP IN GAIN at 40 c.p.s., rather than the actual gain. The drop in gain is important as determining the fidelity of amplification; the actual gain is usually made more than necessary to provide a factor of safety in this respect.

*HIGH-FREQUENCY RESPONSE.* — The high-frequency gain $\alpha_h$ can also be expressed as a certain fraction of the gain at intermediate-frequencies. Thus, referring to Fig. 23(A), the equivalent circuit at high audio frequencies involves $R_p$, $R_L$, and $R_g$, and the shunt capacitance C discussed previously. The series coupling capacitor $C_g$ is not shown because it

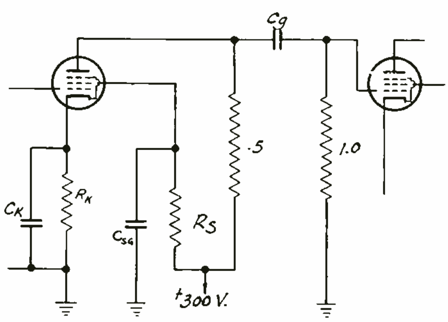



Fig. 22. — Pertinent constants of a resistance-coupled stage used as an example.

is essentially a short circuit in this frequency range.

Once again the circuit can be broken at terminals 1-1 (by Thevenin's theorem) to yield the circuit shown in (B). It is at once noticed that an R-C time constant is involved just as in the low-frequency region. The equivalent source impedance is clearly $R_g$, $R_L$, and $R_p$ in parallel; this was denoted previously as $R'_G$.

The apparent generated voltage $\mu e'_s$ can be readily calculated from $\mu e_s$ in the circuit of (A); however, it is simpler to use the constant-current equivalent circuit shown in Fig. 24, in which the tube is represented as a constant-current generator feeding the constant current $G_m e_s$ into $R_p$, $R_L$, $R_g$, and C, all in parallel.

In the intermediate audio-frequency range, the reactance of C is sufficiently high so that it can be ignored, whereupon the gain becomes that given previously, namely:

$$\alpha_i = G_m R'_G \qquad (12)$$

At the higher audio frequencies, C modifies this in a manner similar to that produced by $C_g$ in

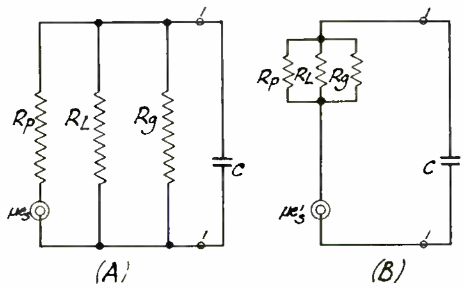



Fig. 23.—Equivalent circuit of the resistance-coupled amplifier at the higher audio frequencies.

the low-frequency range:

$$\alpha_H/\alpha_i = r_H = 1/\sqrt{1 + \omega^2 T_H^2} \qquad (15)$$

The phase shift is

$$\theta_H = \tan^{-1} \omega T_H \text{(lagging)} \qquad (16)$$

where $T_H = C R'_G$

Once again set $\omega T_H = \gamma$, whereupon Eqs. (15) and (16) become

$$r_H = 1/\sqrt{1 + \gamma^2} \qquad (17)$$

and

$$\theta_H = \tan^{-1}\gamma \qquad (18)$$

Then, similarly to Eq. (13), $r_H$ can be plotted against $\gamma$ to provide a generalized curve for facilitating the calculation of the high-frequen-

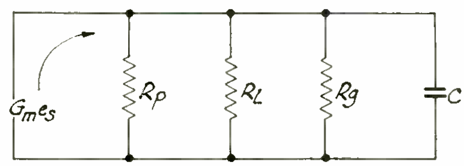



Fig. 24.—Constant current equivalent circuit at high frequencies.

cy response. Instead of using $r_H$, 20 log $r_H$ is used just as in the case of the low-frequency response, and the resulting curve is shown in Fig. 25. It has, of course, a shape exactly similar to a frequency-response curve, since db attenuation is being plotted against a variable $\gamma$ that is proportional to the frequency $\omega$.

To illustrate the use of this curve, let us take the previous example of the 6SJ7 tube. It will be recalled that $R_p$ = 1.0 megohm, $R_L$ was chosen as 0.5 megohm, and $R_g$ as 1 megohm. Assume further that the total capacity C = 25 μμf.

Suppose an attenuation of **0.22** db is permitted at 20,000 c.p.s. Then, from Fig. 25, $\gamma = 0.23 = \omega T_H = 2\pi\, 20000\, T_H$. From this

$$T_H = 0.23/40000\pi = 1.832 \times 10^{-6}$$

But $T_H = CR'_G$, and $R'_G$ is 1.0 megohm, 0.5 megohm, and 1.0 megohm in parallel, or 250000 ohms. Hence $C = 1.832 \times 10^{-6}/.25 \times 10^6 = 7.33$

μμf. This is considerably less than the 25 μμf actually present, and indicates that either a greater db attenuation will have to be accepted at 20,000 c.p.s., or the bandwidth will have to be reduced, or $R'_G$ will have to be reduced.

To see how much db. attenuation 25 μμf. will give at 20,000 c.p.s., first find the corresponding value of $T_H$. It is $25 \times 10^{-12} \times 250000 = 6.25 \times 10^{-6}$ sec. Then $\omega T_H = 2\pi \times 20000 \times 6.25 \times 10^{-6} = 0.785 = \gamma$. For this value of $\gamma$, Fig. 25 indicates that the attenuation will be 2.12 db, which is considerable. If three such stages are employed, for example, the attenuation at 20,000 c.p.s. will be $3 \times 2.12 = 6.36$ db, a sizeable amount.

**To obtain but 0.22 db with 25** μμf capacity, the bandwidth will have to be reduced. Thus, $\gamma$ must equal 0.23. $T_H = 6.25 \times 10^{-6}$ sec. Then, since $\gamma = \omega T_H$, $\omega = \gamma / T_H = 0.23/6.25 \times 10^{-6} = 36{,}800$ rad/sec or

$$f = \omega/2\pi = 36{,}800/2\pi = 5{,}860 \text{ c.p.s.}$$

**Unless the amplifier has many stages, an attenuation of 0.22 db at 5,860 c.p.s. may be acceptable. This depends upon the specifications for the system. If 0.22 db is desired at** 20,000 c.p.s. instead of at 5,860, and C must remain 25 μμf, then the third possibility can be investigated, that of reducing $R'_G$ (which is $R_p$, $R_L$, and $R_g$ in parallel.)

Thus, it was found originally **from Fig. 25 that for 0.22 db attenua**tion, $\gamma = 0.23 = \omega T_H$, and that for $\omega = 2\pi\, 20{,}000$, $T_H = 1.832 \times 10^{-6}$. Then

$$R'_G = T_H/C = 1.832 \times 10^{-6}/25 \times 10^{-12}$$

$$= 73{,}280 \text{ ohms}$$

instead of 250,000. Since $R_p$ cannot be very well altered from its value of 1.0 megohm, and $R_g$ should remain high (preferably at its one-megohm value), $R_L$ will have to be reduced.

Such reduction can readily be accomplished; if $R_L$ is reduced to slightly more than 73,280 ohms, (say 75,000), $R'_G$ will be reduced to the desired value of 73,280 ohms. This in turn will affect the intermediate audio-frequency gain and also the low-frequency response and increase the attenuation in the latter region. However, by increasing $C_g$ by the proper amount, adequate compensation can be obtained.

The phase shift is given by Eq. (18) as

$$\theta_H = \tan^{-1} .23 = 12.97^\circ \text{ (lagging)}$$

$$= 23^\circ 30' \text{ (lagging)}$$

This is the phase shift that will occur at 20000 c.p.s. if either C is reduced from 25 to 7.33 μμf., or $R'_G$ is reduced to 73,280 ohms. Or, this phase shift will occur at 5,860 c.p.s. if C = 25 μμf and $R'_G$ is maintained at 250,000 ohms.

*SCREEN-GRID IMPEDANCE.*—Not only does the control grid cause an a-c or signal component to flow in the plate circuit, but also in the screen grid-circuit. Normally, the screen-grid circuit is bypassed as shown in Fig. 26, but at low frequencies $C_{sg}$ assumes a high reactance, and $R_s$ is normally a high resistance, so that the signal component developes an appreciable signal voltage across these two components in parallel.

This signal voltage is of course 180° out of phase with the input signal voltage applied to the control grid, just as in the case of the plate signal voltage. However, the presence of a signal voltage on

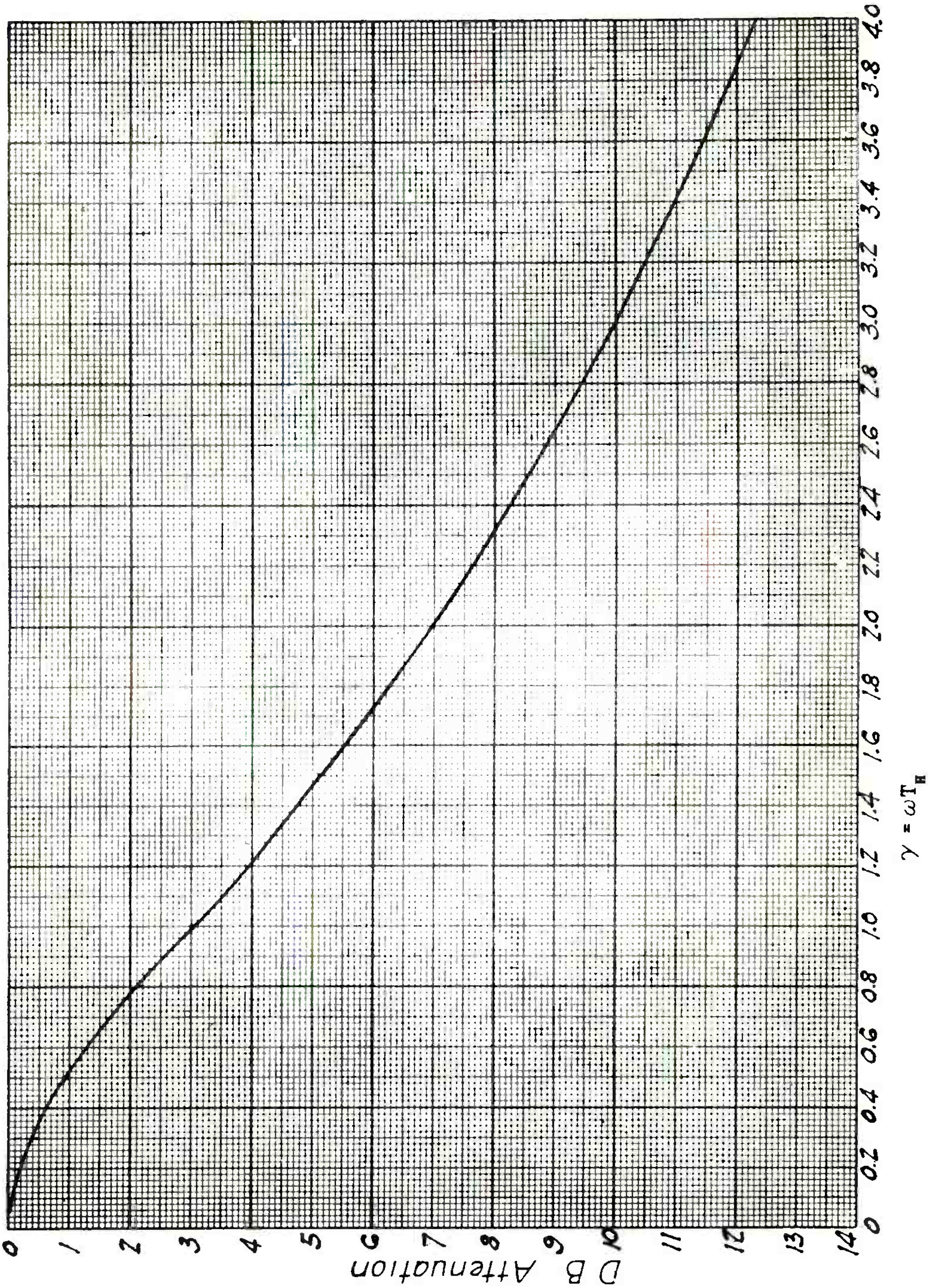



Fig. 25.—Generalized attenuation curve for the high-frequency response of a resistance-coupled amplifier.

the screen grid of reversed polarity tends to oppose the effect of the control grid on the PLATE current, and therefore reduces the output gain. Since the screen signal voltage increases as the signal frequency is decreased, there is present a frequency-selective attenuation; i.e., inadequate bypassing of the screen grid tends to attenuate the lower audio frequencies.

The extent of this low-frequency attenuation has been worked out by Terman and others.* The following formula applies:

$$r_s = \frac{\frac{R_{sg}}{R_s + R_{sg}} + j\,\delta}{1 + j\,\delta} \qquad (19)$$

where $r_s$ is the ratio of the gain at low-frequencies to that at higher frequencies, $R_{sg}$ is the internal cathode-to-screen resistance, and $R_s$ is the external series resistance. The variable $\delta$ is equal to $\omega T_{sg}$, where $T_{sg}$ is the time constant obtained by multiplying the screen bypass capacitor $C_{sg}$ by $R_{sg}$ and $R_s$ in parallel.

As before, Eq. (19) can be plotted to yield generalized information. The quantity $R_{sg}/(R_s + R_{sg})$ can be represented by a single variable $\beta$, whereupon Eq. (19) becomes

$$r_s = \frac{\beta + j\,\delta}{1 + j\,\delta} \qquad (20)$$

In Fig. 27, $r_s$ is plotted against $\delta$, for various values of $\beta$, which is kept constant for any one of the curves. $\beta$ is therefore a parameter in Fig. 27. The phase shift is given by

$$\theta_{sg} = \tan^{-1}\delta/\beta - \tan^{-1}\delta \text{ (leading)} \qquad (21)$$

As a simple example of its use, consider the type 6SJ7 tube previously studied in an amplifier stage. It had an $r_p$ = 1.0 megohm as a pentode. The tube Manual gives its $r_p$

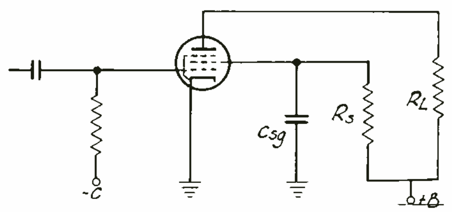



Fig. 26.—R-C circuit elements in the screen-grid circuit affect the low-frequency gain of the stage.

as equal to 7600 ohms when triode connected. As a rough but sufficiently good approximation, assume $R_{sg} = 5r_p = 5 \times 7600 = 38{,}000$ ohms. (The actual value for $R_{sg}$ is unfortunately not given by the tube manufacturer as a general rule.)

The screen potential is given as 100 volts, and the screen current as 0.8 ma. Hence the screen dropping resistor is

$$R_s = \frac{300-200}{0.8 \times 10^{-3}} = 250{,}000 \text{ ohms.}$$

Then $\beta$ = 38000/(250,000 + 38000) = 0.132. Suppose 20 log $r_s$ is to equal 0.2 db at 40 c.p.s. (This attenuation is in addition to that caused by the grid-coupling time constant.) Now refer to Fig. 27. There is no curve for $\beta$ = 0.132, but for 20 log $r_s$ = 0.2 db, all the

*Terman, Hewlett, Palmer, and Pan, "Calculation and Design of Resistance-Coupled Amplifiers Using Pentode Tubes," Trans. A.I.E.E. Vol. 59, p. 879, 1940.

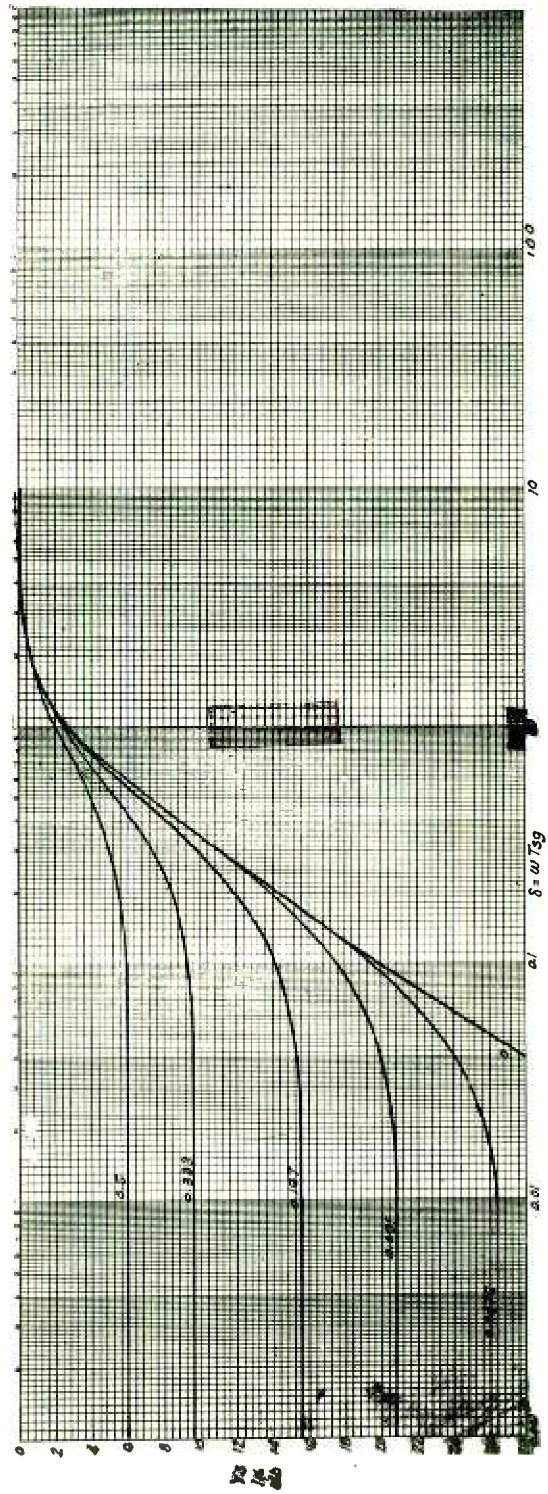



Fig. 27.—Generalized curves giving the low-frequency attenuation produced by the screen bypass capacitor.

curves are practically coincident, and yield a value of $\delta$ of about 3.0, within the accuracy to which the curves can be read. Then $\omega T_{sg} = 5$, and since $\omega = 2\pi\ 40$ rad/sec.,

$$T_{sg} = 3/(2\pi\ 40) = 0.01193 \text{ sec.}$$

Since the resistance factor in $T_{sg}$ is $R_{sg}$ and $R_s$ in parallel, or

$$R = \frac{38,000 \times 250,000}{250,000 + 38,000} = 33,000 \text{ ohms,}$$

then

$$C_{sg} = T_{sg}/R = 0.01193/33,000$$

$$= 0.361\ \mu f.$$

A 0.5 $\mu f$ capacitor should be used.

The phase shift is given by Eq. (21) and is equal to

$$\theta_{sg} = \tan^{-1}(3/.132) - \tan^{-1} 3$$

$$= \tan^{-1} 22.8 - \tan^{-1} 3$$

or

$$\theta_{sg} = 87°30' - 71°30' = 16°$$ (leading)

*SELF-BIAS IMPEDANCE.*—In fig. 28 is shown a cathode self-bias resistor, $R_k$. The plate current $i_p$ flows through the load resistor $R_L$ and $R_k$ in series. Across $R_L$ it develops the output voltage $e_L$, and across $R_k$, it develops a voltage $e_f$. Assuming an electron flow up through $R_k$ to the cathode, it is clear that the cathode will be POSITIVE to ground, and hence to the grid, which is connected to ground through its resistor $R_g$ or similar circuit element. In other words, the grid is biased NEGATIVE to the cathode.

