HIGH POWER LINEAR AMPLIFIERS

INVERSE FEEDBACK

finished Feb. 19, 1959 21RC

NATIONAL RADIO INSTITUTE
ESTABLISHED 1914
WASHINGTON, D. C.

STUDY SCHEDULE NO. 21

For each study step, read the assigned pages first at your usual speed. Reread slowly one or more times. Finish with one quick reading to fix the important facts firmly in your mind, then answer the Lesson Questions for that step. Study each other step in this same way.

☐ 1. How the Class B Linear Amplifier Works . . . . Pages 1-8

Here you learn how it is possible to amplify a modulated signal where low-level modulation is used. The problems of getting linear response are fully covered. Answer Lesson Questions 1, 2, and 3.

☐ 2. Typical Linear Amplifiers and How to Adjust Them . . Pages 9-15

Typical single-ended and push-pull linear amplifiers; the advantages of push-pull operation; adjustment procedure. Answer Lesson Question 4.

☐ 3. A High Efficiency Linear R.F. Amplifier . . . . Pages 16-20

One of the drawbacks of linear amplifiers is their low efficiency. The Doherty circuit described here has far better efficiency than the standard types, and hence is widely used. Answer Lesson Question 5.

☐ 4. Inverse Feedback in Radio Transmitters . . . . Pages 20-27

Inverse feedback is widely used in transmitters for the purpose of reducing distortion, hum, and noise that arises in the modulator and in a.f. amplifiers. Here you study the basic circuits, and learn how the feedback of an out-of-phase signal can reduce the undesired components in a signal. Answer Lesson Question 6.

☐ 5. Typical Inverse Feedback Circuits . . . . Pages 27-36

Although all inverse feedback circuits are just variations on the original idea, nevertheless they are sufficiently different to require some study. Answer Lesson Questions 7, 8, 9, and 10.

☐ 6. Answer Lesson Questions.

☐ 7. Start Studying the Next Lesson.

FM1M457
1957 Edition
Printed in U.S.A.

HIGH-POWER LINEAR AMPLIFIERS

INVERSE FEEDBACK

How the Class B Linear Amplifier Works

EARLIER Lessons showed how amplitude modulation could be obtained by applying the modulating signal to the plate or one of the grids of an r.f. amplifier, and also showed that this modulation could occur in the final r.f. amplifiers or in an intermediate amplifier if a sufficient number of buffer stages were used between the modulating point and the master oscillator.

When the plate voltage to the final r.f. amplifier stage is modulated, the system is called "high-level" modulation, because the modulation occurs at a high power level.

To plate-modulate a high-power final amplifier, however, requires a large amount of audio-frequency modulator power. Indeed, the modulator must be capable of supplying in watts 50% of the power fed to the plate of the final r.f. stage. This means that if 5000 watts is being supplied to a final transmitter stage, then the modulator must be capable of delivering 2500 watts of undistorted audio power for 100% modulation. Such audio-frequency power is not easy to get.

An alternative method of obtaining an amplitude-modulated r.f. output is through the use of "low-level" modulation, in which modulation occurs in the grid circuit of the final r.f. amplifier, or else in the grid or plate circuit of an intermediate amplifier. In low-level modulation, far less audio power is needed than is the case with high-level modulation. However, there is considerable difference in the design of the r.f. amplifiers that have to handle the modulated signal, as compared to the standard class C stages used as r.f. amplifiers when high-level modulation is employed.

When the grid of the final r.f. amplifier is modulated, the stage must be adjusted so that its output power is a linear function of its grid voltage—the stage must operate as a class B "linear" amplifier. If the modulation occurs in a preceding stage, then all stages from the point of modulation up to and including the final stage, must be class B linear stages.

At first glance, it would appear that the saving of modulator power would cause low-level modulation to be universally employed. However, the linear r.f. amplifier has some "quirks" and limitations, one of which is a far lower efficiency than that of a standard class C amplifier. Let's now see how the linear amplifier operates, how it is adjusted, and what its further advantages and disadvantages may be.

BASIC PRINCIPLES OF THE LINEAR R.F. AMPLIFIER

You have already studied the grid-modulated final stage, so let's now

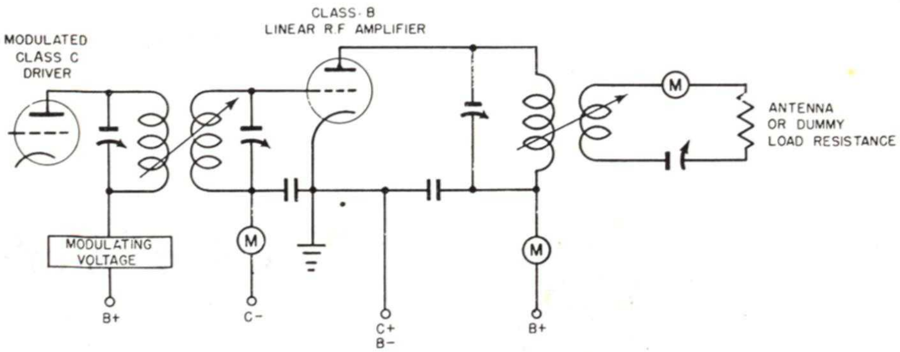

FIG. 1. Simplified circuit of the class B linear r.f. amplifier.

study the type of class B linear stage that follows a modulated stage.

In appearance, this class B linear amplifier resembles the ordinary class C amplifier studied earlier. In Fig. 1 is shown a representative diagram of a single-tube class B stage. For clarity, all neutralizing condensers, r.f. chokes, etc., have been omitted in this drawing.

There is a basic difference between this amplifier and a grid-modulated final stage—the amplifier in Fig. 1 is operated as a *true* class B stage, with the C bias set near plate-current cutoff. In actual practice, since the E_g-I_p characteristic of a vacuum tube is not a straight line near plate current cutoff, the linear amplifier usually is set to the "extended" cut-off bias. This is illustrated in Fig. 2. It can be seen here how the linear or "straight-line" portion of the tube characteristic curve is extended in a dotted line until it crosses the grid voltage axis. Grid bias voltage indicated at this crossing point is the usual operating C bias.

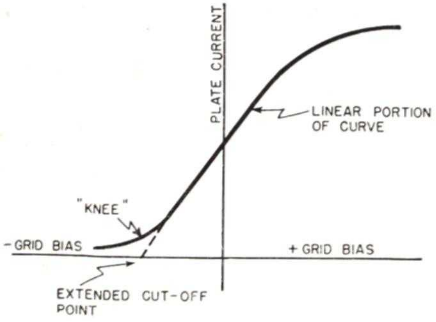

FIG. 2. Dynamic characteristic of a vacuum tube showing how the straight-line portion is extended to give the "extended" cut-off grid bias used in class B operation.

Action with Excitation. With grid bias just high enough to prevent appreciable plate current (when there is no grid excitation), the amplifier action when a modulated r.f. signal is applied is illustrated in Fig. 3. This figure shows a dynamic characteristic curve of a vacuum tube. As shown, the modulated r.f. input signal is superimposed upon the cut-off bias. When the r.f. signal is negative, therefore, it makes the control grid even more negative, and no plate current at all flows for each negative half-cycle. When the r.f. signal is positive, however, this *subtracts* from the grid bias, and plate current flows for the duration of each positive half-cycle. If the r.f. signal voltage is small, then only small half-cycle plate current pulses flow through the tube. If the r.f. signal voltage is large, then large half-cycles of plate current are allowed to flow.

Observe that the plate current pulses, large or small, flow only during the positive half-cycles of the r.f. input voltage. Except for the slight curvature near the extended cut-off bias, the plate current pulses would exist for exactly 180° of each r.f. cycle. Furthermore, if the dynamic characteristic is linear, as it should be, whenever the modulated input signal peaks double in value as for 100% modulation, the linear amplifier plate current peaks also exactly double in value. In other words, as indicated in Fig. 3, the envelope of the plate current pulses is an enlarged duplicate of the positive side of the grid excitation envelope, faithfully following the peaks and troughs of modulation.

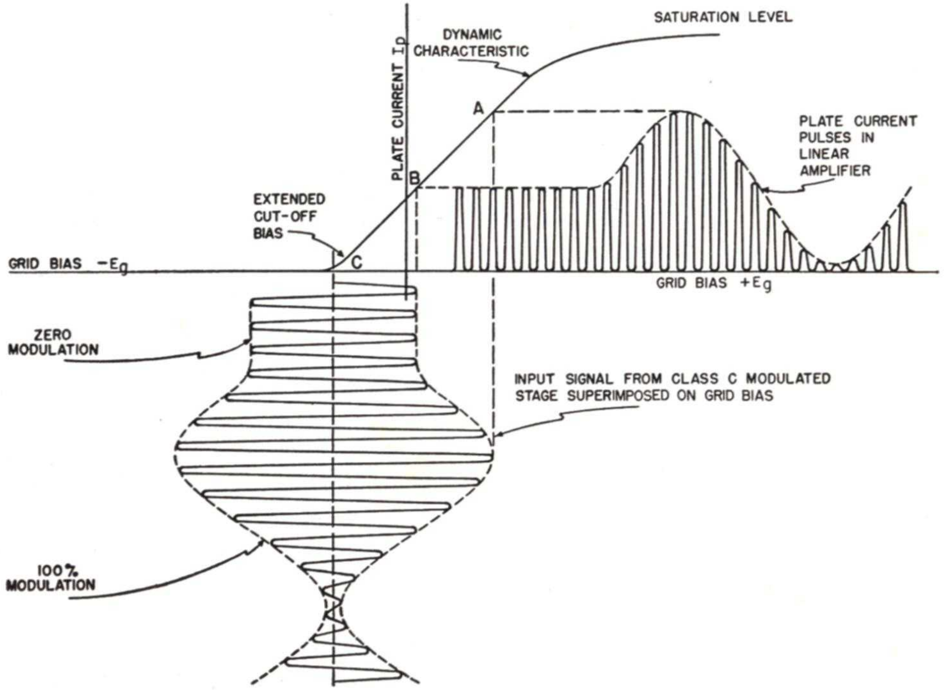

FIG. 3. In class B operation, the modulated excitation is superimposed upon the cut-off grid bias so that plate current pulses of varying height flow only during the positive half-cycle (180°) of each r.f. alternating cycle.

The Flywheel Effect. The varying plate current pulses shown in Fig. 3, of course, deliver energy to the output tank only half the time. During the half-cycle when no plate current is flowing, just as with the class C amplifier studied earlier, the "fly-wheel effect" of the tank circuit takes over to supply the missing negative half-cycle of r.f. current. Fig. 4 is an illustration of what occurs. The positive half-cycles of tank current, generated by plate current pulses, are drawn in heavy lines. The negative half-cycles, as supplied by the tank circuit flywheel effect, are shown in dotted lines. Large plate current pulses, of course, cause a large reverse flywheel current because the tank condenser is given a large charge; small plate current pulses, on the other hand, result in a corresponding smaller flywheel tank current. The tank current, therefore, follows *both* positive and negative sides of the excitation modulation envelope.

Since it is the *tank* current and not exclusively the plate current that is coupled into the antenna or load circuit, the output power in the load is very nearly a perfectly modulated sinusoidal wave, in spite of the fact that the linear amplifier tube feeds power only during one half-cycle and "coasts" the other half.

HOW THE CLASS B AMPLIFIER IS MADE LINEAR

In the ideal linear amplifier, the output voltage should always be a constant factor times the input grid voltage. We find, however, that unless adjusted properly the class B r.f. amplifier will not be exactly linear. There are two important factors that determine the degree of linearity in such an amplifier: the load impedance presented to the tube, and the fixed bias applied to the grid.

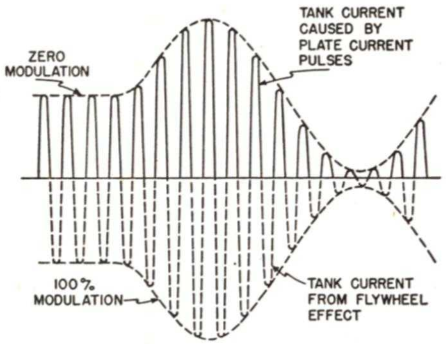

FIG. 4. Even though plate current of a single-tube class B amplifier flows only half the time, the "flywheel" effect of the tank circuit supplies the missing half-cycle to produce a very nearly perfect sinusoidal wave.

