SPECIALIZED TELEVISION ENGINEERING

TELEVISION TECHNICAL ASSIGNMENT

VACUUM TUBE AMPLIFIERS

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INTRODUCTION

Since amplifiers consist of tubes and associated circuits, a discussion of amplifiers must deal with the behavior of the tubes and with the behavior of their associated circuits. As regards the tube, it can be operated fundamentally in one of three ways:

1. *Class A.*—In this mode of operation the output voltage is a faithful copy of the input grid signal voltage, and in general a modest amount of power output is obtained. Class A operation is primarily used for audio and television amplifiers because any appreciable distortion will immediately be noticeable to the ear or eye, and it is also employed in small r.f. and i.f. amplifiers used in radio receivers because it, in general, gives the most voltage amplification of all three ways of operations.

2. *Class B.*—In this mode of operation the tube is operated over a greater part of its characteristic, including cutoff. Much greater power output is possible than in the Class A mode of operation, but the distortion is very much greater. This can be minimized, however, by employing two tubes in push-pull whereupon acceptable performance, even from audio amplifiers, is obtained. However, satisfactory performance is obtained in r.f. work, using but a single tube, by having a tuned circuit as the plate load. The tuned circuit, often called a *tank circuit,* filters out the distortion products generated by the tube, and gives an r.f. output essentially of the same waveshape (and frequency) as the input signal. Such an amplifier is often called a *linear amplifier,* for reasons to appear. As such, it is employed in many television transmitters; notably those manufactured by General Electric and DuMont.

3. *Class C.*—In this mode of operation the tube is operated over an extremely large part of its characteristic. The plate current flows in narrow pulses, which shock-excite a plate tank circuit. The tank circuit thereupon furnishes an essentially pure sine wave of the same frequency as the grid input signal, in spite of the high distortion in the plate current. If, on the other hand, the tank circuit is tuned to a multiple of the input signal frequency, then its output will be a sine wave of a frequency that is the same multiple of the input frequency. The amplifier stage in this case is known as a *frequency-multiplier* stage; frequency doubling, tripling, and even quadrupling, is often employed.

A Class C amplifier can furnish much greater power output than either a Class B or Class A and at a higher efficiency, but as indicated above, it can amplify only a single frequency (carrier frequency). It is particularly suited for transmitter installations, although it must be noted that the local oscillator in a superheterodyne receiver is essentially a Class C amplifier with the addition of regenerative feedback. FM transmitters are an important example of the use of frequency doubling and Class C stages in general.

There is a fourth mode of operation known as *Class AB*. In this mode the tube is operated over a range of its characteristic that is intermediate between Class A and Class B operation. Intermediate results in practically all respects are obtained: the power output and plate efficiency is intermediate between that obtained in Class A and that obtained in Class B; the distortion is intermediate in amount, etc.

The circuit associated with the tube is an important consideration because of the two basic types of amplifiers used, as regarded from a frequency viewpoint:

1. *Broad-frequency-band amplifiers.* Ordinary audio (speech) and video (television) amplifiers are called upon to amplify *simultaneously* a number of frequencies. The complex wave representing the audio or video signal can be resolved into a series of sine waves of the proper amplitude and phase (relative to a particular moment of time chosen). In the case of an audio amplifier, the relative amplitudes must be preserved, although the phase shifts are not important unless excessive. In the case of a video amplifier *both the amplitude and phase relations must be preserved.*

Since the tube, at these relatively low frequencies, can be assumed to amplify almost instantaneously, and also may be regarded as having an internal impedance that is practically resistive in nature, any frequency discrimination as to phase and amplitude must be owing to the plate load impedance rather than to the tube itself. Hence, the load impedance must be correctly designed to function properly over the frequency band desired.

2. *Narrow-frequency-band amplifiers.* Reference is here made to amplifiers that are called upon to amplify but one frequency--such as a carrier wave--; or at most, a narrow band of frequencies, such as a carrier and its side bands. In this case a tuned circuit of some sort can be employed as the plate load, since the latter need have the correct or optimum impedance only over the narrow band to be amplified.

It is clear, however, that the mode of operation of the tube is also involved. In broad-band amplifiers, the tube must be operated Class A, or at most, Class B in a push-pull arrangement, in order that the tube does not generate spurious frequencies within the frequency band that will appear in the output circuit. Such spurious frequencies (not present in the input signal) are considered distortion products, and are highly undesirable.

On the other hand, in the case of a narrow-band amplifier, the distortion products generated by the tube will be filtered out by the tuned plate load circuit (as mentioned previously) and hence will not affect the output wave-shape. As a result, Class B and C operation (for one tube, or two in push-pull) are feasible and desirable for narrow-band operation because of the greater operating efficiency and the greater output possible from a given size tube.

VOLTAGE AMPLIFICATION

Vacuum-tube amplifiers are employed to amplify weak signal powers to a magnitude sufficient to actuate the final equipment, such as a loudspeaker or picture tube. In most cases the source generates a signal of but a fraction of a volt. This is far too small to produce any appreciable output power even when applied to a vacuum-tube power amplifier; it is first necessary to amplify the *voltage* until it is finally strong enough to actuate a vacuum-tube power amplifier and furnish the required power output.

There is, therefore, a need for amplifiers that amplify primarily the voltage, and a need for amplifiers that furnish the required power output. These are known respectively as voltage and power amplifiers, and will be discussed in turn in this assignment.

FUNDAMENTAL CONSIDERATIONS. In Fig. 1 is shown a voltage amplifier in perfectly general form.

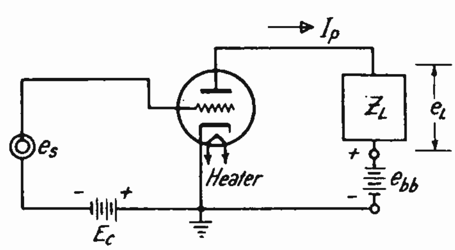


Fig. 1.--Voltage amplifier in general form.

The input signal e_s is applied to the grid of the tube in series with the bias source E_c. (Grid bias will be discussed in greater detail farther on.) In the plate circuit is interposed a plate load impedance Z_L. This may be of any form desired; resistance, inductance, tuned circuit, etc., but must fulfill one requirement: it must be able to pass direct current. Hence, a purely capacitive circuit cannot be employed as a plate load.

The signal e_s on the grid causes corresponding variations in the plate current I_p, i.e., cause the plate current to pulsate. These pulsations, or a.c. component, as they are called, produce an a.c. voltage e_L across Z_L. The voltage e_L is, in general, many times the amplitude of e_s, *but of the same wave-shape*, and is therefore an amplified copy of e_s. It can be used to actuate the grid of a following tube, from which a still larger voltage of the same wave-shape can be obtained, and so on until a magnitude of voltage is obtained sufficient to actuate a power tube and furnish the power output required.

Each stage except the last is known as a *voltage amplifier* stage, and differs from a power output stage in that the load impedance Z_L is chosen of a magnitude and type best suited to obtain maximum *voltage* output e_L, whereas in the case of a power output stage Z_L is suitably chosen to furnish maximum *power* output.

GENERAL VOLTAGE AMPLIFICATION FORMULAS.--In evaluating the magnitude of e_L for a given e_s, tube, and Z_L, advantage is taken of a certain fundamental theorem known as the Equivalent Plate Circuit Theorem. This states that as far as the external plate load resistor is concerned, the tube acts as if it were a generator whose generated voltage is μ times the input signal (μe_s), and whose internal impedance is a resistance of a value R_p (plate resistance) depending upon the tube. This is illustrated in Fig. 2.

Actually the tube is a kind of variable resistance whose magnitude is electrically controlled by a signal voltage applied to its

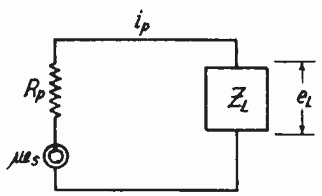


Fig. 2.—Equivalent plate circuit and plate load impedance.

grid, in series with a source of d.c. and the plate load resistance. The Equivalent Plate Circuit Theorem states that such a device can be equally well represented by the more conventional form of circuit shown in Fig. 2, which is more amenable to analysis and calculation. As shown, only the a.c. component i_p of the current and the voltage e_L can be calculated, but since these are the only components of interest, the circuit is satisfactorily general in its application. Thus, by Ohm's Law:

$$i_p = \frac{\mu e_s}{R_p + Z_L} \qquad (1)$$

and

$$e_L = i_p Z_L = \frac{\mu e_s Z_L}{R_p + Z_L} \qquad (2)$$

For a given tube, μ and R_p are known, and if Z_L and e_s are given, e_L can be found from Eq. (2). Usually, the item of interest is the *amplification factor of the stage*, or the *stage gain*, α. This is the ratio of the output to the input voltage, e_L/e_s. From Eq. (2) it is seen that

$$\alpha = e_L/e_s = \mu Z_L/(R_p + Z_L) \qquad (3)$$

An inspection of this equation reveals many important practical facts:

1. The higher the μ of the tube, the greater is the stage gain α.

2. The higher R_p is, the larger is the denominator, and the lower is the fraction denoting α.

3. On the other hand, the greater Z_L is, the greater α will be. This can be seen more clearly by dividing the numerator and denominator of the right-hand side of Eq. (3) by Z_L, and obtaining

$$\alpha = \mu/\left[\frac{R_p}{Z_L} + 1\right] \qquad (3a)$$

If Z_L is increased, R_p/Z_L is decreased, as is also true for $([R_p/Z_L] + 1)$. Since this is the denominator, a decrease in this factor increases α; hence, α increases with Z_L.

4. As Z_L increases, R_p/Z_L approaches zero, and Eq. (3a) reduces to

$$\alpha \cong \mu \qquad (3b)$$

or α approaches the tube μ *as a maximum or limit value* as Z_L increases to infinity. This indicates, in general, the value of making Z_L as high as possible (particularly relative to R_p), and also indicates that the stage gain in normal practice is lower than the tube μ.

Referring to items (1) and (2), it is clear that the tube best suited for voltage amplification is one having a high μ and a low R_p. This can be alternatively expressed as follows: the best tube is one having the maximum ratio of μ/R_p, since as this ratio is larger, the greater μ is, and/or the smaller R_p is. However, $\mu/R_p = G_m$, the transconductance of the tube, hence for voltage amplification a high transconductance tube is desirable.

It will be seen later that a high transconductance is also desirable for high *power output* from a tube, hence this property is not unique to voltage amplifier tubes. Indeed, in the latter case, it is really based on the assumption that Z_L is a given fixed quantity, so that the smaller R_p is, the more nearly will α equal μ, the maximum value.

However, in normal practice Z_L can be chosen as high as desired, within reasonable limits. Eq. (3a) shows that if Z_L is made large enough, R_p/Z_L, and hence $(R_p/Z_L + 1)$, can be made small enough *even if R_p is large*, so that Eq. (3b) obtains. In other words, for voltage amplification the major requirement of a tube is that it have a high μ. If it also has a high R_p, so that $\mu/R_p = G_m$ is but a normal value (2,000 μmhos), then all one need do is to choose a relatively high Z_L compared to R_p, and a value of α approaching the value of μ will be obtained.

Thus, for voltage amplification, a pentode (which has an inherently high μ) or at least a high-μ triode tube is preferred. In this way a value of α between 50 and 100 is easily obtained. The only limitation to the value of R_p (which is normally high if μ is high), and hence to the corresponding value of Z_L, is the band width over which the gain must be uniform. In audio work, the band width is at most from 30 to 20,000 c.p.s., and is sufficiently narrow so that it imposes no practical limitation on the magnitude of Z_L and hence of R_p.

On the other hand, a video amplifier used in television may be called upon to have a band width of from 30 c.p.s. to 4.5 mc or possibly higher. Such an enormous band requires correspondingly *low values of Z_L*, whereupon the lower R_p is, the higher α will be. Hence, in such broad-band amplifiers a tube with as high a G_m as possible is desirable, just as in the case of power tubes. However, even here R_p can be high, provided μ is sufficiently large so that the ratio $\mu/R_p = G_m$ is large.

This will be discussed in greater detail when specific types of voltage amplifiers are considered. At this point it is desirable to discuss another formula for voltage amplification that is used possibly more often than Eq. (3). It is very easily derived from the latter:

$$\alpha = \mu Z_L/(R_p + Z_L) \qquad (3)$$

Thus, multiply numerator and denominator by R_p, and separate the factors so as to bring out the physical significance of the resulting expression. There is obtained:

$$\alpha = \frac{\mu}{R_p}\left(\frac{R_p Z_L}{R_p + Z_L}\right)$$

$$= G_m\left(\frac{R_p Z_L}{R_p + Z_L}\right) \qquad (4)$$

Note that this states that the

stage gain is equal to the product of the transconductance of the tube by an impedance *consisting of* R_p *and* Z_L *in parallel.* This formula is based on a theorem sometimes called Norton's Theorem, and which applies *to all circuits, even those that do not contain vacuum tubes.*

Norton's Theorem states that a circuit having a generator of internal impedance Z_G and a constant-generated voltage E_G, in series with a load impedance Z_L, is equivalent to a constant-current generator which feeds its current into Z_L and Z_G in parallel, and whose current is of a magnitude E_G/Z_G. This is illustrated

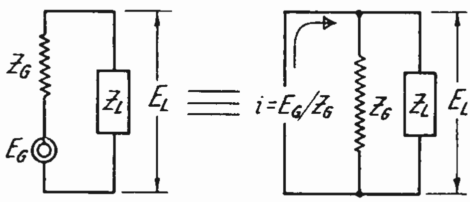


Fig. 3.--Constant-voltage and constant-current equivalent circuits.

in Fig. 3. Notice that in the right-hand circuit the constant-current is E_G/Z_G, i.e., the current that would flow if the generator terminals were directly short-circuited.

The concept of a constant-current generator may require some discussion. Ordinarily, if a source generates say, 10 volts, and has an internal resistance of 5 ohms, then if it be short-circuited, 10/5 = 2 amperes flow. If it be connected to a load of 5 ohms, then 10/(5 + 5) = 1 ampere flows; if the load is 20 ohms, then 10/(5 + 20) = 0.4 ampere flows, and so on. Thus, the current *decreases* from its short-circuit value as the load impedance is increased.

Suppose, however, that the source generated 10,000,000 volts, and had an internal resistance of 5,000,000 ohms. Then, upon short-circuit, 10,000,000/5,000,000 = 2 amperes would flow, just as in the above case. If now 5 ohms be placed in series with the generator, the current that flows is

$$10{,}000{,}000/(5{,}000{,}000 + 5)$$

or practically 2 amperes, the same as on short circuit. Indeed, even if 100,000 ohms is placed in series with the generator, the current is reduced to 10,000,000/5,100,000 = 1.961 ampere or only by about 2 per cent. Thus, the current is practically constant for a range of load impedances from 0 to 100,000 ohms, which is a range sufficient to cover practically most load impedances encountered in vacuum-tube work.

Such a source can be considered a constant-current generator. Theoretically, the source impedance and generated voltage would have to be infinite *in the same ratio* to fulfill the constant-current requirement exactly, but in practice a high-impedance source, *such as a pentode tube,* can be considered a constant-current generator.

Returning to Fig. 3, it is to be noted that the left-hand source generates a constant voltage E_G; it is equivalent to a constant-current generator that generates a current E_G/Z_G. The series internal impedance of the left-hand source becomes a parallel impedance Z_G in the right-hand circuit; the current

E_G/Z_G flowing through Z_G and Z_L in parallel, sets up the same voltage E_L across Z_L as is produced in the left-hand circuit. Thus, as far as the external load Z_L is concerned, the two generators, or rather circuits, produce equivalent results.

As an illustration, suppose $E_G = 36$ volts, $Z_G = 6$ ohms, and $Z_L = 12$ ohms, both resistive. From the left-hand constant-voltage circuit, the current flow I_L is, by Ohm's law,

$$I_L = 36/(6 + 12) = 2 \text{ amperes}$$

Then the voltage across Z_L is

$$E_L = I_L Z_L = 2 \times 12 = 24 \text{ volts}$$

Now use the right-hand constant-current circuit. The constant current is

$$I = E_G/Z_G = 36/6 = 6 \text{ amperes}$$

This current flows through Z_G and Z_L in parallel, a joint impedance of

$$Z = Z_G Z_L/(Z_G + Z_L)$$

$$= 6 \times 12/(6 + 12)$$

$$= 4 \text{ ohms}$$

Then

$$E_L = IZ = 6 \times 4 = 24 \text{ volts}$$

which is exactly the same value as obtained from the left-hand circuit.

When this theorem is applied to a vacuum-tube circuit, the short-circuit (a.c.) current of the tube is that which flows when its plate is connected directly to B+, and a certain signal voltage e_s is applied to the grid. It will be recalled from the Equivalent Plate Circuit Theorem that the apparent generated voltage is μe_s. Hence the short-circuit current is

$$i_s = \mu e_s/R_p = e_s G_m \qquad (5)$$

On a per volt basis ($e_s = 1$ volt), the short-circuit current is G_m amperes, milliamperes, or microamperes, depending upon whether G_m itself is expressed in mhos, millimhos, or—as is more usually the case—in micromhos.

Eq. (4) is then easy to interpret. For a grid signal e_s of 1 volt, the short-circuit current is G_m in magnitude, and the output voltage is that developed across R_p and Z_L in parallel when the current of magnitude G_m flows through it. It is therefore

$$e_L = G_m R_p Z_L/(R_p + Z_L) \qquad (6)$$

But since the circuit gain $\alpha = e_L/e_s$, if e_s is chosen as 1 volt, then $\alpha = e_L/1 = e_L$. Hence, in the above Eq. (6), e_L represents α since it is the output for 1-volt input. Thus, the gain may be written in one of two forms:

$$\alpha = \mu Z_L/(R_p + Z_L)$$

or

$$\alpha = G_m \left(\frac{R_p Z_L}{R_p + Z_L}\right) \qquad (4)$$

As an example, suppose $R_p = 10{,}000$ ohms, $Z_L = 40{,}000$ ohms (resistive), and $\mu = 20$. Then, by definition, $G_m = \mu/R_p = 20/10{,}000 = .002$ mhos $= 2{,}000$ μmhos. The gain, by Eq. (3), is

$$\alpha = \frac{20 \times 40000}{10000 + 40000} = 16$$

and by Eq. (4), is

$$\alpha = .002 \left[\frac{10000 \times 40000}{10000 + 40000}\right]$$

$$= .002 \times 8000 = 16$$

Thus, either equation and circuit gives the same value for the circuit gain. The constant-current form [Eq. (4)] is particularly valuable when pentode tubes are employed. These tubes have a very high-plate resistance R_p, usually 1 megohm or more. The load impedance Z_L is usually considerably less, normally not more than 250,000 ohms at the most. In such a case the shunting effect of R_p on Z_L is practically negligible, so that R_p and Z_L in parallel is essentially equal to Z_L alone. Eq. (4) then reduces to the relatively simple form of

$$\alpha \doteq G_m Z_L \qquad (4a)$$

This formula is particularly useful in the case of television (video) amplifiers, where Z_L is on the order of 1,000 ohms and is very low compared to R_p.