Now when an a-c or signal voltage $e_s$ is impressed between the grid and GROUND, a similar a-c component is developed in $i_p$. As a result an a-c or signal component is developed in $e_L$, and ALSO IN $e_f$. This signal component of $e_f$ is OPPOSITE IN PHASE to $e_s$, and hence opposes it. This represents negative feedback, which will be discussed more fully in the following assignment.

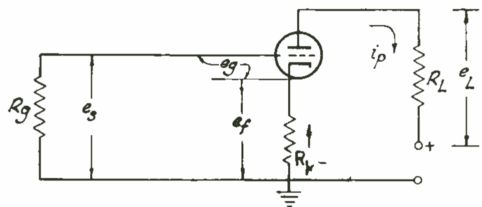



Fig. 28.—Circuit conditions when a self-bias resistor is used in series with the cathode.

As a result of this opposing action, the ACTUAL or NET voltage appearing between the grid and cathode is not the applied signal voltage $e_s$, but the smaller value $e_g$, where $e_g = e_s - e_f$. This in turn means that the output voltage $e_L$ will be less for a given applied $e_s$ than would be the case if $R_k$ were omitted; in short, the gain of the stage has been reduced.

In order to restore the gain to its higher value when $R_k$ is absent, a bypass capacitor can be connected across $R_k$, thereby making the a-c impedance between cathode and ground much lower than the d-c resistance $R_k$. The result is that d-c grid bias is still obtained, but no a-c or signal negative feedback and consequent reduction in gain.

Unfortunately, the reactance of

a capacitor is high at low frequencies. This means that the bypassing effect and reduction in negative feedback will not be as much at the lower audio frequencies as at the higher audio frequencies. The result is an attenuation of the lower frequencies or "drooping" of the frequency response curve at the "low" end.

This "droop" can be reduced to any desired degree by employing a sufficiently large cathode bypass capacitor (call it $C_k$). In the analysis previously cited, in which Terman and his associates calculated the effect of the screen bypass capacitor, he has also calculated the effect of the cathode bypass capacitor.

The results will be given here in slightly modified form. Thus

$$\frac{\text{actual output voltage}}{\text{Output voltage with zero bias inpedance}} = r_b$$

$$= \frac{\sqrt{1 + \omega^2 T_k^2}}{\sqrt{(1 + g_m R_k r_s)^2 + \omega^2 T_k^2}} \qquad (22)$$

where $g_m$ is the transconductance of the tube, $R_k$ is the cathode bias resistor, $T_k$ is the cathode time constant equal to $C_k R_k$, $\omega$ is the angular frequency ($= 2\pi f$), and $r_s$ is the ratio of the actual output voltage to the output voltage with zero SCREEN impedance, as defined in Eq. (20). Note that the required bias bypass capacitor (involved in $T_k$) will therefore depend upon how well the screen circuit is bypassed (to the cathode).

Let the quantity $\omega T_k$ be denoted by $\eta$. Further, let us use the db *attenuation* instead of the voltage ratio $r_b$; i.e., let us use 20 log $r_b$. Then Eq. (22) can be rewritten as

$$A = 20 \log r_b$$

$$= 10 \log \frac{1 + \eta^2}{(1 + g_m R_k r_s)^2 + \eta^2} \qquad (23)$$

and the phase shift produced by $C_f$ as

$$\theta_f = \tan^{-1} \eta - \tan^{-1} \eta/(1 + g_m R_k r_s) \qquad (24)$$

Eq. (23) involves the variables A, $\eta$, and $(g_m R_k r_s)$. In Fig. 29, A has been plotted versus $\eta$, with $g_m R_k r_s$ acting as a parameter, and taking on the values 0.5, 0.8, 1.0, 1.5, 2, 3, 4, 5, 6, 7, 8, and 10, thus giving rise to the family of curves shown. The left-hand family is the plot for values of $\eta$ from 0.1 to 15; the right-hand family is the plot on a larger scale for values of $\eta$ from 15 to 100, so as to improve the accuracy in this region.

An example will make the use of these curves clear. Thus, assume as before the same 6SJ7 tube. It requires a bias of -3 volts, and draws a plate current of 3 ma, and a screen current of 0.8 ma (according to the Tube Manual). The required value of $R_f$ is therefore

$$R_k = \frac{3}{.0038} = 790 \text{ ohms},$$

on the basis that both the plate and screen d-c components flow through $R_k$.

Assume that an attenuation A of 0.2 db is permitted at 40 c.p.s. In the previous example for the screen grid bypassing, the attenuation was also assumed to be 0.2 db; this represented -20 log $r_s$. Hence $r_s$ can be calculated as follows:

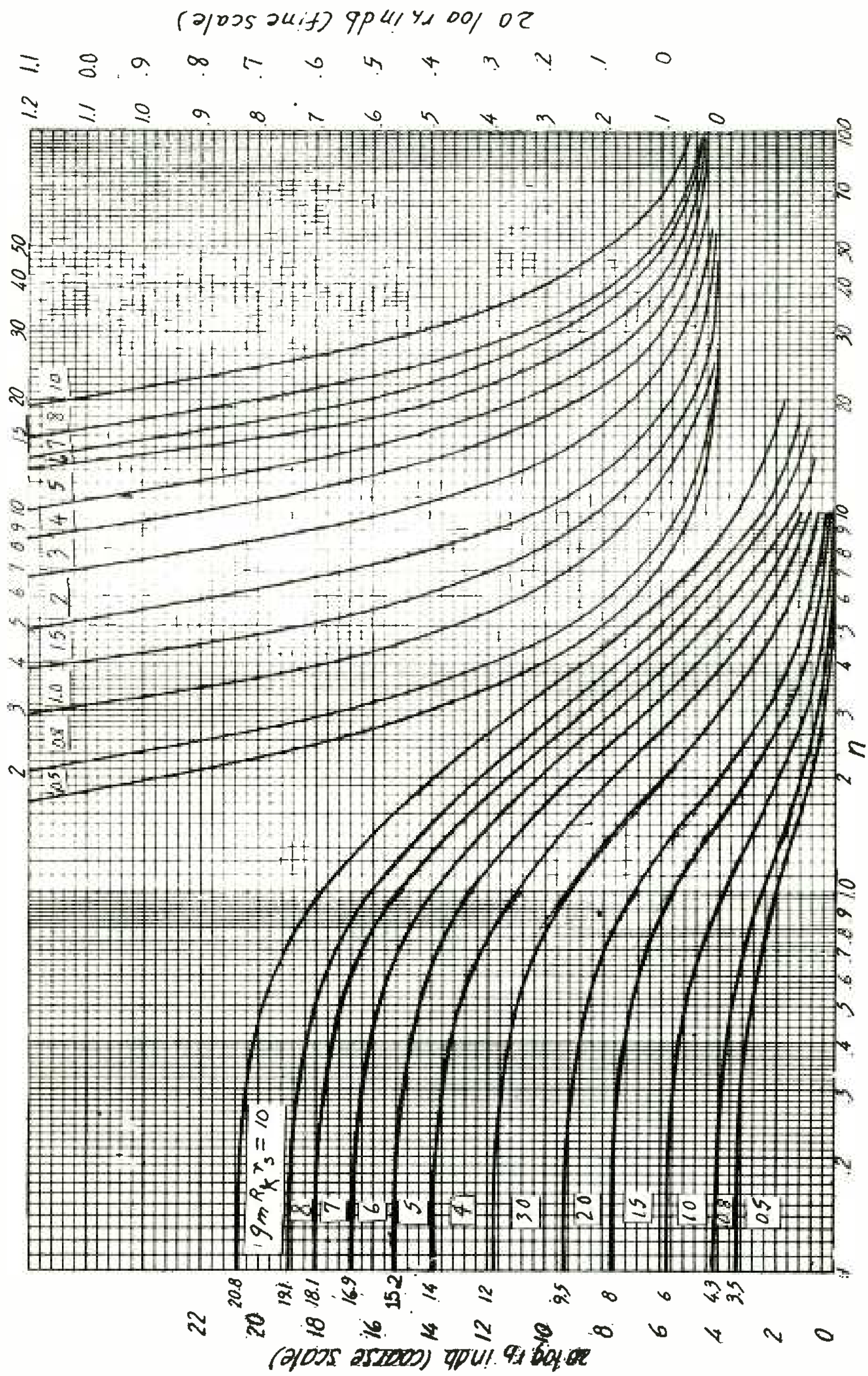


Fig. 29.—Family of curves giving relation between the db. attenuation produced by inadequate bypassing of the cathode bias resistor, and the variable $\eta = \omega T_f$ for various values of a parameter $g_m R_k r_s$.

$$r_s = \frac{1}{\text{anlog}\,\frac{0.2}{20}} = \frac{1}{\text{anlog}\,.01} = 0.977$$

**Now $g_m R_k r_s$ can be calculated; its value is**

$1650 \times 10^{-6} \times 790 \times 0.977 = 1.272$

or approximately 1.25

From Fig. 29, for 0.2 db, one can interpolate about halfway between the right-hand curve labeled 1.0 and the curve labeled 1.5. The value of $\eta$ obtained is 9.6. Since

$$\eta = \omega T_k = 9.6, \text{ and } \omega = 2\pi 40,$$

$$T_k = 9.6/2\pi 40 = 0.0382 \text{ sec.}$$

**Then**

$$C_k = T_k/R_k = 0.0382/790 = 48.3\ \mu f.$$

or about 50 μf.

This is a large value, but since a very low voltage is involved (3 volts), a low-voltage electrolytic capacitor can be employed that is neither too expensive nor too bulky. Electrolytic capacitors of from 500 to 1,000 μf. are employed in video (television) amplifiers for bypassing the cathode bias resistor because very small attenuation, and particularly phase shift, are desired.

The phase shift is, by Eq. (24):

$$\theta_k = \tan^{-1} 9.6$$

$$-\tan^{-1} 9.6/(1 + 1.272)$$

$$= 84°3' - 76°42' = 7°21'$$

This is a small phase shift, in harmony with the small attenuation of 0.2 db specified.

## TRANSFORMER COUPLING

*INPUT AND INTERSTAGE TRANSFORMERS.*—In an earlier assignment on mutual inductance, the iron-core type of transformer was analyzed and its frequency response discussed. A brief review will be given here before taking up the properties of the input transformer.

All transformers involving coupling between two windings can be represented by an equivalent tee configuration, as illustrated in Fig. 30. Here $R_{pw}$ and $R_{sw}$ represent the primary and secondary winding resistances, respectively; $C_p$ and $C_s$, the distributed capacitances of the two windings, $C_m$ represents a mutual capacity between the primary and secondary windings,* which can usually be regarded as so much additional capacity across the secondary. The core losses are represented by $R_c$; the load impedance by $Z_L$; the turns ratio by a; and the mutual inductance, by $L_m$.

The latter represents the flux of the primary which also links the secondary. The primary flux which does *not* link the secondary is the primary *leakage* flux, and is denoted by $L_{pL}$; similarly, the secondary leakage flux is denoted by $L_{sL}$. The ideal transformer of turns ratio 1 : a is designated by I.T.

The equivalent tee representation is admittedly a more complicated appearing configuration than the actual representation of the two

*The presence of $C_m$ makes the equivalent circuit actually a bridged tee rather than an ordinary tee configuration.

winding transformer, but has the advantage of bringing all the hidden effects out into the open, as it were. Furthermore, it can be considerably simplified by studying its frequency response in three separate parts: the low-frequency, the intermediate frequency, and the high-frequency response. This is analogous to the treatment of the resistance-coupled amplifier. However, the transformer exhibits certain parallel and series resonant effects that are not present in the resistance-coupled amplifier, as will be apparent in the discussion that is to follow.

In Fig. 31, (A), (B), and (C) are shown the configurations to which the tee representation reduces at the low-, intermediate-, and high-frequency ends of the spectrum. At low frequencies, shunt capacitances are negligibly high in reactance, series inductances are negligibly low in reactance, and hence both can be omitted. There remains then only the winding resistances and the mutual inductance $L_m$. At intermediate frequencies even $L_m$ has a negligibly high reactance, and can be omitted; this defines the intermediate-frequency range. At the high frequencies the shunt capacitances and series leakage reactances must be taken into account, but $L_m$ can still be omitted.

*LOW-FREQUENCY RESPONSE.* — Consider the low-frequency end first. For convenience, all secondary impedances can be shifted to the primary side by first dividing by $a^2$ to obtain the reflected values.

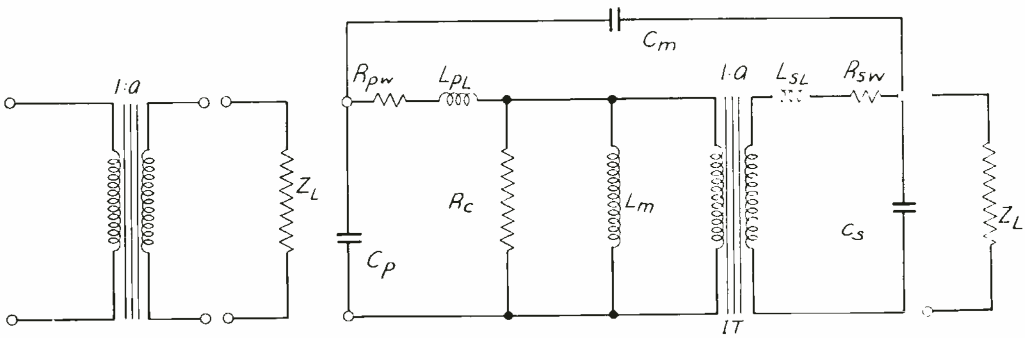



Fig. 30.—Two-winding transformer and its tee equivalent.

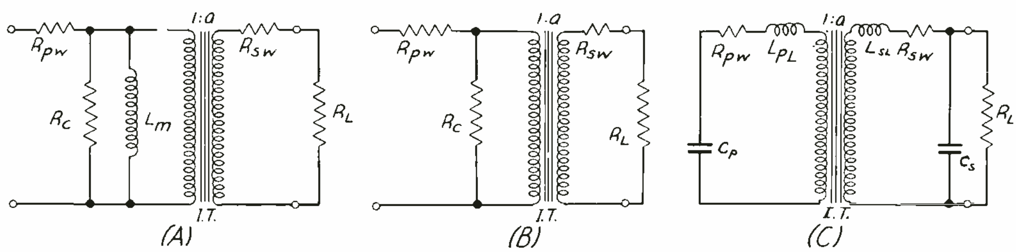



Fig. 31.—The tee representation simplifies down to the above configurations at the low-, intermediate-, and high-frequency ends of the spectrum.

This is shown in Fig. 32 (A). Now Thevenin's Theorem can be applied to combine all the resistors into one representing the internal resistance $R_G'$ of an equivalent generator, whose generated voltage is denoted by $e_G'$ and is a function of the resistance values and of the actual generated voltage $e_G$. This is shown in Fig. 32 (B).

This voltage, being obtained by a voltage divider action of pure resistors from $e_G$, is independent of frequency if $e_G$ is so assumed. But the output voltage $e_o$ across $L_m$ is not, because at low audio frequencies (say from 100 c.p.s. and below), $L_m$ draws appreciable magnetizing current from the source, producing an appreciable voltage drop in the ap-

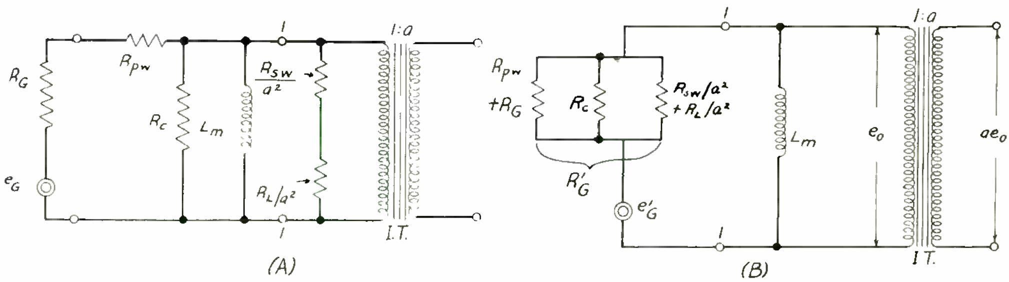



Fig. 32.—A further equivalent circuit at low audio frequencies, and the application of Thevenin's Theorem to this circuit.

Thus, the equivalent source impedance is $(R_{pw} + R_G)$, $R_c$, and $(R_{sw}/a^2 + R_L/a^2)$ in parallel, as is clear from Fig. 32 (A), by interchanging the positions of $L_m$ and $(R_{sw}/a^2 + R_L/a^2)$, and then looking back to the left into terminals 1-1. Call this apparent impedance $R_G'$. The apparent generated or open-circuit voltage $e_G'$ across 1-1 is then the fraction of $e_G$ that appears across $R_c$ and $(R_{sw}/a^2 + R_L/a^2)$ in parallel. Thus

parent source impedance $R_G'$ and thereby reducing $e_o$ below the apparent generated voltage $e_G'$. The output voltage is then $ae_o$.

The resulting frequency response is shown in Fig. 33. The ratio of output voltage $ae_o$ to equivalent generated voltage $e_G'$ has been plotted against frequency. As can be seen, the curve rises from zero to a maximum value of a, the turns ratio. The latter value is reached at a

$$e_G' = e_G \frac{\dfrac{R_c\,(R_{sw}/a^2 + R_L/a^2)}{R_c + R_{sw}/a^2 + R_L/a^2}}{R_{pw} + R_G + \dfrac{R_c\,(R_{sw}/a^2 + R_L/a^2)}{R_c + R_{sw}/a^2 + R_L/a^2}}$$

$$= e_G \frac{R_c(R_{sw}/a^2 + R_L/a^2)}{(R_c + R_{sw}/a^2 + R_L/a^2)(R_{pw} + R_G) + R_c(R_{sw}/a^2 + R_L/a^2)} \qquad (25)$$

frequency which terminates the low-frequency region and starts the intermediate-frequency region.

The range of the low-frequency region depends upon the quality of the transformer; a transformer flat to within 1 db down to 30 c.p.s. may be regarded as having a low-frequency range from zero to perhaps 40 c.p.s., whereas a transformer flat to within 1 db down to 100 c.p.s. might be regarded as having a low-frequency range from zero to perhaps 150 c.p.s. or so.

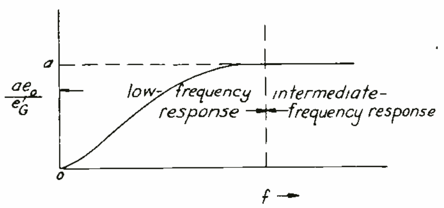



Fig. 33.—Low-frequency response curve of an input transformer.

The low-frequency gain is given by the following formula:

$$ae_o/e_G' = \frac{2\pi fL_m a}{\sqrt{(R_G')^2 + (2\pi fL_m)^2}} \quad (26)$$

$$= \frac{a\omega(L_m/R_G')}{\sqrt{1+\omega^2(L_m/R_G')^2}} = \frac{a\,\omega\,T_T}{\sqrt{1+\omega^2T_T^2}}$$

where $T_T$ is a time constant equal to $(L_m/R_G')$, and also where $e_G'$ in turn is given in terms of $e_G$ by Eq. (25). Eq. (26) indicates how the gain will vary with frequency; at f = 0, the numerator is zero and the denominator becomes $R_G'$, so that the fraction becomes zero, as was shown in Fig. 33, and if f is sufficiently high, so that $\omega^2T_T^2$ is much greater than unity, the gain approaches $a\,\omega T_T/\omega T_T = a$.