Linearity with Load Impedance. If we plot the dynamic characteristic of a tube operating in class B for various values of load impedance, we have a family of curves somewhat like those in Fig. 5. The instantaneous grid volts, actually the result of C bias and r.f. excitation together, are plotted along the horizontal axis, while the plate current, or proportional tank current, is represented along the vertical axis. Since we are concerned with linearity, one interesting fact is apparent—although relative current swing is smaller, *the higher the load impedance, the straighter the dynamic curve becomes.*

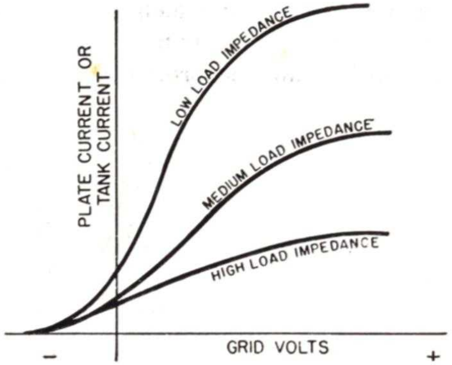

FIG. 5. Linearity curves for a class B amplifier using various values of load impedance.

Linearity with Grid Bias. If we plot the dynamic characteristic of a class B tube for different values of grid bias, we would get a family of curves like those in Fig. 6. The curve for low grid bias is very straight for low values of grid voltage, but quickly folds over into an undesirable flat region. The curve for high grid bias, on the other hand, is concave upward for a low grid voltage because the tube is operating too far down on the "knee" of its characteristic E_g-I_p curve. The curve for medium grid bias is reasonably straight for both low and relatively high grid-excitation voltage. In practice, best medium grid bias is found to correspond closely to the extended cut-off bias described earlier and illustrated in Fig. 2.

Saturation Effects. A very important point shown in Figs. 5 and 6 is that no matter what load impedance is used or what fixed grid bias is applied, if the excitation grid voltage is increased far enough, the dynamic curve no longer is straight, but instead falls over and flattens out so that further increases in excitation have very little effect. This is called "saturation."

But what is saturation? Let us consider the circuit of Fig. 1 for two extreme conditions. First, when the r.f. grid exciting voltage is on a negative half-cycle, the tube is cut off and no plate current flows. With no plate current flowing there is no voltage drop across the load impedance, and the total B-supply voltage is developed across the tube. Next, on a positive grid voltage swing, conditions are reversed. Theoretically, let us say, the positive grid voltage is high enough to reduce the tube plate resistance to zero. If this were so, then no voltage would be lost across the tube, and all the B-supply voltage would be dissipated across the load impedance.

Hence, we see that the theoretical maximum r.f. peak voltage that can be developed across the load impedance is exactly equal to the B-supply voltage, and this is true no matter how large the positive grid swing might become. Hence, increasing the grid swing further cannot produce more output.

Optimum Plate Load Impedance. So far we have learned that there is an optimum grid bias for linearity in a class B amplifier. We have discovered, too, that linearity can be improved by using a high load impedance, but we have said nothing about the power output available, or the average plate efficiencies under various conditions.

Let us suppose, in order to obtain a very high degree of linearity between driving and output voltage, that we use an extremely high value of load impedance. What happens? As you might expect, not only is the linearity good, but also the plate circuit *efficiency* is high. Unfortunately, however, we find the power output is very low, because the load impedance does not match the tube impedance.

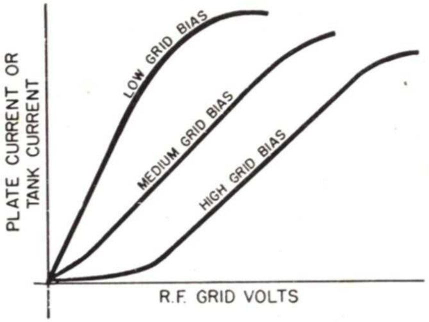

FIG. 6. Curves showing effect of grid bias upon linearity of a class B r.f. amplifier.

Now if we go to the other extreme by trying a very low value of load impedance, we find another set of conditions as the result. If the load impedance is less than the tube plate resistance, most of the voltage due to the flow of plate current is lost across the tube and not across the load impedance. Consequently, the efficiency is very low, and the tube losses reach an undesired high value before satisfactory power output can be obtained. Also, the linearity between input and output voltage has been sacrificed.

As with all generators, the linear amplifier has an optimum output load impedance for which the power output is maximum, and reasonable efficiency and linearity are obtained. In practice, for a triode, it is found that a load impedance that is equal to *twice* the plate resistance is this optimum value.

Conditions for Proper Operation. As we mentioned before, for distortionless output, the amplifier must operate only over the straight-line portion of its dynamic characteristic curve. This means that the highest peak grid voltage should not swing the plate current beyond point A in Fig. 3. But as you know, modulation peaks for a 100% modulated r.f. wave are twice the carrier value without modulation. Point B then, marks the amplifier peak plate current for the no-modulation condition, and represents one-half the current drawn at point A.

When the carrier excitation without modulation is properly adjusted to point B, positive modulation peaks carry the r.f. excitation to point A; negative modulation troughs drop the excitation back down the curve to point C which represents zero, or minimum plate current pulses. The linear variation in size of the plate current pulses is shown in the figure.

Average Plate Current. If the characteristic of the class B amplifier is truly linear, the *average* plate current *rise* during a modulation crest is equal to the *average* plate current *drop* during a modulation trough. This is important. It means that the over-all *average* plate current drawn by the amplifier, and that which is read by a plate current meter, is constant and *will not change from the no-modulation to full 100% modulation condition.* This fact is often used to check the performance of a linear amplifier. If the plate current varies during modulation, it is a sure sign that the amplifier is not operating properly and it is in need of adjustment.

Peak Power Efficiency. With a constant *average* plate current flowing, and constant plate voltage being supplied, the class B amplifier obviously draws an unvarying, constant amount of power from the high voltage supply. On the other hand, we know that for 100% modulation, the output voltage peaks on modulation crests must be twice the value of those for no modulation. Since power in the load is represented by E^2/Z, where E is the load voltage and Z the impedance of the load, the crest peak output power is *four* times the peak power with no modulation.

Where does this increase in power come from? The answer is simple: *The linear amplifier changes its efficiency.* Indeed, the efficiency for 100% modulation excitation is twice as great as it is for excitation with no modulation. For no modulation, the efficiency is about 30% to 35%, which indicates that about two-thirds of the input power is being dissipated by the plate of the tube. For full modulation, an efficiency of 60% to 70% is commonly obtained. This is an interesting fact, for it discloses that the class B tube runs cooler when it is delivering the most power.

Average Plate Efficiency. The conventional linear amplifier, however, is not an efficient device when compared to a class C stage. In the wave forms corresponding to voice and music, for instance, it is found that the highest voltage peaks are often ten times, and maybe as much as twenty times as great as the average voltage level. To handle such peaks without distortion from saturation effects, it is necessary to keep the average excitation to the linear amplifier down to a fairly low level. Amplifier plate circuit efficiency for the modulation *peaks* still will approach 60% to 70%, but the *average* efficiency seldom exceeds 40%. Special means for overcoming this inherent low efficiency have been devised, which we will discuss later.

Supplying the C Bias. One of the problems involved in designing a class B linear amplifier is devising a satisfactory way of supplying the C-bias voltage. As we have already pointed out, there is an optimum C bias for maximum linearity. Furthermore, once this C bias is fixed at the proper value, it should remain substantially constant, and never vary throughout a modulation cycle.

At first glance, since the grid is driven very hard into the positive potential region so that considerable grid current flows, it seems possible that grid-leak self-bias could be used in the same manner as is done with a class C amplifier. Excitation to a class B linear amplifier, however, is *not* constant, and as a result the grid current is small for a low degree of modulation, and very high for modulation nearing 100%. Under these conditions, grid bias obtained by means of a grid leak would change widely throughout a modulation cycle, giving serious distortion.

On the other hand, steady C bias can be obtained by use of the "cathode" resistor self-bias method. This is possible because under proper operating conditions the linear amplifier plate current is constant, and does not change with modulation percentage. Actually, most high-power tubes do not have cathodes. With filament-type tubes, self-bias is accomplished by inserting the bias resistor between the filament center-tap and ground.

By far the most common and most practical way of obtaining C-bias voltage is to use a separate, *low-impedance* C-bias power supply. Low impedance is necessary to prevent voltage changes with different values of grid current. With this system, the separate supply lends itself to easy, independent bias variations so that the proper adjustment of the linear amplifier is most convenient.

Load to the Class B Driver. The varying grid current drawn by the class B amplifier under excitation leads to another important performance consideration: the load the amplifier presents to its driver. On the negative half-cycle of r.f. excitation, the input signal serves to make the grid highly negative, so that the grid draws no current at all, and the input impedance to the amplifier is very high.

But on the positive r.f. half-cycle, conditions are very different. First, as the grid reaches zero potential, it begins to draw a small current; second, as the grid is driven more positive, the current rises sharply; and third, as the positive excitation reaches the peak, the grid draws such a high peak current it may be an appreciable fraction of the total space current of the tube. If current flows, then power is being dissipated. Also, the higher the current, the lower is the impedance presented to the driver circuit.

We see now that over a complete r.f. cycle, the grid impedance may change from an infinite value for negative grid potential to a few hundred ohms for a high positive potential. But what happens to the driver stage during these load-changing excursions? For no grid current and light load, the driver voltage will be high; for high grid current and heavy load, the driver voltage will be lower than it should. The higher the impedance of the driver, the worse this effect will be.

The end result is that positive peaks of excitation are flattened out. This is pictured in Fig. 7. The actual excitation of the class B amplifier grid from a *poorly regulated* driver is shown in heavy line. For comparison, a perfect sinusoidal excitation voltage is drawn in dashed line. Obviously, even though a linear amplifier is operating over a perfectly linear dynamic characteristic curve, excitation such as that in Fig. 7 will result in serious output distortion.

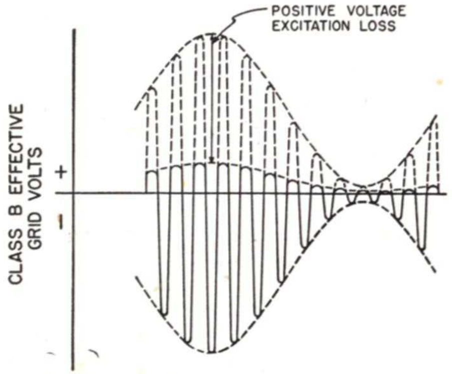

FIG. 7. Class B amplifier excitation distortion which can result from the driver stage having poor voltage regulation.

To minimize this grid-loading effect, the driver regulation must be made as high as possible. This is usually done by deliberately designing the driver stage to be capable of delivering two or three times as much power as that required to drive the linear amplifier grid—thus keeping the driver impedance low. In addition, the input tank of the linear amplifier is shunted by a relatively low resistance. This resistance, of course, absorbs considerable driver power, but it reduces the wide fluctuations in grid circuit impedance. With the resistor in shunt, the input circuit impedance no longer varies from a few hundred ohms to practically an open circuit; instead, it can change from a few hundred ohms only up to the value of the shunting resistance. In this manner the driver is made to work into a more nearly constant load.

Over-all Class B Power Gain. Theoretically, one would expect the class B linear amplifier to be capable of increasing modulated carrier power ten or twenty times. In actual practice, however, this is not generally true. Because of the driver power lost in the grid circuit loading resistor, and the fact that the driver must be considerably overrated to insure good regulation, a power gain of about five-to-one is the value most ordinarily achieved.

Typical Linear Amplifiers and How to Adjust Them

A typical single-ended linear amplifier is shown in Fig. 8. Tuned resonant circuits are used in both grid and plate, L_2-C_2 forming the grid tank, and L_3-C_3 forming the plate tank. Input power from the class C driver stage in this particular circuit is fed in by link coupling; other forms of coupling, however, can be used if desired.

Resistor R_1, shown shunted across the input tank condenser C_2 and most of L_2, is the grid loading resistor used to present a more constant load to the class C driver, thereby improving regulation. As indicated by the tap on the input tank coil L_2, grid neutralization is employed. Plate neutralization is possible and may be used if more convenient.