Thus, if G_m = 2,000 μmhos, R_p = 1,000,000 ohms, and Z_L = 100,000 ohms (resistive), then R_p and Z_L in parallel are

$$Z = \frac{R_p Z_L}{R_p + Z_L} = \frac{10^6 + 10^5}{10^6 + 10^5}$$

$$= 90{,}909 \text{ ohms}$$

and the gain, by Eq. (4), is

$$\alpha = .002 \times 90{,}909 = 181.818 \text{ or } 182$$

By the more approximate method of Eq. (4a),

$$\alpha = .002 \times 10^5 = 200$$

The error is approximately 10 per cent. If Z_L were 50,000 ohms, then the corresponding values would be

$$\alpha = (.002) \left[\frac{10^6 \times 5 \times 10^4}{10^6 + 5 \times 10^4}\right]$$

$$= (.002)\,(47{,}619) = 95.2$$

and

$$\alpha = (.002)\,(5 \times 10^4) = 100$$

or an error of 5 per cent. Usually G_m is not known to any greater accuracy than this, particularly since it varies more than this from one tube to another, and also varies with the tube voltages for any given tube by more than 5 per cent.

RESISTANCE-COUPLED AMPLIFIERS.—The preceding discussion has been very general; no particular type of load impedance was specified. The requirement that Z_L afford a d.c. path for the plate current eliminates the use of a capacitor alone, but such a circuit element can be used in parallel with an inductance or resistance to furnish a suitable plate load impedance.

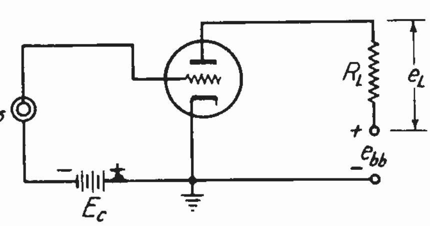


Fig. 4.--Elementary resistance-coupled circuit.

One of the simplest load impedances empolyed in a resistance R_L, as indicated in Fig. 4. The circuit is known as a resistance-coupled amplifier stage, although the complete circuit contains an additional R-C branch to couple the grid of the next stage to the plate of this stage. This type of coupling circuit will be analyzed here from a graphical viewpoint. In a later assignment the frequency response will be derived for this circuit, as well as that for several other types of circuits.

The resistance-coupled stage is almost universally employed today for voltage amplification because of its ability to amplify uniformly a wide range of frequencies, its economy, simplicity, and compactness, and its relative freedom from hum pickup. In a modified form it is employed in the wide-range video circuits used in television, and hence merits detailed study.

METHOD OF GRID COUPLING.--The voltage developed across the plate load resistance, has both the positive d.c. component from the plate supply and the a.c. signal component owing to the action of the grid of the tube. If the following grid is connected to the plate load impedance, it will receive not only the signal voltage, but the d.c. component as well, and will be *positive* with respect to ground. Since the cathode of the following tube (as well as all other tubes) is preferably operated at ground potential (since it must be negative with respect to its plate, which is connected to the B supply) it is clear that the grid will have a positive bias with respect to the cathode.

Grid current will flow, and besides overheating the grid and causing excessive plate current to flow in the tube, this current will present an appreciable shunt load to the previous tube. To prevent these effects, the grid is normally operated at a *negative* bias with respect to its cathode. Some sort of network must be employed that will permit the a.c. signal voltage of the plate load resistor to be impressed upon the following grid, but will block the positive d.c. voltage, thus permitting a separate d.c. negative source to bias the grid negative with respect to its cathode.

The type of network employed is ordinarily an R-C combination,

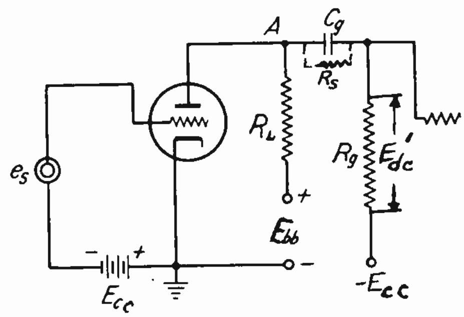


Fig. 5.--Method of coupling grid of one stage to plate load impedance of preceding stage.

and is shown in Fig. 5 as R_g C_g. Across R_L is developed the signal output voltage e_L, and point A is positive to ground by an amount equal to the B supply voltage E_{bb} minus the d.c. voltage drop in R_L. For example, if E_{bb} = 300 volts, and 120 volts is lost across R_L, then the potential of A with respect to ground is

300 - 120 = 180 volts positive

If the grid were directly connected

to A, it would be 180 volts positive to ground and hence to its grounded cathode. By employing C_g, however, this positive d.c. voltage is blocked from the grid, so that the latter may be biased negative to its cathode by means of the C-bias battery shown applied to the bottom end of R_g.

The design of this part of the circuit will be developed in a later assignment also, but some remarks are pertinent at this point. Ordinarily, the larger C_g and R_g are, the flatter is the low-frequency response of the stage. Hence, it would be thought that as large a value of C_g and R_g as possible should be employed. However, there are certain practical considerations that limit the size of R_g and C_g. These are:

1. A flat frequency response down to very low frequencies, particularly 1 or 2 c.p.s. is undesirable because a certain amount of coupling exists between stages via the impedance of the B supply, which is common to the several stages. As a result, motor-boating (low-frequency oscillation) can occur in a multi-stage amplifier.

2. A large coupling condenser C_g generally has appreciable leakage, which acts as a resistance R_s shunting C_g. Suppose this shunt resistance is 100 megohms, and R_g is 1 megohm. The two resistors will act as a voltage divider for d.c., and the voltage across R_g, instead of being zero, will be

$$E'_{dc} = E_{dc} \frac{R_g}{R_s + R_g} \qquad (7)$$

For the value of E_{dc} = +180 volts previously calculated, and for the values of 100 and 1 megohms,

$$E'_{dc} = +180 \frac{1}{100 + 1} = \frac{+180}{101}$$

$$= +1.78 \text{ volts}$$

This means that the grid, in the absence of any other bias, will be 1.78 volts *positive* with respect to its cathode and will draw grid current. If the bias source is made sufficiently negative this initial positive bias can be counteracted, but unfortunately R_s is generally quite variable, so that the positive bias developed tends to vary in an erratic manner with time. It is therefore important to choose as small a value of C_g as is feasible, and to use as high grade a condenser as is consistent with expense and bulk of the unit.

3. A large value of R_g is generally inadvisable. This is because of grid current considerations. The close proximity of the grid to the cathode tends to raise its temperature, and also to coat it with some of the cathode material, whereupon it may emit electrons. This represents a flow of electrons from ground *up through* R_g to the grid, thence across the interelectrode space to the plate and around through the B supply back to ground once more, thus completing the circuit. The flow of electrons from ground up through R_g to the grid makes the grid *positive* with respect to ground and the grounded cathode, and therefore develops a positive grid bias.

Another source of electron flow up through R_g, with the resultant production of a positive bias, is gas current. No tube is pumped to a complete vacuum, and the residual gas is ionized to some extent by the passage of electrons en route to the plate. The positive ions formed travel to the most negative electrode in the tube, namely, the grid, and extract electrons from it, thereby neutralizing themselves into normal gas atoms once more. The electrons extracted from the grid make it go relatively positive, whereupon new electrons travel up from ground through R_g to get to the grid, thus forming a gas current flow. If R_g is high in magnitude, these electrons cause an appreciable voltage drop in R_g such that the grid is appreciably positive to ground and hence to its cathode.

Thus, once again a positive bias is developed and the plate current increases, whereupon the ionization may increase, thereby making the grid more positive, and the cumulative effect, particularly in a larger power tube, may raise the plate current to dangerously high values and wreck the tube. For this reason R_g is limited by the manufacturer to a safe value, generally around 1 megohm, or even as low as 0.1 megohm in a large or high-transconductance tube. Where cathode bias is employed, larger values of R_g may be safely emplcyed, as will be explained subsequently.

METHODS OF OBTAINING GRID BIAS

D.C. BIAS SUPPLY.—An important feature of any amplifier system is the biasing of the control grid negative in order to prevent it from drawing current, at least in Class A operation. It is therefore important to study the various methods of obtaining grid bias in practical amplifiers.

Bias Battery.—This is shown in Fig. 6. The battery is connected in such polarity as to make

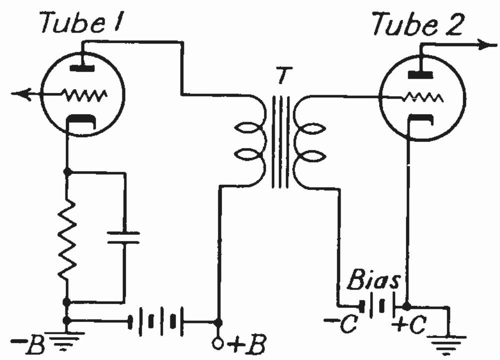


Fig. 6.—Use of C battery for obtaining grid bias.

the grid negative with respect to the cathode by an amount equal to the generated voltage of the battery. If the grid is thus made negative with respect to the cathode, and the signal voltage does not exceed the bias voltage, so that at no time is the grid driven positive with respect to the cathode, then the grid will not draw electrons (current) from the cathode, and therefore will not impose an electrical load on the signal source. This in turn means that there will not be a voltage drop in the signal source during *any* part of the signal cycle, and hence the full, undistorted signal voltage will appear between the grid and cathode.

At present, bias batteries are not used to any great extent because of other methods of obtaining bias. In special cases, as in video (television) amplifiers, it may be desirable to use this source of bias. For the purpose Mallory manufactures a bias cell that is very compact and produces about 1.25 volts. The cell is acorn shaped, has a diameter of 5/8", a height of 11/32", and fits into clips like a small fuse. It is not designed to furnish any current, but as explained above, the grid does not draw any current when it is negative with respect to the cathode, and hence the cell should last for years under such no-load operating conditions.

Another method is to employ a separate power supply for bias purposes. In television amplifiers it is generally of the regulated type and develops about 150 volts. Individual grids obtain their 3 or 4 volts bias from this supply through individual voltage dividers.

Filament Voltage Drop.--Another method of obtaining bias that was employed to a great extent in the early days when filament batteries were used is that of filament voltage drop. This is illustrated by Fig. 7. Resistor R is connected in series with the negative terminal of the filament battery and the filament. The top end of R will therefore be positive with respect to ground, and since the grid circuit is grounded, it will be negative with respect to the top end of R, and hence with respect to the filament.

The actual bias is approximately the voltage drop in R *plus half* of the filament voltage drop. For example, if the filament battery develops four volts, and the

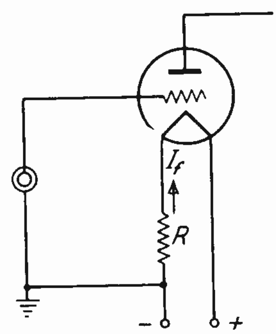


Fig. 7.--Use of filament dropping resistor for grid bias.

filament voltage is 2.0, then the drop in R will be $E_R = 4 - 2.0 = 2.0$ volts, and the bias will be $2.0 + (2.0/2) = 3$ volts. Suppose that the filament current I_f is .06 ampere. Then the value of R is simply

$$R = E_R/I_f = 2/.06 = 33.3 \text{ ohms}$$

While it is true that the plate current also flows through R, this current is normally so small compared to I_f that its effect can be neglected.

Cathode Resistor.--The two methods described above give an essentially fixed bias. Another method that is widely used today is that of utilizing the voltage drop produced in a cathode resistor by the plate current. This is illustrated by Fig. 8. In (A) an indirectly heated cathode is shown, and in (B) a filament type of cathode. In the latter case the cathode bias resistor R is connected between the center tap of the filament secondary and ground, as shown.

In either form of circuit the plate current of the tube flows through R and produces a voltage drop such that the cathode or filament is *positive* to ground. Since the source of a.c. grid signal is connected to ground, the grid is at ground potential *as far as d.c. is concerned;* i.e., in the absence of a.c. signal the grid is exactly at ground potential. It is therefore negative with *net* signal voltage is that *between the grid and cathode,* hence it is equal to the *difference* between the signal voltage of the source and the signal voltage developed across R.

This means that the net signal *voltage* is *less* than that produced by the signal source. For example, if 2 volts are generated by the signal source, and 1.5 volts are developed across the bias resistor R by the a.c. signal component of the plate current, then the effective or

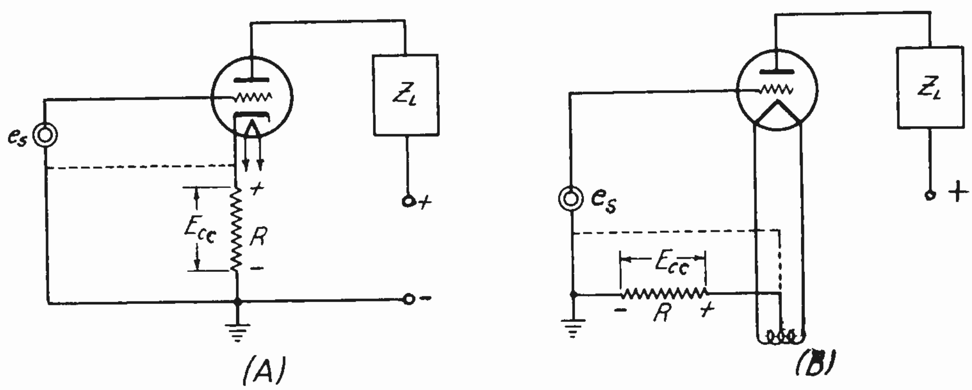


Fig. 8.—**Methods of obtaining self-bias by means of a cathode resistor.**

respect to the cathode or filament, and thus negative bias is obtained.

One difficulty encountered is that the plate current flowing through R has both a d.c. and a.c. signal component. The d.c. component furnishes the desired d.c. bias; the a.c. component introduces a signal voltage between cathode and ground *that is opposite in polarity to that produced between the grid and ground by the signal source.* The *effective* or net signal voltage applied to the tube is 2 - 1.5 or 0.5 volt. This represents an *inverse* or *degenerative feedback* effect, and reduces the stage gain α. This is clear from the fact that α is the ratio of output signal voltage (across the plate-load impedance) to the *applied* signal voltage, but since the output signal voltage is determined by the *net input signal,* it is lower than the value that would be obtained if no inverse feedback were present.

Considering further the above example, suppose that the stage gain in the absence of feedback is 15. Then the output voltage for a net input voltage of 0.5 is 15 x .5 = 7.5 volts. But the actual voltage of the source is 2 volts, even though the net voltage is only 0.5 volt owing to the feedback effect of the cathode bias resistor R. The gain of the stage in the presence of such feedback is therefore

$$\alpha = 7.5/2 = 3.75$$

whereas in the absence of feedback it is 15, or 4 times as great.

In some applications such inverse feedback is desirable, in spite of the large reduction in stage gain, because of certain beneficial effects, such as reduction in distortion of the output signal, stabilization of the gain in spite of B supply and tube variations, etc. Sometimes, in a video amplifier, a stage is required to give the necessary polarity to the signal, yet no particular voltage amplification is required. In such a case an unby-passed cathode bias resistor is desirable to hold the gain down to a reasonable value in spite of the need for the additional stage to give the correct polarity to the signal.

In other applications, however, such marked reduction in gain is undesirable, and means must be employed to obviate this effect. One means would be to connect the bottom end of the signal source to the cathode instead of to ground. This is indicated in Fig. 8 by the dotted line connections. Unfortunately this is possible only where the signal source is a generator like a microphone, phonograph pickup, transformer, or similar device, neither of whose terminals are inherently at ground potential.

Where, however, the signal source is a preceding vacuum tube or iconoscope directly coupled to this stage, such connections are not possible, because the ordinary vacuum tube is a source one of whose terminals is inherently at (a.c.) ground potential. This is apparent from the fact that the plate-load resistor, which is the source of signal for the following stage, has one end connected to B+, which is at a.c. ground potential. Hence, if feedback is to be eliminated in such a case, other means than reconnecting the signal source between the grid and cathode must be employed.

The usual method is to connect a large by-pass condenser across the cathode bias resistor R. This is chosen sufficiently large so that its reactance at the lowest signal frequency is negligibly small compared to the resistance of R, whereupon it substantially short-circuits R as far as the signal component is concerned. As a result, the signal component does not develop any appreciable voltage across the combination, so that there is no feedback voltage produced.

At the same time the condenser is essentially an open circuit or infinite impedance for d.c., so that the d.c. component of the plate current is constrained to flow through R alone. It therefore produces the desired d.c. voltage drop or bias across R, while the a.c.

component, as explained above, is unable to produce the undesired feedback voltage.

The size of the resistor R is determined by the magnitude of the d.c. component of the tube and the required bias voltage. For example, from the manufacturer's Tube Manual it is found that a 6J5 triode requires -8 volts grid bias for a plate potential of 250 volts, and that the plate current is 9 ma. The value of R is therefore

$$R = \frac{8}{.009} = 889$$

or approximately 900 ohms.

The size of the by-pass condenser to minimize feedback with resulting reduction in gain, can be found by the use of a formula which takes into account the permissible reduction in gain at the lowest frequency to be encountered, the permissible phase shift in the case of a video amplifier, the μ and R_p of the tube, and the value of R. However, at this point a simple, rough rule will be sufficient:

The reactance of the by-pass condenser C should be 0.1 R at the lowest signal frequency to be amplified. For example, if R = 900 ohms, as calculated above, and the lowest frequency to be amplified is 30 c.p.s., then

$$1/2\pi fC = 0.1\ R \qquad (8)$$

or

$$1/2\pi\ 30\ C = (0.1)\ (900) = 90$$

from which

$$C = \frac{1}{2\pi\ 30 \times 90} = .000059 \text{ farad}$$

$$= 59\ \mu f$$

a condenser of this size will undoubtedly be of the electrolytic type, and while it is of large capacity, its voltage rating can be very low, since there will be only 8 volts d.c. drop across it, (see Fig. 9). It is customary

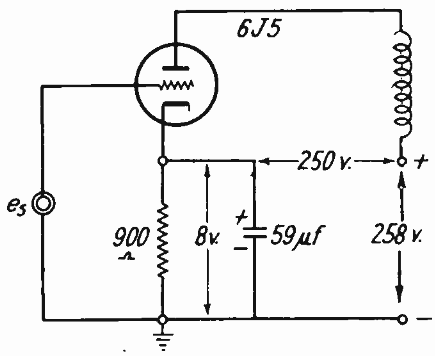


Fig. 9.—Use of cathode by-pass condenser to minimize feedback.

to employ a condenser of 25-volt rating for this purpose, and where bias voltages exceeding 25 volts are required, as in the case of many power tubes, condensers of 50-volt rating are employed.

In passing it is to be noted that the power supply voltage must equal the plate voltage plus the bias voltage. In the above problem, the plate voltage, which is the voltage *between plate and cathode*, is 250. Since the cathode, however, is 8 volts positive to ground, as shown in Fig. 9, the total B supply voltage must be 250 + 8 = 258 volts, as shown.