Eq. (26) therefore indicates that if the gain is to be high (close to a) at low values of $\omega$ (frequency), then $T_T$ must be high; i.e., $L_m$ must be large compared to $R_G'$. For a given apparent source resistance $R_G'$, this can be accomplished either by using a large mutual inductance $L_m$, such as by using a lot of iron of high permeability in the transformer and/or a large number of turns. This of course makes for an expensive transformer.

For a given transformer and magnitude of $L_m$, a better low-frequency response can be obtained by decreasing $R_G'$. Reference to Fig. 32 (B) shows that if the actual source impedance $R_G$ is fixed, $R_G'$ can still be made low by making $R_L$ and hence $R_L/a^2$ low*.

In other words, by putting a sufficiently low resistance $R_L$ across the secondary, the low-frequency response can be improved. Alternatively, a resistor of magnitude $R_L/a^2$ can be connected across the primary to give the same result. Later on it will be shown that the effects on the high-frequency response are exactly opposite, but so far as the low-frequency response is concerned, both connections flatten it. However, one disadvantage of this procedure is that $e_G'$ is reduced, as will be evident from Eq. (25).

*The secondary winding resistance $R_{sw}$, as well as $R_{pw}$, are usually negligible compared to the other resistances associated with them.

where the numerator decreases more rapidly than the denominator as $R_L/a^2$ is reduced. Thus, flatness of frequency response is obtained at the expense of gain, as is shown in Fig. 34.

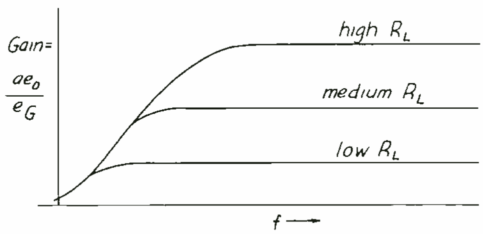



Fig. 34.—The low-frequency response is improved at the expense of gain as the secondary load resistance $R_L$, is reduced.

*INTERMEDIATE-FREQUENCY RESPONSE*.—In the intermediate-frequency range the equivalent circuit is quite simple, and is shown in Fig. 35. Here $e_o$ is equal to $e_G'$, as given by Eq. (25), and the output voltage is therefore $ae_o = ae_G'$, a value independent of frequency. If $R_L = \infty$ (no load across the secondary), and $R_c$ is very high, such as 200,000 ohms or so, then $e_G' \cong e_G$, and the output voltage of the transformer is the *generated* voltage of the source multiplied by the turns ratio, or $ae_G$. In this range the actual transformer approaches most closely to the ideal transformer; the range extends to 3,000 or even 10,000 c.p.s. or better, depending upon the excellence of the transformer.

*HIGH-FREQUENCY RESPONSE*.—At the higher frequencies, the leakage inductances and distributed and shunt capacitances cannot be ignored. Of the latter, the primary capacitance $C_p$ can usually be neglected, but the secondary capacitance $C_s$ (which can also be regarded as including $C_m$) cannot be ignored.

The equivalent tee circuit

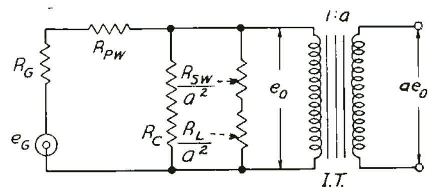



Fig. 35.—Equivalent circuit in the intermediate-frequency region.

therefore approximates that shown in Fig. 36 (A). Usually the core-loss equivalent resistance $R_c$ can be ignored, too, and after referring all secondary impedances to the primary side, Fig. 36 (B) is obtained. Here it is apparent that a series-resonant circuit is involved (ignoring $R_L/a^2$ for the moment). As such, a peak will be obtained in the high-frequency response, whose height will be determined by how low the source resistance $(R_G + R_{pw} + R_{sw}/a^2)$ is.

The resonant frequency, for $R_L = \infty$, is simply

$$f_r = \frac{1}{2\pi\sqrt{(L_{pL} + L_{sL}/a^2)a^2C_s}} \tag{27}$$

If the Q of the circuit at resonance, or $Q_r$, is equal to unity, i.e.,—

$$Q_r = \frac{2\pi f_r(L_{pL} + L_{sL}/a^2)}{(R_G + R_{pw} + R_{sw}/a^2)} = 1 \tag{28}$$

then there is no resonant peak in

the response, and the curve is flat. If $Q_r$ is greater than unity, there is a peak; if Q is less than unity, the response drops off. This is clear from Fig. 37; by proper design a transformer can be made to have a flat response over the range of frequencies of interest.

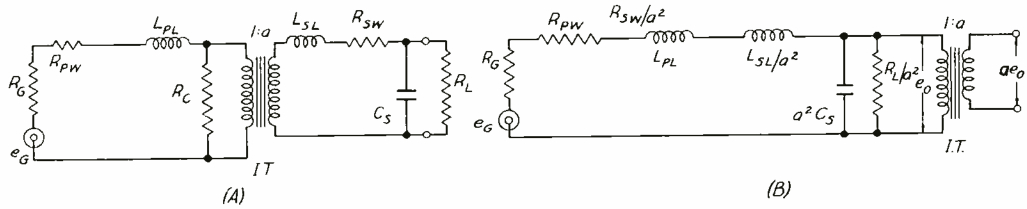



Fig. 36.—Configuration representing the transformer at the high-frequency end of the audio spectrum.

When the secondary is loaded with $R_L$, the capacitor $C_s$ is shunted by $R_L$, and this serves to lower the circuit Q in addition to the lowering of the circuit Q by the SERIES resistance of the circuit. Thus, the unloaded transformer may show a resonant peak, and the loaded transformer a droop in response. To minimize such variations from the unloaded to the loaded case, $C_s$ should be relatively large and the total leakage inductance should be relatively small for a given resonant frequency. However, if the response is to be flat up to a very high frequency, say 20,000 c.p.s. or more, then $f_r$ must be around 20,000 c.p.s., so that from Eq. (27) both the leakage inductance and $C_s$ must be small.

It can now be seen what effect loading the primary with a resistance instead of the secondary will produce. In the low-and intermediate-frequency ranges the effects were the same. But at the high frequency end, if the primary is actually loaded, the circuit assumes the configuration shown in Fig. 38 instead of Fig. 36. (In Fig. 38, $R_L$ should be numerically equal to $R_L/a^2$ in Fig. 36, for the same low-and intermediate-frequency response.)

If Thevenin's Theorem be applied to the circuit of Fig. 38 (A), and all the resistances combined to give an equivalent source impedance $R_G'$, the circuit of Fig. 38 (B) will be obtained. It is immediately evident that the source impedance has been reduced from $R_G'$ to $R_G$ and $R_L$ in PARALLEL, or $R_G'$ is less than $R_G$.

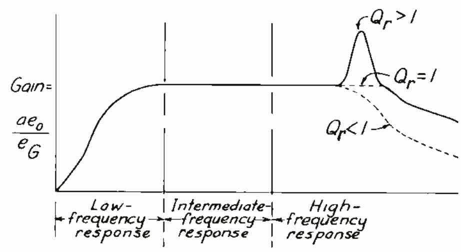



Fig. 37.—The peaking at the high end depends upon the Q of the series resonant circuit.

As a result the $Q_r$ of the circuit is RAISED, and a PEAK in the high-frequency response will be obtained instead of a droop. This is the important difference between loading the secondary or the primary of a transformer; the high-frequency response is oppositely affected.

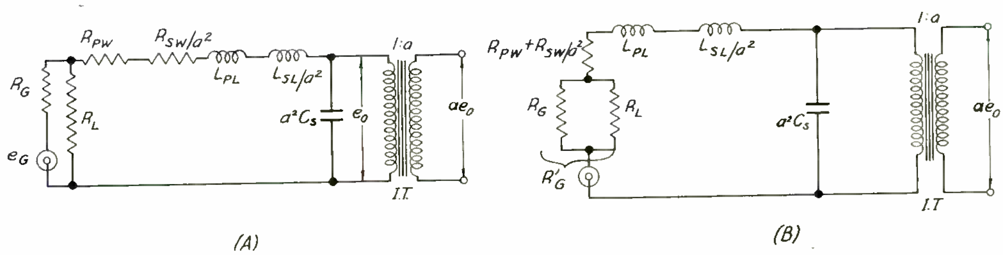



Fig. 38.—Circuit configuration when the primary instead of the secondary side is loaded with a resistance $R_L$.

Note from Fig. 36 that if the secondary is loaded, the reflected resistance $R_L/a^2$ is separated from $R_G$ by the intervening series leakage inductance, and hence cannot be combined with $R_G$, as it can at the low-frequency end of the spectrum where the leakage reactance is negligible. It is this fact that makes the high-frequency behavior for the two kinds of loading different, whereas the low-frequency behavior is identical.

No numerical computations have been given because the transformer constants cannot be very readily changed, and the values can be inferred from the frequency-response curve.

However, one practical case can be cited as an example: the adjustment of the frequency response by judicious primary and/or secondary loading. Suppose the response drops off too much at the low end, and is flat (no peak) at the high end.

This is illustrated by the solid line in Fig. 39. It is desired to raise the response at the low end as indicated by the dotted line. If the primary side were loaded by connecting a suitable resistance across its terminals, the low-frequency end of the curve would be raised, but an unwanted peak would appear at the high-frequency end owing to the increase in the circuit Q by primary loading.

On the other hand, were the

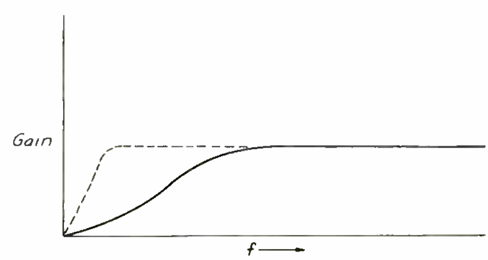



Fig. 39.—Extending the low-frequency response of a transformer by proper loading of the primary and secondary windings.

secondary loaded, the low-frequency response would again be raised, but a droop in the high-frequency response would occur owing to the decrease in the circuit Q. Therefore, it is necessary to connect *higher* resistances across BOTH the primary and secondary, of such relative values that they balance one another and maintain the unity Q that the unloaded transformer possesses, and at the same time cooperate to raise the low-frequency response. This is preferably a matter of cut-and-try, although the proper values for the resistors could be calculated once the transformer constants were measured.

*PRACTICAL EXAMPLES.*—The discussion up to now concerned itself with a source impedance $R_G$ and a generator voltage $e_G$. What do $R_G$ and $e_G$ refer to in practice? The answer is, "To a variety of things." Fig. 40 shows a possible arrangement of components in an audio system.

A ribbon microphone, having an internal impedance of but a fraction of an ohm, feeds a step-up transformer $T_1$, on whose secondary side the internal impedance of the ribbon appears as 500 ohms.

The line connecting $T_1$ and the input transformer $T_2$ is therefore known as a 500-ohm line. Thus, $R_G$ for $T_2$ is 500 ohms, and $e_G$ is the voltage appearing across terminals 1-1 when $T_2$ is disconnected. Transformer $T_2$ then steps up the impedance from 500 ohms to perhaps 150,000 ohms. This means that 500 ohms connected to the primary side appears as 150,000 ohms on the secondary side; conversely, 150,000 ohms connected to the secondary appears as 500 ohms on the primary side.

However, ordinarily the secondary is not loaded, so that looking into the primary of $T_2$, one sees merely the transformer internal impedance, which is very high in the audio-frequency range. If the 500-ohm line were very long, such as a telephone line, it would be necessary to either connect 500 ohms across the primary of $T_2$, or 150,000 ohms across the secondary, in order to terminate the line and prevent reflections from its far end with consequent peaks and dips in the frequency response.* For short runs, however, such as in a studio, no terminating resistance is required.

Indeed, no termination of 150,000 ohms is desired for $T_2$.

*Actually a microphone line can only be run for 50 to 100 feet or so, owing to the noise pickup, which would "swamp out" the weak microphone signal.

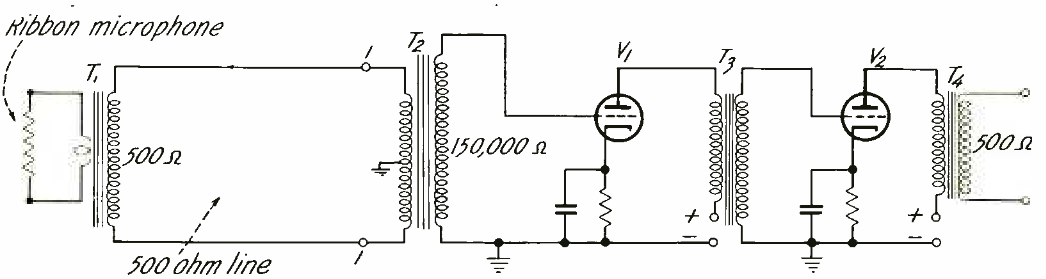



Fig. 40.—Ribbon microphone and pre-amplifier, showing various types of transformers involved.

This is because of signal-to-noise considerations. As will be discussed more fully farther on in this course, all resistors generate noise voltages owing to the thermal (heat) agitation of their free electrons. This noise voltage is proportional to the SQUARE ROOT of the band width, certain constants, the absolute temperature, and the magnitude of the resistance. Or, to put it another way, the SQUARE of the R. M. S. value of the noise voltage is proportional to the resistance R, as well as other factors.

Refer now to Fig. 41. Suppose $R_L$ is matched to $R_G$, which means $a^2R_G = R_L$. Then looking into terminals 1-1 an apparent source resistance of $a^2R_G = R_L$ will be seen; i.e., the apparent source resistance is equal to the load resistance $R_L$.

With $R_L$ disconnected, $e_G$ appears as $ae_G$ across terminals 1-1. When $R_L$ is connected, the voltage across 1-1 drops to $ae_G/2$, since half of $ae_G$ is consumed in the apparent source resistance $a^2R_G$, and half is available across $R_L$. In short, connecting $R_L$ reduces the voltage applied to the grid of the next tube to one-half.

Now consider the noise voltage. When $R_L$ is not connected, the resistance generating thermal noise, as viewed from terminals 1-1, is a $a^2R_G$ (= $R_L$). Let the thermal noise have a magnitude $e_n$. Now suppose $R_L$ is connected. Now the resistance seen looking into terminals 1-1 is $R_L$ and $a^2R_G$ in parallel, or $a^2R_G/2$.

Since the resistance is halved, the thermal noise voltage is reduced by a factor, not of 1/2, but of $\sqrt{1/2}$ or 0.707. It therefore now has the magnitude of 0.707 $e_n$. Consider now the signal-to-noise voltage ratio in each case. With $R_L$ disconnected, it is:

$$S_1/N_1 = ae_G/e_n$$

With $R_L$ connected, it is:

$$S_2/N_2 = \left(\frac{ae_G}{2}\right)/(.707e_n) = \frac{\sqrt{2}}{2}\frac{ae_G}{e_n}$$

$$= 0.707\ (ae_G/e_n)$$

Thus, by connecting $R_L$, the signal-to-noise ratio has been reduced to 70.7 per cent of its value with $R_L$ disconnected.

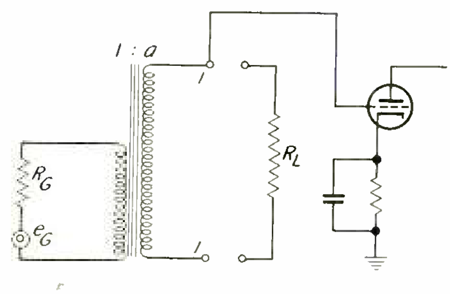



Fig. 41.—If the secondary loading resistance $R_L$ is omitted, the signal-to-noise ratio is improved.

It is therefore desirable to design an input transformer to have the required frequency response without secondary (or primary) loading, and thus to feed it from the source. This applies to amplifiers operating at a low input level, such as the pre-amplifier shown in Fig. 40. The amplifier following the pre-amplifier can have its input transformer loaded if it is desired for flatness of frequency response, since the signal level is usually sufficiently high at this point to override any thermal noise.

Another point that may arise is as to the value of an input transformer. Its value is that of contributing to the gain of the amplifier, as indicated previously, and with practically no contribution of noise and little loss. If the microphone were connected direct to the grid of $V_1$ in Fig. 40, the voltage would be small and considerable amplification would be required.

Unfortunately this results in a reflected secondary capacitance $a^2C_s$ that is so large that the series resonance frequency is decreased to a relatively low value. In short, the high-frequency response is curtailed, so that the transformer cannot cover the band width desired.

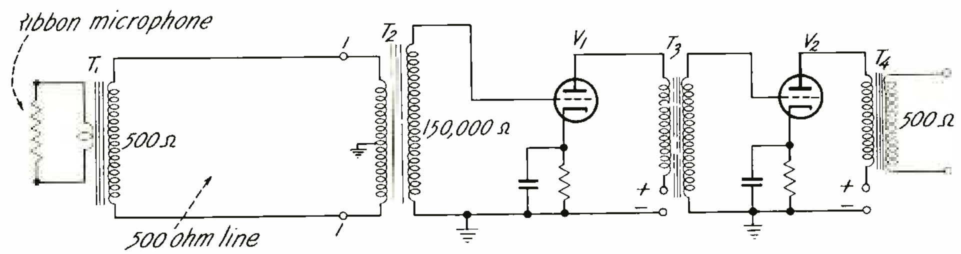



Fig. 40.—Ribbon microphone and pre-amplifier, showing various types of transformers involved.

If a transformer is used, or two transformers such as $T_1$ and $T_2$, the voltage is stepped up. For example, for an impedance step-up of from 500 to 150,000, the turns ratio is

$$a = \sqrt{\frac{150000}{500}} = 17.32:1$$

This means that the signal voltage is stepped up by a factor of over 17 before being applied to the grid of V1; this is practically the gain of an ordinary voltage amplifier stage. Even more step up is obtained if one starts with a very low impedance source, such as the ribbon microphone shown in Fig. 40.

It would therefore appear to be desirable to have as high a step-up a as possible. In other words, transformer $T_2$, for example, should be made to have an impedance step-up of from 500 ohms to perhaps 500,000 ohms, or a turns ratio of $a_f = \sqrt{500000/500} = 31.7:1$, instead of 17.32:1.

For that reason, in practice an audio transformer has a turns ratio such that 500 ohms is stepped up to 150,000 ohms or thereabouts; ordinary transformers do not materially exceed this secondary impedance* and turns ratio.

Referring to Fig. 40 once more note $T_3$. This couples one tube to another, and is therefore known as an interstage transformer. It differs from an input transformer mainly in that it operates from a higher impedance source; i.e., the $R_p$ of a tube, which is in the neighborhood of 10,000 to 20,000 ohms.

The generated voltage $e_G$ is therefore $\mu e_s$, where $e_s$ is the signal voltage impressed on the grid of the tube.

Because the source impedance is relatively high, $L_m$ of the trans-

*Note that this is not the impedance of the secondary winding itself, but rather that of the load connected to the secondary.

former must be high, perhaps 50 to 100 henries. Such a high inductance requires many turns, and since the secondary number of turns is about the same whether the transformer is designed as an input or interstage transformer, the step-up ratio is of necessity small, say 3 or 4 to 1.

This limits the gain, particularly since the transformer requires a source impedance of relatively low value, not to exceed perhaps 20,000 ohms, so that for a given $G_m$ of a tube, if $R_p$ = 20,000 ohms, $\mu = R_pG_m$ will be fairly low. For example, if $G_m$ = 2,000 μmhos, $\mu = 2000 \times 10^{-6} \times 20000 = 40$, and if the step-up is 3, the overall gain is approximately $3 \times 40 = 120$.

Fairly close values to this can be obtained with a high μ (and high $R_p$) tube in resistance coupling, at much less expense, weight, and space. For that reason transformer coupling is not used today as much as in the past; it is mainly used where a low grid resistance is required, or in push-pull operation, although even here resistance-coupled phase inverters are available.