In this amplifier, a separate C-bias power supply is used. The filament supply, therefore, is grounded through the filament center-tapped resistor.

Grid current, drawn by the linear amplifier, is read on d.c. milliammeter MA_1. Similarly, milliammeter MA_2 measures the plate current. The meters are by-passed by condensers C_9 and C_{10} to prevent erroneous readings, or damage to the instruments from stray r.f. currents. Antenna or load current is indicated on the thermo-couple-type r.f. ammeter MA_3.

Meter Indications. The readings on all three meters are indicative of proper linear amplifier operation. If the class B amplifier is adjusted correctly, and it is being driven by a well-

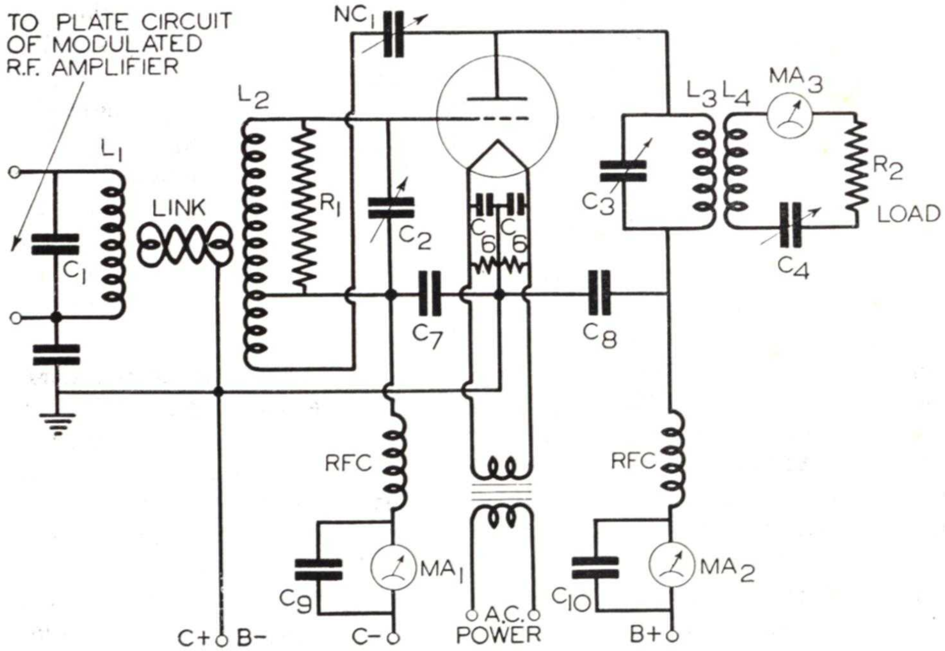

FIG. 8. A typical single-ended class B linear r.f. amplifier.

regulated driver, the meter readings should behave as follows:

With *no modulation,* and the carrier on, the grid current MA$_1$ will be small but steady in value. The plate current MA$_2$ also will be steady, and should be the proper value for recommended power output. Antenna or load current MA$_3$ should read a steady value showing full constant carrier output.

With *full 100% sinusoidal modulation,* the grid current MA$_1$ should rise sharply to a maximum value. Plate current MA$_2$, however, should not change; and will not if the amplifier is truly linear. The load current MA$_3$, of course, should show the normal 22.5% rise due to complete modulation.

With *voice or music modulation,* the grid current MA$_1$ will rise and fall drastically with the peaks and troughs of modulation. Plate current MA$_2$, nevertheless, should never vary from its normal, steady, recommended reading. Because of the low average power of voice or music, the load current MA$_3$ will be only a few per cent higher than the no-modulation value, showing only slight "wiggles," and no marked rise except for loud, sustained passages.

PUSH-PULL LINEAR AMPLIFIER

So far we have studied only class B amplifiers using a single tube. It is possible, however, to build class B linear amplifiers using two tubes, or several pairs of tubes, in push-pull. One reason for push-pull operation, of course, is that twice as much power output can be obtained from two tubes in push-pull. There is another much more important reason.

You will remember that a single-tube class B stage depends upon the flywheel effect of the plate tank circuit to "round out" and supply the missing half-cycle of r.f. energy when the single tube is idle and drawing little or no plate current.

Unfortunately, the flywheel effect of any tank circuit is not perfect. Flywheel effect usually is measured in terms of the circuit "Q," or quality factor. Q, itself, is approximately equal to the inductive reactance of the tuned circuit divided by the resistance in the circuit at resonance. A very important point is: the resistance in the circuit means not only the resistance of the inductance coil, but *also the effective antenna or dummy load resistance coupled into the circuit.*

We see then, in an effort to get more power output, the closer we make the coupling between coils L$_3$ and L$_4$ in Fig. 8, the more effective resistance is coupled into the tank circuit L$_3$-C$_3$. With more effective resistance in the tuned tank circuit, the Q of this circuit is decreased markedly, and the flywheel effect is made correspondingly smaller. The final result is that the dashed half-cycle r.f. pulses in Fig. 4 become smaller and smaller as the Q and the flywheel effect are diminished. This means, of course, that second harmonic and other distortion products in the output begin to reach seriously high values as the coupled resistance is increased.

In actual transmitter installations, a compromise must be made. Coupling between the class B amplifier tank circuit and the antenna or dummy load is adjusted so that reasonable output power is obtained, but enough flywheel effect is left to keep distortion below the maximum that can be tolerated. Even then it may be necessary to insert filters in the antenna transmission line to reduce high-frequency, second-harmonic output.

Eliminating the Need for Flywheel Effect. If two tubes in push-pull arrangement are used instead of a single-ended system, we find that conditions are much better. In the first place, just as with class B *audio* amplifiers, we have each tube working on alternate half-cycles. One tube supplies energy during the *positive* r.f. half-cycle, then its plate current ceases, and tube No. 2 supplies energy during the following *negative* half-cycle. Each tube works one-half the time or exactly 180° of each r.f. alternation. Since tube No. 2 is now supplying energy for the *negative* half-cycle pulses instead of allowing the tank circuit flywheel effect to take over, we find that the flywheel effect isn't necessary at all.

With push-pull linear amplifier operation we find it possible to forget the Q of the tank circuit with its accompanying flywheel effect, and proceed to couple the load for maximum linearity and optimum power output and efficiency. We can actually get increased output with greatly reduced even harmonic distortion.

The Q of the tank circuit under these conditions is very low. This leads to an additional circuit improvement we have not hitherto considered.

Side-Band Clipping. As discussed above, the Q of a tuned resonant circuit is a measure of its flywheel effect, Q being inversely proportional to the effective resistance present. Q also is a measure of the *frequency* selectivity of the circuit. Thus, as shown by the curves in Fig. 9, the maximum impedance of a parallel-tuned circuit at resonance, and the general "sharpness" of the selectivity curve are determined by the circuit Q.

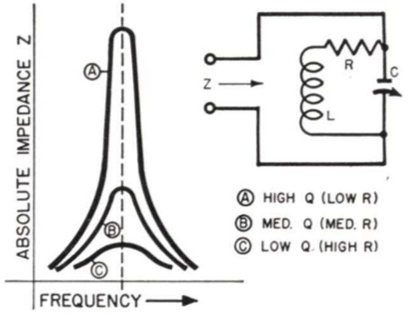

FIG. 9. The selectivity "sharpness" of a tuned tank is determined by the circuit Q; and Q, in turn, is controlled by the effective resistance present.

But let us review the signal that the linear amplifier is amplifying. By mathematics and experiment we know that an amplitude-modulated wave is the same as a carrier frequency plus an upper and a lower set of side-band frequencies. For instance, if a 1-megacycle carrier wave is being modulated by a pure 1000-cycle audio-frequency tone, then the output will be the 1-megacycle carrier, an upper side-band frequency of 1 megacycle plus 1000 cycles, and a lower side-band frequency of 1 megacycle minus 1000 cycles.

Now let us suppose that we attempt to use as load for a class B linear amplifier, a *high-Q* circuit that has a response curve like curve A in Fig. 9. We would find the performance about as indicated in Fig. 10A. The impedance of the tank circuit would be very high at the carrier resonant frequency so power output would be good, and the efficiency high for the carrier without modulation. The circuit is so selective, however, that the load impedance for the upper and lower side bands would be relatively low. We can expect, then, that the side-band frequencies will be "shorted out," very little power being radiated at these frequencies. In addition, the higher the audio-modulation frequency, the greater is the separation between the carrier and the side bands, and the worse this discrimination. The over-all transmission result is audio-frequency distortion by discrimination against the higher frequency components of voice or music. This effect is called "side-band clipping."

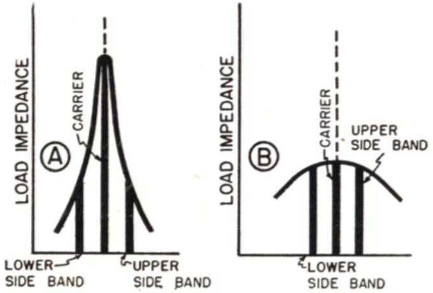

FIG. 10. If the tank circuit of a class B amplifier is too selective, as at left, sideband "clipping" will result because the load impedance for the side bands is too low. A low-Q, broad selectivity curve, as at right, is more desirable because carrier and side bands will receive more nearly equal amplification.

On the other hand, if we now use for the class B linear amplifier load, a *low-Q* circuit with a characteristic like curve C in Fig. 9, performance will be similar to that in Fig. 10B. Although the maximum impedance at the resonant carrier frequency is much lower than with a high-Q circuit, by proper design it still can be made to equal twice the d.c. plate resistance of the amplifier for optimum power output. Because of the very broad selectivity characteristic, the tank circuit impedance is very nearly constant over a wide frequency band. The side bands, therefore, will be transmitted very nearly as well as the carrier itself. Frequency discrimination from side-band clipping will be reduced to a minimum.

In the single-ended class B amplifier, for which at least a medium Q must be maintained for flywheel effect, some side-band clipping is always present. For the push-pull amplifier, the negative half-cycle of the r.f. wave is supplied by the second tube, and not the tank circuit. Tank circuit Q for the push-pull circuit accordingly can be made very low with negligible side-band discrimination. This enhances the high-fidelity characteristics of the transmitter.

► From the foregoing, the push-pull connection of a class B linear amplifier offers these advantages over a single-tube stage:

1. Greater power output.
2. Elimination of need for flywheel effect, allowing more efficient coupling.
3. Freedom from side-band clipping through the use of a lower Q tank circuit.

A TYPICAL PUSH-PULL CLASS B LINEAR AMPLIFIER

A circuit for a push-pull class B linear amplifier, typical of those used in broadcast installations of 1000 watts or more, is given in Fig. 11. The grid tank is comprised of variable inductor L_1 and the capacity of condensers C_2 and C_3 in series. Condenser C_2 is equal to C_3, and the common point between them is grounded through C_8. The tube grids, therefore, are excited 180° out of phase, and the excitation voltages are equal. Condenser C_1 is a coupling condenser used to feed in the driving excitation voltage.

The grid circuit loading resistors necessary for driver regulation are represented by R_1 and R_2. With this particular arrangement of the grid tank circuit, there is no *low-resistance* d.c. return path to ground for rectified grid current. (Resistors R_1 and R_2 are high enough in value to provide considerable unwanted grid bias on the positive modulation swings.) R.F. chokes L_2 and L_3 provide this path.

As the r.f. plate voltages of two tubes operating in push-pull are 180° out of phase, advantage is taken of this fact by using what is called "cross neutralization." Neutralizing condenser C_7, for instance, is connected from the plate of VT_2 to the grid of VT_1. Similarly, neutralizing condenser C_6 is between the plate of VT_1 and the grid of VT_2. When the amplifier stage is properly neutralized, these condensers will be set at the same capacity, approximately, as the grid-plate capacity of each tube

When properly adjusted and in operation, the meter indications of the push-pull class B amplifier are identical to those described for the single-ended stage.