In short, the bias voltage is not obtained free; provision for it must be made in the design of the B supply. While 8 volts additional in the above example is not of any great consequence, in the case of power tubes the additional bias voltage may be appreciable. For example, a pair of 2A3 tubes in push-pull require 300 volts on their plates, and 60 volts bias, so that the power supply must develop 360 rather than 300 volts, or a 20 per cent increase. Nevertheless, cathode resistors are one of the most convenient sources of grid bias.

One advantage of cathode bias, when the cathode resistor is by-passed, is that while degenerative feedback of the signal component is obviated, such feedback of the d.c. component still exists. Thus, if the d.c. component of the plate current tends to increase for any reason, more counteracting bias is obtained. In such a case the resistance of the grid circuit can be safely increased from the 1-megohm value previously given, because any tendency of the grid gas current to increase the plate current will be counteracted by the increase in cathode bias thereby produced. The manufacturer's specifications for any particular tube should be consulted where a high value of grid circuit resistance is required. The use of an extremely high-grid resistor may occasionally be encountered in some special industrial electronic application.

Bias Obtained From Power Supply Resistor.— Another method of obtaining bias for several tubes is to employ a resistor between ground and the negative terminal of the power supply. The arrangement is shown in Fig. 10. Two transformer-coupled stages are shown for convenience. The plate current for the first stage is i_1; that for the second stage, i_2, and the direction of flow shown is that for electron motion. Thus, a total electron flow of $i_1 + i_2$ flows into the B+ or positive

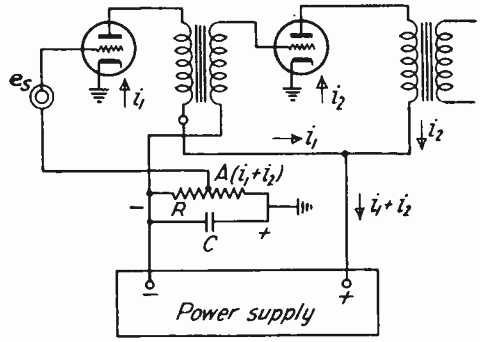


Fig. 10.— Method of obtaining grid bias by voltage drop in resistor between ground and B-.

terminal of the B supply, and emerges from the B- terminal. It thereupon flows through a common bias resistor R, adequately by-passed by a condenser C, and then splits into its component values i_2 and i_1 as shown to flow to the respective cathodes of the tubes. The total current $i_1 + i_2$, in flowing through R, makes its grounded end positive with respect to the B- end (since electrons flow from minus to plus in the external circuit, such as R).

There is thus available in R a source of negative potential with respect to the *grounded* cathodes of the tubes. Hence, the grid of the second tube can be returned to the B- side of R, whereupon it is at a maximum negative potential with respect to its cathode, and if less bias is required for the first stage, its grid may be returned to a tap A on R, which is less negative with respect to ground.

If other stages are connected to the power supply, their plate currents will flow through R, thus adding to the voltage drop in it. Thus, suppose a total of 100 ma. flows through R, and the maximum amount of bias required is 25 volts. Then R must have a value of

$$R = \frac{25}{.1} = 250 \text{ ohms}$$

The by-pass condenser C must then be chosen as in the previous case; i.e., its reactance at the lowest frequency to be amplified should be about 0.1 R, in order that no appreciable signal voltage appears across R.

This indicates one disadvantage of the system. Where many stages are employed, connecting the various grids to a common bias resistor invites the danger of feedback. In particular, the plate current of the last stage has the greatest (amplified) signal component, and the grid of the first stage is most susceptible to the signal voltage set up across R by the above component of current. Depending upon the number of stages, the signal IR drop produced across R by the a.c. component of the last stage may act on the first grid so as to increase the same plate current component, whereupon regeneration and possible oscillation occurs; or, the voltage may be of such phase as to produce the opposite effect on the plate current producing it, whereupon degenerative feedback may occur. In a multi-stage amplifier regenerative feedback and oscillation may occur between any two stages—not necessarily the first and last stages—and it is always desirable to eliminate such an effect unless it is desired and deliberately introduced *in controlled amounts.*

The condenser C tends to reduce such a feedback voltage, but in a high-gain amplifier exceedingly minute amounts of feedback can produce undesired effects, such as oscillation, because of the high forward gain of the amplifying system. One way to cut down such

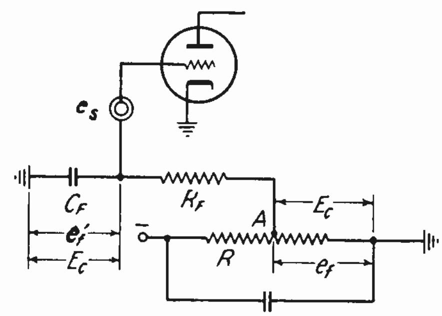


Fig. 11.—Use of R-C filter to decouple a grid from the bias resistor R.

feedback effects is to use simple R-C filter circuits between each grid and the resistor R. Consider tap A in Fig. 10. Suppose that instead of the grid of the first stage connecting directly to A, it connects through an R-C filter as shown in Fig. 11 by R_F and C_F.

The feedback voltage that acts upon the grid in question is between ground and tap A and is denoted by e_f. This acts upon R_F and C_F in series. The two form a special kind of voltage divider: if C_F and R_F are sufficiently large, most of e_f is consumed across R_F and very little is available across the low reactance of C_F. Thus but a small fraction of e_f is fed to the grid as e_f'.

On the other hand, with regard to the d.c. bias voltage, C_F has infinite reactance, so that no d.c. flows through R_F and C_F in series, there is no d.c. voltage drop in R_F, and hence all of the d.c. bias voltage appears across C_F. This is the voltage between the grid and cathode of the tube, as is clear from an inspection of Fig. 11. Hence, the final result is that R_F C_F divides down a.c. voltages such as the signal voltage, e_r, to a negligibly small value e_r', particularly at the higher frequencies, whereas it applies the full desired d.c. bias voltage to the grid without any reduction.

This type of filter is very effective, and is employed in many other places in an amplifier circuit. Representative values are C_F = 0.1 μf and R_F = 100,000 ohms. Each grid may be connected to an appropriate tap of R through an individual R-C filter, and in addition, a by-pass condenser across R is usually employed to reduce the initial magnitude of the feedback voltage.

GRAPHICAL CONSTRUCTIONS

It will be shown in a later assignment that the frequency response of an amplifier can be expressed in terms of the gain at some mid-frequency point, such as 1,000 c.p.s. for an audio amplifier. However, the determination of the gain at the mid-frequency point by means of a formula may be inaccurate owing to variations in the tube parameters, such as μ, R_p, and G_m, from the values given by the manufacturer for a certain power supply and bias voltage.

It is therefore preferable to determine the mid-frequency gain by some more accurate method, and in addition—if possible—to calculate the distortion characteristics of the tube and the maximum signal that it can handle. This can be done graphically, as will be explained in the following section.

THE LOAD LINE.—In a previous technical assignment the tube characteristics were exhibited in the form of a family of curves; in the case of the plate family, the plate current i_p was plotted against plate voltage e_p for different values of grid voltage. Each e_p - i_p curve was for a fixed value of grid voltage; the latter is known as the *parameter* of the family of curves.

In Fig. 12 is shown the characteristic curves for a 6J5 triode. Suppose it is connected in series with a 50,000-ohm plate resistor R_L, and a 250-volt supply as indicated in Fig. 13. The problem is to determine the proper bias, maximum signal swing e_s, maximum voltage output e_L, and the gain in the mid-frequency portion of the spectrum, by means of a graphical construction.

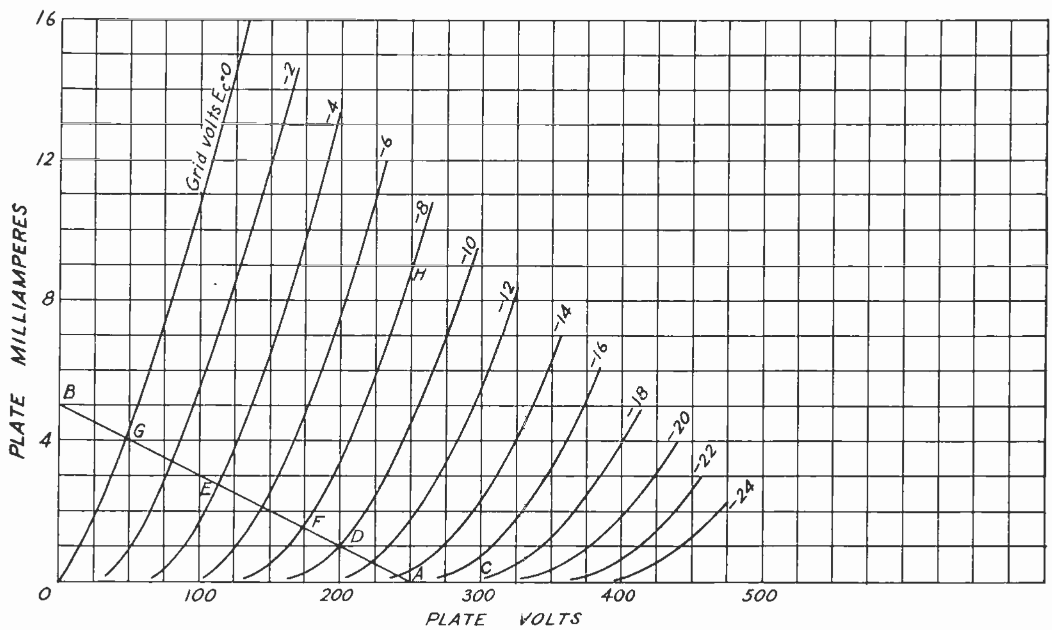
First some remarks must be made concerning the action of the circuit. Suppose the grid were biased so negative that no plate current flowed. Then there would be no voltage drop in R_L, and all of the applied 250 volts would be developed between the plate (A in Fig. 13) and ground.

On the other hand, suppose it were possible to drive the grid so positive that no plate voltage were required to draw plate current through the tube. (In an actual tube a very positive grid would tend to divert electrons from the plate, thus decreasing the plate current.) If, however, it were possible to arrange to pass plate current through the tube without any voltage drop, then the current in the plate circuit would still be limited by R_L, across which all of the 250 volts supplied would appear. The plate current would be

$$i_{b\ max} = \frac{250}{50000} = 5\ ma$$

Thus, the maximum possible variation in plate current is from 0 to 5 ma.

Refer now to Fig. 12. If the tube is biased to cutoff, the plate voltage will be 250 volts,



(Courtesy RCA)

Fig. 12.—Plate family of characteristic curves for a 6J5 triode tube.

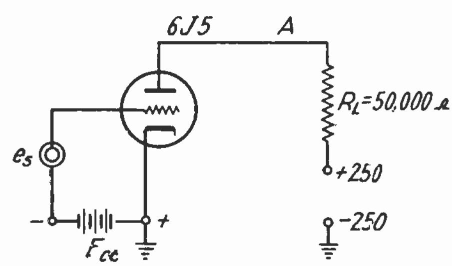


Fig. 13.—Resistance-coupled stage employing a 6J5 tube.

and the plate current equal to zero. This locates point A on the figure. It is also to be observed that the grid curve passing through A appears to be -16 volts when prolonged. This means that if the tube is biased to -16 volts, the tube will be at cutoff, whereupon $i_b = 0$ and $e_b = 250$ volts.

On the other hand, for 5 ma. through the tube, $e_b = 0$ volts. Hence go up along the current axis to 5 ma; this is point B in the figure. While now the plate voltage e_b is clearly zero, and the plate current is 5 ma., it is to be noted at the same time that the voltage across R_L is now the full 250 volts. This can be interpreted from the figure as follows: all voltage measurements to the *right* from the origin represent voltages between the cathode and plate of the tube, or plate voltage as measured at the tube socket. All voltages measured from the supply voltage point, such as point A in Fig. 12 (to the left), represent voltages across R_L.

To check these statements, take points A and B themselves. Thus, for point A, the plate voltage is OA = 250 volts, and the voltage across R_L must therefore be zero. On the other hand, for point B, the plate voltage measured from 0 is clearly zero, while the voltage across R_L is from A to B, or 250 volts. This is apparent from physical considerations in that the total voltage across both the tube and R_L must equal E_{bb}, the supply voltage. If the tube requires a certain amount of this voltage for its operation, then the difference between this required amount and the total supplied must be what is left for R_L. This fact will be further employed in the discussion that follows.

Join A and B as shown. Line AB is called the *load line* or *terminal characteristic* for the resistor R_L. Its superposition on the tube characteristics (or tube terminal characteristic) enables the current flow through the two in series to be determined for any value of plate and grid voltages. In this problem the supply voltage has been set at 250 volts, hence the load line for R_L starts at point A. If a supply voltage of 300 had been chosen instead, the load line would have started at the 300-volt point (C), and similarly for any other supply voltage.

For the 250-volt value chosen, consider the conditions *when the grid voltage is fixed at -10 volts.* The plate voltage and plate current must be so related that they represent one or other points on this tube curve labelled -10 volts. At the same time the current must be such that it lies also on the *load-line for R_L*. Therefore, the *intersection* of AB with the -10 volt curve, or point D in the figure, locates the values of plate voltage, load-resistance voltage, and current flowing through the two in series. From Fig. 12, the current is found to be 1 ma.; the plate voltage (measured from the origin 0 to the right is 200 volts), and the voltage across R_L, (measured from A to the left), is 250 - 200 = 50 volts. Clearly the sum of the two voltages is 200 + 50 = 250 volts, that of the supply.

As another example, suppose the grid voltage changes to -4 volts. The intersection of the -4 volt curve with the load line is point E in Fig. 12. The plate current is now higher, namely, 2.7 ma.; the plate voltage is 115 volts; and the load-resistance voltage is 250 - 115 = 135 volts. Thus, as the grid is made less negative, the plate current increases, the plate voltage decreases (the conductivity of the tube is in effect increased), and the load voltage increases.

If the grid voltage is made more negative than the cutoff value of -16 volts, the plate current remains zero, the plate voltage remains at 250 volts, and the load voltage at 0 volts. This is because the plate current cannot be less than zero; i.e., it cannot be negative and hence a reverse current, because the tube conducts in one direction only. As a consequence, no change in output voltage (across R_L) will be obtained for grid voltages more negative than the cutoff value, or the output voltage will fail to follow in wave-shape the grid-voltage wave-shape beyond the cutoff value. This can be expressed alternatively by saying that the plate current must be allowed to flow at all moments during the grid voltage cycle. Such operation is known as *Class A*, and the following considerations relate to this mode of operation.

QUIESCENT POINT.--It is therefore clear that for the conditions illustrated in Fig. 12 the grid voltage cannot vary by more than 16 volts: from 0 volts to -16 volts, if the output voltage is to be a faithful copy of it. Suppose the grid signal voltage e_s is a sine wave. Then 16 volts is the peak-to-peak value, as is clear from Fig. 14. A d.c. bias of -8 volts is required so that on the positive excursion of the grid signal voltage, the grid will just come up to 0 volts, and not actually go positive, whereas on the negative excursion the grid will

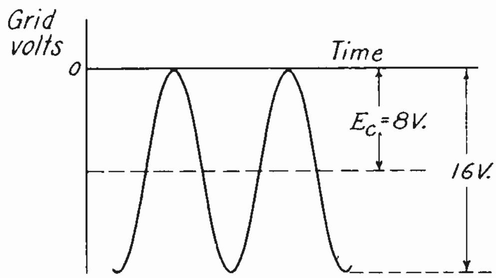


Fig. 14.—Relation of bias and signal voltage for 6J5 tube.

not become more than 16 volts negative (will not exceed cutoff).

The reason why the grid should not go positive is that it thereupon draws grid current, and thus acts like a low-resistance load on the previous tube. If the latter is a high-resistance source, then the sudden dropping of the

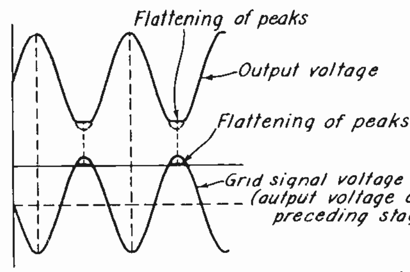
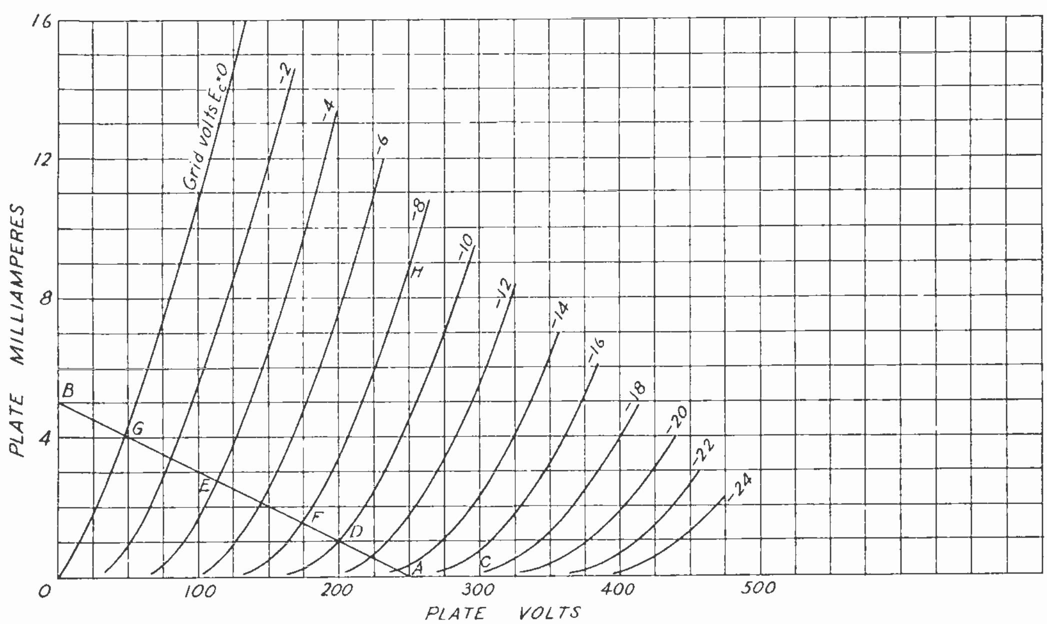


Fig. 15.--Distortion produced both in grid signal voltage and in output voltage of a tube by driving the grid positive.

load impedance, represented by the sudden conductivity of the grid at the peak of the cycle, will tend to flatten the output voltage wave of the preceding tube at its peak, as illustrated in Fig. 15. In short, the input signal voltage to the tube under consideration will be distorted, and in turn will produce a similar distortion in its output wave.

The reason why the grid should not be driven beyond cutoff has been discussed previously. Hence, a bias of -8 volts is required. With no signal input, the grid voltage is -8 volts, and the plate current and other values can be found from Fig. 12 just as described previously. Thus, where the -8 volt curve intersects the load line at F is the so-called "quiescent" point, i.e., the point determining tube current, etc., when no signal is impressed and the plate current is steady or d.c. in nature. For point F it is clear that the plate current is 1.5 ma. d.c.; the plate voltage is 173 volts d.c., and the d.c. voltage drop across R_L is 250 - 173 = 77 volts. If a d.c. voltmeter were connected between the plate and cathode socket terminals, a reading of 173 volts would be had; if across R_L, a reading of 77 volts would be noted.