Finally, referring to Fig. 40, note $T_4$. This is known as a tube-to-line transformer. It is essentially an output transformer, although it is not usually called upon to handle any large amount of power, but merely as a break in the amplifier chain. Thus, suppose the pre-amplifier shown will be called upon to feed a so-called studio amplifier in another part of the broadcast station. Then a tube-to-line transformer $T_4$ will be used to lower the high source impedance (the $R_p$ of the tube) to a lower and more reasonable value of say 500 ohms, before the signal is transported 100 feet or more to the studio amplifier.

At the studio amplifier an input transformer (not shown in Fig. 40) similar to $T_2$ will be employed, so as to step up the signal from 500 ohms to as high a value as possible. The overall step up from $V_2$ to the grid of the first stage of the studio amplifier includes first the step down in $T_4$ say from 10,000 ohms (the $R_p$ of $V_2$) to 500 ohms and then the step up in the input transformer from 500 ohms to say 150000 ohms.

The overall step-up is therefore from 10000 ohms to 150000 ohms or $\sqrt{150000/10000} = 3.88{:}1$ which is about that of an interstage transformer. This means that in spite of the step down to 500 ohms in $T_4$ in order to be able to run a long line to the studio amplifier, the gain is as if the studio amplifier were adjacent to the pre-amplifier and coupled to it by a single interstage transformer.

The tube-to-line transformer $T_4$ therefore acts as a special output transformer, which operates from a relatively small voltage-amplifier tube $V_2$. Nevertheless its design is essentially that of an output transformer; at the high audio frequencies the secondary capacity has little effect, and the leakage reactance acts mainly to reduce the response.

*OUTPUT TRANSFORMER.*—The output transformer is usually of the step-down type in order to match the relatively high $R_p$ of a tube to the low resistance of the usual load, namely, the 6-15 ohms of the voice coil of a loudspeaker. Where a step-down rather than a step-up is involved, the secondary winding capacity is shunted by such a low (load) resistance that the Q of the equivalent series resonant circuit at the high-frequency end is much less than unity.

As a result a droop in the high frequency response practically always occurs, and the capacities of the windings can be ignored. The low- and intermediate-frequency equivalent circuits of the output transformer are practically identical with those of the interstage and input transformers, but the high-frequency equivalent circuit is simpler and appears as shown in Fig. 42 (A) and (B). All circuit constants are referred to the primary side.

rent, just as in the case of the input and interstage transformer. To maintain a flat response down to a

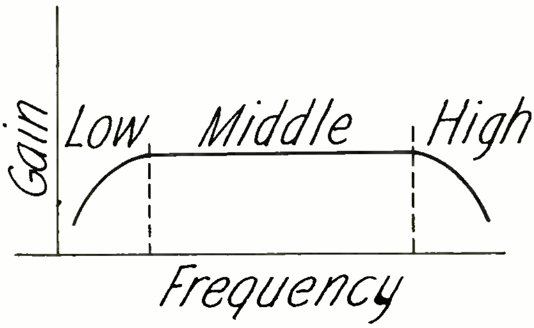



Fig. 43.—Frequency-response curve for an output transformer.

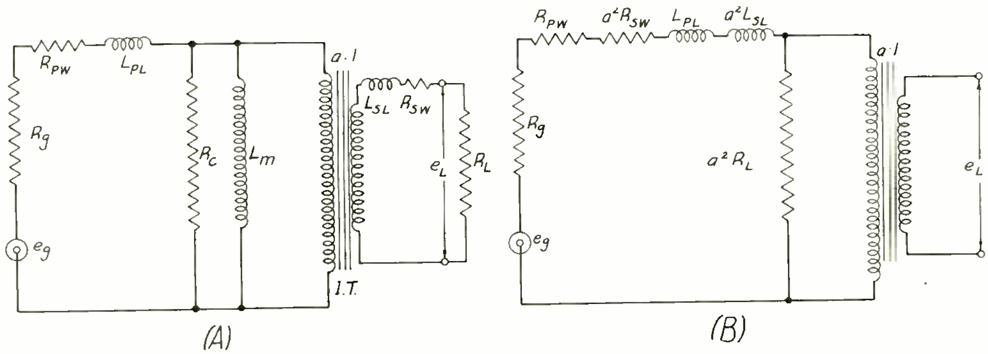



Fig. 42.—High-frequency equivalent circuits for an output transformer having a primary-to-secondary step down turns ratio of a:1.

The current (refer to (B)), produced by $e_g$ has to flow not only through $R_g$, $R_{pw}$, and $a^2R_{sw}$, but also through leakage reactance $L_{pL}$ and $a^2L_{sL}$. The voltage drops in the latter increase with frequency, so that less current can flow through them to get to $a^2R_L$ (the reflected load) as the frequency is increased. The result is that the output power decreases as the frequency goes up yielding the frequency response shown.

Fig. 43 shows the complete frequency response. At the low end, attenuation occurs because the primary draws a large magnetizing current, just as in the case of the input and interstage transformer. To maintain a flat response down to a very low frequency is more difficult for an output transformer because it has to have a high primary open-circuit or mutual inductance even though the output tube is large and draws a large d-c as well as a-c component of current through its primary. A large d-c component particularly tends to cause saturation of the core, with its attendent decrease in inductance as well as production of distortion products.

This is a particularly difficult matter in the case of a single tube output.

Where push-pull output tubes are employed, the two d-c components flow in opposite directions in their windings and tend to cancel each other magnetically. This in turn decreases the tendency of saturation very markedly, and for the same low-frequency response, the push-pull output transformer can be built with less core material and/or turns as compared to the single-ended output transformer.

In the intermediate-frequency range the response is flat just as in the case of the interstage and input transformers, and no further comments are necessary. At the high-frequency end, a droop occurs for the reason cited previously: the voltage drop produced by leakage reactance.

This can be minimized by breaking up the primary and secondary windings into many sections, and interleaving these sections on the core.

The result is an expensive construction, but a superior output transformer, with a flat high-frequency response as far out as desired. In the succeeding assignment a variation in design employed by McIntosh will be described that results in an exceedingly low leakage reactance and exceptionally wide high-frequency response.

The above discussion will indicate to the student the application and behavior of the various types of audio transformers employed. A knowledge of their behavior is of value not because the student can expect to modify the design, but because he will know for example how to modify the associated circuit elements so as to adapt the inherent frequency response of the transformer to the requirements of the particular problem confronting him.

## PUSH-PULL AUDIO AMPLIFIERS

The principal advantages of push-pull operation are an output greater than that obtainable from two tubes in parallel, cancellation of all even order harmonics, and elimination of danger of core saturation in the output transformer.

*PUSH-PULL ANALYSIS.* — A push-pull circuit is shown in Fig. 44. When a signal $e_g'$ is impressed upon the primary of the input transformer, equal and opposite voltages $e_g$ and $-e_g$ are impressed upon the grids of the two power output tubes. The plate current of one tube is caused to increase, while that of the other is decreased, and if the grid swing is sufficiently great, one current will increase to a maximum value, while the other will drop to zero cutoff). Design considerations are practically always based on peak grid swing and resultant maximum power, hence in the analysis that follows it will be assumed that either tube's plate current is alternately driven to cutoff.

Note that the load resistor is represented by $R_L$, extending from plate-to-plate; and the output transformer by L, a center-tapped choke. In actual practice L would be the primary of the output transformer, and the load resistance would be connected to the step-down secondary of this transformer. This condition can be reduced to the circuit shown in Fig. 44 very simply. Thus, suppose the step-down ratio of the output transformer is 20 : 1, and the connected load is 15 ohms.

Then the equivalent plate-to-plate resistance shown as $R_L$ in Fig. 44 is $15 \times (20)^2 = 6000$ ohms.

In the method to be described, as well. The result is that the load seen by either tube is variable or nonlinear over a cycle of signal voltage, and this introduces a major

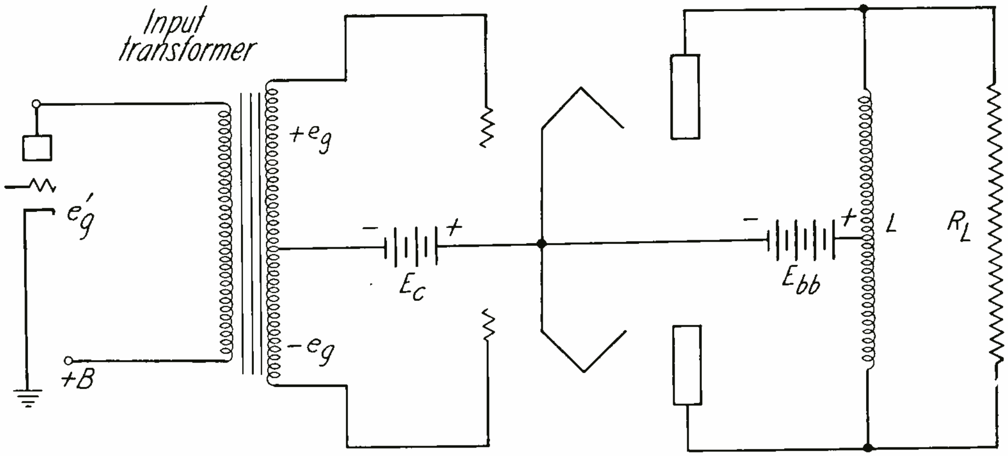



Fig. 44.—A typical push-pull output stage.

$R_L$, the plate-to-plate resistance, will be determined, and then the step-down ratio can be calculated that will make the actual given load resistance look like the desired plate-to-plate resistance. Thus, suppose it is found that a plate-to-plate resistance of 3800 ohms is optimum for the pair of tubes chosen, and the actual load resistance is 500 ohms. The step-down ratio for the output transformer is

$$a = \sqrt{\frac{3800}{500}} = 2.76 : 1$$

The output of each tube in a push-pull stage depends upon the load impedance it faces, just as in any other power circuit. However, the load seen by either tube is not only the reflected value of $R_L$ as it appears across its half of the primary, *but the reflected value of the plate resistance of the other tube*, complication.

A graphical analysis is possible that enables the load line to be plotted as it appears to either tube. The result will be as shown in Fig. 45. Here the load line as it appears to either tube is C A Q B D.

Before explaining this, first note that the impressed plate voltage is $E_{bb}$; the bias is $E_c$, and the zero-signal (d-c) plate current is $I_b$, as determined by plate dissipation requirements. Specifically, if the permissible plate dissipation is $P_p$ watts, then the permissible value of $I_b$ is simply

$$I_b = P_p/E_{bb} \qquad (29)$$

For this value of $I_b$, a certain bias $E_c$ is required. By proceeding vertically upward from $E_{bb}$ to the value $I_b$, as indicated by point Q in Fig. 5, the corresponding bias $E_c$ is determined. Point Q is called the *quiescent* point; in the absence of grid signal the plate current remains fixed at point Q. The following example will make this clearer.

*EXAMPLE.*—Two 2A3 tubes are to be operated at 300 volts plate potential. The maximum plate dissipation is 15 watts per tube. Then the maximum d-c plate current per tube is

$$I_{bo} = \frac{15}{300} = 50 \text{ ma.}$$

Refer now to the tube characteristics shown in Fig. 46. At 300 volts proceed vertically upward to a value of 50 ma. By interpolation, the tube curve passing through this point would have a bias designation of about -58 volts, since the point is closer to the -60-volt curve than it is to the -50-volt curve. Hence the required bias is -58 volts.

Smaller values of $I_b$ can be employed, with a reduction in plate dissipation and an improvement in the all-day operating economy. Thus, if the bias is increased to -60 volts, the zero-signal plate current $I_b$ drops to 40 ma., so that the plate dissipation is but

$$P_p = (.040)(300) = 12 \text{ watts}$$

instead of 15 watts.

The penalty is slightly increased distortion, since the tubes will be operating closer to Class B conditions (to be explained). This is usually of no consequence, and moreover helps to reduce the plate dissipation when maximum signal is applied to the grid. Often the plate dissipation at maximum signal may exceed the permissible value, when that for a zero signal is within safe limits. The reason for this will be discussed later in this section.

*CURVED LOAD LINES.*—Refer back to Fig. 45 once more. Suppose an a-c grid signal is now impressed. During the positive half cycle, this signal will reduce the grid voltage from the bias value $E_c$ to a less negative value, such as $e_1$, $e_2$, $e_3$ (= 0), or even up to $e_4$ or $e_5$, if the signal is strong enough to drive the grid positive. Then it will retrace these values back to $E_c$ once more at the end of the positive half cycle.

During the negative half cycle, the grid will go more negative, namely, from its bias value $E_c$ to $e_1'$, $e_2'$, $e_3'$, $e_4'$, and $e_5'$, and then back through these values to $E_c$ at the end of the negative half cycle.

The plate current will rise to a maximum value $I_{pm}$ at the peak of the positive half cycle. This peak value is denoted by point C in Fig. 45, and CD = $I_{pm}$. During the negative half cycle, the plate current will decrease to zero. Observe that this occurs at a bias $e_2'$ (point B).

As the grid swings more negative, the plate current cannot decrease any further, since a negative crease would mean a *reverse* current, and the tube cannot pass current in the reverse direction. Hence the plate current remains zero, and the path of operation is simply along the voltage axis, namely BD.

The path of operation over the rest of the signal cycle when current is flowing is CURVED; it is path B Q A C. The reason for this was indicated previously: the presence of the other tube in the push-pull circuit makes the tube under consideration think it is seeing a nonlinear or variable load resistance, and for such a resistance the load line is curved.

To see the reasonableness of such a curved load line, consider the moment in the grid-signal cycle when the signal voltage is passing through zero. At this moment the instantaneous grid voltage is simply the bias voltage $E_c$; both tubes are at the quiescent point Q in their path of operation.

If the tubes are well balanced they will share the load equally between them at this moment. To better understand this, suppose one tube were suddenly removed at this moment. Then if $R_L$ is the PLATE-TO-PLATE resistance, it would appear to the

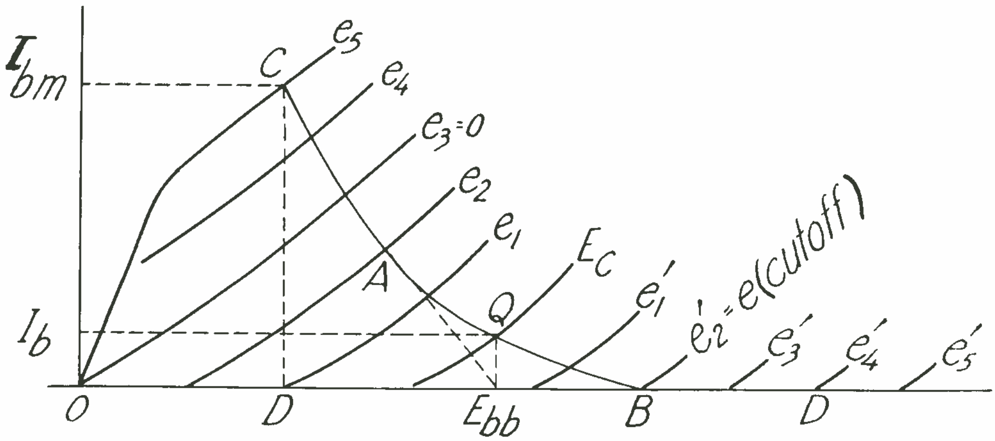



Fig. 45.—Load resistance as it appears to either tube in a push-pull circuit.

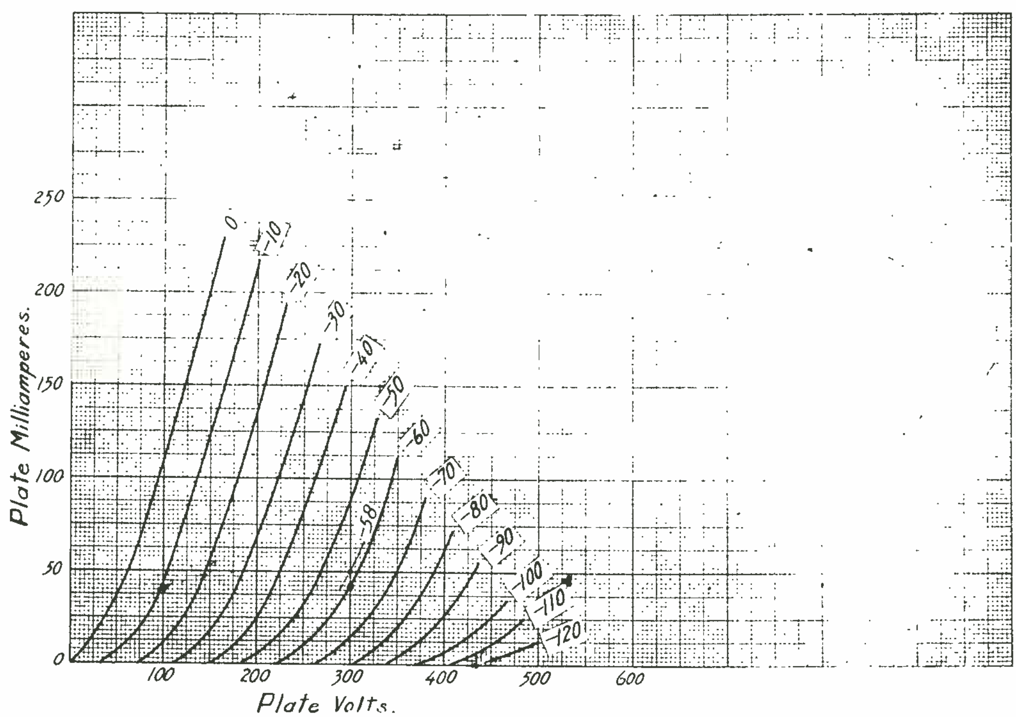



Fig. 46.—Tube characteristics for the Type 2A3 triode.

other tube as $R_L/4$, since that tube is viewing it through HALF the primary, so that there is a step-down in impedance of $(1/2)^2 = 1/4$, and hence $R_L$ appears as $R_L/4$.

The tube would deliver a signal current of i amperes to its half of the primary. Now suppose the other tube is not removed. It should share the plate-to-plate load $R_L$ EQUALLY with this tube. Hence either tube can force only HALF the current into its half of the primary that it otherwise could if the other tube were removed, or the current in either half of the primary is now i/2, instead of i into one-half of the primary alone.

This in turn means that either tube views the plate-to-plate load resistance $R_L$ as $R_L/2$ instead of $R_L/4$ at and around the quiescent point Q. Hence the SLOPE of the curved load line at Q must be that for a load $R_L/2$; i.e., the TANGENT to the curve at Q has the slope corresponding to $R_L/2$.

Consider next the conditions for an instantaneous grid swing that drives one tube to cutoff. Call this tube TUBE B. The path of operation for it is from Q to B in Fig. 45, where $e_2'$ is the cutoff value. Simultaneously the other tube, call it TUBE A, is being driven in a positive direction to a corresponding deviation from the bias value; call this $e_2$.

The path of operation for this tube is from Q to A. Thus Fig. 45 shows the conditions for BOTH tubes at any moment in the signal cycle, and from the symmetry of the circuit, it is clear that the tubes interchange their paths of operation in the two halves of the signal cycle.

The grid swing has been assumed greater than that producing cutoff alternately in either tube. Thus, as tube B's path of operation proceeds from B to D; that of Tube A proceeds from A to C. But, since during this part of the signal cycle, tube B is at cutoff, it is effectively out of the circuit. Hence tube A "sees" $R_L$ as $R_L/4$ from A to C and back to A, in short, the portion AC of the load line has the slope corresponding to a load resistance of magnitude $R_L/4$.