HOW TO ADJUST A LINEAR AMPLIFIER

Let us suppose that we are setting up the single-ended class B linear am-

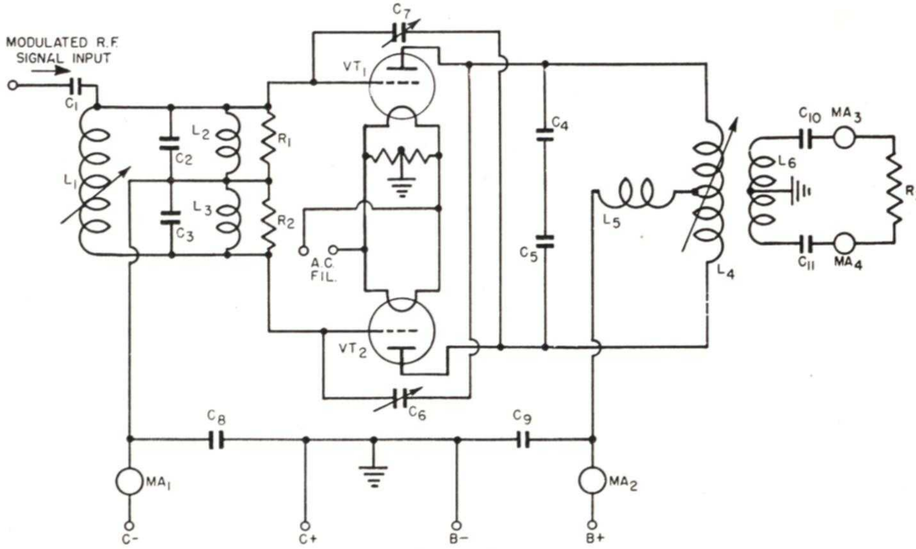

FIG. 11. A typical push-pull class B linear r.f. amplifier that uses variable-inductance fixed-capacity tuning in both grid and plate circuits.

plifier shown in Fig. 8. In adjusting this lass B stage and setting it into operation, the first step to consider is neutralization of the tube grid-plate capacity, just as with the familiar class C amplifier. This is accomplished, you will remember, by reducing the plate voltage to zero, then with the filament lighted and the C bias normal, applying a small amount of r.f. input excitation. With a neutralizing indicator or flashlight bulb with pick-up loop coupled to the output tank coil L_3, the neutralizing condenser NC then is varied until the r.f. present in the plate circuit drops to zero. Neutralization can be checked by rotating the tank condenser C_3 while watching the grid current meter MA_1; proper adjustment, of course, is obtained when the grid current shows no change for any setting of the plate tank condenser.

Use of One-half Voltage Values. The next step is to determine the optimum C-bias adjustment and the best plate load value for maximum linearity and proper power output. This is best done experimentally.

A convenient method is to run the linearity curve using only *one-half* the proper C-bias voltage, and *one-half* the intended plate-supply voltage. Saturation of the class B stage then can be accomplished with only one-quarter the driver power. After good linear operation has been obtained, both voltages can be increased to normal values, and the linearity will remain the same.

Bias voltage equal to one-half the approximate extended cut-off value is first applied. This bias voltage may be calculated by taking one-half the voltage determined by dividing the *full* plate-supply voltage by the tube mu or the tube's amplification factor.

Now, a low value of *unmodulated* r.f. excitation should be provided. This is increased slowly by tightening the coupling between the link coils and tank coils L_1 and L_2 (Fig. 8) until grid current starts to flow through meter MA_1. Both driver tank and amplifier grid tank circuits should be re-adjusted to resonance by tuning each for maximum grid current. Plate voltage of one-half value is next applied, and the plate tank *quickly* tuned to resonance as indicated by a sharp minimum reading of plate meter MA_2.

The load circuit is now coupled to the amplifier output by slowly increasing the coupling between coils L_3 and L_4. At each step, the load condenser C_4 should be adjusted for maximum load current on MA_3. The tank condenser C_3 will also need re-adjustment each time for a minimum reading on MA_2. This is continued until the plate current minimum is about three times greater than that minimum without load.

Check of Meter Readings. Check the plate current reading on MA_2; this current multiplied by the *full* normal plate voltage should give the proper maximum d.c. input power recommended for the amplifier tube.

If the readings on MA_2 and MA_3 are too high, then excitation is excessive and should be reduced accordingly. If only MA_2 reads high and MA_3 reads low, perhaps the load coupling is not sufficient. In this case, reduce the excitation, increase the load coupling, and then again increase excitation. If excessive plate current still persists, it is possible that the C bias is too low. Bias should be increased by 10% to 20%.

Optimum C Bias for Linearity. Once satisfactory input and output power values for the class B stage have been obtained, optimum C bias for linearity can be found, if the correct value is unknown, by plotting load current against grid excitation for various bias values.

Curves should be run for two or three bias values slightly above and below the estimated cut-off point. One of these curves will be straighter than the others, and represents the bias voltage to be used.

If maximum linearity is achieved for a C bias not very different from the calculated value, and proper input and output power is being obtained at full excitation, the linear amplifier can be considered in proper adjustment. On the other hand, if excessive plate current is being drawn, the load impedance is too low. In the reverse condition, too high a load impedance will result in power input and output below normal. In either case, the load coupling should be reduced or increased as required, and a new set of linearity curves run. *Each different value of load impedance will have a different optimum C-bias value.* One combination of these will give the best amplifier adjustment.

Applying Full Power. After proper adjustment of the class B amplifier has been completed, the one-half values of C bias and plate supply voltages are removed, and *full* normal operating voltages applied. Grid, plate, and output currents should be very nearly the same as before. If the driver stage is now modulated for the first time, all amplifier meter reading behavior should be as described earlier, the grid current changing rapidly with modulation, the plate current always remaining steady, and the load current flickering or showing the 22.5% rise, depending upon the modulation signal quality.

Other Performance Checks. In addition to the plate current meter indications, final operating performance of the linear amplifier can be observed further by means of an oscilloscope or a carrier shift indicator exactly as is done for a class C transmitter stage.

If no adjustment of the linear amplifier results in proper power output without distortion, it is advisable to check the driver stage regulation. If poor regulation is the case, then either the driver power must be increased, or the amplifier grid tank loading resistance must be decreased, or both.

A High-Efficiency Linear R.F. Amplifier

We have already found that the typical class B amplifier has a low *average* efficiency of about 40%, but on modulation crests when excitation is high, the instantaneous *peak* efficiency rises and approaches the high value of say, 70% near saturation. Now if we could run the typical class B amplifier near saturation at all times, its efficiency would be high. We could do this for the unmodulated carrier alone, but saturation would set in for positive modulation crests and the output power would be very distorted.

Suppose, however, that we use two tubes instead of one, and that we arrange them so that tube No. 1 ordinarily works near saturation for carrier alone, tube No. 2 being inoperative; and then for positive modulation peaks, although tube No. 1 begins to saturate, tube No. 2 begins to function so that additional power is added to keep the output power rising in a linear manner. The output power then would not be distorted and each tube when performing would be operating at high excitation levels and giving good power efficiency.

Such a system has been devised and in practical form is called the "Doherty High-Efficiency R.F. Amplifier."

THE DOHERTY AMPLIFIER

In Fig. 12 is given a simplified version of the Doherty linear amplifier. Note that both tubes VT$_1$ and VT$_2$ are driven by the same modulated driver stage. Tube VT$_2$ feeds power into the resonant output tank circuit L$_2$-C$_3$ by being connected directly across it. VT$_1$ supplies power to the same load through the quarter-wave phase-shifting section L$_1$-C$_1$-C$_2$. The antenna pick-up coil L$_3$ is inductively coupled to the plate tank.

Resting Operation. Although not shown here, VT$_1$ and VT$_2$ may have equal plate supply voltage, but very different C-bias voltages are used. VT$_1$ is biased so that for excitation *without modulation* it operates close to the saturation level, and delivers its maximum output voltage at high efficiency. VT$_2$ is given a greater bias, and under no-modulation conditions, draws little or no plate current.

Modulated Excitation. Now with modulated carrier excitation, as the input signal increases in a positive direction, tube VT$_1$ begins to saturate, and its output voltage levels off. The bias of VT$_2$, however, is just beginning to be overcome by the increasing positive grid voltage, and the VT$_2$ plate current starts to rise in proportion to the signal. Finally, at the excitation peak, VT$_2$ supplies half the power to the load.

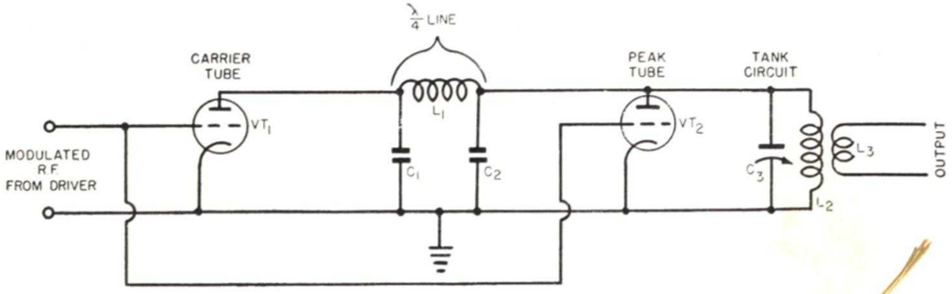

FIG. 12. Simplified version of the Doherty high-efficiency linear r.f. amplifier.

On a negative modulation half-cycle, the modulated r.f. excitation decreases. Being cut off with high grid bias, tube VT$_2$ remains inoperative and supplies no power at all. VT$_1$, however, operates in a true linear class B fashion, and although near saturation for the unmodulated carrier, it can *decrease* its power output in a straight-line manner.

The Peak and Carrier Tubes. We see then, that VT$_1$ supplies the average carrier power and lesser power needed during negative modulation swings, while VT$_2$ adds additional power only for the positive modulation surges. Because of their peculiar manner of operation, tube VT$_1$ is commonly called the "carrier tube," since it supplies the carrier or "resting" energy, and tube VT$_2$ is named the "peak tube" because of its ability to provide additional power for modulation peaks.

The voltage wave forms present on the plates of each tube, with and without modulation, will be similar to those in Fig. 13. Notice that for zero modulation the carrier tube VT$_1$ is supplying nearly all carrier power. Observe, too, that for 50% modulation the carrier-tube voltage never rises, the peak tube VT$_2$ instead delivers the additional power; and for negative modulation voltage the peak tube VT$_2$ cuts off, leaving the reduced power to be handled by the carrier tube alone. The same conditions are true to a greater degree for 100% modulation.

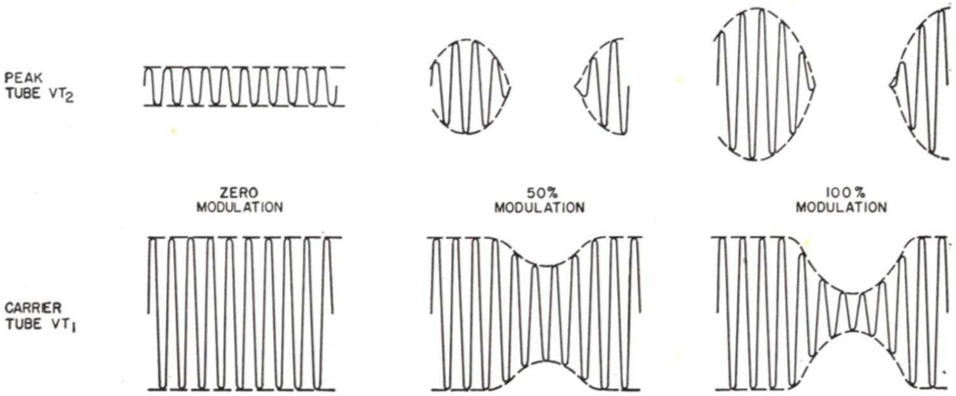

FIG. 13. The approximate voltage wave forms present on the plates of carrier and peak tubes in a Doherty amplifier for different degrees of modulation.

Up to this point, we have seen that the peak tube supplies half the energy to the load for positive modulation peaks. At first glance, it would seem that VT$_2$, operating "in parallel" with VT$_1$, would give only *twice* the peak power obtainable with one tube alone. But for 100% modulation we know that the peak power must be *four* times the average carrier power. Where does this extra power come from if the peak tube does not supply it? The answer is *the carrier tube VT$_1$ also increases power because its load impedance is not constant.*

Action of the Quarter-Wave Line. If we look back at Fig. 12 we see that the carrier tube works directly into the quarter-wave phase-shift network L_1-C_1-C_2.