(Courtesy RCA)

Fig. 12.--Plate family of characteristic curves for a 6J5 triode tube.

DETERMINATION OF STAGE GAIN.—If now a sine-wave signal voltage of 8 volts peak is introduced in the grid circuit, the plate current, plate voltage, and load voltage will all vary in essentially a sinusoidal manner. At the positive peak of the signal voltage, the actual grid voltage is 8 volts peak plus -8 volts bias, or zero volts, as indicated previously in Fig. 14. Thus the 0-volt curve is used, and where it intersects the load line at G, is the point giving the peak current, minimum plate voltage, and maximum load voltage. These are, respectively, 4 ma., 47 volts, and 250 - 47 = 203 volts.

On the negative peak excursion, point A is reached, for which the values are, respectively, 0 ma., 250 volts, and 0 volts. Note that the so-called "path of operation" is along the load line. The relation between grid voltage, plate current, plate voltage, and load voltage is illustrated in Fig. 16. The plate current varies from 1.5 ma. (quiescent value) up to 4 ma.

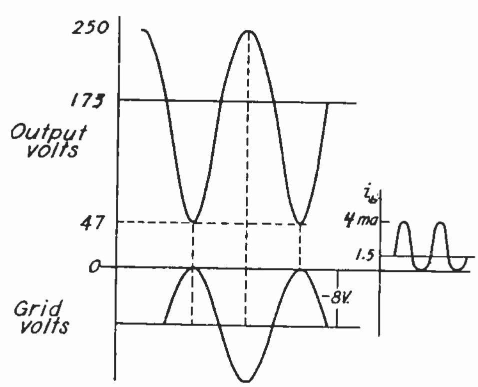


Fig. 16.—Plate, load, and grid voltage relations, as well as plate current variations.

and down to 0 ma. The plate voltage varies from a quiescent value of 173 volts to a minumum of 47 volts and a maximum of 250 volts; the load voltage varies simultaneously from a quiescent value of 250 - 173 = 77 volts to a maximum of 173 - 47 = 126 volts. The *peak to peak value* is clearly $e_L = 77. + 126 = 203$ volts.

This can be found very simply by subtracting the minimum value of plate voltage, 47 volts, from the maximum value, 250 volts, or

$$E_L = 250 - 47 = 203 \text{ volts}$$

In a similar manner the peak-to-peak value of the plate current variation can be found, $i_{b\,max} - i_{b\,min} = 4 - 0 = 4$ ma.

Inspection of Fig. 16 shows, however, that the positive peak of e_L is but 77 volts, and the negative peak is 126 volts, or the wave is far from being sinusoidal in shape and hence a faithful copy of the input grid signal voltage. This is a result of the distortion produced in the tube, and will be analyzed in greater detail farther on. For the moment assume that the wave is undistorted, and take its peak-to-peak value, namely, 203 volts. The peak-to-peak value of the input signal voltage, from Fig. 14, is seen to be 16 volts. Hence, the stage gain can be calculated simply as

$$\alpha = 203/16 = 12.7$$

Thus, the graphical method can be employed to calculate the stage gain under actual operating conditions, wherein the actual G_m, R_p, and μ of the tube are automatically employed rather than values taken from a tube manual that are correct for one set of operating conditions only.

As just stated, the output voltage wave e_L, is asymmetrical in that the positive peak is less than the negative peak, as measured from the 173-volt average or quiescent value. An examination of Fig. 12 shows why this is so: the more negative grid curves draw together more rapidly near the plate-voltage axis, thus foreshortening the positive peaks of e_L. This is a common occurrence in vacuum tube characteristics, and indicates that if excessive distortion is to be obviated, the region near the plate-voltage axis must be avoided.

In the case of the 6J5 tube, this means that plate-current values below about 1.5 ma., where the characteristic curves bend over sharply, had better not be used. This in turn means that instead of using the cutoff -16 volt grid curve as the maximum negative grid swing, a value of -8 volts had better be employed, since this intersects the load line at F where i_b = 1.5 ma. The peak-to-peak grid swing will therefore be 0 - (-8) = 8 volts, and the bias will be half of this, or -4 volts. Hence, the peak grid swing will now be 4 volts, or half of its previous value.

This locates the quiescent point as E in Fig. 12. The quiescent plate voltage is 115 volts, the corresponding drop across R_L is 250 - 115 = 135 volts, and the quiescent plate current is 2.7 ma. The maximum plate voltage occurs where the grid swings to -8 volts (point F), and is equal to 173 volts. The corresponding minimum plate current is 1.5 ma., as stated previously. The minimum plate voltage and corresponding maximum plate current are given, as before, by point G and are 47 volts and 4 ma., respectively.

The peak-to-peak output voltage is now

$$e_L = e_{b\,max} - e_{b\,min} = 173 - 47$$
$$= 126 \text{ volts}$$

The negative peak is

$$E_{bo} - e_{b\,min} = 115 - 47 = 68 \text{ volts}$$

The positive peak is

$$e_{b\,max} - E_{bo} = 173 - 115 = 58 \text{ volts}$$

Note that the negative peak but slightly exceeds the positive peak, or the distortion, while still in the same direction, is small.

The stage gain under such reduced bias and grid swing is

$$\alpha = \frac{126}{8} = 15.8$$

or appreciably higher than the previously calculated value of 12.5. This is because the flattening of the positive peaks of the output voltage (distortion) has been reduced; such flattening represented a limiting effect in the tube owing to a reduction in its G_m at high negative grid swings.

It will be of interest to compare this value with that calculated from the values given by the manufacturer. These are R_p = 7,700 ohms, μ = 20. If these be substituted in Eq. (7), there is obtained

$$\alpha = 20\,\frac{50000}{7700 + 50000} = 17.34$$

This is clearly too high. The reason is not hard to find. The tube manual states that the grid bias is -8 volts, and the plate current is 9 ma., at 250 volts supply. These values are consistent with one another; they represent point H on the tube curves given in Fig. 12. Point H, however, is clearly off the 50,000-ohm load line, and represents a mode of operation in which there is no voltage drop in the plate load impedance to the d.c. component of the plate current. This will be taken up under the discussion of power amplifiers.

Hence, the values given in the tube manual do not often refer to the type of operation contemplated in some particular application—here, resistance coupling. On the other hand, the data given in the Resistance Coupled Amplifier Chart of the RCA Tube Handbook checks the value found here quite closely. Thus, for a 6J5 tube, having a plate supply of 300 volts; a load resistance of 50,000 ohms; a grid resistance of 250,000 ohms (not considered here in the graphical construction), the output voltage is 60 volts peak, and the gain is 14.

The average peak output voltage found graphically was (58 + 68)/2 = 63 volts, and the gain was 15.8. Since the construction did not take into account the shunting effect of the grid resistor R_g on the plate load resistor R_L, the agreement is satisfactory. The shunting effect of R_g is not serious, however, if R_g is several times R_L. Thus, for the values cited above, R_g and R_L in parallel represents the following:

$$R'_L = \frac{(50000)\ (250000)}{50000 + 250000} = 42{,}000 \text{ ohms}$$

This is the impedance presented to the a.c. component of the plate current, and is not very much less than the 50,000 ohms resistance presented to the d.c. component.

Correction, however, can be very simple for the effect of R_g. First note from Fig. 17 that the coupling condenser C_g blocks the d.c. component of the plate current from flowing through R_g,

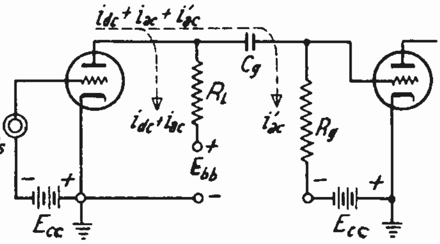
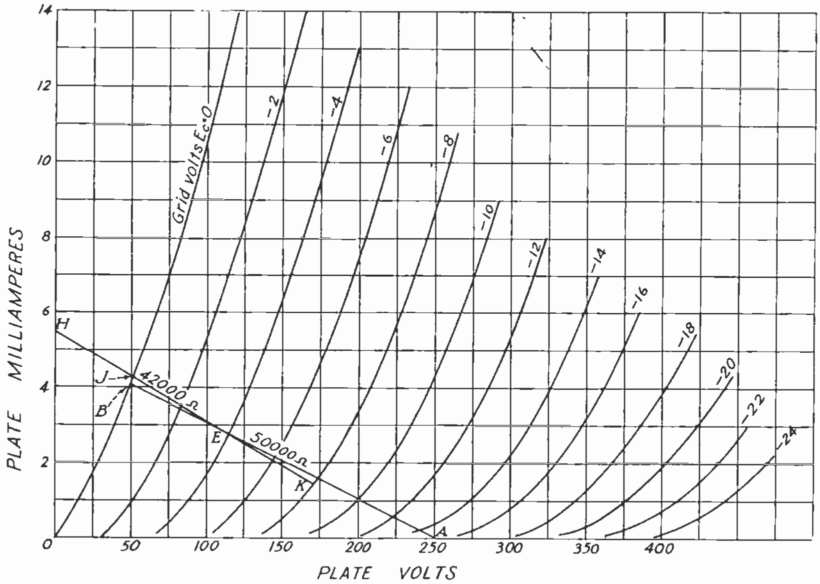


Fig. 17.—Path for d.c. and a.c. components of plate current.

whereas the a.c. component can readily flow through both R_L and R_g in parallel, particularly at frequencies in the normal audio range (mid-frequency range) where the reactance of C_g is negligibly small compared to R_g.

This means that the load line for the d.c. component must correspond to R_L (50,000 ohms in the above example), whereas the load line for the a.c. component must correspond to R_L and R_g in parallel, or R'_L, equal to 42,000 ohms. To obtain this effect graphically, proceed as in Fig. 18. Here AB is the 50,000-ohm load line for R_L, just as in Fig. 12. It had been decided to employ a 4-volt grid bias so as to minimize the distortion in the output voltage as far as that was compatible with adequate voltage output. This locates point E in the figure. E and prolong to K, or as far as desired.

The path of operation is now from E (-4 volts bias) to J (0 volts bias) back to E, down to K (-8 volts bias), and back to E once more, for one signal voltage cycle. The output voltage is (peak-to-peak)



(Courtesy RCA)

Fig. 18.— Graphical construction for case where a.c. load impedance is lower than d.c. load impedance.

Through E draw a load line HK representing 42,000 ohms or R_L'. This is done very simply as follows: point E represents a (d.c.) plate current of 2.7 ma. and a (d.c.) plate potential of 115 volts. If the voltage across R_L' changed by 115 volts, the current through R_L' would change by

$$\frac{115}{42000} = 2.74 \text{ ma.}$$

Add this 2.74 ma. to the d.c. component of 2.7 ma. corresponding to point E, and obtain 5.44 ma. Lay off this distance along the current axis, obtaining point H. Join H to

$$e_{b\,max} - e_{b\,min} = 171 - 50$$

$$= 121 \text{ volts}$$

The negative peak is

$$E_{bo} - e_{b\,min} = 115 - 50 = 65 \text{ volts}$$

and the positive peak is

$$e_{b\,max} - E_{bo} = 171 - 115$$

$$= 56 \text{ volts}$$

The gain is $121 \div 8 = 15.1$ which checks the above-mentioned chart fairly well.

The discrepancy between the two peaks is not much different from that found previously (68 and 58 volts). However, if R'_L is much lower than R_L, then it will in general be found—at least for triodes—that the discrepancy between the positive and negative peaks, and hence the distortion, is increased. Hence, if possible, R_g should be much greater than R_L, so that R'_L will be as nearly equal to R_L as possible, and the a.c. and d.c. load lines will differ from one another to a minumum degree.

For highest gain and minimum distortion in the case of a triode, R_L should be as many times greater than R_p as possible, and yet as small compared to R_g as possible, in order that R_L and R_g in parallel or R'_L, is not much less than R_L itself. Where R_p equals about 10,000 ohms and R_g can be made as high as 1 megohm, R_L can be about $5R_p$ or 50,000 ohms and yet be but a small fraction of R_g, so that R'_L is not much less than R_L. In such a case it makes little difference whether R_L or R_L and R'_L are used for the graphical construction.

There is another factor, however, indicated in Fig. 18, and that is that the greater slope of the a.c. load line for R'_L tends to reach cutoff sooner than the d.c. load line for R_L, thus limiting the magnitude of the grid swing and hence signal voltage to a greater extent. This may be particularly pronounced in the case of a pentode tube, for which R_L is high and comparable to R_g. For example, suppose R_L = 100,000 ohms and R_g = 100,000 ohms. The a.c. resistance will then be obtained as follows:

$$R'_L = 100{,}000/2 = 50{,}000 \text{ ohms}$$

or half of R_L. The construction will thereupon yield results as indicated in Fig. 19. If the d.c. load line is used as the path of operation for the a.c. component, then a peak-to-peak grid swing of from 0 to about −8 volts, or 8 volts, is permissible before the sharp lower bend in the tube characteristics is reached.

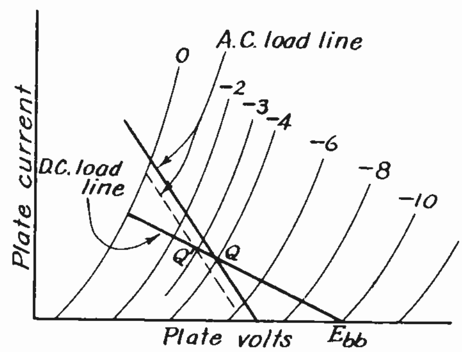


Fig. 19.—Case where R'_L is half of R_L, showing limitation in grid swing.

On the other hand, if the a.c. load line is employed (which is the correct procedure), then the peak-to-peak grid swing is but from 0 to −6 volts, or 6 volts before the lower bend in the tube characteristics is encountered. Actually, for sine-wave or other symmetrical-wave excitation the grid swing would be from −4 to −6 volts or 2 volts negative peak, and correspondingly the positive swing would have to be from −4 to −2 volts, or a total of but −6 − (−2) = 4 volts peak to peak.

Some increase in grid swing could be obtained by reducing the bias somewhat, say to -3 volts, as indicated by Q' and the dotted line for R_L', but the increase in output level would be small. The considerations presented above are of importance where a small tube is required to furnish not only appreciable voltage gain, but also appreciable output voltage. This means that not only is the *ratio* of output voltage to input signal voltage required to be high, but the *actual magnitude or level* of the output voltage, and hence of the input signal voltage, is required to be large. This occurs in the case of the voltage amplifier stage feeding the power amplifier stage of an amplifier, since the latter requires considerable grid swing to furnish the desired power output.

SECOND HARMONIC DISTORTION.— In the previous example, it was found that the positive and negative peak values of the output voltage, as measured from the quiescent value, were unequal. For example, for the 50,000-ohm load line and -4 volts bias, the positive peak was 68 volts, and the negative peak was 58 volts. This represents distortion in that the wave departs from a sinusoid in shape, and hence is different from the input sine-wave signal voltage. In particular, as measured from the quiescent value of 115 volts, the plate voltage dips too much compared to the amount by which it rises above the 115-volt value.

Note that the dip in plate voltage is due to the grid swinging in the *positive* direction and *increasing* the plate current, thus increasing the voltage drop in R_L and leaving less of the supply voltage for the plate. On the other hand, when the grid swing is in the negative direction, the plate current decreases, the drop in R_L is decreased, and more of the 250-volt supply is available at the plate. Hence, the plate voltage variations are 180° out of phase with the grid signal voltage, as has already been indicated in Figs. 15 and 16. This consideration of reversal of phase, as well as amplification of the input voltage by a vacuum-tube stage, is of importance in feedback connections and oscilloscope and television amplifiers, since in the latter case, for example, the polarity of the signal output to the picture tube determines whether a positive or negative image will be obtained.

Returning once more to the question of distortion, the reason for the disparity in the two peak amplitudes is that the plate current rise from the quiescent value is greater (for a positive grid swing) than the plate current fall (for a negative grid swing). This is chargeable to the tube; the latter is the cause for the distortion.

If the distortion is not excessive, and is of the asymmetrical type, then it can be regarded as essentially second-harmonic in nature. This will be made clearer by reference to Fig. 20. In (A) is shown a sine wave of the same frequency as the input signal wave. If (A) were the output obtained, then the stage would exhibit distortionless amplification.

Now suppose a second-harmonic (double-frequency) wave, as shown in (B), is added to the fundamental wave of (A). The object is to obtain a final wave shape that is essentially identical with the output wave actually obtained. To add the two waves, add instantaneous values or ordinates. Thus, at the start, wave (A) has zero amplitude (point A);

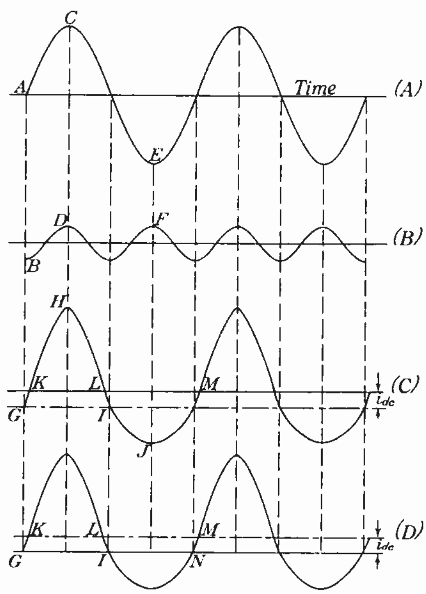


Fig. 20.— Analysis of a distorted wave showing fundamental, second-harmonic, and additional d.c. components.

whereas wave (B) has the negative amplitude (point B). The sum is simply the negative instantaneous value of the second-harmonic, wave (B), and is shown as point G in the resultant wave (C). One-quarter of a fundamental cycle later, wave (A) has the amplitude C, wave (B) has the amplitude D, and their sum yields point H of wave (C). At the end of a half fundamental cycle, wave (C) has the same negative amplitude I as it had at the beginning (point G).

At the end of three-quarters of a cycle fundamental wave (A) has peak negative amplitude E, but second-harmonic wave (B) has positive peak amplitude F once more, so that the *algebraic sum* is less than amplitude E, and is denoted by point J in wave (C).

In this manner the resultant wave (C) can be plotted for all instants of time, and has the shape shown. The positive half-cycles have a peaked form; the negative half-cycles have a flattened form, and the shape therefore appears to resemble that of the plate current of a vacuum tube—at least a triode—very closely.

There is, however, one discrepancy, as may be noted from an inspection of wave (C). For a sinusoidal grid signal wave, the time for each half cycle is the same. On the other hand, note that wave (C) has a positive excursion during time KL, and a negative excursion during time LM, and clearly LM is greater than KL. To make the wave fit the observed shape, it is necessary to lift it above the axis by an amount *equal to the peak of the second-harmonic com-*

ponent. This amount is denoted in (C) by i_{dc}.