Since the load line appears as $R_L/2$ at Q, and as $R_L/4$ (steeper) at A, the path of operation must be a curve that sweeps upward from Q to A, and then remains at a FIXED slope from A to C. On the other hand, to tube B, $R_L$ appears to approach an infinite value (at cutoff) as its grid is driven negative.

Actually, it is incapable of providing its share of the power during this negative part of its cycle because its plate resistance is increasing, but to the tube it appears instead that $R_L$ is increasing, until from B to D $R_L$ appears as infinite to the tube. These rather remarkable results occur because of the nonlinear internal resistances of the tubes.

Another odd fact is that if CA be prolonged downward, it passes through the point $E_{bb}$, as suggested by the dotted line A $E_{bb}$. This forms the basis of a simplified push-pull construction, as will be explained shortly. Before discussing this, it will be of value to analyze the various modes of operation.

*MODES OF OPERATION.*—From the data given in Fig. 45, the so-called transfer or dynamic characteristic can be plotted. This, it will be recalled, is the relation between the grid voltage and the plate current. The transfer characteristic for Fig. 45 is shown in Fig. 47. The grid voltage $e_g$ is plotted as abscissa, and the plate current $i_p$ as ordinates.

Thus, for the bias voltage $E_c$, the current magnitude $QE_{bb}$ in Fig 45

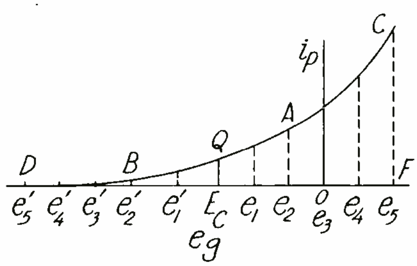



Fig. 47.—Transfer characteristic for a tube in a push-pull amplifier.

is laid off as $QE_c$ in Fig. 47. The other currents are laid off in similar manner; points C, A, Q, B, and D refer to the same current values in the two figures. Cutoff occurs at point B in Fig. 47; for less negative instantaneous grid voltages the current rises in the manner shown to the peak value CF.

However, two tubes are involved, and hence two transfer characteristics should be shown. But it must be remembered that as one grid swings in a positive direction, the other swings in a negative direction, so that one plate current increases as the other decreases.

This is readily shown by combining the two characteristics in the manner illustrated in Fig. 48. The top characteristic is that of tube A, the bottom, that of tube B. As one proceeds from the bias value $E_c$ to the right, the grid of tube A is driven in a positive direction, whereas that of tube B is driven in a negative direction.

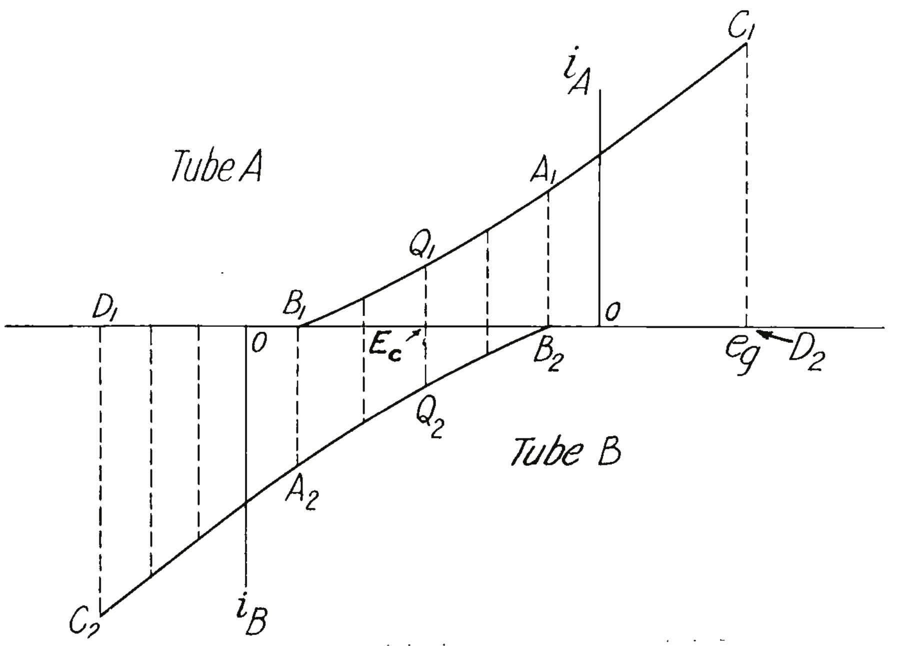



Fig. 48.—Push-pull transfer characteristics.

As a result the current of tube A rises from Q to $A_1$ to $C_1$; the current of tube B simultaneously drops from $Q_2$ to $B_2$ (cutoff) and remains at the zero value to $D_2$.

On the other half cycle the two tubes reverse their roles: the grid of tube A goes negative, and that of tube B goes positive. The current of tube A drops from $Q_1$ to $B_1$ and proceeds to $D_1$, whereas that of tube B rises from $Q_2$ to $A_2$ to $C_2$. In short, the transfer characteristic for tube B is simply that of tube A inverted and aligned so that the bias or quiescent point $Q_2$ is directly underneath $Q_1$.

The two plate currents flow from the outer ends of the output transformer to the center tap; in short, they flow in opposite directions through the two half-primaries of the output transformer. Since the currents also vary in opposite directions from their quiescent values, the net effect is that the two currents are additive; i.e., the two tubes actually aid each other in inducing voltage in the primary and thus cooperate in delivering power into the load $R_L$.

The plate-to-plate current $I_L$ flowing in $R_L$ is therefore at any instant the difference between the currents of the two tubes, A and B; i.e.,

$$i_L = i_A - i_B$$

This is illustrated in Fig. 49, where the two transfer characteristics of Fig. 48 are shown once more, together with their difference, $i_L$. The latter is portrayed as a dotted line between $A_2$ and $A_1$. As an example, point E on it is obtained by subtracting the instantaneous value of $i_B$ = GH from the corresponding instantaneous value of $i_A$ = GF. The difference is

$$EG = GF - HG$$

and this is one value of $i_L$.

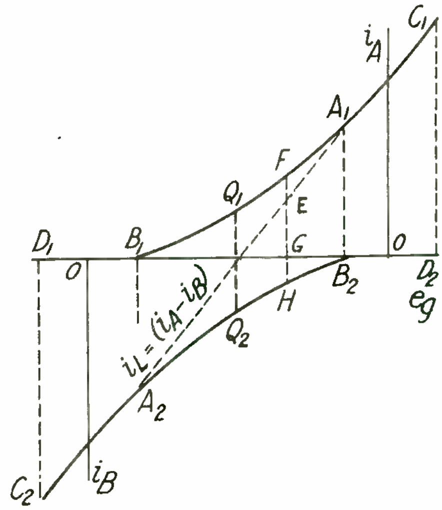



Fig. 49.—Push-pull transfer characteristics and resultant load current $i_L$.

At the instant where $i_A$ reaches $A_1$, $i_B$ reaches $B_2$ (cutoff) and is thereafter zero. Hence from $B_2$ to $D_2$ $i_B$ = 0, and $i_L$ becomes identical with $i_A$. Similar considerations hold for $A_2$ where $i_A$ reaches cutoff at $B_1$; $i_L$ becomes identical with $i_B$, and the total characteristic for $i_L$ is $C_2A_2E\ A_1\ C_1$.

Observe that the operation is such that for sufficiently large grid swings, the two tubes act together over only part of the audio cycle, and at the positive and negative peaks of the cycle alternately cutoff, so that during these intervals only one tube is in operation at a time.

Thus, if the peak-to-peak grid swing were only from $B_2$ to $B_1$, both tubes would be operative throughout the cycle: current would flow in each tube for the full 360°. Where current flows throughout the cycle, the operation is designated as CLASS A.

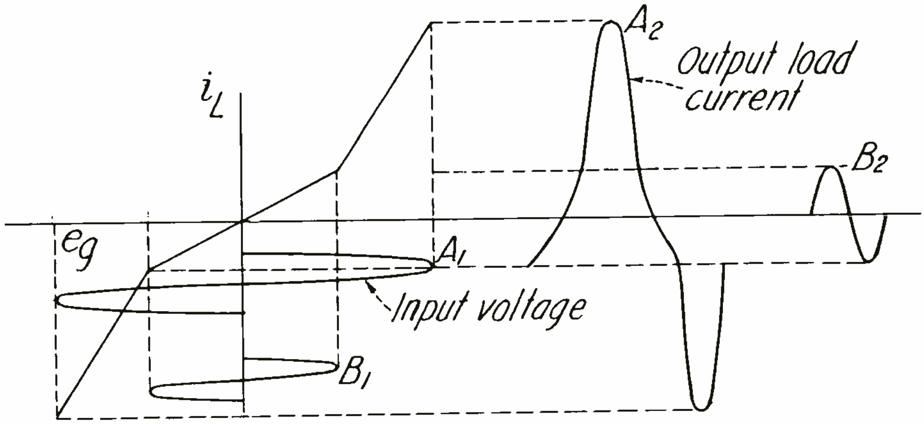



Fig. 50.—Distortion produced by nonlinear relation between input grid voltage and output load current.

On the other hand, where the peak-to-peak grid swing is from $D_2$ to $D_1$, the tubes alternately cutoff and current therefore flows in either for less than a full 360°. Such operation is known as CLASS AB; one way to obtain it is to drive the grids sufficiently hard.

If Class AB operation can be obtained without having to drive the grids positive, it is known as CLASS $AB_1$ operation. Where the grids are driven positive and hence draw grid current, the operation is known as CLASS $AB_2$. This is the case illustrated in Figs. 45, 47, 48, and 49. Note that by suitable adjustment of the tube voltages and load (as will be further discussed), Class $AB_1$ operation can be obtained, and then—if one desires—the grids can be driven still more positive until they draw grid current, whereupon the operation becomes Class $AB_2$.

*DISTORTION CONSIDERATIONS.*—An important question is that of distortion. In the most general terms, distortion occurs when the effect is not in direct proportion to the cause. Here the cause is the grid signal voltage, the effect is either the output load current $i_L$ or the load voltage $e_L$ across $R_L$. (Note that $e_L$ is simply $i_L$ multiplied by $R_L$.)

Referring to Fig. 49, the effect is $i_L = (i_A - i_B)$ plotted against $e_g$. If $i_L$ is directly proportional to $e_g$, the graph $C_2\ A_2\ E\ A_1\ C_1$ will be a STRAIGHT line. Usually the portion $A_2\ E\ A_1$ is straight; it represents the part of the audio cycle when both tubes are operating, or one might call it the Class A portion of the cycle.

The other two portions $A_1\ C_1$ and $A_2\ C_2$, where tube A and tube B, respectively, are operating alone, may be nearly straight, too. But it may come about that these two portions will not be IN LINE WITH the center Class A portion. In that case a very definite distortion will occur.

Its effect is illustrated in Fig. 50 in somewhat exagerrated form. First it will be observed that for a grid swing $B_1$ that is not excessive, only the center linear portion of the push-pull characteristic (Class A part) will be traversed, and the output load current will be $B_2$, an essentially undistorted copy of $B_1$.

However, if the grid swing is $A_1$, then the entire push-pull characteristic is traversed, including the two corners, and the output wave is $A_2$. Clearly $A_2$ is a distorted copy of $A_1$. Note that the distortion is symmetrical on both half cycles, and results from the symmetry of the push-pull circuit. It can be shown mathematically that the distortion can contain only *odd* harmonics when such symmetry is obtained. Hence a push-pull amplifier is said to have only odd harmonic distortion, providing the tubes and circuit are perfectly balanced.

In order to obtain a reasonably straight-line characteristic over the entire range of operation, it is necessary to choose the proper tubes, and employ the proper bias so that the three branches of the characteristic shown in Fig. 50 be in one straight line.

Fortunately for most tubes this is not to difficult to obtain. The reason is that if the tube characteristics are of a form known as a parabola (which is the shape of an automobile headlight reflector), then the center portion of the push-pull characteristic shown in Fig. 50 is tangent to the two outer portions and hence flows smoothly into them.

The result is that the output current $i_L$ does not have the sharp breaks shown in Fig. 50 by $A_2$, and this in turn means that the distortion is moderate and composed mainly of the lower harmonics (third and fifth), which are not so displeasing to the ear. Actual tubes have a characteristic that approximates the parabolic shape, and hence the distortion is in general not excessive. This is true even in the more extreme case where the overlap of the two tube transfer characteristics is small, so that the center portion in Fig. 50 is short.

*CLASS A OPERATION.*—Mention was made previously that by suitable choice of load resistance and tube voltages, one mode of operation or another can be obtained. This will now be examined in greater detail.

First Class A operation will be briefly reviewed. Note 1) the grids are not driven positive, and 2) they are not driven on the negative half of the signal cycle beyond plate-current cutoff. The path of operation is illustrated in Fig. 51. The load current $i_L$ is in general essentially a straight line,

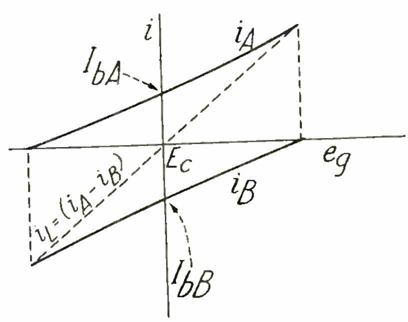



Fig. 51.—Class A operation.

which means that the distortion is a minimum. This is a major advantage of Class A operation.

On the other hand, since the zero-signal (d-c) plate currents $I_{bA}$ and $I_{bB}$ for tubes A and B, respectively, are fixed by plate dissipation considerations, and since the grids must not go positive on the one hand, or beyond cutoff on the other hand, $R_L$ is fixed (as will be shown) and cannot be selected so as to afford a somewhat higher output. Moreover, the grid swing is limited and this prevents higher outputs from being obtained.

One advantage, however, of Class A operation is that the d-c component of the plate current does not materially change from the zero-signal to the full-signal value, whereas in Class AB and Class B operation it increases. This means that a simpler and cheaper "B" power supply, of higher permissible regulation, may be used for a Class A amplifier.

*CLASS AB OPERATION.*—Suppose however that the bias is increased. This decreases the zero-signal d-c plate currents $I_{bA}$ and $I_{bB}$, and also causes either tube to reach cutoff before the peak of the negative grid swing is attained. This—as explained previously—produces Class AB operation, and is illustrated in Fig. 52.

If it is desired to swing the grid up to zero volts, as before, then the grid swing to either tube has to be increased because it starts from a greater negative bias.

Note also that cutoff is reached sooner on the negative swing, and hence that the overlap between the two tubes (shown by the dotted line) is shorter. This mode of operation gives rises to somewhat higher distortion, but it also has certain advantages which will be described farther on.

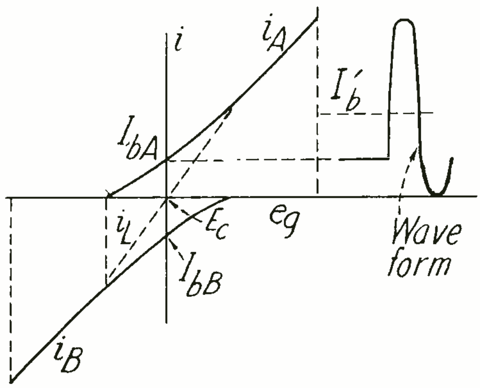



Fig. 52.—Class AB operation obtained by increasing the grid bias.

It is to be observed at this point, however, that the wave form for either tube (depicted at the right in Fig. 52) is considerably peaked, because it can rise to a high value on the positive grid swing, but cannot decrease below zero on the negative swing. This means that the AVERAGE value of the wave, denoted by $I_b'$, is higher than the original d-c value, such as $I_{bA}$, or the d-c component drawn by the tube increases from zero to full signal.

This increase is due to a phenomenon known as self-rectification, although Fig. 52 shows very simply and clearly how it comes about. The bias having been increased, and the d-c zero-signal plate current $I_b$ having been thereby decreased, the d-c input power = $I_b E_{bb}$ is insufficient to cover the a-c output under full signal conditions as well as the inevitable plate dissipation. Hence, in order to satisfy the principle of the Conservation of Energy, the d-c plate current increases to a higher value $I_b'$ under full signal conditions, and thereby enables the d-c input power to cover both the a-c output and the plate dissipation.

The increase in d-c component from zero-to full-signal swing means that the power supply must be designed to maintain an essentially constant voltage over a rather wide range of current drawn; i.e., its voltage regulation must be low. Otherwise, as the current drawn increased, the voltage would drop and this would tend to prevent the current from increasing as much. In turn this would decrease the power output.

Another disadvantage of varying d-c current occurs when self-bias is employed. As shown in Fig. 53, if a

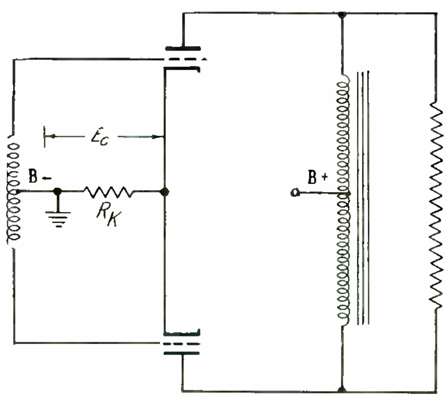



Fig. 53.—Usual form of self-bias.

**cathode bias resistor $R_k$ is employed, a bias $E_c$ is obtained owing to the drop produced in $R_k$ by the flow of the plate current from both tubes. Note that normally no by-pass capacitor is employed across $R_k$ because since one plate current is increasing** when the other is decreasing, their sum tends to be constant or d.c. and produces a d-c bias voltage.

A more exact analysis shows that all the even harmonics flow **through $R_k$, and do not materially affect the fundamental output com**ponent (which is the first or an odd harmonic). However, if the d-c component increases with signal, the bias voltage rises when signal is applied. This in turn tends to prevent the current from rising to such high values, and therefore limits the output power. It must not be construed that self-bias and internal resistance in the "B" supply cannot be permitted; all that is intended to be conveyed here is that these factors tend to reduce the power output from the maximum otherwise available from the tubes.

A second means of converting from Class A to Class AB operation is to reduce the plate-to-plate load resistance $R_L$. Fig. 54 shows how this converts the Class A operation portrayed in Fig. 51 to Class AB operation. The load line for either tube in push-pull operation with the other tube may be curved, but it essentially has the same characteristic as the straight load line for single-ended (single-tube) operation; namely, the lower the load resistance the steeper the load line.

Thus, in Fig. 54 the bias $E_c$ and hence $I_{bA}$ and $I_{bB}$ are the same as in Fig. 51, but $R_L$ is less, so that the $i_A$ and $i_B$ curves are steeper and reach higher maximum values on the positive signal swings. By the same token, however, they reach cutoff sooner on negative signal swings, so that the overlap is less, and Class AB rather than Class A operation is obtained. The significance of this is that in the process of tube design, it may not be possible to design a tube so that the quiescent point is high enough on the tube characteristic because of plate dissipation limitations. On the other hand, the load resistance for maximum output may be relatively low, and give rise to a premature cutoff as shown in Fig. 45.

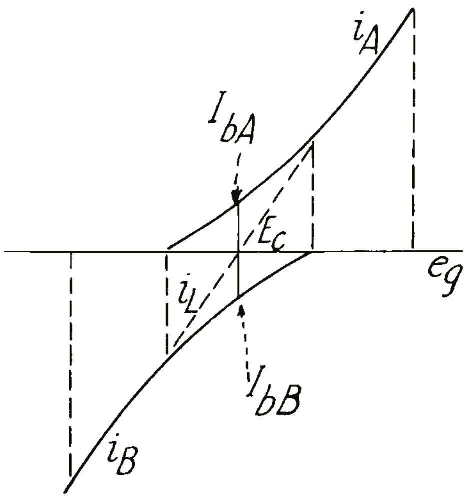



Fig. 54.—Class A operation can be changed into Class AB operation by reducing the plate-to-plate resistance.