It is a peculiar property of such a network that the impedance at one end is always inversely related to the impedance used as load at the other end. In other words, the behavior of the peak tube VT_2 has a pronounced effect on the load impedance for the carrier tube VT_1. If the effective network impedance determined by VT_2 should *rise*, then the load impedance presented to VT_1 *drops;* conversely, if VT_2's action makes the impedance across the phase-shift network *drop*, then the plate load impedance for tube VT_1 is *increased* accordingly.

And what is the end result? First, with unmodulated excitation, VT_2 is not operating, and the impedance present at the end of the phase-shift network is of moderate value. The network is designed, under these conditions, to present to the carrier tube VT_1 just *twice* as great a load impedance as that optimum value needed for maximum output. The carrier tube, therefore, delivers only *half* as much power as it could under better conditions, but its efficiency is very high and its linearity good.

When modulation is applied, the peak tube VT_2 begins to operate on the positive modulation swing. In so doing, by supplying energy at the end of the phase-shift network, the effective impedance seen by the network is *increased*. Because of the inverse impedance relation inherent in the network, the plate load impedance for VT_1 correspondingly *drops*. In proper operation, this load impedance will continue to drop until, for a modulation crest, it is *half* the value for the unmodulated radio-frequency carrier signal alone.

With its load impedance halved, carrier tube VT_1 draws twice as much plate current and consequently delivers twice as much power as before. The peak tube VT_2, however, also is delivering the same amount of power under these conditions. This corresponds to a total instantaneous peak power of four times the resting or carrying power—a necessary condition for distortionless transmission.

Undesired Phase Shift. The fact that the quarter-wave phase-shift network also shifts the current phase by 90° in addition to changing the load for carrier tube VT_1 is an undesirable one. It is a natural characteristic of the network, however, and cannot be avoided.

Power delivered by the carrier tube, nevertheless, will not add properly to the power delivered by the peak tube unless the two load currents are in phase. To offset the resultant phase shift in the carrier tube network, it is therefore necessary either to introduce additional phase shift in the *opposite* direction in the *carrier tube* grid circuit, or to introduce phase shift in the *same* direction in the *peak tube* grid circuit. Either system is satisfactory.

PRACTICAL DOHERTY CIRCUIT

A representative version of a practical Doherty amplifier is given in Fig. 14.

Power excitation from the driver stage tank L_1-C_1 is supplied by the link coupling L_2-L_3 to the resonant grid tank L_4-C_2. Full excitation voltage is fed directly to the peak tube VT_2. Since the carrier tube VT_1 usually requires less grid voltage, its excitation is secured by tapping down on the grid tank inductance.

The Grid Phasing Network. Note that the carrier tube excitation is fed through the grid phasing circuit C_4-L_5-L_6. This introduces a *leading* current phase of 90° to offset the *lagging* phase of the quarter-wave plate line so that power from the two tubes will add properly. The resistor R_1 is used to terminate the grid phasing circuit; this resistance also imposes a more constant load on the driver, thereby improving regulation.

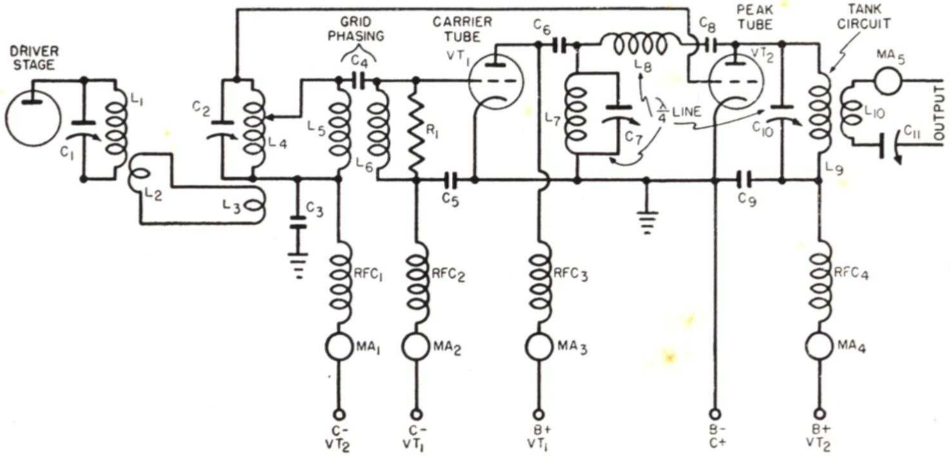

FIG. 14. A practical Doherty high-efficiency amplifier which gives an over-all efficiency, with or without modulation, of about 70%.

The Quarter-Wave Line. In Fig. 12, condenser C_2 of the quarter-wave line actually is in parallel with the tank condenser C_3. In the practical circuit of Fig. 14, this quarter-wave line condenser is omitted; it is replaced by the tank C_{10}-L_9. The quarter-wave line then is composed of L_7, C_7, L_8, L_9, and C_{10}. The resonant circuits L_7-C_7 and L_9-C_{10} are made capacitive by detuning these circuits a slight amount.

Other Components. The resonant circuit L_9-C_{10}, in addition to being part of the quarter-wave line, is also the power output tank circuit. It is inductively coupled to the antenna or load circuit through inductance L_{10}. Condensers C_6 and C_8 are simply blocking condensers, and if they are of sufficiently large capacity, they have no effect on the operation of the quarter-wave line.

Plate supply shunt feed is used for the carrier tube VT_1, series feed for the peak tube VT_2. Provision is made for two separate high-voltage supplies, although operation with a common supply is feasible. Plate currents of VT_1 and VT_2 are indicated on milliammeters MA_3 and MA_4, respectively. Two separate C-bias supplies are indicated, and in general, it will be found that the peak tube VT_2 requires a much higher bias than the carrier tube VT_1. Rectified grid currents for each tube flow through their respective milliammeters MA_1 and MA_2.

Adjustment Procedure. In setting up and tuning the circuit, the first step is to adjust the quarter-wave phase-shift line. This is done by first shortcircuiting the tank circuit L_9-C_{10}, then with plate supply and excitation applied only to VT_1, adjusting the condenser C_7 for minimum plate current on MA_3. In like manner, the resonant circuit L_7-C_7 is next shorted, plate supply and excitation applied only to VT_2, and the tank circuit L_9-C_{10} tuned for minimum plate current on MA_4. If the inductance L_8 has the proper value, this procedure not only provides proper tuning, but makes the phasing line exactly one-quarter wave long.

Proper exciting and C-bias voltages next must be determined. As a first step, the peak tube VT_2 is made inoperative. The bias and excitation of the carrier tube VT_1 are then adjusted so that for *no modulation* VT_1 is just approaching saturation. With VT_2 now turned on, its separate bias and excitation are adjusted so that this peak tube is just beginning to show plate current. Excitation that is twice the carrier level is then applied, VT_2 should operate in a normal manner, and deliver half the peak power to the load. In general, the average plate current of VT_2 for peak excitation will be about 80% of the plate current drawn by VT_1.

Average Efficiency. As both tubes operate principally under their most efficient conditions, the over-all efficiency of the circuit will be high, approaching usually 60% to 70%. And furthermore, this high efficiency is almost constant, changing very little from no-modulation to full-modulation conditions. This is almost double the average efficiency of the typical class B linear amplifier described earlier.

Inverse Feedback in Radio Transmitters

In any radio transmission system, the prime purpose is to deliver to the receiver readable signals that should be perfect duplicates of those originating at the transmitter microphone, facsimile scanner, or TV camera. Unfortunately, the vacuum tube is not a perfect device. Any audio-frequency amplifier, r.f. amplifier, detector, modulator, etc., always has a certain amount of distortion and noise present in its output.

Any means we can use to reduce these spurious disturbances will improve the performance of our electronic equipment. The application of inverse feedback is extremely effective in this respect. Although this Lesson is concerned primarily with inverse feedback in transmitters, since all electronic equipment has many characteristics in common, it should be understood that the principles involved can be applied to electronic devices in general.

WHAT INVERSE FEEDBACK IS

As the word suggests, "feedback" means taking part or all of the output of an amplifier, and feeding it back into its own input circuit. The means we use to do this, however, will determine what the results will be. The most important factor we have to consider is the phase relationship between the original input signal and the output signal fed back.

Regeneration. Suppose we feed back output energy so that it is *in phase* with the input signal. We would expect then, since they rise and fall together, that the output voltage would aid and increase the input voltage. Such is the case. To go a step further, if the feedback voltage is gradually increased, the resultant output will get larger and larger, until finally the amplifier will be entirely independent of the original excitation, and promptly burst into oscillation.

This feeding back of voltage *in phase* with the input is commonly called regeneration or *positive* feedback. It is used, of course, in all oscillators, and is useful, too, in the regenerative detector. In general, however, *amplifiers* with regeneration are noisy, usually unstable, hard to control, and have bad frequency discrimination characteristics—something to avoid in wide-band amplified circuits.

Inverse Feedback. Now suppose that we take the amplifier output voltage but feed it back into the input circuit so that it is 180° *out of phase* with the input signal. This means that whenever the input voltage is rising in a positive direction, the feedback voltage is rising in a negative direction. What happens? Since the feedback voltage is now "bucking" instead of aiding the input voltage, we can expect the resultant input voltage to be reduced. This is true. The over-all gain of the amplifier, consequently, will be correspondingly reduced.

We find, however, that certain benefits are obtained. First, factors tending to change the gain of the amplifier, such as variations in filament or B-supply voltage, will not be as effective as before. Although lower, the gain is more constant; and the more we increase the feedback, the lower the gain will become, and the more constant it will be. Furthermore, we find that all types of distortion, noise, and hum normally present in the amplifier also are reduced in proportion to the amount of feedback we use.

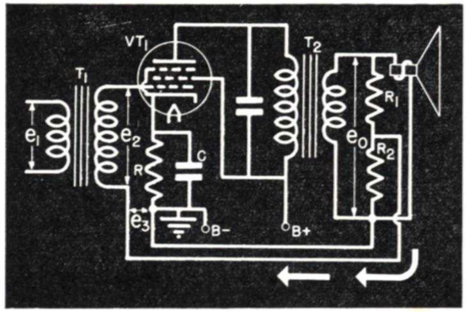

FIG. 15. Basic inverse feedback circuit.

This process of feeding back output voltage out of phase with the grid voltage is called degeneration, negative feedback, or *inverse* feedback. Although it always involves a sacrifice in amplifier gain, better stability, improved frequency response, and less noise and distortion are the characteristic results. Let us examine the process in greater detail.

HOW INVERSE FEEDBACK WORKS

In Fig. 15 is given a basic single-stage feedback circuit. Except for minor changes, it is a conventional triode power amplifier. Although it is shown here driving a loudspeaker, the load could just as well be a low-impedance transmission line, or the grid circuit of a following amplifier. The action is the same.

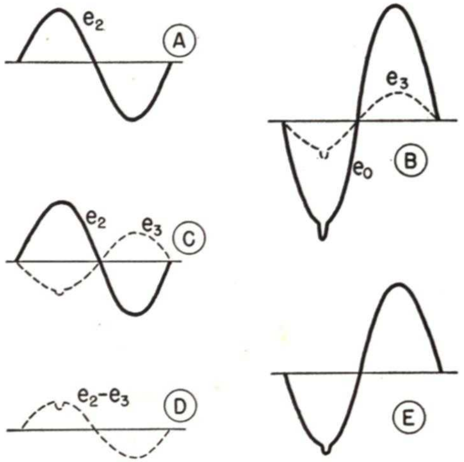

FIG. 16. Curves showing result of inverse feedback.

Observe that the lower end of the secondary of transformer T_1 is not grounded as customarily, but is connected to the voltage divider R_1-R_2 across the load winding of output transformer T_2. (R_1 and R_2 in series are much higher in resistance than the loudspeaker so that they absorb very little power from the circuit.) With this arrangement, part of the output voltage is fed back into the grid circuit; the exact amount of voltage actually fed back can be adjusted by changing the values of divider resistors R_1 and R_2.

Now what occurs as an input signal is applied to the basic circuit? First, the input voltage e_1 is stepped up by the input transformer to e_2, which after being amplified, appears in the output circuit of the amplifier as e_o.