When this is done, as shown in (D), then the crossing of the axis occurs in equal time intervals; i.e., GI = IN, etc., and wave (D) very closely resembles the actual output of an amplifier stage. However, wave (D) contains not only a fundamental and a second-harmonic component, *but additional d.c. as well*, over and above the normal d.c. plate current that flows even when no signal voltage is impressed. Indeed, a test for second harmonic distortion is to insert a d.c. milliammeter in series with the plate circuit, and observe the normal d.c. component that flows in the absence of a signal voltage. Then a signal voltage is impressed, preferably of maximum permissible amplitude. It will be found that the plate current reads somewhat higher; the increase is the additional d.c. component mentioned above, *and its magnitude indicates the magnitude of the peak second-harmonic component.* Since the output voltage across the plate load resistor R_L is directly proportional to the plate current, its wave-shape will be similarly distorted, although its phase (with respect to ground) will be opposite to that of the plate current and grid voltage, as mentioned previously.

It is possible to compute the percentage of second harmonic distortion from the graphical construction described previously. Referring first to Fig. 20, note that the second-harmonic current lifts up both the positive and negative peaks of the fundamental by an amount equal to its peak amplitude. The additional d.c. component acts similarly and to equal degree. This means that the distance between the two peaks of waves (C) or (D) still remains equal to the peak-to-peak value of the fundamental. In short,

$$i_{b\,max} \text{ to } i_{b\,min} = 2I_{pm} \text{ (peak)} \qquad (9)$$

On the other hand, the average value of wave (D) is raised above the axis by an amount equal to twice the additional d.c. component or the peak second-harmonic wave. Hence, if the initial d.c. is subtracted from the average value, the difference represents twice the additional d.c. component or twice the peak second-harmonic amplitude. In mathematical terms

$$i_{dc} = I_{2pm} \text{ (2nd harmonic)}$$

$$= \frac{1}{2}\left[\left(\frac{i_{b\,max} + i_{b\,min}}{2}\right) - i_{bo}\right] \qquad (10)$$

The percentage second-harmonic distortion is clearly

$$\frac{I_{2pm}}{I_{pm}} \times 100$$

$$= \frac{\frac{1}{2}\left[\left(\frac{i_{b\,max} + i_{b\,min}}{2}\right) - i_{bo}\right]}{\frac{i_{b\,max} - i_{b\,min}}{2}} \times 100$$

$$= \frac{(i_{b\,max} + i_{b\,min}) - 2I_{bo}}{2\,(i_{b\,max} - i_{b\,min})} \times 100 \qquad (11)$$

This formula is quite useful as a measure of the distortion in the ordinary triode tube. Consider the 6J5 tube used for the graphical construction. Referring back to Fig. 18, the quiescent plate current was 2.7 ma., and for the 50,000-ohm load line, $i_{b\,max} = 4$ ma., and $i_{b\,min} = 1.5$ ma. Substitution in Eq. (11) gives

(Percentage 2nd harmonic distortion)

$$= \frac{4.0 + 1.5 - 2 \times 2.7}{2(4.0 - 1.5)} \times 100$$

$$= \frac{5.5 - 5.4}{5} \times 100$$

$$= \frac{0.1}{5} \times 100 = 2 \text{ per cent}$$

Whether this percentage distortion is excessive depends upon several factors:

1. The band width; i.e., whether the system is one of average or of high fidelity. For telephone conversation, where merely intelligibility of speech rather than naturalness of reproduction is desired, an audio band width of from about 250 to 3,500 c.p.s. is sufficient. In that case possibly as high as 7 to 8% distortion is permissible although not desirable. For the average radio or phonograph reproduction a band width of at most 100 to 5,000 c.p.s. is employed, in which case the distortion should not exceed possibly 5 per cent. Less distortion is permissible for wider frequency bands.

2. The quality of the system. Usually this is tied up with the band width; a high-quality or high-fidelity system should have a band width of from 50 to 10,000 c.p.s. or wider and should also have a minimum of distortion, background noise, etc. A possible limit to the distortion is 2%

3. The nature of the service. Since a few hundred broadcast stations or phonograph recording studios furnish the signal for millions of radios or phonographs, it is clear from economic considerations that for a given maximum permissible *overall* distortion, more of it should be allotted to the large number of receivers or phonographs, and less of it to the broadcast stations or recording studios. For example, if 6% overall distortion is not excessive, then possibly 5% should be permitted in each receiver, making it less expensive to build, and only 1% in the broadcast system.*

The value of 2% calculated above may therefore seem reasonable, but is actually rather high. Usually most of the distortion occurs in the last or power output stage, and the preceding voltage amplifier stages are generally free of appreciable distortion. The high value obtained here is owing to the large grid swing employed and the correspondingly large output voltage obtained, namely around 60 volts. If the power tube does not require so large a voltage as this for its grid swing, then the input to this stage can be decreased, say, to 3 volts peak, whereupon both the voltage output and particularly the distortion will be less.

If the grid swing on the 6J5 is only 3 volts peak, then $i_{b\,max}$ is about 3.7 ma. (interpolating by eye between the -2 and 0-volt curves shown in Fig. 12); $e_{b\,min} = 65$ volts; $i_{b\,min} = 1.75$ ma.; and $e_{b\,max} = 160$ volts.

*It is interesting to note that in video applications the above distortion is not as important as that of phase distortion (non-uniform time delay).

Substitution of the current values in Eq. (11) yields

(Percentage 2nd harmonic distortion)

$$= \frac{3.7 + 1.75 - 2 \times 2.7}{2(3.7 - 1.75)} \times 100$$

$$= 1.28\%$$

or roughly 1.3 per cent. A slight error in reading can easily change the results appreciably; i.e., the accuracy is not very good unless a large drawing is employed. However, the decrease from 2 per cent to 1.3 per cent is noticeable and indicates that the distortion goes down when the tube is not forced to deliver a large output. The a.c. output is now

$$e_{b\,max} - e_{b\,min} = 160 - 65$$

$$= 95 \text{ v. peak-to-peak}$$

or 95/2 = 47.5 volts peak, which is, in general, sufficient to drive the grid of the ordinary small power tube.

If 60 volts output is desired, a higher supply voltage, such as 300 volts should be employed. This shifts the load line parallel to itself *to the right*, thereby requiring a greater grid cutoff voltage, and hence permitting a greater grid swing with a corresponding greater voltage output.

At this point it is of interest to note that the graphical method reveals a great deal of information concerning the operation of the tube. Provided actual tubes do not deviate too greatly from the curves published, it is possible to predict the gain, maximum output voltage, percentage distortion, required supply voltage, optimum bias, and the plate current drawn. The results thus obtained will be found to check very closely those experimentally determined in an actual setup.

THIRD HARMONIC DISTORTION.—The distortion in the output voltage is not exclusively second-harmonic. In the case of a triode this is a fairly good approximation, but measurements show some appreciable third, fourth, and even higher harmonic content. Often the higher harmonics, even though weak, are more objectionable to the ear than the second harmonic.

In the case of a pentode tube, under certain load conditions the second-harmonic distortion may be very low, but the third-harmonic distortion may be considerable. Hence, a method for calculating this from the graphical construction will be of value. Before presenting the method, however, it will be desirable to discuss wave-shape distortion in some greater detail.

Two simple rules are the following:

1. If the wave has *mirror* symmetry about the axis, then only odd harmonics are present. This is illustrated in Fig. 21. In (A) and (B) are shown two waves having mirror symmetry. The test is that if a mirror be imagined placed below the positive half-cycle, for example, the reflection in the mirror (shown by the dotted line) will be an exact duplicate of the negative half-cycle. Below (A) and (B) are shown the components of the waves. The components consist of a fundamental and a third harmonic; the waves

differ in shape because the third

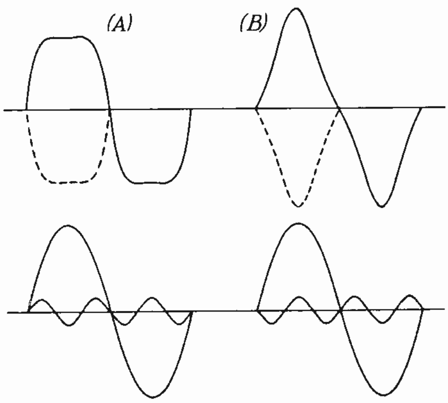


Fig. 21.—Examples of third-harmonic distortion showing mirror symmetry.

harmonic of (B) is 180° out of phase with that of (A).

2. If the wave does not have mirror symmetry, then even harmonics are present. This is illustrated in Fig. 22. In (A) the wave is

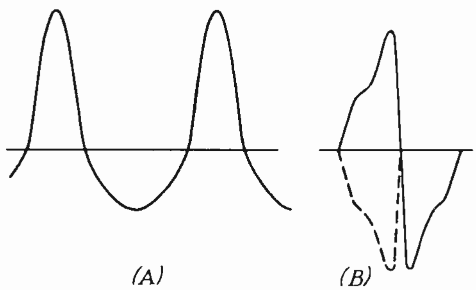


Fig. 22.—Presence of even harmonics owing to lack of mirror symmetry.

clearly lacking in mirror symmetry, it therefore has even harmonics. Indeed, it will be recognized as the wave discussed in connection with Fig. 20, and represents the typical output waveshape of an overdriven triode amplifier.

In (B) the wave has symmetry about the axis, *but it is not of the mirror type.* This is seen by comparing the mirror image of the positive half-cycle, shown in dotted lines, with the negative half-cycle itself. The two shapes differ in exactly the same manner as the right and left hands. Hence this wave has even harmonics too.

In general, if the wave lacks mirror symmetry, then even harmonics are present, *but this does not mean that odd harmonics may not be present, too.* The even harmonics counteract the tendency of the odd harmonics to produce mirror symmetry and thus mask their presence as regards this method. On the other hand, if the wave has mirror symmetry, then only odd harmonics, and no even harmonics are present.

With this in mind, consider the pentode characteristics illustrated in Fig. 23. Although pentode tubes will be discussed in a subsequent assignment, the tube characteristics can be studied here without it being necessary to know why the curves have the shape shown. Suppose the tube has a load resistance R_L in its plate circuit in series with a supply voltage E_{bb}, and assume further that R_L can be varied as desired.

The graphical construction is exactly the same as that described for the triode tube. The length E_{bb} is laid off on the voltage axis. From E_{bb} a line is drawn at the proper slope to represent the plate load resistance. Two lines are shown: one for a low

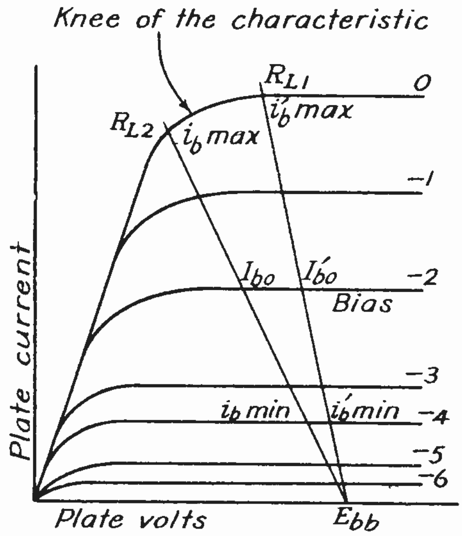


Fig. 23.—Pentode tube characteristics, showing how variation of the load line can produce even or odd harmonics, or both.

value of load resistance R_{L1}, and one for a higher value R_{L2}.

Let the bias be -2 volts, as shown. Referring to R_{L1}, it is clear from an inspection of the figure that for a positive grid swing from -2 to 0 volts, the rise in plate current from the quiescent value I_{bo} to the peak $i'_{b\,max}$ is much greater than the decrease for a negative grid swing from -2 to -4 volts, for which the plate current varies from I'_{bo} to $i'_{b\,min}$. If the current variations be plotted against time, they will appear as in Fig. 24 (A), and clearly the lack of mirror symmetry indicates the presence of even harmonics (principally the second).

On the other hand, for a higher load resistance R_{L2}, the corresponding maximum and minimum values $i_{b\,max}$ and $i_{b\,min}$ are about equally spaced from the average value I_{bo}, and therefore can give rise to a practically symmetrical wave-shape as shown in Fig. 24 (B). Such a wave will have pronounced odd harmonics, principally the third, as indicated in Fig. 24 (C), where the third harmonic is of such phase relative to the fundamental that the resultant wave (shown in dotted lines) is flattened on both positive and negative peaks.

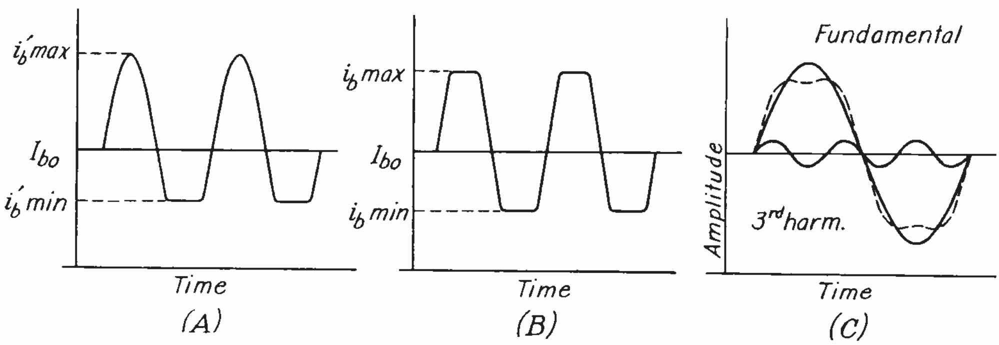
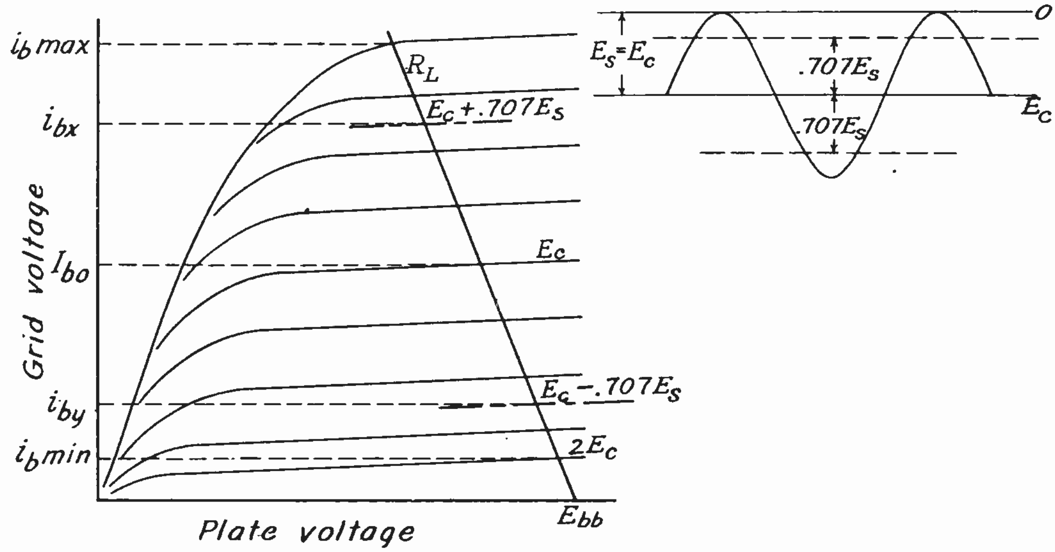


Fig. 24.—Pentode current wave-shapes for high and low plate load resistance.

The reason for the flattening

on the negative peaks is the same as that for Fig. 24(A) produced by R'_L of Fig. 19; namely, the grid curves draw together more and more as one goes to more negative grid voltages. The reason for the flattening of the positive peaks is also apparent from Fig. 23; namely, R_{L2} cuts into the so-called "knee" of the tube characteristic, a peculiarity of pentode and similar tubes at low plate voltages. In short, as the grid swings in a positive direction from the -2 volt bias value, the plate current rises less and less rapidly, producing the flattening of the positive peaks.

The amount of third-harmonic distortion can be calculated in a manner similar to that for the second-harmonic distortion, but additional and suitable instantaneous current values are required. These are found as indicated in Fig. 25. The load line is drawn through the supply-voltage point E_{bb}, as described previously. Since maximum Class A operation is intended, the peak grid signal swing E_s is equal to the bias voltage E_c. In the figure to the right is shown the sinusoidal signal-voltage wave superimposed on the bias.



(Courtesy RCA)

Fig. 25.—Required instantaneous current values to furnish information as to distortion content.

Note the two dotted lines in the figure, each .707 E_s volts to either side of the bias (zero-signal) line. The grid voltage passes through each of these instantaneous values twice per cycle. During the positive half-cycle the instantaneous grid voltage corresponding to the above value is E_c + .707 E_s; during the negative half-cycle the instantaneous grid voltage is $E_c - .707\ E_s$. In the left-hand diagram the curves corresponding to these two values are shown in broken lines; in general they will have to be drawn in by visual or other interpolation between their adjacent neighbors furnished by the manufacturer.

Where these two curves intersect the load line chosen, two values of current are obtained—i_{by} and i_{bx}, as shown. At the peaks of the grid swing there are similarly obtained $i_{b\,max}$ and $i_{b\,min}$, and where the bias curve intersects the load line gives the quiescent or no-signal d.c. component I_{bo}. Under signal excitation, the d.c. current may change to a higher (or lower) value A_0 owing to the effect of self-rectification described previously. In addition, an alternating wave will be generated that has a fundamental component A_1, that is, of course, a faithful copy of the input grid signal voltage, plus possibly a second-harmonic component A_2, a third-harmonic component A_3, a fourth-harmonic component A_4, etc.

The following formulas evaluate the peak amplitudes of these components in terms of the current values, $i_{b\,max}$, i_{bx}, I_{bo}, i_{by}, and $i_{b\,min}$:

$$A_0 = \frac{(i_{b\,max} + i_{b\,min}) + 2(i_{bx} + i_{by} + I_{bo})}{8} \qquad (12)$$

$$A_1 = \frac{\sqrt{2}\,(i_{bx} - i_{by}) + i_{b\,max} - i_{b\,min}}{4} \qquad (13)$$

$$A_2 = \frac{i_{b\,max} + i_{b\,min} - 2I_{bo}}{4} \qquad (14)$$

$$A_3 = \frac{i_{b\,max} - i_{b\,min} - 2A_1}{2} \qquad (15)$$

$$A_4 = \frac{2A_0 - i_{bx} - i_{by}}{2} \qquad (16)$$

It will be instructive to use these formulas in an example. For this purpose consider the 6J7 tube whose characteristics are shown in Fig. 26. Assume a 300-volt supply, and consider a 40,000- and a 50,000-ohm plate load resistor. Their load lines are shown in the figure. Use a -2 volt bias for the 40,000-ohm resistor, and a -2.5 volt bias for the 50,000-ohm resistor. The corresponding maximum grid swings are 2 volts and 2.5 volts peak, respectively.

Consider the 40,000-ohm load first. To locate i_{bx}, the instantaneous grid swing is

$$-2 + (2 \times .707) = -2 + 1.414$$

$$= -0.59 \text{ volts}$$

The corresponding curve can be sketched in (interpolated) between the -0.5 and -1 volt curves as follows:

In the vicinity of the load line, the -0.5 and the -1 volt curves are separated by a distance of 9 divisions. Therefore, the -0.59 volt curve must be

$$9\left(\frac{.59 - .5}{1 - .5}\right) = 1.62 \text{ divisions}$$

or approximately 1.6 divisions below the 0.5-volt curve. It is so drawn in Fig. 26 in a dotted line. It crosses the 40,000-ohm load line at $i_{bx} = 5.95$ ma.