If only Class A operation were permitted, then a less steep load line would be necessary, in order to extend the cutoff point of each tube to the negative peak of the grid swing. This in turn would necessitate a load resistance higher than the optimum value, and as a result, the power output would be less.

The decrease in output, however, is in general not very marked; nevertheless, the removal of the restriction that operation be only within cutoff that a push-pull amplifier permits, can allow the output in general to be increased somewhat.

Another consideration arises that is also of importance. If a push-pull circuit permits operation beyond cutoff without appreciable distortion, why not permit the grids to be driven positive, as well, and thus permit really appreciable increases in output?

When only Class A operation was known, the bias had to be half-way between cutoff and the maximum positive or least negative value of the grid voltage. Plate dissipation considerations, however, limited the bias to a certain negative value; less than this value would give too large a d-c zero-signal plate current and hence excessive plate dissipation. The grid swing then could not exceed the difference between this bias value and cutoff; on the positive half cycle this brought the grid just about up to zero volts by a suitable choice of $R_L$ that was somewhere between $2R_p$ and perhaps $4R_p$.

In Class AB operation, $E_c$, $R_L$, and the grid swing or signal voltage are in large measure independent of one another. Thus $E_c$ can be chosen to keep the plate dissipation within the permissible limits. $R_L$ can be chosen, in a manner to be described, to give maximum output, although plate dissipation considerations at full signal enter in and affect its choice. Finally, the grid swing can be chosen to give as much power output as is desired, consistent with such ultimate limiting factors as grid power drawn from the preceding so-called driver tube, plate dissipation, grid dissipation, and power-supply limitations. These will be discussed presently.

*CLASS B OPERATION.*—If a push-pull circuit can be operated beyond cutoff, why not proceed to the limit and have the tubes biased to cutoff, whereupon one tube is operative during say, the positive half cycle; and the other tube is operative during the negative half cycle? Such operation, or rather a close approach to such operation is called Class B.

Actually, owing to the curvature of the tube characteristics, operation at cutoff leads to excessive distortion in practical cases. This is illustrated in Fig. 55. The curved dynamic characteristics for $i_A$ and $i_B$ become identical with that for the load current $i_L$, and at the origin or bias $E_c$ point, the curvative is excessive and gives rise to the distortion shown in the $i_L$ wave shape. The English call this "join-up" distortion, since it is a function of the manner in which the $i_A$ and $i_B$ curves merge.

As a result, true class B operation is not employed. Instead, extreme Class AB operation is used and called Class B, since it is essentially Class B operation, at least from a practical viewpoint. Fig. 56 shows this mode of operation; it is clear that the overlap in the two characteristics is made just sufficient to iron out any marked distortion in this region.

Although Class A operation is most free of distortion, and Class B is least free, nevertheless the latter mode of operation has one important advantage over Class A, and to a lesser extent, over Class AB operation.

This results from the fact that the d-c current and hence d-c input power drawn during zero-signal periods is very small. In large modulator stages, the saving in power consumption from this item can be considerable, since in normal broadcast practice the quiet periods perhaps exceed the loud periods when totalled over a day's time. Of course, on the other hand, the power supply must have low regulation, but this is more readily obtained in large heavy-duty rectifier systems.

At the other end of the scale, a small battery-operated power output stage will permit longer life from the batteries if it is operated

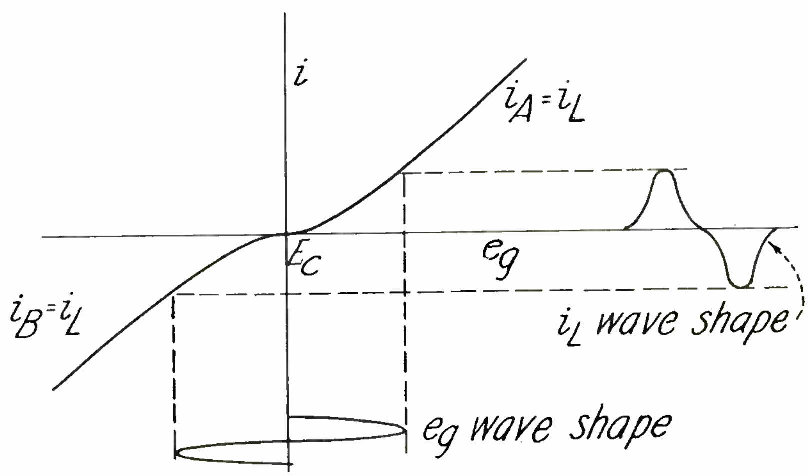



Fig. 55.—Extreme Class B operation leads to the so-called "join-up" distortion shown.

Class B. Furthermore, batteries inherently have low internal resistance and hence low regulation under variable current drain.

*GRAPHICAL ANALYSIS.*—From the preceding discussion it is now possible to formulate a graphical procedure which will enable the proper grid swing, load resistance, power output, etc., to be determined. It was shown in Fig. 45, which is repeated here, that the load line for either tube varies in slope from a value of zero at cutoff, B, corresponding to an apparent load resistance equal to infinity, to a value corresponding to a load resistance of $R_L/2$ at the quiescent point, Q, and finally to a value corresponding to a load resistance of $R_L/4$ at a point A at which the other tube has reached cutoff.

From A to C, the load line for the tube is straight and remains at the slope corresponding to $R_L/4$, because the other tube is beyond cutoff and hence "out of the picture", so that the tube under consideration sees the plate-to-plate resistance $R_L$ through a 2 : 1 step down and hence as $R_L/4$. Furthermore, CA, when prolonged downward, passes through the point $E_{bb}$.

For maximum output, $R_L$ should be of such value that its apparent value to the tube should equal the tube's $R_p$ at all points in the signal cycle. In other words, the tube's $R_p$ is variable to some extent; if possible, the load presented to it should equal this $R_p$ as nearly as possible at all times.

A push-pull circuit tends to approximate this desirable result. For example, at and near cutoff the $R_p$ of a tube appears extremely high. But at and near cutoff $R_L$ appears to the tube to be extremely high, too, and hence an impedance match tends to take place.

At the peak of the swing, point C in Fig. 45, $R_L$ appears as $R_L/4$.

If the slope of the $e_5$ tube characteristic through C also has

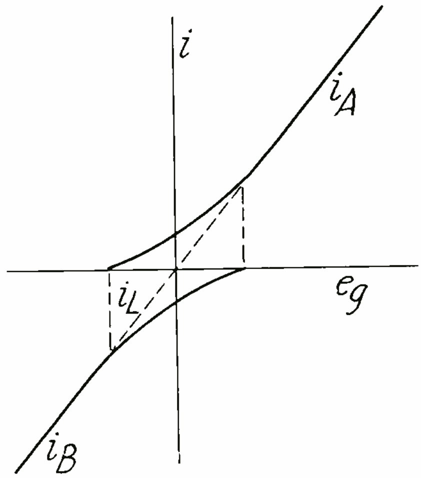



Fig. 56.—A slight overlap of the two tube curves, representing actually extreme Class AB operation, is usually called Class B operation.

this slope, matched conditions will be obtained. In other words, $R_L$ can be so chosen that $R_L/4 = R_p$ at the peak of the swing. If this is done, maximum power output can be expected because either tube supplies the most power during the signal cycle at the moment of peak current flow, and hence the load resistance should preferably be matched to the tube at this moment of the cycle.

Whether $R_L$, appearing as $R_L/2$ at the quiescent point Q, matches the $R_p$ of the tube at this point is a question of the curvature of the tube characteristics, however, the mismatch cannot be too great, and in any event it is not serious so far as affecting the power output is concerned.

The rule for triode tubes is therefore to measure the average slope of the tube characteristics in the upper left-hand positive grid region. This represents the average $R_p$ of the tube in this region. Then choose a value of plate-to-plate load resistance $R_L$ such that

$$R_L = 4R_p \qquad (30)$$

This gives an optimum value for $R_L$ so far as maximum power output is concerned; plate dissipation considerations at maximum signal may require this value for $R_L$ to be modified.

The amount of grid swing depends upon whether or not it is desired to drive the grid positive, the amount of power output, and again plate dissipation considerations. If it is desired to drive the grid positive, the extent of the positive swing is determined as shown in Fig. 57.

Here the average slope of the tube curves is tan $\alpha$, where the first positive grid curve has been extended to the $e_p$-axis by a dotted line and makes the angle $\alpha$ with the axis. Then through $E_{bb}$, the given plate-supply voltage, a line $E_{bb}C$ is drawn so as to make the same angle $\alpha$ with the $e_p$-axis. Then tan $\alpha$ = $R_L/4$, or the plate-to-plate resistance $R_L$ will be 4tan $\alpha$ and hence $4R_p$ as demanded by Eq. (30).

Now if it were not desired to drive the grid positive, the peak current would be given by point D, which is on the zero-grid-volt curve. This could represent Class $AB_1$ operation.

Remember, however, that this is true only if for $e_g$ = 0, the other tube has been driven sufficiently negative to be at or beyond cutoff, for only then does the tube under consideration face $R_L$ alone and "see" it as $R_L/4$. In short, this construction applies only where the tubes are driven sufficiently to reach cutoff. Since, however, this is the case for maximum grid swing and hence maximum power output, it furnishes the proper construction even for Class A operation, where

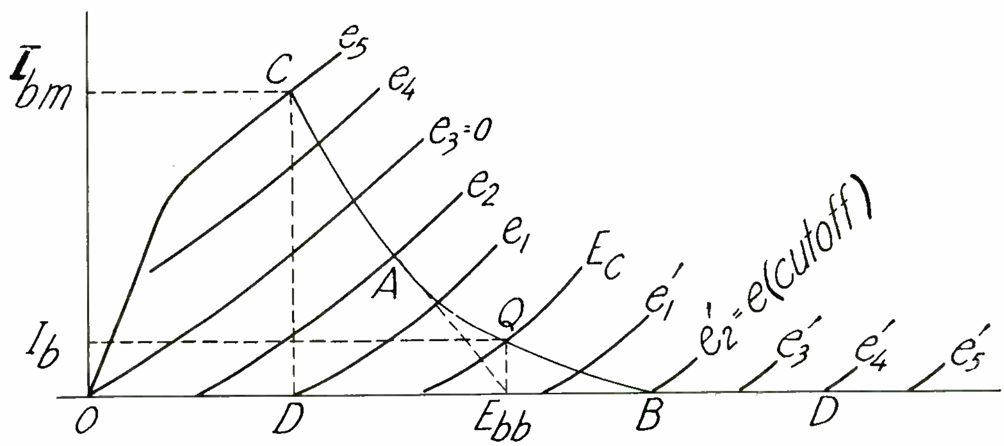



Fig. 45.—Load resistance as it appears to either tube in a push-pull circuit.

maximum grid swing alternately brings either tube practically up to cut-off.

Now consider the case where the grid is to be driven positive. If this is to be a small amount, then the path of operation will go beyond D say, to F. A still greater positive swing will be to G. If the positive swing reaches C, it is clear that a limit has been reached, because the higher positive grid curves all slope into point C, as is evident from Fig. 57.

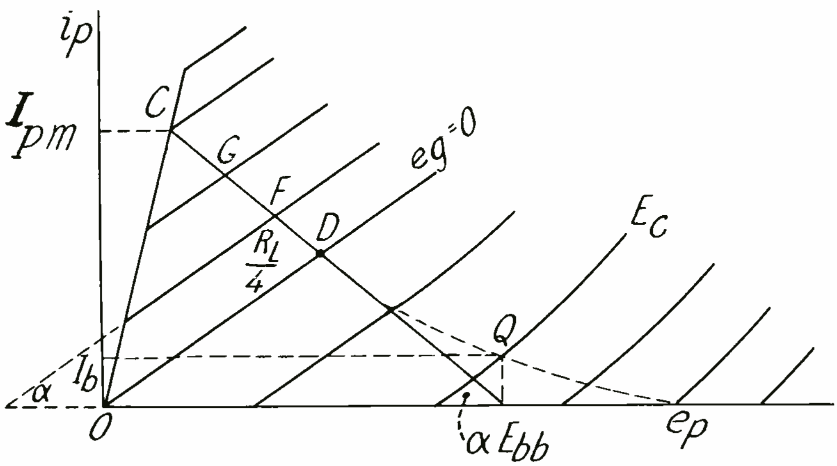



Fig. 57.—**Maximum grid swing is up to the knee of the characteristic.**

In short, for the optimum value of $R_L$ chosen, the maximum permissible grid swing is up to point C. If it is attempted to swing the grid above this value, the plate current will not rise above the value $I_{pm}$ shown, and will instead develop a flat top, which means considerable distortion.

Of course the actual positive grid swing permitted may be less than up to point C if the grid is overheated by the resultant grid current drawn. Or, it may not be desired to use a preceding driver stage large enough to furnish the requisite power with a minimum of distortion of the signal waveform by the grid current. A final consideration is that the plate dissipation for this amount of grid swing may be excessive, and this is the next point to be taken up.

*POWER CONSIDERATIONS.*—As mentioned previously, when the tubes are operated Class AB or B, so that they do not draw current over the entire signal cycle, appreciable self-rectification takes place, and the d-c component rises. This in turn means that the d-c input power rises, and after the a-c output power is subtracted from it the difference, which is the plate dissipation, may be greater than at zero-signal, and perhaps may even be excessive.

It is therefore necessary to calculate the power output and d-c current and power input at full signal. The power output of BOTH tubes is given very simply by

$$P_o = \frac{I^2_{pm} R_L}{8} \qquad (31)$$

where $I_{pm}$ is the peak current (see Fig. 57), and $R_L$ is the plate-to-plate load resistance (normally the *reflected* value of the actual load.

An approximate formula for the increased d-c component $I'_b$ at full signal, *per tube*, as follows:

$$I'_b = \frac{k}{2}\left(\frac{I_{pm}}{2} + I_b\right) \qquad (32)$$

where $I_b$ is the d-c component at zero-signal of each tube, and k is a constant depending upon the ratio of $I_{pm}$ to $I_b$. For $(I_{pm}/I_b)$ equal to 10 or less, k is about 1.05; for $(I_{pm}/I_b)$ between about 10 and 50, k is about 1.12; and for larger values of $(I_{pm}/I_b)$, k is about 1.20.

The d-c input power at full signal *per tube* is then, analogous to Eq. (1),

$$P_i = I'_b E_{bb} \qquad (33)$$

and the plate dissipation per tube is therefore

$$P_p = P_i - \frac{P_o}{2} \qquad (34)$$

The output power $P_o$ is divided by two to get the output per tube, and this is then subtracted from the input power $P_i$ per tube to give the plate dissipation of each tube.

Suppose (as will be shown very shortly in an example) that the plate dissipation is excessive. What modifications in the operation may be made? One obvious thing is to reduce the signal swing. This will reduce $I_{pm}$, and thereby the d-c component $I'_b$, as well as the power output $P_o$; the difference, or $P_p$, will in general decrease.

Another modification is to increase $R_L$. This will cause the load line to be less steep, and thus reduce $I_{pm}$. Once again $I'_b$ and $P_o$ will be reduced, and in general, the difference $P_p$ will be decreased. Usually both modifications are employed in Class $AB_2$ operation, although it is preferred to increase $R_L$ rather than decrease the grid swing because the power input in the former case is decreased at a greater rate than the power output, and hence $P_p$ decreases faster.

This means that the power output need not be decreased as much. In the case of Class $AB_1$ operation, the grid is normally driven up to zero volts, and it is undesirable to reduce the swing below this value. Instead, as in Class $AB_2$ operation, more output can be obtained from the same dissipation by increasing $R_L$.

*ILLUSTRATIVE EXAMPLE.*—The above discussion will now be illustrated by an example. Two 2A3 tubes will be employed in Class $AB_1$ operation. A plate potential of 300 volts will be employed as in the previous example, and for a permissible plate dissipation of 15 watts, the quiescent d-c plate current is 40 ma., for which a bias of -60 volts is required. This gives rise to the quiescent point marked as Q in Fig. 58.

In Fig. 58 a straight line has been drawn tangent to the zero-volt curve. As described in a previous assignment, the resistance represented by this line can be found by choosing any two points, and dividing the difference in voltage values by the difference in current values. This is preferred in actual computations to finding the cotangent of the angle of slope and correcting it for the different scales employed for the voltage and current axes.

Suppose the extreme ends of the line are chosen. Then the difference in voltage values is 175 - 48 = 127 volts; and the difference in current values is 250 - 0 = 250 ma. = 0.25 ampere. The resistance of the tube is then

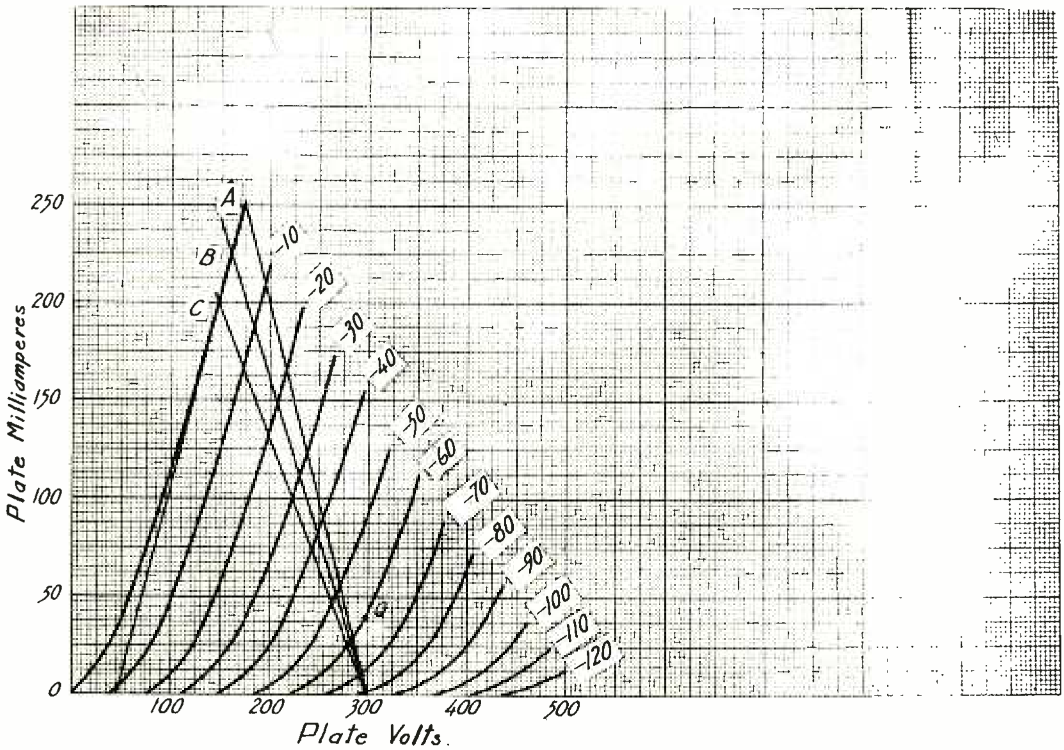



Fig. 58.—Graphical constructions for two 2A3 tubes in push-pull.

$$R_p = \frac{127}{.25} = 508 \text{ ohms.}$$

Then, by Eq. (30), the plate-to-plate resistance should be

$$R_L = 4R_p = 4 \times 508 = 2032 \text{ ohms.}$$

Through $E_{bb}$ = 300 volts, a load line of 508 ohms is drawn as shown. It meets the zero-volt curve at A. From past experience we know that the load line is too steep, and $I_{bm}$ is too high—250 ma. for point A, which is actually beyond the highest value shown by the manufacturer. However, let us continue the problem to a conclusion, and then correct the results.