Let us assume now that our input voltage e_1 is a pure sine wave such as that in Fig. 16A. Suppose also, that instead of a sine-wave output we get for e_o something like Fig. 16B. The "bump" on the curve may represent noise, hum, distortion, or all three as they are generated within the tube itself. Certainly, the bump was not present in our original input signal, and represents something that should be eliminated.

By means of the feedback divider circuit, a portion of e_o is fed back into the grid circuit in series with e_2. This feedback voltage is designated as e_3 and is shown in dotted lines in Figs. 16B and 16C.

We can assume for our purpose that all voltages are generated and appear instantaneously; there is no time delay between the instant e_2 is applied to the tube grid, and the time e_3 is generated and fed back into the input circuit.

If we are careful to connect the windings of output transformer T_2 correctly, feedback voltage e_3 can be made to arrive at the grid exactly 180° out of phase with the input voltage e_2. This is the essential requirement for inverse feedback. The phase relations are shown in Fig. 16C.

Since e_2 and e_3 are of opposite polarity, they oppose or "buck" each other. The resultant input voltage, therefore, is actually equal to their difference. This is illustrated by subtracting e_3 from e_2, and obtaining the resultant input voltage e_2-e_3 in Fig. 16D.

The output voltage will be an amplification of this new reduced input voltage, and is represented in Fig. 16E. This obviously, is a much better reproduction of the original input than that obtained in Fig. 16B—the "bump" is far less prominent.

By means of inverse feedback we have taken noise or distortion products present in the output, and have fed them back into the input circuit in such a manner that they tend to cancel themselves.

Of course, the power output of the amplifier stage and its over-all gain have been reduced. This is a direct result of the input voltage being cancelled to some extent by the feedback voltage, and is the price that must be paid for better signal-to-noise ratio, or lower per cent harmonic distortion. The gain can never be restored, but if the input voltage is increased, the power output can be brought back to its former level, leaving distortion and noise products substantially reduced.

Improvement of Frequency Response. Suppose we have a poor amplifier with a frequency-gain characteristic curve like curve 1 in Fig. 17. The bad drop-off at low frequencies may be caused by coupling condensers that are too small, or by transformers with insufficient primary inductance. The high peak at 3000 cycles may be due to resonant effects in the same transformers, and the poor response above 10,000 cycles may be due to shunting effects of tube capacities. Such an amplifier would not give high-fidelity reproduction as a result.

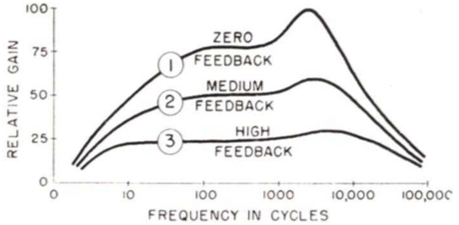

FIG. 17. Curves showing how inverse feedback improves frequency response.

Now if we arrange the amplifier circuit to feed back to the input in proper phase a medium amount of the output voltage, we will get a new response like curve 2 in Fig. 17. This shows considerable improvement. Note that the over-all gain has been reduced at every point. The gain at the peak at 3000 cycles, however, has been cut more than anywhere else; also, the gain at extremely low or extremely high frequencies has been cut less.

How this is brought about may be visualized as follows: Below 100 cycles and above 10,000 cycles, the amplifier's gain without feedback was low. This tended to result in a low output voltage. Since the feedback voltage is proportional to the output voltage, the feedback voltage for these frequencies also was low, and did not cancel the input voltage as much as it might have otherwise. For the peak at 3000 cycles, the amplifier's inherent gain was high, and normally would result in high output voltage. In tending to make the output voltage high, however, the feedback voltage also is increased, and this in turn cancels more of the input voltage, thereby cutting the effective gain considerably.

Even greater improvement in frequency response can be obtained if the feedback voltage is increased in magnitude. See curve 3 of Fig. 17. Observe, however, that additional feedback is always accompanied by a further decrease in gain. In general, feedback can be increased as much as desired so long as sufficient driving voltage for satisfactory output can be supplied.

Improvement of Stability. The effects of any factor that tends to decrease or increase the gain of an amplifier can be reduced by inverse feedback. Feedback, for instance, is quite effective in minimizing gain variations due to filament or B-supply voltage changes.

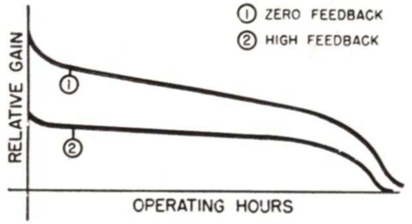

FIG. 18. Effect of inverse feedback in reducing change of gain with B battery voltage decline.

As an example, the over-all gain of a battery-operated amplifier rapidly decreases as the filament and B-battery voltages drop, and the batteries exhaust themselves. The variation in gain of such an amplifier is about like curve 1 of Fig. 18. This is a direct result of battery-voltage decay and cannot be avoided.

If, however, we re-arrange the amplifier to include inverse feedback, and adjust the feedback value until the gain with fresh batteries is about one-half its former value, we will get a gain versus operating-hours characteristic similar to curve 2 of Fig. 18. The gain now, though only half its former value, is very nearly constant throughout battery life. At a cost of one-half the gain we have improved the operating stability by negative feedback.

HOW INVERSE FEEDBACK AFFECTS GAIN

The actual voltage gain or amplification obtained from an amplifier after negative feedback has been applied is determined by the amount of output voltage fed back, and the gain of the amplifier before applying feedback. This is usually expressed in equation form as:

$$\text{Voltage amplification with inverse feedback} = \frac{A}{1 - A\beta}$$

where A is the voltage amplification without feedback, and β is that fraction of the output voltage fed back into the input circuit. When the feedback is out of phase with the input so that degeneration occurs, β is considered as a *negative* quantity, so this formula can be written as $\frac{A}{1 + A\beta}$ for inverse feedback.

A, of course, is some number like 10, 50, 200, or 1000—whatever the amplifier gain without feedback may be. β, however, is a fraction like 0.02, 0.08, 0.3, or 0.9, and is always less than one. β can never be greater than unity for it is the ratio between feedback voltage and full output voltage, and we cannot feed back more voltage than is actually present in the output.

Let us take a few examples to see what this equation really means.

Case I. Let's assume that we have an amplifier with a gain of 100 without feedback, and we feed back inversely 0.01 or 1% of the output into the input. What will be the resultant voltage amplification?

We have, therefore: $A = 100$, and $\beta = 0.01$.

The gain with feedback then will be:

$$\frac{100}{1 + (100 \times 0.01)} = \frac{100}{1 + 1} = 50$$

We find that the gain is cut to 50 or one-half its former value.

Case II. A second amplifier has an original gain of 1000 and we still use 1% inverse feedback. What will be the result?

Here we have: $A = 1000$, and $\beta = 0.01$, and the gain with feedback will be:

$$\frac{1000}{1 + (1000 \times 0.01)} = \frac{1000}{1 + 10} = 91$$

It is clear that *the higher the gain without feedback, the more this gain will be reduced upon application of a given amount of feedback.*

Case III. Now let us investigate what will happen to the two cases above if we use a *very large amount* of inverse feedback, 50% for instance.

For our first example where gain without feedback was 100, we now have:

$A = 100$, and $\beta = 0.5$, and the gain will be:

$$\frac{100}{1 + (100 \times 0.5)} = \frac{100}{1 + 50} = 1.96$$

An original gain of 100 without feedback is dropped to slightly less than 2.

Case IV. In Case II where gain without feedback was 1000 and we now use 50% feedback, we have:

$A = 1000$, and $\beta = 0.5$

Resultant gain for this condition is:

$$\frac{1000}{1 + (1000 \times 0.5)} = \frac{1000}{1 + 500} = 1.99$$

And *again* we find that the gain with such high feedback is slightly less than 2.

This brings to light an interesting characteristic: *For large amounts of inverse feedback, the resultant gain is determined principally by the feedback ratio, and is almost entirely independent of the original gain of the amplifier.*

REDUCTION OF NOISE, HUM, AND DISTORTION

Any noise, hum, or distortion developed in the stages employing inverse feedback will be reduced the same amount as the gain. Thus, if this gain with feedback is only 25% of that without feedback, then the distortion, etc. will be only 25% of the former value also.

Now if we assume in Case I (described above) that the distortion without feedback was 20%, then with feedback this distortion will be reduced by one-half or 50% along with the gain. The resultant distortion then will be only 10%. Furthermore, if we now double the input excitation voltage to make up for the loss in gain, the original full power output will be obtained with *only half as much distortion as before.*

For Case II, the gain was reduced by approximately 90%. However, distortion, hum, and noise also are reduced to about 10% of their original values. It will now take ten times as much input excitation voltage to bring the power output up to normal, but if we can supply this increased voltage, we will find the distortion products still reduced by approximately 90%. If the distortion originally was about 20%, with feedback it will be only 1.81%.

AMPLIFIER PHASE SHIFTS

In all discussions of inverse feedback, we have made the assumption that the feedback voltage always arrives at the grid exactly 180° out of phase with the input voltage. Unfortunately, this is not always the case.

In any amplifier, the coupling-condenser-grid-leak networks or the interstage transformers can introduce undesired phase shifts. These phase shifts usually are small for medium frequencies, but for extremely high or extremely low frequencies they can become quite serious. In fact, for extreme frequencies the phase shift per stage can approach plus or minus 90°. To make matters worse, in any amplifier, the phase shift in one stage will add to that in another, the over-all phase shift between input and output being the sum of the phase shifts in each stage.

If we are trying to incorporate inverse feedback in a three-stage amplifier, we may find that the phase shift per stage at *medium* frequencies is so low that it is not effective, and the system performs satisfactorily. At extremely high or low frequencies, however, the phase shift per stage may increase enough to change the over-all phase shift by more than plus or minus 180°. If this occurs, and the amplifier gain is sufficient at these frequencies, a very undesirable situation may develop.

First, the feedback voltage, instead of arriving at the grid 180° *out of phase* and thereby opposing the input voltage, now arrives *in phase* with the input, and actually increases or aids it. *Because of low- or high-frequency phase shift within the amplifier itself, inverse feedback has been changed to positive feedback or regeneration.*

We can expect all the bad effects of positive feedback. If the gain of the amplifier is high enough at these extreme frequencies to overcome the losses in the circuit, it will quickly burst into oscillation, or "motorboat."

How is this condition taken care of by our basic equation for amplification with feedback? If the internal phase shift of an amplifier becomes plus or minus 180°, the algebraic sign of the feedback ratio β is changed from negative to positive. This changes the sign in the denominator of the basic equation. The denominator now becomes a *difference,* not a *sum.* For *positive* feedback, the amplifier gain now is:

$$\text{Voltage amplification with } \textit{positive} \text{ feedback} = \frac{A}{1 - A\beta}$$

Where A is low, or β is so small that the product $A\beta$ is far *less* than unity (1), no trouble is encountered, as the resulting voltage amplification, even with positive feedback, is not excessive. However, as either A or β increases so that their product becomes nearly equal to one, when this product is subtracted from one, the denominator gets smaller, approaching zero. As the denominator gets smaller, the resultant voltage gain gets larger. Thus, positive feedback can increase the gain to a value far higher than the initial value of A. When the resultant voltage gain gets high enough, the amplifier will oscillate.

For medium frequencies and a given amount of inverse feedback, the undesired phase shift in each amplifier stage usually is so low that we can make the amplification A as great as we please. At very high or low frequencies, however, these extra phase shifts begin to be effective. Since we can do nothing to eliminate these phase shifts, in order to keep the amplifier stable we must be sure that the gain goes down rapidly with changing frequency. In this way, when inverse feedback is finally changed to positive feedback by the internal phase shift, the feedback voltage is made so small that oscillation cannot take place. This corresponds to making A decrease as an extremely high or low frequency is approached, thus keeping the product $A\beta$, for undesired positive feedback, always much less than unity.

Typical Inverse Feedback Circuits

There are many ways to obtain inverse feedback. In most systems, the feedback loop circuit may be made to include one, two, or more stages of amplification. Feedback around a single stage, however, is the simplest, and since the maximum high or low frequency phase shift possible with a single coupling circuit is 90°, this arrangement usually gives no motorboating trouble.