In a similar manner i_{by} is obtained. The grid voltage is

$$-2 - (2 \times .707) = -3.41 \text{ volts}$$

The distance between the -3 and

-4 volt curves in the vicinity of the load line is 12 div. Therefore, the -3.41 volt curve is

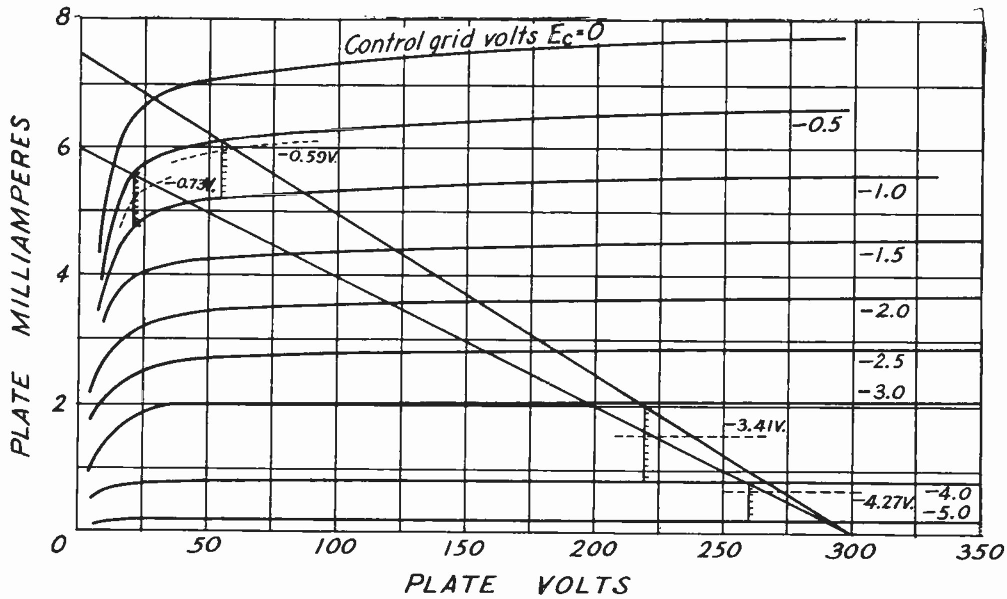
$$\left(\frac{3.41 - 3.00}{4.00 - 3.00}\right) 12 = 4.9 \text{ div.}$$

below the -3 volt curve, and is so drawn in dotted lines in Fig. 26.

Substituting these values in Eqs. (12) to (16), there are obtained:

$$A_0 = \frac{6.75 + 0.8 + 2(5.95 + 1.51 + 3.6)}{8}$$

$$= \frac{29.67}{8} = 3.71 \text{ ma.}$$



(Courtesy RCA)

Fig. 26.—Graphical constructions for determining the distortion for a 6J7 pentode tube for two values of load line.

Where this line crosses the 40,000-ohm load line gives i_{by} = 1.51 ma. Furthermore, from Fig. 26 it is found that $i_{b\,max}$ = 6.75 ma.; I_{bo} = 3.6 ma.; and I_{min} = 0.8 ma., corresponding to the intersection of 0, -2, and -4 volt curves, respectively, with the 40,000-ohm load line.

$$A_1 = \frac{\sqrt{2}\ (5.95 - 1.51) + 6.75 - 0.8}{4}$$

$$= \frac{12.23}{4} = 3.06 \text{ ma.}$$

$$A_2 = \frac{6.75 + 0.8 - 2 \times 3.6}{4}$$

$$= \frac{0.35}{4} = 0.0875 \text{ ma.}$$

$$A_3 = \frac{6.75 - 0.8 - (2 \times 3.06)}{2} = \frac{-0.17}{2}$$

$$= -.085 \text{ ma.}$$

$$A_4 = \frac{2 \times 3.71 - 5.95 - 1.51}{2} = \frac{-.04}{2}$$

$$= -0.02 \text{ ma.}$$

The negative sign in front of A_3 and A_4 signifies merely a reversal in phase; the components line up with A_1 so as to start out in a negative direction when A_1 starts out in a positive direction.

The percentage distortion for each harmonic is found by taking the ratio of each component to A_1. Thus

$$\% \text{ 2nd harm.} = (A_2/A_1)\,100$$
$$= 0.0875 \times 100/3.06$$
$$= 2.86\%$$

$$\% \text{ 3rd harm.} = (A_3/A_1)\,100$$
$$= .085 \times 100/3.06$$
$$= 2.78\%$$

$$\% \text{ 4th harm.} = (A_4/A_1)\,100$$
$$= 0.02 \times 100/3.06$$
$$= 0.653\%$$

Note that in the case of this pentode tube, the third-harmonic distortion almost equals the second-harmonic distortion in magnitude, and that the fourth-harmonic distortion is appreciable.

Consider next the 50,000-ohm load line and the -2.5 volt bias condition. The grid-voltage for i_{bx} is now

$$-2.5 + 2.5(.707) = -0.73 \text{ volt}$$

The spacing between the given curves in the vicinity of the 50,000-ohm line is 8 div., hence

$$\frac{.73 - .50}{1.00 - .50} \times 8 = 3.68 \text{ div.}$$

is the distance the -0.73 volt curve is below the -0.5 volt curve. The corresponding value of i_{bx} is 5.42 ma.

The grid voltage for i_{by} is now

$$-2.5 - 2.5(.707) = -4.27 \text{ volts}$$

There are 6 div. between the -4 and -5 volt curves, hence

$$\frac{4.27 - 4}{5 - 4} \times 6 = 1.62 \text{ div.}$$

is the distance the -4.27 volt curve is below the -4.0 volt curve. The value of i_{by} is found to be .64 ma. In addition, $i_{b\,max}$ = 5.75 ma., i_{bo} = 2.8 ma.; and $i_{b\,min}$ = 0.2 ma. Therefore,

$$A_0 = \frac{(5.75 + 0.2) + 2(5.42 + .64 + 2.8)}{8}$$

$$= 2.71 \text{ ma.}$$

$$A_1 = \frac{\sqrt{2}(5.42 - .64) + 5.75 - .2}{4}$$

$$= 3.08 \text{ ma.}$$

$$A_2 = \frac{5.75 + 0.2 - 2 \times 2.8}{4}$$

$$= 0.0875 \text{ ma.}$$

$$A_3 = \frac{5.75 - 0.2 - 2 \times 3.08}{2}$$

$$= -0.305 \text{ ma.}$$

$$A_4 = \frac{2 \times 2.71 - 5.42 - .64}{2} = -0.32$$

Therefore

% 2nd harm. = (.0875/3.08) 100

= 2.84%

% 3rd harm. = (.305/3.08) 100

= 9.9%

% 4th harm. = (.32/3.08) 100

= 10.38%

Note how the shift in bias, and the use of a higher load resistance, whose load line cuts farther to the left into the knee of the tube characteristics, has increased the distortion, particularly the third and fourth, which is especially undesirable. Of course, such high values of distortion are encountered only at maximum grid swing. If the tube were employed in the first stage, where the grid swing might be only a fraction of a volt, the knee of the tube characteristics could easily be avoided even if a higher load resistance were employed, and the distortion could be kept down to acceptably low values.

Although the distortion is best given in percentages of the different components, often a single overall value is desired. This is obtained by employing either the geometrical or the algebraic sum. The geometrical sum is the square root of the sum of the squares of the various percentages, and corresponds to a kind of r.m.s. value. Thus, for the 50,000-ohm load discussed above, the following is obtained,

Total percentage distortion

$$= \sqrt{(2.84)^2 + (9.9)^2 + (10.38)^2}$$

= 14.63%

This is the value most frequently used. The arithmetic sum is simply the sum of all the percentages, and gives a higher value than the geometrical sum. The former is, therefore, a more conservative rating for the performance of the stage, or for the whole amplifier. For the above example

Total percentage distortion

= 2.84 + 9.9 + 10.38 = 23.12%

POWER AMPLIFIER STAGES

The preceding discussion dealt mainly with voltage amplifier stages, although the method of calculating distortion applies more generally. In this section the power output stage will be analyzed, and design features considered. As has been indicated previously, the input signal voltage from a microphone, photocell, iconoscope, phonograph pickup, etc., is in general far too small to swing the grid of a tube sufficiently to cause it to produce any appreciable power output. It is necessary to amplify the signal voltage to a magnitude sufficient for this purpose.

The amplified signal is then impressed on the grid of a tube large enough to furnish the requisite power output. If the proper tube and load impedance are chosen, and the grid signal is adequate, the power output will be that desired, the distortion will be acceptably low, and the plate dissipation of the tube will be within the limits prescribed by the manufacturer. Correct design must therefore take note of all these features.

Circuit Considerations.— In Fig. 27 is shown the fundamental method of coupling the load impedance to the tube. In (A) a

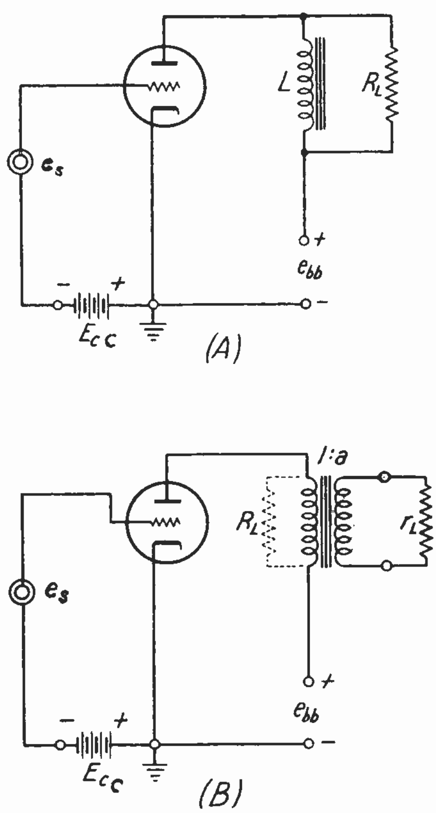


Fig. 27.— Fundamental and practical output circuits.

high-inductance choke L is used in parallel with the load resistance R_L. The latter is assumed to be the optimum value (ordinarily between 2 and 3 times R_p in the case of a triode tube), and L should have enough inductance so that its reactance is at least 2 or 3 times the magnitude of R_L and R_p in parallel, at the lowest frequency it is desired to amplify. This insures a flat low-frequency response for audio purposes.

In most practical cases, however, the actual load resistance r_L is not equal to the desired optimum value R_L. For example, the R_p of the tube may be 1,800 ohms, and the optimum value of R_L may be a little more than $2R_p$, or 3,900 ohms. On the other hand, the actual load may be a dynamic loudspeaker voice coil, whose impedance r_L is only 15 ohms. In order to couple this value of impedance efficiently to the tube, a step-down transformer is required, as shown in Fig. 27(B) by T. If the step-down ratio is 1 to a, such that

$$a = \sqrt{R_L/r_L} \qquad (17)$$

then the actual impedance r_L will be reflected to the tube as the optimum value R_L, and maximum power transfer will occur from the tube through the transformer to the load. In the example just cited, the required step-down ratio would have to be by Eq. (17),

$$a = \sqrt{3900/15} = 16.13$$

Viewed *from the tube*, r_L appears as $a^2 r_L = (16.13)^2\ 15 = 3{,}900$ ohms, the desired value, represented by the dotted lines and designated as R_L in Fig. 27(B).

When a transformer is employed, the inductance of the primary is determined in exactly the same way as the choke L in Fig. 27(A); i.e., it should have a reactance at the lowest frequency to be amplified, that is between 2 and 3 times the value of a^2r_L (or R_L), and R_p in parallel. Note that in any event the inductance is that to the a.c. component, *in the presence of the d.c. component of the tube.* The choke or transformer must therefore be designed so that the core does not saturate owing to the d.c. component, as this would drastically reduce the inductance to the a.c. component. Ordinarily, this is a design problem for the transformer engineer.

The *resistance* of the choke or of the transformer primary is usually negligible compared to R_L or R_p. In view of the fact that the d.c. component flows through it, the graphical construction for this circuit must be suitably altered from that previously described to take this fact into account. (In the resistance-coupled amplifier both the d.c. as well as a.c. components flowed through R_L, so that one common load line sufficed for both components.)

In the case of Fig. 27(A), the a.c. component—on the other hand—cannot flow to any appreciable extent through the choke L because its reactance has been deliberately chosen very high to preclude such flow. Hence, the a.c. component must flow through R_L (in parallel with L), where its power is expended as desired. Thus, no d.c. power is wasted in R_L; only the a.c. power is absorbed in it. This makes for higher efficiency of operation, which is very desirable in the output stage, where the greatest amount of d.c. power is drawn from the power supply, and where the largest and most expensive tube is to be found.

In the case of Fig. 27(B), the action is exactly similar. the transformer reflects the resistance r_L on its secondary side as R_L on the primary side, and this reflected resistance R_L appears effectively in *parallel with the open-circuit inductance of the primary winding.* Thus, the latter corresponds to L in Fig. 27 (A), and the two circuits behave practically identically. Hence, Fig. 27(A) can be used as the basis for further work.

Before discussing the graphical construction, it will be of value to make a preliminary physical analysis of the action of the stage. Assume that a steady d.c. component I_{bo} has been set up in the choke L, and that therefore no signal e_s is applied to the grid. Since under these conditions there is no appreciable voltage drop in L, all of the d.c. input power is consumed in the tube. The amount is simply $I_{bo}E_{bb}$. What actually happens is that the electrons whose motion represents I_{bo}, are accelerated to the plate by the voltage E_{bb}, and strike the plate with considerable force thereby heating it up. In short, the d.c. power $I_{bo}E_{bb}$ is converted into heat at the plate, and this energy must be radiated at a rate equal to that at which it is produced. The effect is known as *plate dissipation,* and the manufacturer sets an upper limit to the amount of power the plate of a given tube can dissipate.

When a signal e_s is applied to the grid, one may say that the internal resistance of the tube is varied, thus varying the plate current and producing an a.c. component superimposed on the d.c. component. This is illustrated in Fig. 28. Note that if the rise

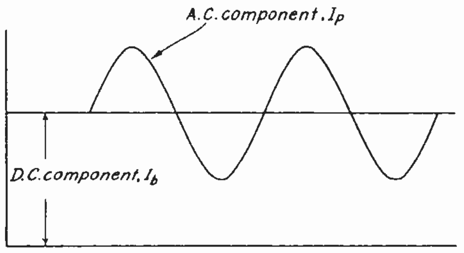


Fig. 28.—The production of an a.c. component in a Class A amplifier does not markedly alter the magnitude of the d.c. component.

and fall of the current are equal, the *average, which is the d.c. component*, is practically unchanged. Previously it was shown that owing to self-rectification in the tube, the average or d.c. component during signal excitation does change somewhat, but the illustrative examples showed that the change was small.

Hence, it can be assumed that the d.c. component remains substantially constant. This means that the power input from the supply source is still $I_{bo}E_{bb}$, since an a.c. current times a d.c. voltage gives zero *average power*. However, the a.c. component of peak amplitude I_p, in flowing through R_L produces an expenditure of power equal to $I_{pm}^2 R_L/2$, where the 2 arises from the fact that the effective value $I_{pm}/\sqrt{2}$ rather than I_{pm} itself, must be squared to get the power expended. This *output* power is not necessarily converted into heat; it may actually be converted into acoustic energy radiated from a loudspeaker, or as light, or to operate a motor.

Nevertheless, $I_{pm}^2 R_L/2$ represents energy delivered to a load, yet the d.c. input $I_{bo}E_{bb}$ has not changed. The only conclusion is that less of $I_{bo}E_{bb}$ is dissipated at the plate, and the difference is delivered to R_L. To put it another way, when a signal is applied to the grid, the plate dissipation *is reduced* from the maximum no-signal value of $I_{bo}E_{bb}$ to the lower value $I_{bo}E_{bb} - I_p^2 R_L/2$, *or the plate actually runs cooler when the tube is delivering power to a load than when it is not.* Hence, at least for Class A operation, the plate dissipation must be checked under *no-signal conditions;* if it is not excessive in this case, it will be satisfactory under signal conditions.

The next point to consider is the action of the plate current

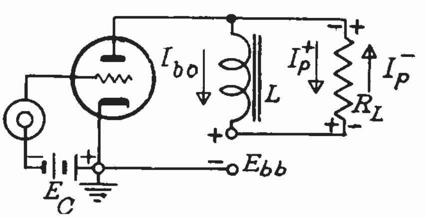


Fig. 29.—Path of a.c. component of plate current in an amplifier circuit.

during various instants in the signal cycle, Suppose the grid voltage swings in the positive direction thus causing the plate current to increase. The increase in plate current cannot flow through L because the high inductance of the latter prevents the current in it from changing appreciably from its initial d.c. value I_{bo}. Hence, the excess of plate current must flow through R_L as I_p^+ in Fig. 29, and since this is an electron flow, it makes the top end of R_L negative to the bottom end, as shown. The result is that *the plate voltage is less than the supply voltage* E_{bb} *by the amount* $I_p^+ R_L$; i.e., the plate voltage drops below the supply voltage.

Now suppose that the grid swings in a negative direction, thereby decreasing the plate current by an amount I_p^-. Since the tube now passes less current than the d.c. value I_{bo}, the choke L seeks an alternative path for this deficit of current I_p^- that the tube will not pass. Such a path is clearly through R_L as shown, but since the current I_p^- flows *up* through R_L, it makes the top end *positive* to the bottom end. The plate voltage now *rises* above E_{bb} by the amount $I_p^- R_L$.

This rise in plate voltage above the supply voltage is owing to the inductive rise in voltage produced in a choke when it is attempted to reduce the current flowing through it. In particular if the grid is driven to plate-current cutoff, then the decrease in plate current I_p^- is from I_{bo} to zero, or a total change of I_{bo}, and maximum rise in plate voltage is obtained — namely, to a value $E_{bb} + I_{bo} R_L$. It is this inductive rise in voltage during negative grid swings that makes this circuit more efficient than the ordinary resistance-coupled amplifier, as will be shown.

GRAPHICAL CONSTRUCTIONS.—The above discussion indicates the procedure to be followed in making a graphical construction to determine the power output, the distortion, etc. For the purpose consider a 2A3 tube operating at 250 volts plate supply. From the manufacturer's data it is found that the maximum permissible plate dissipation is 15 watts, and that the optimum load resistance is R_L = 2,500 ohms.

Consider first the plate dissipation of 15 watts. At no signal, all of the d.c. input power, or $E_{bb} I_{bo}$, is consumed at the plate. Since here E_{bb} = 250 volts, I_{bo} can be found:

$$250 I_{bo} = 15 \text{ watts}$$

or

$$I_{bo} = 15/250 = .06 \text{ amp.} = 60 \text{ ma.}$$

This is the maximum d.c. plate current that can be permitted to flow.

Now refer to Fig. 30, where the tube characteristics are given. Since I_{bo} flows through a practically zero resistance choke, its load line should be *vertical* and hence correspond to the *ordinate* through 250 volts. This is shown in Fig. 30, the ordinate is drawn to a height of 60 ma.; this locates the quiescent point Q. The grid bias can now be ascertained; it is the label of the tube curve passing through the quiescent point.