The power output is given by Eq. (31) as

$$P_o = \frac{(.25)^2\ (2032)}{8}\quad 15.9 \text{ watts.}$$

The d-c component under full signal $I_b'$ will next be calculated from Eq. (32). First note that $I_{pm}/I_b$ = .250/.040 = 6.25, so that the constant k in Eq. (30) will be 1.05. Hence

$$I_b' = \frac{1.05}{2}\left(\frac{250}{2} + 40\right) = 86.6 \text{ ma.}$$

Then, by Eq. (33), the d-c power input is

$$P_i = (.0866)(300) = 26 \text{ watts.}$$

The plate dissipation is therefore, from Eq. (34).

$$P_p = 26 - \frac{15.9}{2} = 26 - 7.95$$
$$= 18.05 \text{ watts.}$$

Since this exceeds the permissible value of 15 watts per tube, the operation will have to be modified by increasing $R_L$. Suppose $R_L$ is increased to 2500 ohms. Then $R_L/4$ = 625 ohms, and when the load line for this is drawn from $E_{bb}$, it intersects the zero-bias curve at B.

Now $I_{bm}$ = 225 ma., so that

$$P_o = \frac{(.225)^2(2500)}{8} = 15.81 \text{ watts.}$$

Since $I_{bm}/I_b = .225/.040 = 5.63$, k still equals 1.05, and

$$I_b' = \frac{1.05}{2}\left(\frac{225}{2} + 40\right) = 80.1 \text{ ma.}$$

Hence

$$P_i = (.0801)(300) = 24.0 \text{ watts}$$

and the plate dissipation is

$$P_p = 24.0 - \frac{15.81}{2} = 16.09 \text{ watts.}$$

This, too, is excessive, hence a still higher value of $R_L$ = 3,000 ohms will be used. When the load line $i_{bm}$ for 3000/4 = 750 ohms is drawn, it intersects the zero-bias curve at C. The corresponding values are

$$i_{bm} = 200 \text{ ma.}$$

$$P_o = \frac{(.2)^2(3000)}{8} = 15 \text{ watts}$$

$$I_{pm}/I_b = .2/.04 = 5.; \quad k = 1.05$$

$$I_b' = \frac{1.05}{2}\left(\frac{200}{2} + 40\right) = 73.5 \text{ ma.}$$

$$P_i = (.0735)(300) = 22.05 \text{ watts}$$

$$P_p = 22.05 - 7.5 = 14.55 \text{ watts.}$$

Since this is within the permissible limit of 15 watts, $R_L$ will be taken as 3000 ohms plate-to-plate, and $P_o$ will be 15 watts.

*DISCUSSION OF RESULTS.*—Note first that the power output has decreased very little as $R_L$ is increased; from 15.9 to 15 watts. On the other hand, the plate dissipation has decreased materially; from 18.05 to 14.55 watts.

In the second place, the reader may question the rule first given to the effect that $R_L/4$ should equal the $R_p$ of the tube in the region of $i_{bm}$. When this was tried for the 2A3 tube, the plate dissipation was found to be excessive, and so a higher value of $R_L$ than that suggested by the rule had to be employed.

The answer is that the rule suggests a tentative value that is optimum for the tube and will give maximum power output. If the tube, however, is not physically designed to dissipate the heat developed on the plate, a higher value of $R_L$ will be required. Fortunately, this does not reduce the power output by very much, but does materially reduce the plate dissipation, as was shown above.

The third point to note is that this graphical analysis also furnishes information as to the magnitude of grid swing required. Thus, for the 2A3 tubes, if the bias is -60 volts, and each grid is to be driven up to zero volts, a total of 2 × 60

= 120 volts is required from grid to grid.

**If the push-pull input transformer has a step-up of 1: 4 from** the primary to the entire secondary, then the voltage that has to be developed across the primary by the preceding tube is 120/4 = 30 volts. If the input voltage to the amplifier is to be 0.706 volts peak, then the voltage gain of the VOLTAGE AMPLIFIER stages is simply 30/0.706 = 42.5.

In short, from the given input level, the input voltage can be calculated. For the desired output power, the output tubes can be chosen and their output, plate dissipation, and grid-driving voltage determined graphically. Then the necessary voltage gain from the input up to the grids of the power output tubes can be calculated, and thus the number and type of voltage-amplifier stages determined.

*PUSH-PULL PENTODES.*—The preceding discussion and example had to do with a triode tube. When pentode tubes are employed, the graphical analysis has to be modified in practically the same manner as in the case of the single-ended tube. The procedure is, if anything, simpler than for the triode tube.

Fig. 59 shows the tube characteristics for a 6L6 beam power tube, which are of practically the same form as those for an ordinary pentode tube. The plate voltage will be taken as 360, and the screen voltage as 250. The maximum permissible plate dissipation is 19 watts, hence the zero-signal plate current is, from Eq. (29):

$$I_b = 19/360 = 52.8 \text{ ma.}$$

From Fig. 59 it is seen that the bias must be somewhere between -20 and -15 volts. Since the quiescent point is not very important so far as push-pull operation is concerned, a bias of -20 volts will be taken, **for which the d-c component $I_{b1}$ is 45 ma. The zero-signal plate dissipation will therefore be**

$$P_p = (.045)(360) = 16.2 \text{ watts}$$

which is decidedly on the safe side.

Suppose Class $AB_1$ operation is contemplated. Then a grid swing up to zero volts will be required. From the 360-volt point on the axis, a line is drawn up to the zero-volt curve just to the right of the knee of the characteristic. This is line A $E_{bb}$ in Fig. 59, its slope corresponds to a resistance equal to $R_L/4$.

The numerical value of $R_L/4$ is obtained by noting that $I_{bm}$ = 160 ma., and $e_{bmin}$ = 47 volts, or the change in voltage is 360 - 47 = 313 volts. Hence

$$R_L/4 = \frac{313}{.160} = 1955 \text{ ohms}$$

so that $R_L = 4 \times 1955 = 7820$ ohms. The actual value recommended is 6600 ohms; by locating point A a little **to the right, $R_L$ is reduced, and the distortion is decreased.**

**Using 7820 ohms, however, the power output is, by Eq. (31)**

$$P_o = \frac{(.160)^2(7820)}{8} = 25 \text{ watts.}$$

**The d-c component at full signal is, by Eq. (32)**

$$I_b' = \frac{k}{2}\left(\frac{160}{2} + 45\right) = \frac{1.05}{2}(80 + 45) \; I_b' = 65.6 \text{ ma.}$$

because k = 1.05, since $I_{bm}/I_b$

= 160/45 = 3.55 which is less than ten. Then, by Eq. (33), the d-c power input is

$$P_i = (360)(.656) = 23.6 \text{ watts}$$

from which, by Eq. (34), the plate dissipation is

$$P_p = 23.6 - 25/2$$
$$= 11.1 \text{ watts per tube}$$

or well within the safe limit of 19 watts.

$$P_o = \frac{(.167)^2(6600)}{8} = 23 \text{ watts, or}$$

somewhat less than before.

This is to be expected, because in the case of a pentode tube, $R_L/4$ is inherently far less than the $R_p$ of either tube at peak positive grid swing, and therefore the impedance mismatch is very great. Under these conditions, the greater $R_L$ is chosen, the closer is $R_L/4$ to $R_p$, and hence

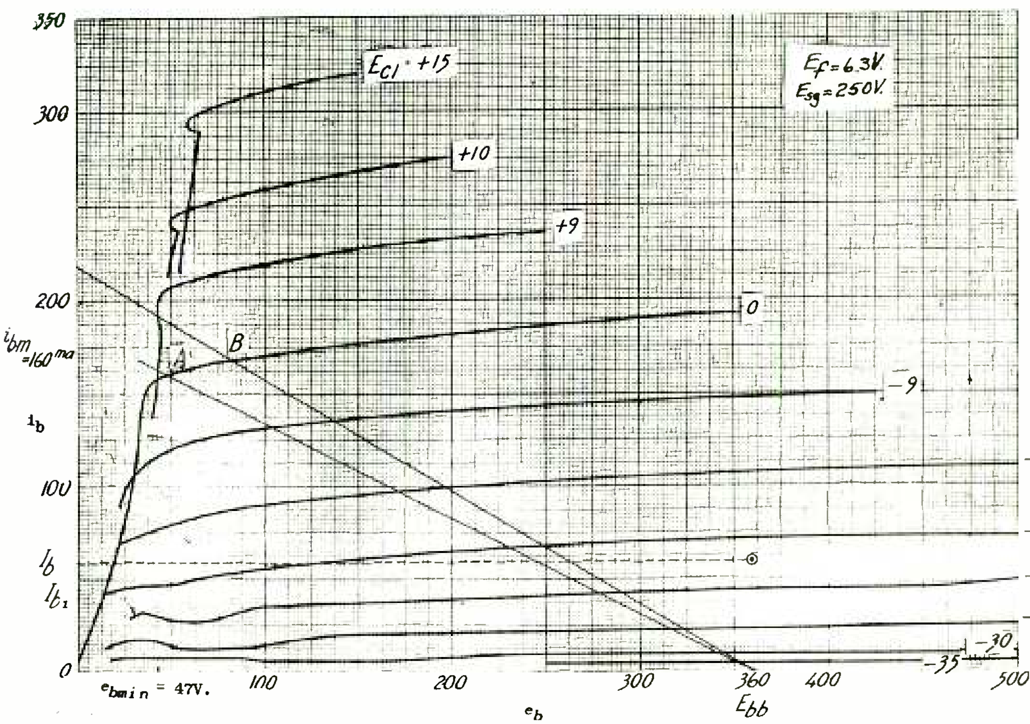



Fig. 59.—Graphical constructions for two 6L6 tubes in push-pull.

Suppose a value of 6600 ohms were employed for $R_L$. This corresponds to line $E_{bb}$ B in Fig. 59; it is obtained by dividing 360 volts by 6600/4 to obtain 218 ma., which is laid off on the current axis, and joined to $E_{bb}$ = 360 volts.

Its intersection with the 0-volt curve is point B, and corresponds to $i_{bm}$ = 167 ma. Then

the closer does the operation approach maximum power output conditions. Thus, $R_L$ = 7820 ohms yields 25 watts output, whereas $R_L$ = 6600 ohms yields only 23 watts output.

$$I_b' = \frac{1.05}{2}\left(\frac{167}{2} + 45\right) = 67.5 \text{ ma.}$$

$$P_i = (.0675)(360) = 24.3 \text{ watts, and}$$

$$P_p = 24.3 - 23/2 = 12.8 \text{ watts,}$$

which is slightly greater. This indicates that the efficiency of operation is somewhat lower for $R_L$ = 6600 ohms than for $R_L$ = 7820 ohms, but the distortion is somewhat less.

*POSITIVE GRID OPERATION.*—Now suppose it were desired to drive the grids positive. A safe value, as indicated by the manufacturer, is +15 volts; this does not cause excessive control-grid dissipation nor excessive screen-grid dissipation.

However, as indicated in Fig. 60, compared to the original load lines chosen, namely $E_{bb}$ A for $R_L$ = 7820 ohms and $E_{bb}$ B for $R_L$ = 6600 ohms, now a steeper load line $E_{bb}$ C must be used in order to avoid the knee of the $E_{c1}$ = +15-volt curve. The corresponding value of $R_L/4$ is given by $I_{bm}$ = 305 ma. and $e_b$ min. = 75 volts. Then

$$R_L/4 = \frac{360 - 75}{.305} = 935 \text{ ohms, or}$$

$R_L = 4 \times 935 = 3740$ ohms. This compares with the value of $R_L$ = 3740 ohms given in the tube manual for a screen voltage of 270 instead of the 250 volts employed to give the characteristic curves of Figs. 59 and 60.

$$P_o = \frac{(.305)^2(3740)}{8} = 43.5 \text{ watts,}$$

$$I_b' = \frac{1.05}{2}\left(\frac{305}{2} + 45\right) = 103.8 \text{ ma.}$$

$$P_o = (.1038)(360) = 37.4 \text{ watts.}$$

$$P_p = 37.4 - 43.5/2 = 15.65 \text{ watts,}$$

which is still below the limit of 19 watts.

It is to be observed that also in the case of a triode whose grid is driven positive, $R_L$ must be decreased as the grid swing is in-

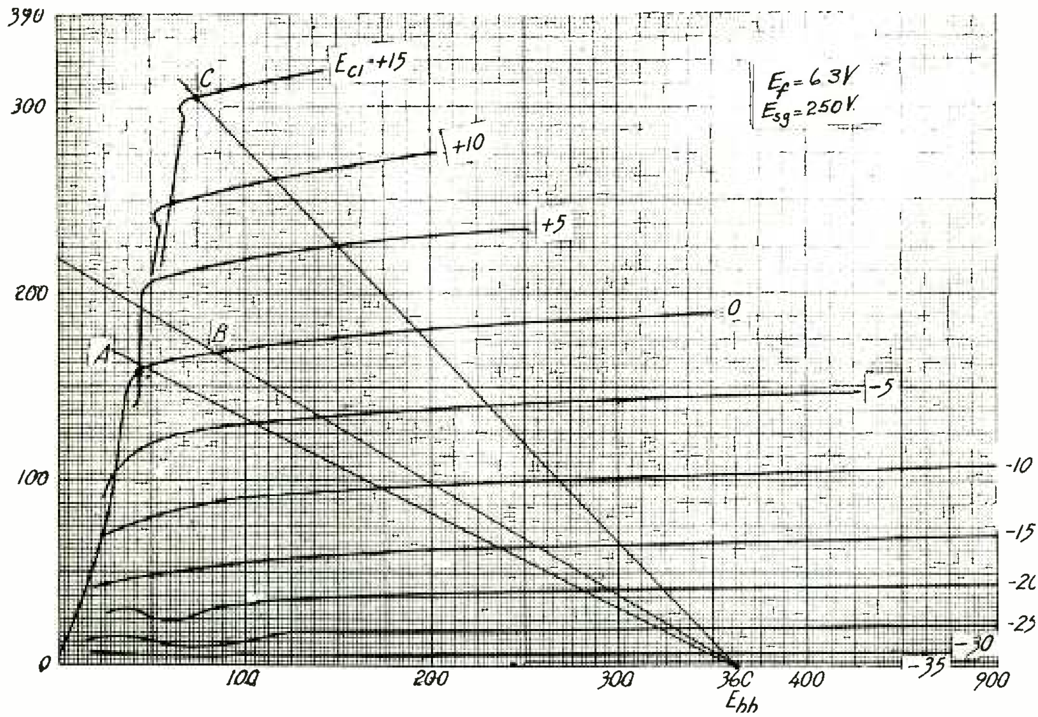



Fig. 60.—Two 6L6 tubes in Class $AB_2$ operation.

creased. This is because triode tubes also develop a knee in each positive grid curve, and for the same reason as in a tetrode or pentode tube: at low plate voltages the positive control grid or screen grid, as the case may be, robs the plate of electrons and causes the plate current curve to shoot downward.

*DRIVER-STAGE DESIGN.*—When the grids are driven positive, they draw current, and thus a sudden load is placed on the preceding driver stage at the peak in each half cycle. This load suddenly applied produces a voltage drop in the driver stage at the peak of the cycle and tends to flatten the voltage appearing at the grids of the class $AB_2$ stage.

The effect is illustrated in Fig. 61. The driver stage is represented by the a-c source generating voltage $e_G$ in series with its internal resistance $R_G$. These are the values as they appear to the grid of either tube, and usually represent $\mu e_g$ and $R_p$, respectively, AS VIEWED from the secondary terminals of a driver transformer normally intervening between the driver tube or tubes and the push-pull grids.

The grid resistance is represented by $r_g$; it is very nonlinear, since it is an infinite resistance or open circuit until the grid reaches a positive potential relative to the cathode, whereupon $r_g$ drops to a fairly low finite value. The current $i_g$ in the right-hand diagram of Fig. 61 flows as shown only during the peaks of the cycle. At such times it produces a voltage drop in $R_G$, whereupon the terminal voltage $e_T$ drops below $e_G$. The result is the flattened wave as shown; it occurs on both halves of the signal cycle.

Such symmetrical distortion of the grid driver voltage indicates the production of *odd* harmonics. Thus the signal input to the 6L6 stage is not sinusoidal in shape,

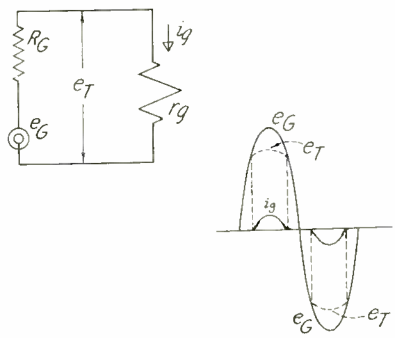



Fig. 61.—Equivalent circuit of a driver stage, and flattening of the terminal voltage.

as previously assumed, but is itself distorted owing to the flow of grid current. Such distortion is in addition to that produced in the output (plate) circuit of the 6L6 tubes. Since the output is also symmetrical in connection, the distortion there is also of the odd-harmonic type.

A fair approximation to the distortion both in the grid and plate circuits is to assume that it is mainly third harmonic, (although appreciable fifth, and even higher odd harmonics may be present). It may be that the third harmonic produced in the plate circuit is opposite in polarity to that produced in the grid circuit, and equal in effect, so that the two cancel. In such an event the driver stage is easier to design.

On the other hand, the two harmonics may be additive for certain tubes, so that the problem becomes more difficult. It is proposed to furnish a method here whereby the driver stage may be designed so that the *net* third-harmonic distortion is a certain permissible percentage of the fundamental current required for a fundamental power output.

The method, although approximate, gives a fairly good idea of what driver tube or tubes are required, what driver input impedance is necessary, and what type of driver transformer is required. It even indicates whether or not it is possible to design the stage in the event that an unreasonably low distortion is demanded for a required power output.

First, the actual grid current flow must be determined. The grid current is a function of the instantaneous (positive) grid voltage and the instantaneous plate voltage. Its plot is most conveniently represented by a grid family of curves as shown in Fig. 62. Each curve is for a different fixed value of positive grid voltage, and shows how the grid current varies with plate voltage.

In the dynamical operation of the stage, i.e., when a plate-to-plate load resistance is present, the plate voltages as well as the plate currents of the two 6L6 tubes vary with the instantaneous grid voltages. The instantaneous grid current, for every instantaneous value of positive grid voltage and hence plate voltage, can be found as follows.

The path of operation for the plate current and plate voltage of each tube is shown in Fig. 62 by the dotted line which blends into the lower part of the straight line ABCD. Our interest is mainly in the

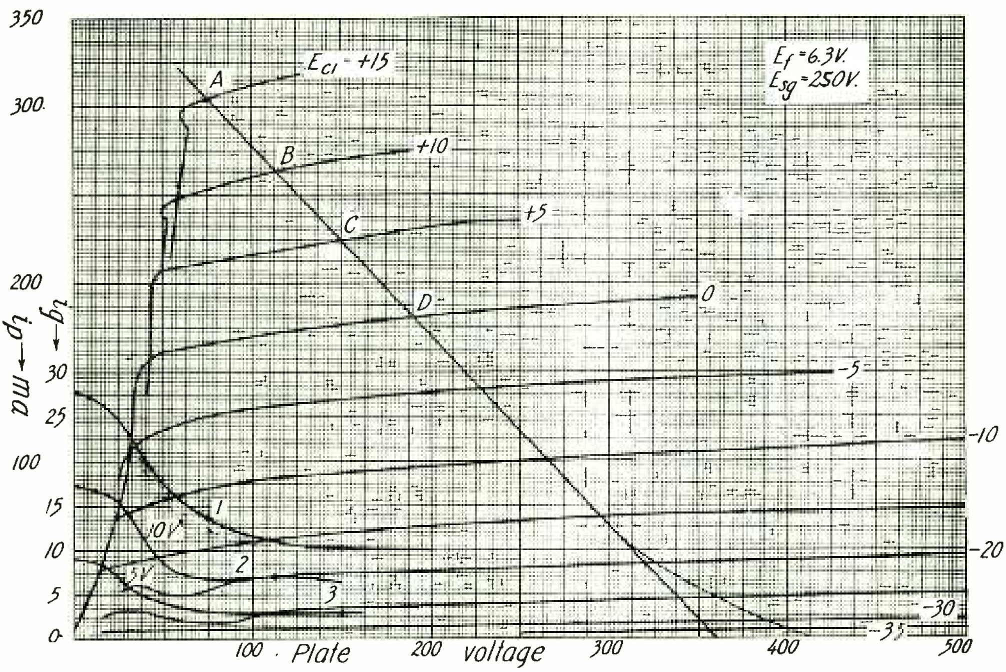



Fig. 62 —Determination of grid current drawn by each tube of a 6L6 push-pull stage.

straight line portion that involves the larger grid swings. This line has the slope $R_L/4$ and when prolonged, intersects the plate-voltage axis at point $E_{bb}$. (This was discussed previously).