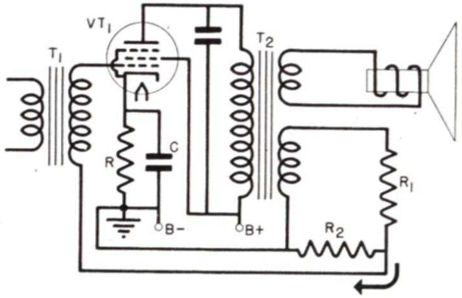

FIG. 19. Circuit using extra winding for feedback.

It is seldom necessary that the total output voltage be fed back to the input, and for this reason most feedback circuits use a voltage divider arrangement of some sort. In the basic feedback circuit of Fig. 15, the divider is represented by the resistors R_1 and R_2. In the circuits shown in Figs. 19 to 21, the divider resistors also are marked R_1 and R_2. Output voltage in each case is applied to R_1 and R_2 in series, and the feedback voltage is that fraction of the output developed across R_2 alone. The feedback ratio can be regulated by varying the resistance values.

In the one case of Fig. 15, however, because of the transformer step-down ratio and the low impedance of the speaker or transmission line, sometimes sufficient feedback voltage cannot be developed across the load. To overcome this, an extra winding with a greater number of turns can be added to the transformer as in Fig. 19. This increases the available feedback voltage with correspondingly better results.

In Fig. 20, high feedback voltage is not taken from a transformer secondary winding; instead, it is obtained directly from the plate of the tube. The condenser C_1 serves only as a blocking condenser to prevent shorting the d.c. plate voltage, and to prevent application of the plate voltage to the tube grid. The capacity of C_1 should be high enough so that the reactance of the condenser is small compared to the resistance of R_1 and R_2, otherwise the feedback will not be constant for different frequencies.

Feedback voltage may be fed in *parallel* with the input as well as in series. This is illustrated in Fig. 21. With resistance-coupled stages this forms a very convenient method. C_1, as before, is merely a blocking condenser.

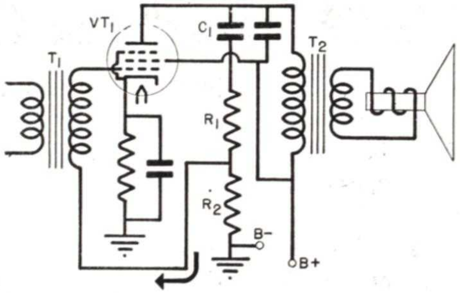

FIG. 20. Circuit using a plate voltage divider.

CURRENT-CONTROLLED FEEDBACK

Up to this point, all the circuits shown have been of the *voltage-control* type, that is, the feedback voltage has been a certain fraction of the output or load *voltage*. In fact, we used a voltage divider to adjust the feedback voltage to the proper value. With this method, it is the *voltage* gain of the amplifier that is stabilized; and it is the output *voltage* wave form that is improved by reduction of distortion.

It is just as feasible to get output *current* stabilized by inverse feedback. This is done by making the feedback voltage proportional to the output *current* instead of the voltage. In so doing, it is the output or load current that is stabilized and rendered more constant for a given value input voltage.

Let us examine the circuit in Fig. 22. This is another basic feedback circuit. Note that no voltage divider is used across the output transformer secondary winding. Instead, a feedback resistor R_2 is placed in series with the load resistor R_3. The load current i_o, therefore, flows through both resistors, developing the output voltage e_o, and the feedback voltage e_3. In turn, e_3 is fed back into the input circuit to "buck" the original input signal e_2.

Usually, R_2 is much smaller than the load resistor R_3 so that little power is wasted in the feedback resistor. The exact amount of feedback voltage e_3 can be adjusted by varying the resistance values. Since the feedback is proportional to the output current, Fig. 22 is the basic circuit for *current-controlled* feedback.

Cathode-Resistor Feedback. In Fig. 23 we have a special form of current feedback. Except for the omission of the cathode by-pass condenser, this is a conventional voltage amplifier. Without the by-pass condenser, the plate-load impedance is not only that due to the output transformer but also the cathode resistor R_2 as well, and a small part of the output voltage e_3 is developed across this resistor. Since the plate voltage is 180° out of phase with the input voltage, e_3 is out of phase with e_2 and an inverse feedback is obtained. Since the same plate current flows through the feedback resistor R_2 and the primary winding of the output transformer, the feedback voltage is proportional to the output current. It is current feedback effects, therefore, that are obtained.

This is another very convenient method of introducing feedback into an amplifier. Furthermore, it can be applied to each stage individually in a multi-stage amplifier without danger of phase shift difficulties. A disadvantage of this system will be brought out later.

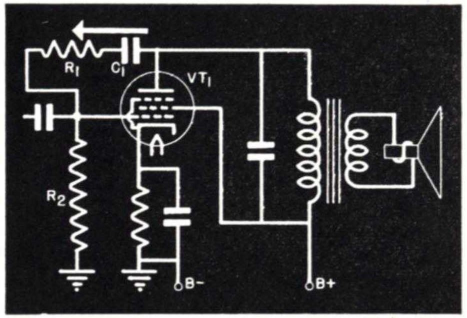

FIG. 21. A parallel feedback circuit.

MULTI-STAGE FEEDBACK

In any multi-stage amplifier it is theoretically possible to introduce feedback by coupling part of the output voltage directly to the input circuit, thus bridging the feedback loop across any number of stages. With resistance or impedance coupling, however, feedback should include an *odd* number of stages. This is imperative because each stage shifts the phase by 180°, and it is necessary that the feedback voltage arrive 180° out of phase with the input signal. An *even* number of stages would shift phase by multiples of 360° or 0°, and would result in regeneration. With transformer coupling in one stage of the amplifier, any number of stages may be bridged, odd or even, for connections to one transformer winding can be reversed, if need be, for proper phasing.

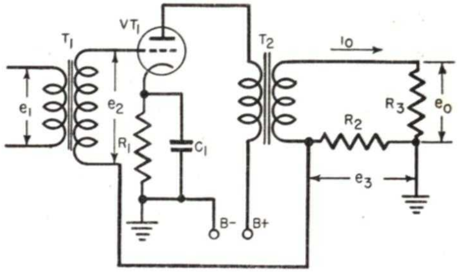

FIG. 22. Basic current-controlled inverse feedback circuit.

In general, the more stages included in the feedback circuit, the more difficult it is to achieve a stable amplifier system. This is due, of course, to the additive phase shifts in each coupling network that change inverse feedback to positive feedback or regeneration for extremely high or low frequencies. Unless carefully done, feedback around more than two stages usually results in a pronounced tendency to "motorboat."

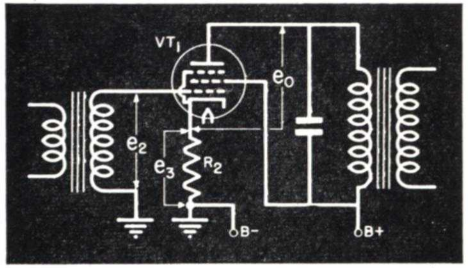

FIG. 23. Cathode degeneration circuit.

Interstage Feedback. Because of the natural phase characteristics that must be considered, and the increased difficulty in obtaining stability, feedback over a complete multi-stage amplifier is not often attempted. An easier method is to divide the amplifier into sections of one or two stages each, and incorporate feedback around these individually. In this way each section is stabilized, and the over-all performance improved.

For feedback over two stages, the circuit in Fig. 24 is often used. Observe that the feedback voltage is inserted in the cathode circuit. Feedback could not be applied to the grid of the first tube because we have a phase shift of 180° in each tube, and with two tubes this makes a total shift between input and output of 360°. The voltage at the grid of the first tube, therefore, is exactly in phase with the output voltage, and coupling the feedback to the first grid would result in positive feedback.

Feedback, however, can be applied to the grid of the first tube in a two-tube amplifier section if an output transformer winding is used. Fig. 25 is an example of this. In this case, the connections to one transformer winding may be reversed, if necessary, for proper phasing. Resistor R_3 serves only to load the transformer winding properly, and has no direct effect on the feedback voltage.

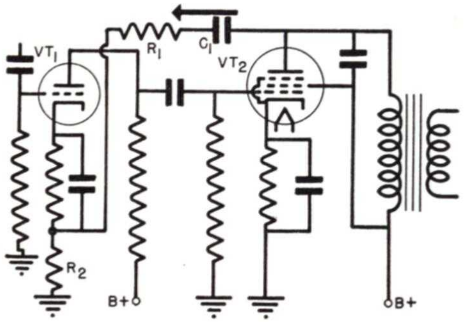

FIG. 24. Plate to cathode feedback over two stages.

AMPLIFIER OUTPUT IMPEDANCE

Since feedback has such a pronounced effect on the performance of an amplifier, it may be expected that it would change the effective characteristics of the vacuum tubes. It does exactly that.

Voltage Feedback Lowers the Plate Resistance. In any *voltage*-controlled feedback circuit, one of the effects of feedback is to *lower* the apparent plate resistance of the output tube. This is true because the feedback tends to keep the output voltage constant regardless of what the load impedance and load current may be. In other words, voltage feedback endeavors to make up for the voltage that is lost in the tube generator itself, and in so doing, lowers the effective tube resistance.

Current Feedback Raises the Plate Resistance. Conversely, *current*-controlled feedback results in an apparent *increase* of output tube plate resistance. In this arrangement, current feedback tends to keep the output current constant in spite of load impedance changes. The tube plate resistance, as a result, is raised so that the load current is more independent of load values.

Both Types of Feedback Reduce Distortion and Noise. Broadly speaking, voltage and current-controlled feedback are equally effective in reducing distortion, hum, and noise. Where the output load impedance is constant and the difference in apparent plate resistance is taken into account, either type of feedback may be employed with about equal results. This means, for example, in the initial stages of an audio amplifier system, either voltage or current feedback can be used effectively.

Where the load impedance is *not* constant, however, such as with an amplifier driving a loudspeaker, or the grid circuit of a class B audio amplifier, the two feedback methods will not perform identically. Current feedback minimizes changes in the current amplitude, but this does not necessarily correct the frequency response, because the *same* value of current at different frequencies can cause different load voltages if the load impedance changes with frequency.

For example, if current feedback is applied to an amplifier with loudspeaker load, it generally is found that, due to the varying load impedance, high frequencies are accentuated and low frequencies are suppressed. This is not a desirable situation, hence, voltage feedback more often is used instead because it can be made to operate directly from the load, and hence cancels load variations. Also, the lower apparent plate resistance with voltage feedback is desirable in order to improve loudspeaker damping and reduce "hangover" effects.

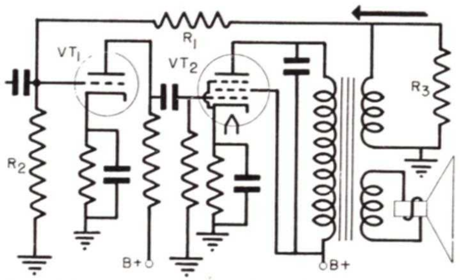

FIG. 25. A parallel two-stage feedback circuit.

► For the driver stage of a class B audio amplifier, current feedback is never used. The variable load presented by the class B stage in drawing grid current requires good driver regulation. If the apparent driver plate resistance is made even higher by current feedback, the regulation is made worse instead of better. On the other hand, voltage feedback is commonly used in this situation because the lowered plate resistance improves regulation in addition to the reduction of distortion and noise.

SPECIAL FEEDBACK CIRCUITS

In all the feedback circuits previously discussed, we have assumed that the feedback voltage was constant and equal for all frequencies. It is possible, however, to insert inductances or capacities in the feedback circuit so that feedback will change with frequency. By this means very unusual amplifier performance can be obtained.

Selective Amplification. Suppose that instead of a resistance voltage divider across an amplifier output, we use a resistance-capacitance divider as in Fig. 26A to obtain feedback voltage. What is the result?