Quite often the particular curve is not furnished by the manufacturer, in which case it must be interpolated visually.

In Fig. 30 it is noted that it is the -43.5 volt curve. A word of caution is here necessary. In the case of a filament type tube, the manufacturer usually employs d.c. to light the filament in the test setup, and connects his grid

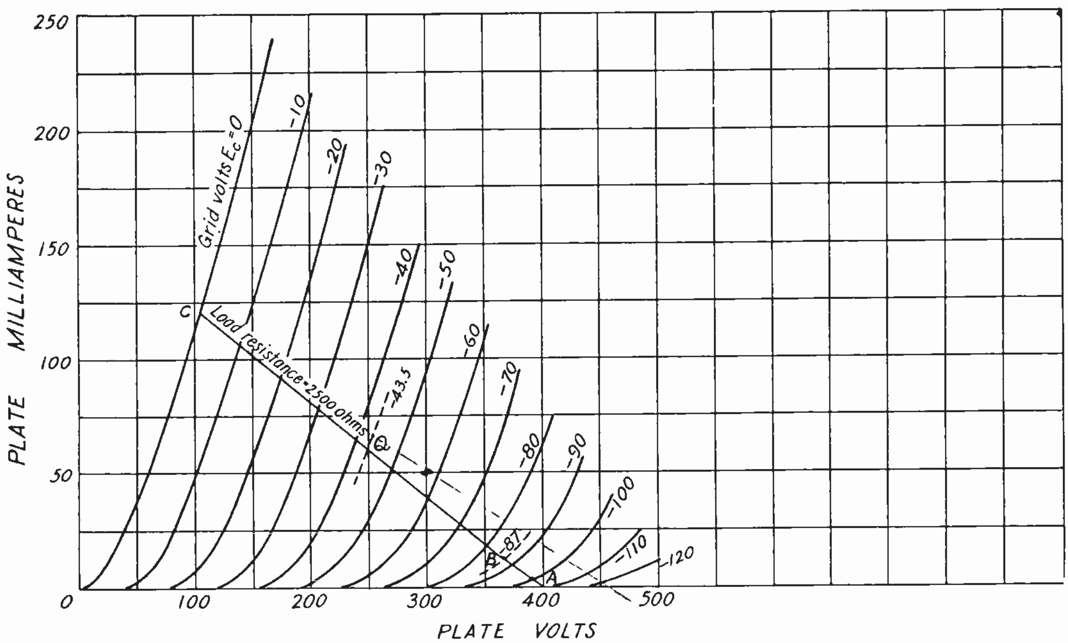


Fig. 30.—Graphical constructions for a 2A3 power tube.

bias battery to the negative side of the filament. The *true bias* is then the grid bias plus one-half the filament voltage. In the case of the 2A3 the latter is 2.5 volts, hence the true bias is -43.5 - 2.5/2 ≅ -45 volts.

While the difference is small, the effect on the current can be large and cause excessive plate dissipation if the wrong bias is employed. For example, if in an actual amplifier the filament is lighted on a.c. and the bias battery or resistor is connected to the center-tap of the filament transformer winding, then the bias will have to be -45 volts instead of -43.5 volts to keep I_{bo} down to 60 ma.

Once the quiescent point has been located, the load line for R_L can be drawn. Suppose the grid bias is driven to cutoff. Then the instantaneous plate voltage rises to $E_{bb} + I_{bo} R_L$. In this example, the rise is to

$$e_{b\,max} = 250 + (.06)(2,500) = 400 \text{ volts}$$

Thus, at the moment when i_b is zero, e_b is 400 volts. This locates a point A on the voltage axis, as shown. If A is joined to Q and prolonged, the load line for R_L is obtained, and it passes through Q, as required. Any other point on the load line can be found, if desired, by similar triangles.

Now, if the grid voltage varies from its bias value, the plate current varies above or below its quiescent value I_{bo}, and the instantaneous value of the plate current is found, as previously, by the intersection of the load line with the tube curves.

POWER OUTPUT CONSIDERATIONS.—The maximum possible grid swing for Class A operation is clearly equal to the bias voltage, or 43.5 volts peak in this example. On the negative half-cycle the grid therefore reaches a negative voltage of 2 x 43.5 = -87 volts. As can be seen from the figure, the 2,500-ohm load line intersects the -87 volt tube curve at point B, where the tube curve is beginning to bend over considerably, and can cause considerable distortion for a greater grid swing. Indeed, it is clear from the figure that operation below about 12.5 ma. is inadvisable because of distortion considerations, regardless of the position of the quiescent point and of the slope of the load line (value chosen for R_L).

Therefore 12.5 ma. is the minimum plate current $i_{b\,min}$ that can be permitted, and it occurs at a grid potential of -87 volts. Since the plate current has decreased from its initial value of 60 ma. to 12.5 ma., or 60 - 12.5 = 47.5 ma., the plate voltage rises to a value $e_{b\,max}$ = 250 + (.0475) (2,500) = 369 volts, as can be checked from Fig. 30.

On the other hand, when the grid swings positive up to zero volts, the plate current rises to a maximum value $i_{b\,max}$ = 119 ma., and the plate voltage drops by an amount (.119 - .060) (2,500) = 147.5 volts, to a minimum value $e_{b\,min}$ = 250 - 147.5 = 102.5 volts, or approximately 103 volts, as is also clear from the figure, (point C). The current and voltage relations across R_L are therefore as in Fig. 31.

From Fig. 31 it is clear that the *peak-to-peak* value of the current is

$$i_{b\,max} - i_{b\,min}$$

and of the voltage across R_L, it is

$$e_{b\,max} - e_{b\,min}$$

The *peak* values are therefore half of the above or

$$I_{pm} = (i_{b\,max} - i_{b\,min})/2$$

and

$$E_{pm} = (e_{b\,max} - e_{b\,min})/2$$

The effective values are $I_{pm}/\sqrt{2}$ and $E_{pm}/\sqrt{2}$, and the power in R_L is the product of these two quantities, so that finally the power output is

$$P_0 = \frac{I_{pm}}{\sqrt{2}}\frac{E_{pm}}{\sqrt{2}} = \frac{I_{pm}E_{pm}}{2}$$

$$= \frac{(i_{b\,max} - i_{b\,min})(e_{b\,max} - e_{b\,min})}{8} \quad (18)$$

This formula is somewhat approximate, but is fairly accurate if the distortion is fairly low and consists mainly of second harmonics. In the case of the 2A3,

$$P_0 = \frac{(.119 - .0125)(369 - 103)}{8}$$

$$= 3.54 \text{ watts}$$

which checks the manufacturer's value of 3.5 watts very closely.

Since this is a triode tube, Eq. (11) for second harmonic distortion should be satisfactory, as very little third or higher harmonics need be expected. From Eq. (11)

(Percentage 2nd harmonic distortion)

$$= \frac{119 + 12.5 - 2 \times 60}{2\,(119 - 12.5)} \times 100 = 5.4\%$$

which is normally regarded as not excessive for a power tube.

An important point to note from Fig. 30 is that the load line appears on the tube curves just as it does in Fig. 11 for a resistance-coupled amplifier, but with the following important difference:

In Fig. 12 the load line meets the voltage axis at point A corresponding to the plate-supply voltage E_{bb}; in Fig. 30 the load line meets the voltage axis at point A corresponding to a value *greater than* E_{bb}. Specifically, point A in Fig. 30 corresponds to 369 volts, and E_{bb} to 250 volts.

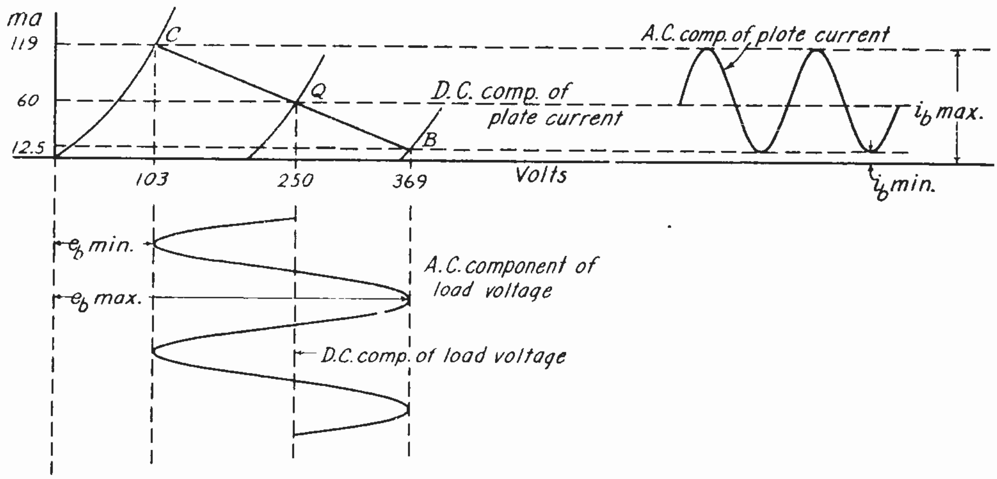


Fig. 31.—Diagram showing current and voltage components for load resistor.

Thus, the inductive feed employed in a power output stage *makes that stage perform like a resistance-coupled stage to which a much higher supply voltage is applied;* in the above example the power stage with 250 volts applied behaves like a resistance-coupled stage to which 369 volts are applied.

The reason for such equivalent performance is that there is no d.c. power wasted in the load resistance in the case of the power output stage, so that it can furnish as much output as a resistance-coupled stage with higher supply voltage, in which an appreciable part of the d.c. power is wasted in the load resistance. This indicates why inductive plate feed is employed for a power output stage; more a.c. output power is obtained for a given amount of d.c. input power, since the efficiency of operation is higher.*

OPTIMUM VALUE OF LOAD RESISTANCE.—The factors determining the value of load resistance for an actual tube are often at variance with those specified for a hypothetical tube. Thus, if the tube characteristics are all straight lines, parallel to one another and equidistant in spacing, then their common slope represents a fixed value of plate resistance R_p.' For such a hypothetical tube, maximum power output is obtained when the load resistance R_L equals twice R_p, *and the bias is adjusted to a value* $-3/4\ (E_b/\mu)$. No consideration is given to permissible plate dissipation, and the grid swing is assumed to be just sufficient to drive the grid to zero volts on the positive half-cycle, and to cutoff on the negative half-cycle. Since the tube curves are assumed to be straight lines, their entire length can be used, and hence $i_{b\,min}$ can be made actually zero, in contrast to the 12.5 ma. minimum in the case of the 2A3 tube just analyzed.

However, it is evident from an inspection of Fig. 30 that the 2A3-curves are not straight lines, and the bias cannot be chosen as $-3/4\ (E_b/\mu)$, but rather on the basis of plate dissipation, as discussed previously; in short, the quiescent point is determined by the permissible plate dissipation and applied B-voltage.

But if the quiescent point is determined, then the grid swing is determined too, since this is numerically equal to the bias voltage. On the negative half-cycle, the grid will swing to twice the bias voltage, or $2E_c$. The plate current thereupon decreases, but the plate voltage rises, and as a result a point such as B in Fig. 30 is reached. The position of B is somewhere along the $2E_c$ curve (where $2E_c = -87$ volts in Fig. 30).

How high up B is depends upon the slope of the load line; i.e., upon R_L. If B is high up on the curves, then the slope of the load line is small, R_L is high, and $i_{b\,min}$ is large, as is indicated by line AQB in Fig. 32. As is clear from the figure, $i_{b\,min}$ = BG; $i_{b\,max}$ = AH, and $i_{b\,max} - i_{b\,min}$ is relatively small. At the same time, $e_{b\,min}$ = OH; $e_{b\,max}$ = OG; and $e_{b\,max} - e_{b\,min}$ = GH is relatively large. The product

$$(i_{b\,max} - i_{b\,min})(e_{b\,max} - e_{b\,min})$$

is proportional to the power output P_o [see Eq. (18)].

On the other hand, if R_L is low in resistance, then its load line is relatively steep, such as CQD. But D corresponds to zero plate current, or $i_{b\,min} = 0$, and this value is reached before the grid has completed its peak negative excursion to $2E_c$. This means that the output wave will have the peak of its negative cycle flattened—or "clipped", as it is often called—, and such clipping represents very definite distortion.

This is highly undesirable

*Unfortunately, wide-band transformers are not available for video purposes, at least in high-level (high-power) stages. In such a case a compensated resistance-coupled amplifier is required, as for modulation purposes, and the efficiency of operation is correspondingly low.

in the case of an audio amplifier, hence such a low value of R_L is undesirable. An intermediate value and the load resistance corresponding to EQF will be the preferred value. If the distortion is ex-

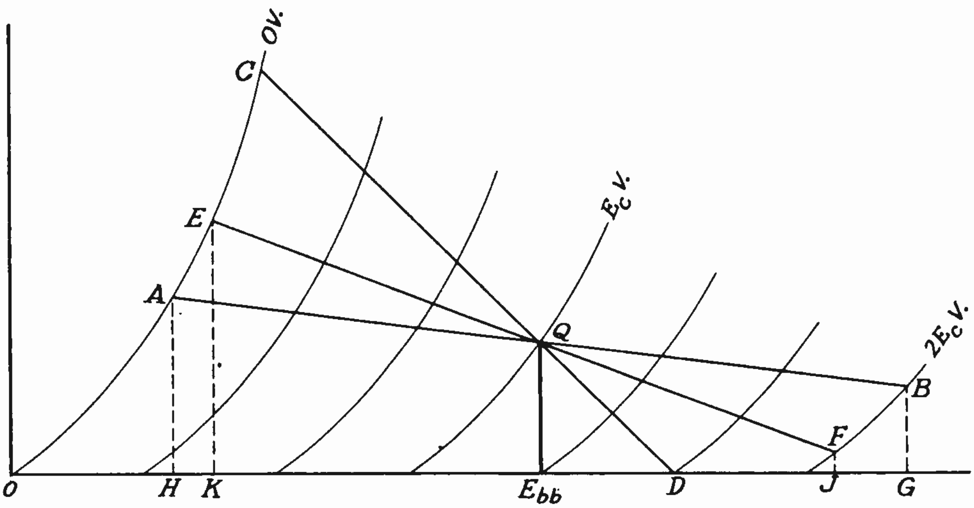


Fig. 32.—Determination of optimum value of load resistance.

is represented by EQF. Here F is lower on the tube curve and hence in the lower bend region. $i_{b\,min}$ = FJ, and is less than BG, but $i_{b\,max}$ = EK and is greater than AH. On the other hand, $e_{b\,min}$ = OK and is greater than OH, while $e_{b\,max}$ = OJ and is less than OG. Thus ($i_{b\,max} - i_{b\,min}$) is greater than that for the load line AQB, but ($e_{b\,max} - e_{b\,min}$) is less, and the product, hence power output, may or may not exceed that for AQB.

On the other hand, there is no doubt that the distortion for load line EQF will be greater than that for AQB because F is on the more curved portion of the tube characteristics. If the power output for EQF comes out greater than that for AQB, and the distortion is within the specified limits, then EQF will be preferred. cessive, however, or if the power output, for AQB is greater, then the load resistance corresponding to AQB will be the preferred value.

Hence, the optimum value of R_L for an actual tube is determined by the permissible distortion, and the maximum power output that can be obtained, *under the condition that the quiescent point is fixed by supply voltage and plate dissipation considerations.* Unless the supply voltage is inordinately low, so that E_{bb} is close to 0, the minimum slope of the load line (which just hits cutoff—$i_{b\,min}$ = 0—at peak negative grid swing) will be more than twice R_p as a general rule for a triode. Under those conditions the lower R_L can be made without exceeding the specified distortion, the greater will the power output be. In Fig. 32 this means that EQF will furnish greater power output than AQB, and will therefore be preferred *if the distortion is within the permissible limits.*

In the case of the 2A3 tube, for E_{bb} = 250 volts, and the plate dissipation not to exceed 15 watts, the quiescent point was shown to be Q (Fig. 30) where I_{bo} = 60 ma., and E_c = -43.5 volts. The plate resistance R_p at this point is found by drawing the tangent to the -43.5 volt tube curve, and calculating its slope. It will be found that this corresponds to about 800 ohms. Since R_L is specified by the manufacturer as 2,500 ohms, it is clear that R_L is more than $3R_p$ and hence considerably greater than the optimum value of $2R_p$ or 1,600 ohms specified for a linear tube.

Hence, if more than 5% distortion can be tolerated, R_L can be made somewhat less than 2,500 ohms, whereupon the power output will be somewhat greater than 3.54 watts. The increase, however, will not be large before excessive distortion is obtained. On the other hand, if 5% distortion is permitted, there is no point in making R_L greater than 2,500 ohms, since the distortion is within 5% for this value of load resistance. For R_L larger than this the power output will be decreased, with an unnecessary decrease in distortion, since 5% is permissible.

To summarize the above: the quiescent point, grid bias, and grid swing are determined by the plate-supply voltage and permissible plate dissipation. Then, after the quiescent point is determined, the optimum load line furnishing maximum power output consistent with the amount of distortion permitted, can be found.

In passing, it is to be noted that since the optimum value of R_L is in the neighborhood of two to three times R_p (and in the case of a pentode tube is actually a fraction of R_p, as will be discussed in a subsequent technical assignment), maximum voltage gain will not be obtained, nor is this the object of a power output stage. Indeed, the reason for the relatively low value of R_L is to obtain maximum power *output* rather than maximum voltage *gain.*

It is clear that the higher the G_m of the tube, the greater will be the power output. This is because:

1. $G_m = \mu/R_p$ and if either μ is high or R_p is low G_m is large.

2. If μ is high, then the apparent generated voltage of the stage (μe_s) can force a higher a.c. component through R_p and R_L in series, and thus give more power output, or

3. If R_p is low, then R_L will be low, too, as was mentioned in the preceding paragraph, so that the given value of μ will be able to force more a.c. component through the lower impedance of R_L and R_p in series, and hence give more power output.

As a simple example of (3), suppose two tubes, A and B, each have a μ = 10, and a grid swing of 20 volts peak, but A has an R_p of 5,000 ohms, and B has an R_p of 2,500 ohms. Suppose in either case the optimum value of load resistance is twice the R_p of the tube. Consider Tube A. The apparent generated voltage is $\mu e_s = 10 \times 20 = 200$ volts. The peak a.c. component is

$$I_p = \mu e_s/(R_p + R_L)$$

$$= 200/(5{,}000 + 10{,}000)$$

$$= 13.33 \text{ ma.}$$

and the output power is

$$P_o = I_{pm}^2 R_L/2 = (.01333)^2 \; (10{,}000)/2$$

$$= .888 \text{ watt}$$

Next consider Tube B. Since its R_p is half of that of Tube A, its G_m is twice as great. The power output is found similar to that for Tube A, and comes out to be

$$P_o = \left(\frac{200}{2500 + 5000}\right)^2 \times \frac{5000}{2}$$

$$= 1.778 \text{ watts}$$

or twice as great as that for Tube A. As a result, power tubes are designed to have as high a transconductance as possible, and this quantity represents "a figure of merit" for a tube.