Consider a grid swing of +15 volts. This intersects the path of operation at point A. The plate current at that instant is 305 ma., and the plate potential is 75 volts. The ordinate through A intersects the +15-volt grid-current curve at point 1. The grid current is then given as 14 ma. This is the instantaneous grid current for that grid swing and the corresponding plate voltage.

Furthermore, the *load current* flowing through the *plate-to-plate resistor* $R_L$ is half of the plate current, owing to the 2:1 step down in current from the half winding to the entire primary winding across which $R_L$ is assumed to be connected. The load current therefore has the value of 305/2 = 152.5 ma.

Next take point B corresponding to the +10-volt grid swing. The plate current is 263 ma; the plate potential is 112.5 volts; and the grid current (point 2) is 7 ma. The load current is 263/2 = 131.5 ma. In similar manner the load and grid currents for point C (and corresponding point 3), can be found; they are respectively 225/2 = 112.5 ma. and 3.0 ma.

Other values of *load* current for smaller grid swings (that do not draw grid current) are also determined. In the present example values down to that for a +10-volt grid swing will be sufficient. This will embrace the straight-line portion ABCD and yet avoid the bottom curved portion, which is more difficult to determine since it represents that part of the cycle where both tubes are operative. (Straight-line portion ABCD represents the part of the cycle where one tube is beyond cutoff and hence inoperative.)

The values of load current and grid current are now plotted against grid swing, AS MEASURED FROM THE BIAS POINT. The plot is shown in Fig. 63. Since the bias is -20 volts, a positive grid swing of +15 volts, for example corresponds to a total grid swing of 15 + 20 = 35 volts. The corresponding load cur-

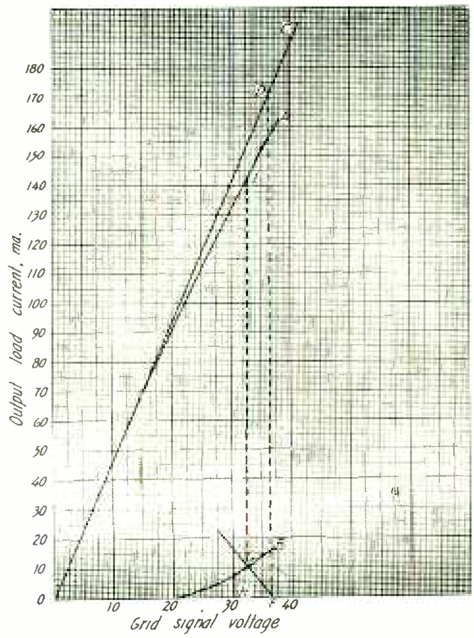



**Fig. 63.—Graphical constructions for driver-stage design.**

rent was found to be 305/2 = 152.5 ma., and the grid current, 14 ma.: these are plotted as shown in Fig. 63.

In a similar manner the other points are plotted for instantaneous grid swings of 30, 25, 20, 15, and 10 volts. For zero grid swing the load current is obviously zero, so that the load current plot OA in Fig. 63 passes through the origin. The grid-current curve OB reaches zero at a 20-volt swing, since this just cancels the -20-volt bias and brings the grid up to zero volts. For lesser swings the grid is negative relative to the cathode and therefore draws no current.

Now suppose that 40 watts *fundamental* power output is desired, and that the permissible distortion is 5%, a typical value. In the prior calculation of 43.5 watts, no grid voltage distortion was assumed; indeed, no distortion of any kind was assumed. This means that curve OA in Fig. 63 was assumed to be a straight line; i.e., the output load current was directly proportional to the input signal voltage.

Actually, OA is not a straight line, but curves in this case below a straight line OC drawn tangent to it at the origin. This means that even without grid-voltage distortion, a distorted output will be obtained. If, however, the dropping characteristic of OA relative to OC, or "undershoot", is not excessive, additional undershoot produced by grid current can be permitted, and yet keep the total distortion within the 5% specified.

In the case of some tubes, or perhaps for lower values of $R_L$, an "overshoot" can be obtained; i.e., OA will lie above OC. In this case the flattening effect of grid current will be to lower OA down to OC, and then, if sufficiently great, to cause OA to lie below OC. In other words, extreme grid flattening can convert an overshoot into an undershoot.

In the case at hand, the analysis starts with an undershoot and hence under a handicap. First determine the peak fundamental current $I_f$ required to produce 40 watts in a 3600-ohm plate-to-plate load resistance. This is

$$I_f = \sqrt{\frac{2P_o}{R}} \qquad (36)$$

or

$$I_f = \sqrt{\frac{2 \times 40}{3600}} = 149.5 \text{ ma}$$

If in the general case the percentage permissible third-harmonic distortion be denoted by n, then the following two quantities have next to be calculated: $(1 + 3n)I_f$ and $(1 - n)I_f$. In the problem at hand these are, for n = .05,

$$(1 + 3 \times .05)(149.5)$$

$$= (1.15)(149.5) = 172.0 \text{ ma.}$$

and

$$(1 - .05)(149.5) = 142.0 \text{ ma.}$$

Now refer to Fig. 63. On the tangent straight line OC lay off 172.0 ma., (point D). Next locate 142 ma. on the load-current curve OA (point E). Now project point D down to the grid-signal-voltage axis, point F; and project E down to the grid-current curve OB, meeting it in point G.

Join F to G. Then FG is the load line for the maximum permissible driver resistance *as viewed from the grid terminals*, that will permit 40 watts output with but 5% third-harmonic distortion.

The value of this resistance is simply (to scale)

$$R_D = \frac{HF}{GH} = \frac{4}{.01} = 400 \text{ ohms.}$$

The driver circuit must therefore be so designed as to appear as a 400-ohm source to either grid. However, the actual driver is a vacuum tube (or tubes) whose internal (plate) resistance is on the order of thousands of ohms. In order for it to appear as only 400 ohms, a step-down transformer is required between it and the 6L6 grids.

This step down transformer is generally called a driver transformer, and it is necessary to determine not only its step down ratio, but its winding resistances as well.

First a driver tube or tubes must be selected. No rule can be given as to the choice, but after a tube has been selected, it can be checked by the following procedure to see if it is suitable. In general, a triode is preferred because it has a lower internal resistance compared to a pentode and is therefore not as sensitive to load resistance changes, such as the extreme variations produced by a grid circuit which draws current only at the peaks of the signal cycle.

Suppose a single 6J5 tube is used as the driver. From the Tube Manual the following data is obtained:

| | |
|---|---|
| $E_b$ = 250 volts | $\mu$ = 20 |
| $E_c$ = -8 volts | $R_p$ = 7700 ohms. |

If the tube fed the grids directly, they would be fed by a source having an internal resistance of 7700 ohms. This is much too high, and would cause an excessive undershoot and hence excessive amount of third- (and higher) harmonic distortion. The proper source impedance is 400 ohms. Hence a stepdown driver transformer must be interposed between grid and the 6J5 tube.

The circuit is shown in Fig. 64. The driver transformer has a primary winding resistance $R_{pw}$, and a secondary winding resistance of $R_{sw}$ for each HALF of the secondary. Furthermore, there is a stepdown of a : 1 from the entire primary to each HALF of the secondary; in this way the source resistance can be made to appear as 400 ohms to either grid. It remains to determine a, and also incidentally $R_{pw}$ and $R_{sw}$, which contribute to the source impedance.

It was shown in a previous assignment that in transformer coupling, the voltage across the primary is practically mu times the input voltage $e_s$. Then the voltage appearing across the secondary is $\mu e_s/a$ in the case of the driver transformer. For the 6J5 tube, where $\mu$ = 20, this is $20e_s/a$.

The maximum grid swing of the 6J5 tube is equal to the bias voltage, $E_c$, —here 8 volts. Hence the maximum voltage that can be delivered to either 6L6 grid is 8 × 20/a = 160/a. Since from Fig. 63, 36.5 volts are required to swing either grid 15 volts positive, a must be at least 160/36.5 = 4.38 : 1.

However, a must also make the tube and transformer resistances appear as 400 ohms to either grid; this is the more exacting requirement and hence a will be determined to bring this about. It can then be checked to see that it does not exceed 4.38 : 1, otherwise there will be insufficient drive for either 6L6 grid.

The $R_p$ of the 6J5 tube is known:

it is 7700 ohms; but the winding resistances $R_{pw}$ and $R_{sw}$ are not known. As a reasonable value, assume that $R_{pw} + a^2R_{sw}$ = 10% of 7,700 ohms, or 770 ohms. Then the total resistance on the primary side of the driver transformer is 7,700 + 770 = 8,470 ohms.

This must be stepped down to 400 ohms, the desired driver resistance $R_D$. Hence

$$a = \sqrt{\frac{1.1R_p}{R_D}} \qquad (37)$$

or

$$a = \sqrt{8470/400} = 4.6$$

If it is further assumed that the primary and secondary losses are about equal, then $R_{pw} = a^2R_{sw}$, so that either equals half of 770 ohms or 385 ohms. Thus $R_{pw}$ = 385 ohms, and $a^2R_{sw}$ = 385 ohms, or

$$R_{sw} = 385/(4.6)^2 = 18.15 \text{ ohms.}$$

However, a = 4.6 exceeds the highest value permitted for full grid swing of 36.5 volts; namely, 4.38. Hence the design is inadequate; a larger driver stage is required. Either a larger tube can be used, or two 6J5 tubes can be used in parallel or in push-pull. Suppose two are employed in push-pull.

Then they act as two generators in series. The peak grid swing is twice that of one tube alone, or 2 × 8 = 16 volts; the mu remains unchanged at 20; and the total internal resistance is $2R_p$ = 2 × 7700 = 15400 ohms. Now the primary winding resistance $R_{pw}$ is that of the entire primary fed by both 6J5 tubes, and a refers, as before, to the step-down ratio between the entire primary and one-half secondary.

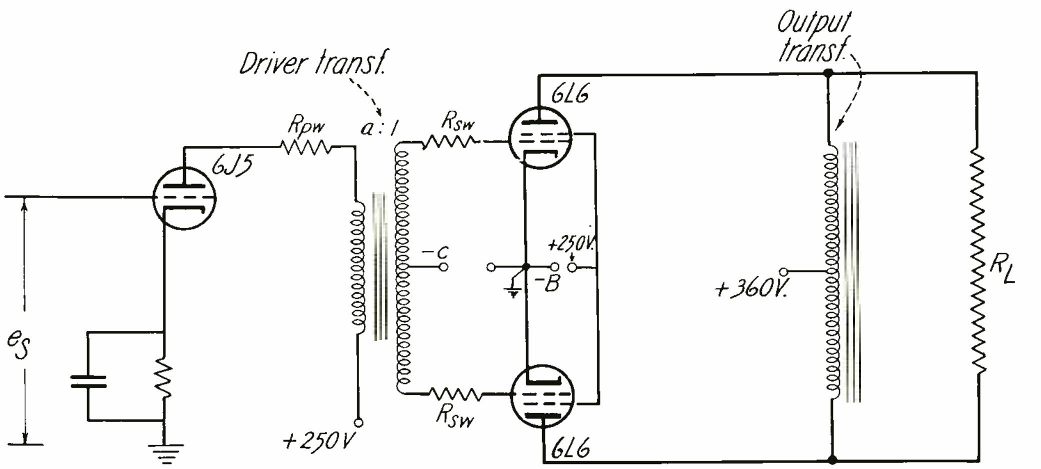



Fig. 64.—Use of single 6J5 tube to drive the grids of a 6L6 push-pull Class $AB_2$ stage.

With these facts and figures, substitution in Eq. (8) yields

$$a = \sqrt{\frac{1.1 \times 15400}{400}} = 6.5$$

The step down is greater than before; specifically it is $\sqrt{2}$ as great. But the apparent generated voltage $\mu e_s$ is double, hence it appears as $\sqrt{2}$ times as great at the grid terminals. It is

$$2 \times 8 \times 20/6.5 = 49.2 \text{ volts.}$$

This is clearly in excess of the 36.5 volts that is required to appear at either grid, hence this push-pull combination is more than adequate as a driver stage.

It is desired to point out once again that the open-circuit voltage appearing across either half of the secondary will be 36.5 volts peak, which the grid current will reduce to 35 volts, or 15 volts positive, and give rise to the undershoot OEA shown in Fig. 63. This will correspond to a permissible value of 5 per cent third-harmonic distortion.

This concludes the section on push-pull amplifiers. Normally Class $AB_1$ is preferred to Class $AB_2$ because of the difficulties involved in preventing grid-current distortion; nevertheless where the occasion arises for increased power output, Class $AB_2$ is a perfectly feasible means for obtaining it.

## RESUME'

This concludes the assignment on Part I of Audio Amplifiers. After a discussion of the meaning and use of decibels, a general, preliminary analysis was made of the amplifier gain required for a given input and output power. The number and kind of stages were calculated, but no details of the circuits worked out.

Then, after a discussion of the significance of frequency response characteristics, the various types of voltage amplifiers were analyzed, particularly resistance-coupled stages, and the design considerations pertaining to low-, intermediate-, and high-frequency response worked out. Following this, the design formulas and curves for screen grid and cathode self-bias bypass capacitors were presented, and the effect on the low-frequency response illustrated.

The assignment then continued with a discussion of the various types of audio transformers encountered, and the design factors that determine their behavior. No attempt was made to discuss the design of such transformers, since this is a very specialized subject, but practical methods of modifying the response characteristics were discussed in detail.

The concluding topic was push-pull amplifiers. The operation of triodes and pentodes in push-pull were analyzed in detail for the various modes of operation, and the driver design taken up for the case where the power tube grids are driven positive.

*AUDIO FREQUENCY AMPLIFICATION PART I*

## EXAMINATION

Note: In all these problems, 0 db = 1 milliwatt.

1. The input power to an amplifier is $5 \times 10^{-3}$ μwatt. The output power desired is 20 watts.

   a). Express the input level in db.

   b). Express the output level in db.

   c). What is the db gain?

   d). What is the power ratio?

2. a). A microphone has an output level of -73 db at a sound pressure of 10 bars. What is the power in watts?

2. b). The voltage developed across a dummy load resistor of 8 ohms is 15 volts. What is the power consumed in the resistor?

c). What is the db level?

3. In the frequency-response test setup of Fig. 7 in the text, suppose the db meter reads +8 db at the audio oscillator terminals, and +15 db at the dummy load resistor, and suppose further that this resistor has a value of 5 ohms. Assume the attenuation box is set for -54 db attenuation, and that the input impedance of the amplifier is the same as that of the attenuator box, namely, 500 ohms.

a). What is the input level to the amplifier?

b). What is the output level of the amplifier?

c). What is the db gain of the amplifier?

4. **(Reference Question 3.)**

a). What is the power output of the audio oscillator in watts?

b). What is the power input to the attenuation box in watts?

c). What is the power input to the amplifier in watts?

d). What is the power output of the amplifier in watts?

5. Take the example in the text of the 6SJ7 tube resistance coupled. The constants are: $R_p$ = 1 megohm, $G_m$ = 1650 μmhos, $R_L$ = 250,000 ohms, $R_g$ = 500,000 ohms. Suppose a 1 db drop is permitted at 30 c.p.s.

a). Find the low-frequency time constant permitted.

5. b). Find the coupling capacitor required.

6. Refer to the example in the text on the resistance-coupled amplifier using a 6SJ7 tube. Assume this tube has a total shunt capacitance of $C = 35\ \mu\mu f$. What will be the attenuation in db at 15,000 c.p.s.? Use $R_g' = 250{,}000\Omega$.

7. Reference to Question 6. Suppose the low-frequency attenuation is specified as 0.5 db at 30 c.p.s. Find the screen time constant and screen bypass capacitor, using the various circuit constants given in the text.

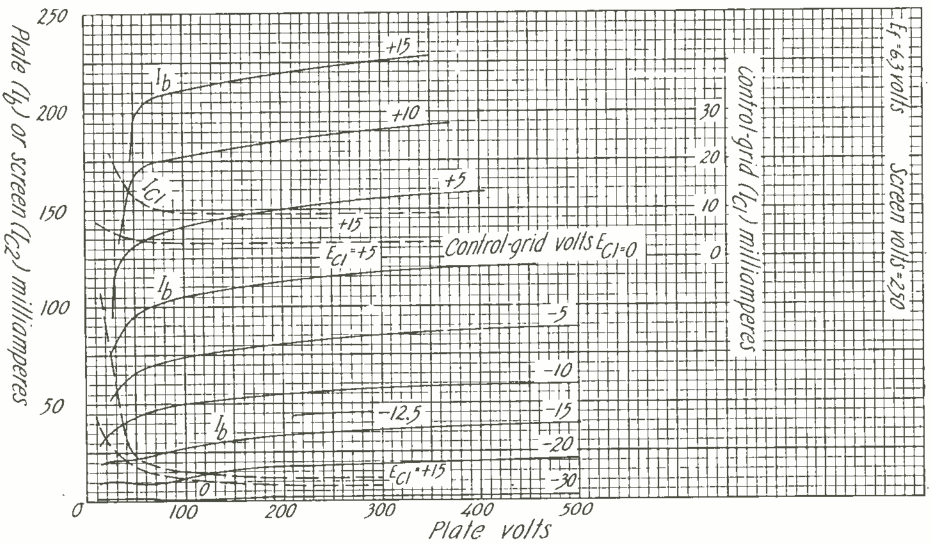
8. Refer to the example in the text on the self-bias bypass capacitor. Use the various circuit constants given, but assume the attenuation permitted is 0.5 db at 30 c.p.s. Find the magnitude of the bypass capacitor.

9. a). In an interstage audio transformer, what is the effect of connecting a resistor across the primary terminals with regard to the low-, intermediate-, and high-frequency response?

b). What is the effect of connecting an equivalent resistance across the secondary terminals?

c). Suppose, instead, that a resistance equal to the plate resistance of the tube is connected IN SERIES with the plate and the primary of the transformer. How will this affect the frequency response?

10. Given the pentode tube characteristics for a 6V6 tube as shown in the accompanying figure. The plate potential is 300 volts, the screen potential is 250 volts, and the maximum plate dissipation is 12 watts. Class $AB_1$ operation is contemplated.




a). Find the zero-signal plate current and bias voltage per tube. (Interpolate for bias voltage).

b). What is the plate-to-plate load resistance, $R_L$?

c). What is the power output?

d). What is the peak signal voltage?

e). What is the d-c plate current per tube at full signal?

f). What is the plate dissipation at full signal? Note: This must be equal to or less than 12 watts per tube. If in excess, $R_L$ must be increased, or the grid swing reduced, or both.