RESUMÉ

This concludes this assignment on vacuum tube amplifiers. In it were discussed voltage amplifiers of the broad-band type, such as for audio and video applications, graphical constructions, and power amplifiers of the broad-band type. Analysis of Class A, r.f. amplifiers, push-pull or balanced amplifiers, Class B amplifiers, and design considerations concerning Class C amplifiers, will be developed in later technical assignments.

V-9174-33

VACUUM TUBE AMPLIFIERS

EXAMINATION

1. A given tube has a mu of 18 and an R_p of 10,000 ohms. It is connected to a plate resistor having a value of 44,000 ohms?

   (A) What is the G_m of the tube?

   (B) What is the voltage gain?

2. (A) A television pentode amplifier tube has a G_m of 9,000 µmhos, and is connected to a video load whose impedance is 2,000 ohms. What is the voltage gain of the video stage?

   (B) Suppose a tube of G_m = 5,000 µmhos is substituted for the above tube, but the load impedance can be increased to 3,000 ohms. What will be the voltage gain of the stage under these conditions?

3. Refer to Fig. 5. In a certain amplifier stage, both R_L and R_g are 100,000 ohms in value, and C_g = 0.25 µf. At 1,000 c.p.s. its reactance is negligible in comparison to R_g.

   (A) What is the plate load impedance to the a.c. component for a 1,000 c.p.s. grid input wave?

   (B) If the tube has a mu of 20, and an R_p of 15,000 ohms, what is the voltage gain at 1,000 c.p.s.?

4. (A) Why is a flat frequency response down to—say, 5 c.p.s., undesirable in an audio amplifier?

   (B) In certain special amplifiers that are required to amplify uniformly down to about 1 or 2 c.p.s., large coupling condensers of several microfarads capacity are required. These are chosen to have a relatively high voltage rating in order to obtain a low leakage conductance. Explain why this is necessary.

5. A certain vacuum tube is to be operated with a plate potential of 300 volts and a grid bias of -20 volts, under

5. *(Continued)*

which conditions it will draw a plate current of 20 ma.

(A) Calculate the size cathode bias resistor required to furnish the above -20 volts.

(B) What size by-pass condenser should be used with this bias resistor? (Calculate at 30 c.p.s.)

6. (A) Refer to Fig. 11. The total current through R from all amplifier stages is 80 ma., and the voltage drop desired is 60 volts. Calculate the value of R necessary.

(B) Calculate the by-pass condenser required in parallel with R. (Calculate at 30 c.p.s.)

7. (A) Reference Fig. 11. Suppose $e_f = 2$ volts, $C_F = 0.1$ µf and $R_F = 100$ K ohms. What will be the value of e'_f at 60 c/s?

(B) The bias E_c required for the stage shown in Fig. 11 is 12 volts. What resistance is required between point A and ground?

8. Given the plate characteristics for the 6SQ7 as shown in the accompanying graph. The power supply voltage is 300, the bias is -1.5 volts, and the plate load resistance R_L is 250,000 ohms.

(A) What is the maximum peak-to-peak grid signal voltage?

(B) What is the corresponding peak-to-peak signal output voltage across R_L?

(C) What is the voltage gain α of this stage, as determined graphically?

9. Refer to Problem 8.

(A) What is the d.c. *plate* voltage when no grid signal is applied?

9. (B) What is the corresponding d.c. plate current?

(C) What is the d.c. voltage drop across R_L when no grid signal is applied?

10. Given the triode characteristic curves for a 6L6 tube as shown in the accompanying graph. The tube is to be operated as a *power* amplifier. The power supply voltage is 250, and the maximum permissible plate dissipation is 10 watts, and the load resistance is 5,000 ohms (as reflected between plate and B+).

(A) Locate the quiescent point.

(a) What is the d.c. component of plate current under zero-signal conditions?

(b) Estimate to the nearest whole number the grid bias required, and use this value in subsequent calculations.

(B) What is the maximum grid swing possible under Class A operation?

(C) **Draw the** load line through the quiescent point for a value of R_L = 5,000 ohms. Interpolate the tube curve required for the maximum negative grid excursion, by the method given in the text for distortion measurements.

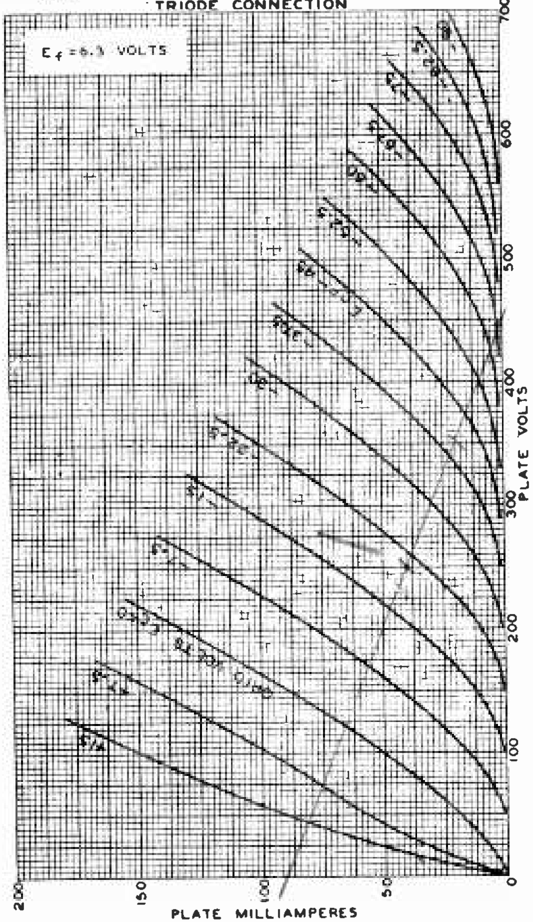
(D) Calculate the power output.

(E) Calculate the per cent second-harmonic distortion.

NOTE:—Some discrepancy between the values determined here and those given by the manufacturer will be found owing to the fact that the load line is not corrected for self-rectification effects. The discrepancy, however, will be small.

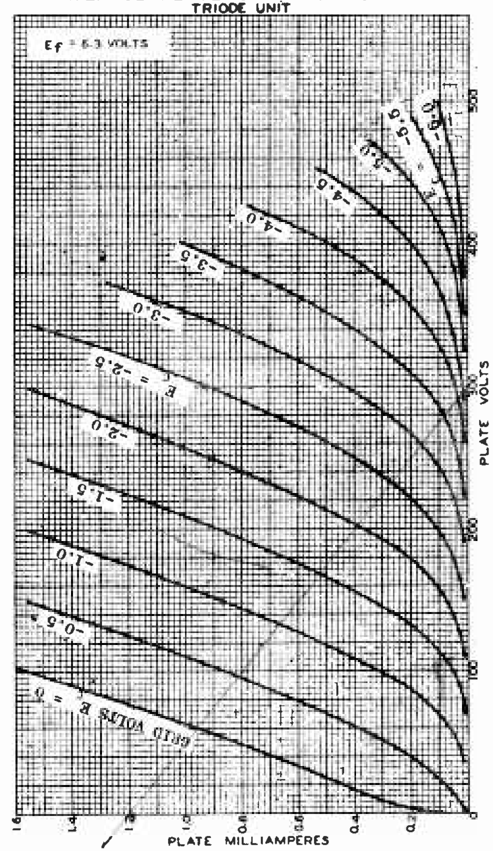
6L6

6L6
AVERAGE PLATE CHARACTERISTICS
TRIODE CONNECTION



6SQ7

6SQ7
AVERAGE PLATE CHARACTERISTICS
TRIODE UNIT



AUG. 11, 1941

RCA RADIOTRON DIVISION
RCA MANUFACTURING COMPANY, INC.

92C-4975R1

V-9174-33

1. A. $G_m = \mu / R_p = \frac{18}{10,000} = \underline{\underline{1800}}$ micro-mhos.

B. $\alpha = \mu Z_L / R_p + Z_L = \frac{18 \times 44,000}{10,000 + 44,000} = \underline{\underline{16.1}}$ X → 14.67

2. A. $\alpha = G_m Z_L = 2,000 \times .009 = \underline{\underline{18}}$

B. $\alpha = G_m Z_L = 3,000 \times .005 = \underline{\underline{15}}$

check it.

3. A. $Z_L = \frac{100,000}{2} = \underline{\underline{50,000}}\ \Omega$

$\alpha = \mu Z_L / R_p + Z_L = \frac{20 \times 50,000}{15,000 + 50,000} = \underline{\underline{6.66}}$ X → 15.4

4. A. This is undesirable because of tendency to motorboat, due to coupling between stages through the common B supply.

B. The leakage acts as a resistor shunting the condensor. This in conjunction with R_G acts as a voltage divider for D.C. thus causing the grid of the following tube to be positive with respect to the filament, so grid current will flow.

5. A. $E_p = 300V \quad E_g = -20V \quad I_p = 20$ ma

$R_c = \frac{E_g}{I_p} = \frac{20}{.020} = \underline{\underline{1000}}$ ohms

B. $\frac{1}{2\pi FC} = .1 R_c \quad R_c = 1000\ \Omega \quad F = 30$ cps.

$\frac{1}{6.28 \times 30 \times C} = 100 \quad C = \frac{1}{628 \times 30} = \frac{1}{18,840} = .000053$

or $\underline{\underline{53}}\ \mu fd$

6. A. $I_p = 80$ ma $E_g = 60$ V

$R = \frac{E_g}{I_p} = \frac{60}{.08} = 750$ ohms

B. $\frac{1}{2\pi FC} = .1R = 75$ $F = 30$ cps.

$C = \frac{1}{6.28 \times 30 \times 75} = .000071$ farad.

or 71 ufd

7. A. $e_f = 2$V $C_F = .1$ ufd $R_F = 100$ k ohms $F = 60$ cps

$X_c = \frac{1}{2\pi Fc} = \frac{1}{6.28 \times 60 \times 10^{-7}} = \frac{10^7}{6.28 \times 60} = 26,600$ ohms.

see back of opposite page

$e'_F = \frac{26,600}{100,000 + 26,600} \times 2 = .42$ volts ← $X_c + R$ cannot be added directly since the voltage drops across R + C are not in phase!

B. $12 : 100,000 = x : 126,600$

$x = \frac{126,600}{100,000} \times 12 = 15.2$ volts

8. with no voltage drop in tube $I_b = \frac{300}{250,000} = 1.2$ ma.

A. Max. peak to peak grid signal voltage $= 1.5 \times 2 = 3$ volts

B. Signal output voltage, P to P $= 258 - 60 = 198$ volts.

(C) Voltage gain $\alpha = \frac{198}{3} = 66$

9. A. DC plate Volt. with no grid signal = 168 volts

B. DC plate current with no grid signal = 0.53 ma

C. Voltage drop across R_L with no grid signal $= 300 - 168 = 132$ volts

10. A. (a) DC Plate current at zero signal

$= \frac{10}{250} = .04$ or $\underline{40}$ ma.

(b) Grid Bias = $\underline{-20}$ volts closer to -21V

B. Max grid swing possible under class A is 20 volts peak each way.

C. Where the max. negative grid voltage of -40 volts intersects the load line is still above the curve in the characteristic curves. Therefore it should be satisfactory to use full swing of 20 volts.

solution for more exact values!

D. $E_{max} = 353$ V. $I_{min} = 20$ ma. (17.5)

$E_{min} = 122$ V $I_{max} = 67$ ma. (65)

$$P = \frac{(I_{max} - I_{min})(E_{max} - E_{min})}{8} = \frac{.047 \times 231}{8}$$

$= \underline{1.36 \text{ watts}}$

E. $\% \text{ Dist.} = \frac{(I_{max} + I_{min}) - 2 I_o}{2(I_{max} - I_{min})} \times 100 = \frac{67 + 20 - 80}{2(67 - 20)} = \frac{7}{94} \times 100$

$= 7.45\%$ high

(The power checks with data in tube manual but the distortion is approx. 50% high)

solution

(over)

#7(A) $X_c \doteq 26{,}500\,\Omega$

$$Z = \sqrt{R^2 + X_c^2} = 103{,}500\,\Omega$$

$$I = \frac{e_f}{Z} = 19.3\,\mu A$$

$$e_f' = I X_c = \boxed{0.513V} \text{ ans.}$$

$$e_f = \sqrt{e_f'^2 + e_{r_f}^2} = 2V$$

7(B) $R = \frac{E}{I} = \frac{12}{.08} = 150\,\Omega$

VACUUM TUBE AMPLIFIERS.

Problem 8. To draw load line:- One point where $E_b = 300$ and $I_b = 0$. The second point where $E_b = 0$ and $I_b = (300/250{,}000) = 1.2$ ma. (See pages 19 and 20).

(a) From the tube manual we see that this type of tube may only be operated class A, hence the grid must not be allowed to swing positive as grid current would flow and distortion will appear in the output. In addition, it can be seen that the grid swing should not be allowed to fall below -3 volts. Otherwise the tube would be working into the lower bend of the E_pI_p curves and distortion would be the result. The maximum peak-to-peak grid signal voltage is then equal to twice the bias. $E_g = 2 \times 1.5 = 3$ volts. answer.

(b) The peak-to-peak signal voltage across R_L is $E_L = 260 - 60 = 200V$. (See page 23).

(c) Gain = 200/3 = 66.7

Problem 9(a).D-C voltage = 168 volts. (See page 22).

(b) $I_b = .52$ ma. (See curve).

(c) If the plate voltage is 168 and the power supply voltage is 300, then the difference (300 - 168) must be dropped across R_L. $E_L = 300 - 168 = 132$ volts.

Problem 10. (a) Under quiescent conditions the plate current I_b is:- I_b = 10 watts/250 volts = 40 ma. (See page 43). From the curves the grid bias is -21 volts. The maximum grid swing 2 x 21 = 42 volts. To draw the load line; one point is where the 40 ma plate current intersects the 250 plate voltage line. The second point is where I_b is zero and $E_b = 250 + (.040 \times 5000) = 450$ volts. (See page 44).

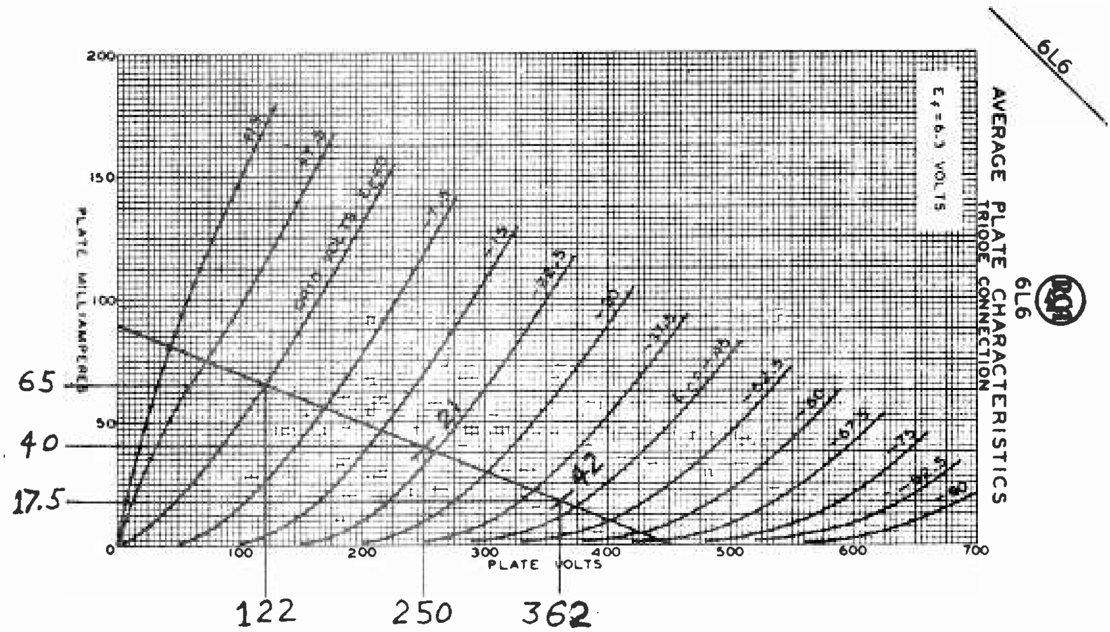
(d) The power output is found by using equation (18) page 45.

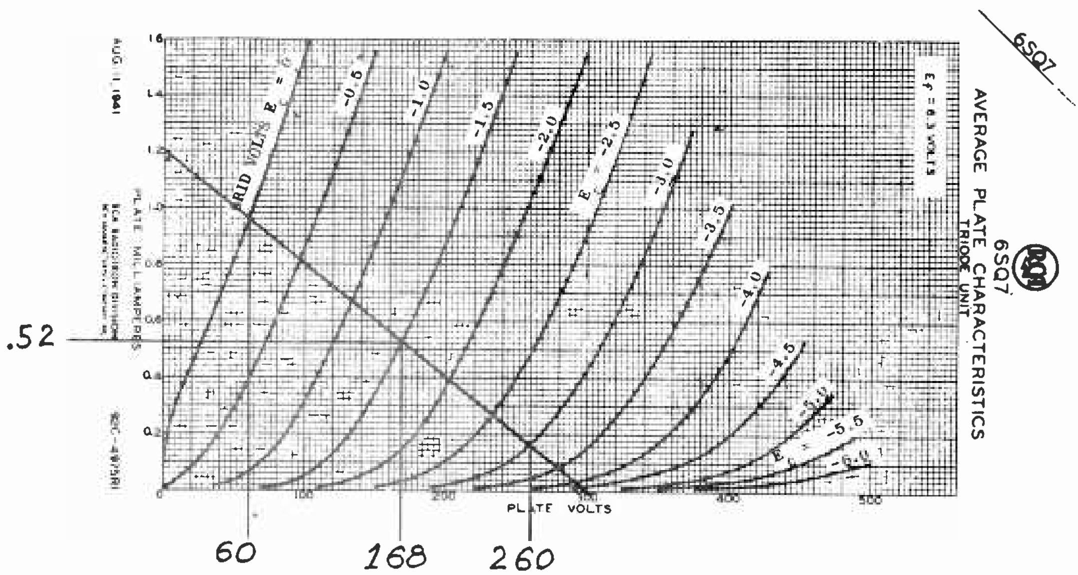
$P_o = (.065 - .0175)\ (362 - 122)/8 = 1.425$ watts.

(e) The per cent second-harmonic distortion may be found by using equation (11) on page 30.

Distortion = (.065 + .0175) - (2 x .04) x 100 / 2(.065 - .0175)
= 2.64 per cent second harmonic distortion.

The tube manual, for these operating conditions, gives the power output as 1.4 watts, and the total harmonic distortion as 5 per cent.


6L6
AVERAGE PLATE CHARACTERISTICS
6L6
TRIODE CONNECTION
Ef = 6.3 VOLTS
PLATE MILLIAMPERES
PLATE VOLTS
65
40
17.5
21
42
122
250
362


6SQ7
AVERAGE PLATE CHARACTERISTICS
6SQ7
TRIODE UNIT
Ef = 6.3 VOLTS
PLATE MILLIAMPERES
PLATE VOLTS
AUG. 1, 1941
92C-4975R1
.52
60
168
260