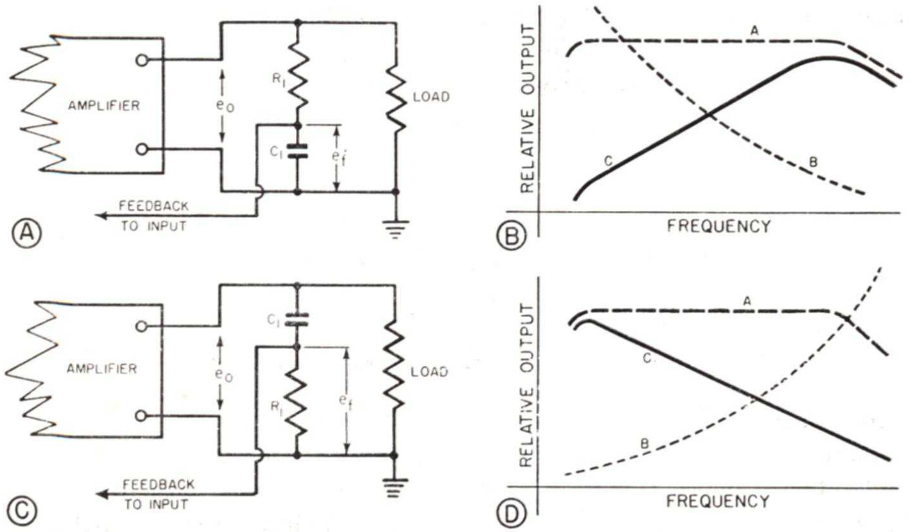

FIG. 26. Bass or treble peaking with frequency-selective feedback.

A small current proportional to the output voltage now flows through the resistor R_1 and capacity C_1 in series. The feedback voltage e_f is that voltage developed across the capacitive reactance alone. The reactance of the condenser, however, is not constant with frequency. At low frequencies, the reactance will be large, and a large feedback voltage will be developed. At high frequencies, the reactance will be small, and only a small feedback voltage will be made available.

We can expect under such circumstances that our amplifier response will be very different at high and low frequencies. As shown in the curves of Fig. 26B, if the amplifier response without feedback is as drawn in the dashed curve A, and the feedback voltage varies as curve B, then the amplifier response with selective feedback will be something like curve C. This is so because at those frequencies where feedback is low, the feedback has little effect and almost full amplifier gain is obtained.

The reverse situation will be the result if the resistor R_1 and condenser C_1 of the feedback divider are interchanged. This is illustrated in Fig. 26C. Here the feedback voltage e_f is developed across the resistor R_1. At low frequencies, the reactance of condenser C_1 is so large that very little current flows through the divider, consequently, at low frequencies there is little feedback voltage. As the frequency is increased, the reactance of C_1 goes down rapidly. More and more current flows through the divider, and at high frequencies almost all the output voltage is developed across R_1 as feedback voltage. We have then, very little feedback at low frequencies, but a high feedback ratio at high frequencies.

As before, where feedback is high, the amplifier gain will be cut sharply; where feedback is small, the gain will be reduced but little. With an amplifier response without feedback as in curve A in Fig. 26D, and feedback voltage varying with frequency as in curve B, the resultant response will be changed to a curve like C.

Obviously, selective feedback is a means of achieving bass or treble peak response. Similar circuits have been used extensively for tone control purposes.

Cathode Circuits. Selective feedback also may be brought about by using cathode current-controlled feedback. Such a method is simpler, but usually is not as effective since only one stage is included in the feedback loop.

Fig. 27A shows a treble-peaking circuit. Resistor R_1 is the customary cathode resistor for obtaining proper grid bias. The condenser C_1, however, is made too small to by-pass low frequencies. As a result, when the variable resistor R_2 is zero, high frequencies will be by-passed satisfactorily by condenser C_1, and full stage gain will be given for these frequencies; but low frequencies will not be by-passed, and the cathode degeneration will reduce them in the output. The over-all response of the stage for different frequencies will be similar to curve C of Fig. 26B. The degree of treble-peaking can be adjusted by changing the variable resistor R_2 in the cathode circuit.

For bass boost, the circuit of Fig. 27B can be used. An inductance L_1 is inserted in series with the conventional cathode resistor R_1, which is by-passed by condenser C_1. The effectiveness of L_1 can be changed by adjusting the shunting variable resistor R_2. At low frequencies, the inductive reactance of L_1 is very small, hence, little feedback voltage is developed across it, and nearly full stage gain is obtained. At high frequencies, the inductive reactance of L_1 increases, and by raising the feedback voltage, reduces the gain accordingly. Response of this circuit is similar to curve C in Fig. 26D.

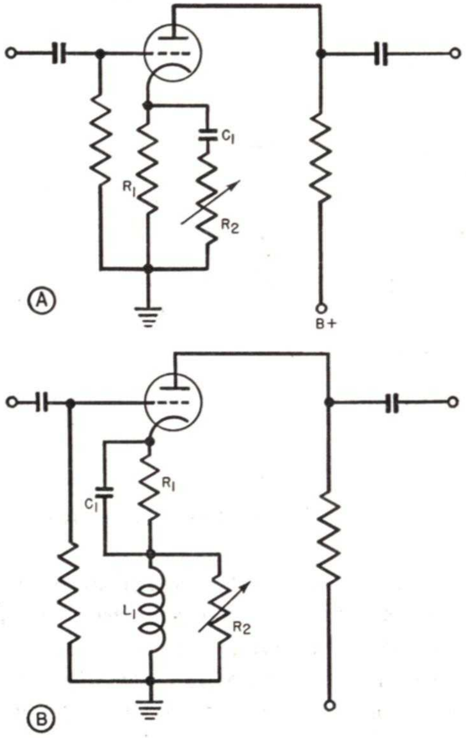

FIG. 27. Frequency peaking with cathode degeneration.

The Scott Selective Amplifier. It is possible to use inverse feedback to produce a selective amplifier—one that amplifies only a narrow frequency band. An example is the Scott Selective Amplifier incorporated in the General Radio Wave Analyzer.

The method of obtaining feedback in this analyzer is outlined in Fig. 28A. Instead of a voltage divider, a filter network is inserted in the feedback lead. This filter, composed of three condensers and three variable resistors, is so arranged that it passes with little loss all frequencies but one. For this one "null" frequency, however, the attenuation through this filter is very high, and if the circuit is balanced properly, the transmission will be zero.

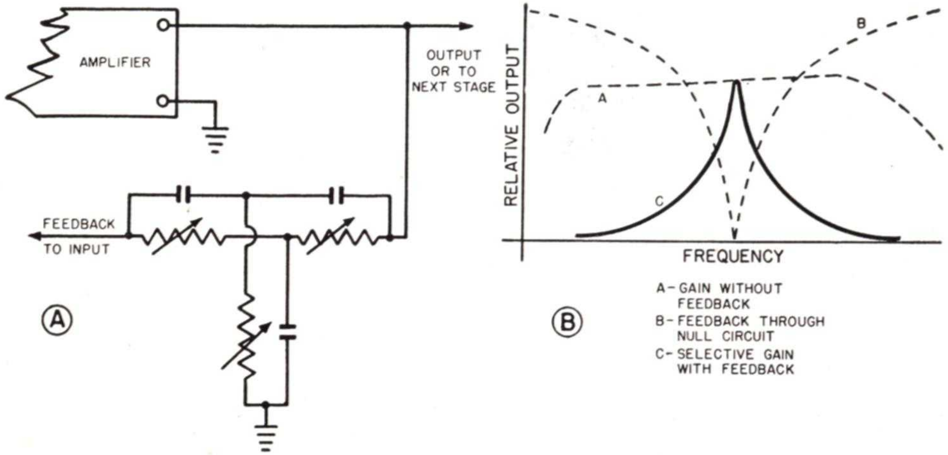

FIG. 28. Effect of the bridge-T null circuit in the Scott Selective Amplifier.

The feedback voltage will be determined by the transmission characteristics of the filter network. This is shown in curve B of Fig. 28B. It is apparent that for all frequencies but one, nearly 100% feedback is developed; but at the null frequency, the feedback drops to zero. Hence, if the amplifier without feedback had a response like curve A of Fig. 28B, response with such feedback would appear as curve C. At frequencies away from the null, nearly 100% feedback cuts the amplifier gain to very nearly one; exactly on the null, no feedback voltage is passed, and the full high gain of the amplifier is brought into play.

The response looks like an ordinary resonance curve. Thus, frequency selectivity is obtained with no L-C tuned circuits whatsoever.

MODULATOR CORRECTION

All inverse feedback systems we have seen up to this point have concerned only audio-frequency amplifier or modulator equipment. In a transmitter it is possible to include the grid or plate-modulated r.f. stage, and the class B linear r.f. amplifier stage as well in the same feedback loop. In doing this, the noise, hum, and distortion originating in these radio-frequency stages also can be cut down remarkably.

Over-all r.f. to a.f. feedback is accomplished as shown in Fig. 29. A small diode rectifier, or detector, is coupled through a tuned circuit to either the final output tank circuit, the antenna transmission line, or directly to the antenna itself. The modulated signals are detected, and after the r.f. energy is removed by filters, the resulting audio frequencies are fed back into the speech amplifier or modulator driver circuits as shown. If the audio feedback voltage arrives in proper phase, very good results can be obtained.

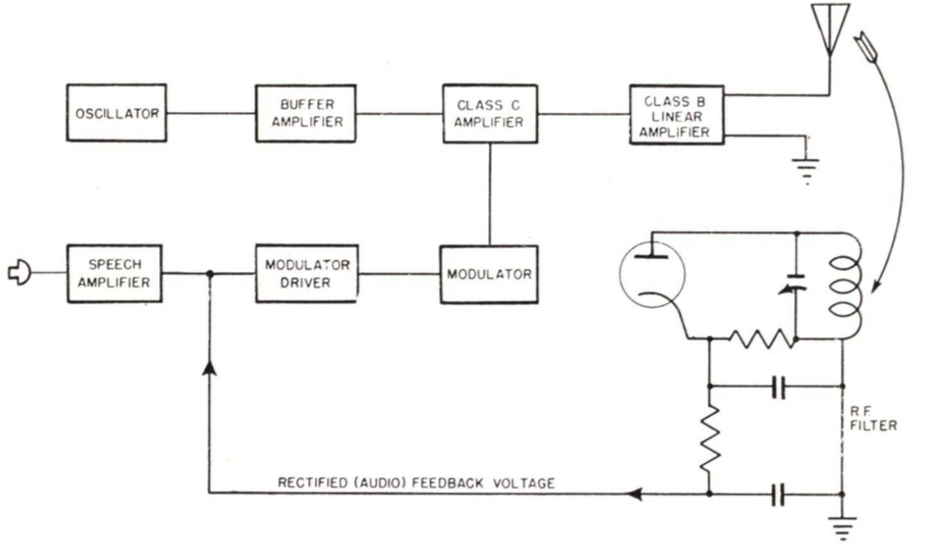

FIG. 29. Over-all r.f. to a.f. feedback that reduces distortion, hum, and noise in the modulated class C and class B linear r.f. amplifier stages as well as in the audio-frequency amplifiers and the modulator.

Noise or distortion in the modulator or modulator driver, and even those distortion components that are generated in the modulation process can be reduced to almost insignificant values. This system is so effective that the class B linear amplifier stage can be operated with a.c. heated filaments, for the transmitted a.c. hum will be brought down to a reasonable value by the feedback.

UNDESIRED FEEDBACK

Upon constructing a high-gain, multistage amplifier, it is common experience to find that on turning up the gain, the unit becomes unstable, and the output breaks into a series of howls, squeals, or "putt-putt-putts" in the customary motorboating fashion. This is the result of accidental, undesired positive feedback.

If the amplifier output contains howls or squeals, it is probably caused by high-frequency feedback. This may be the result of inadequate shielding or of having components in separate stages placed too close together. High-frequency feedback may be minimized by proper shielding, keeping each stage separated, and keeping the input and output leads or cables as far apart as possible. If this doesn't help, the gain at high frequencies of one or several stages should be decreased by employing by-pass condensers or by other means. In this way, although high-frequency feedback is still present, the feedback voltage is made too low to maintain oscillations.

On the other hand, if the amplifier motorboats, the most probable cause of feedback is coupling from output to input through the common power-supply system. This is low-frequency feedback, and occurs because the output stage, in drawing heavy plate current, actually drops the supply voltage on current peaks. Whenever the slight supply voltage drop reaches the plate of the input stage, it is treated like an input signal, and is amplified down the line back to the output stage. The process repeats itself, giving the characteristic "putt-putt-putt."

Quite often motorboating can be stopped by increasing the capacity of the output filter condenser of the power supply filter. This improves the by-pass effect, and lowers the power-supply impedance that is common to the output and preceding stages.

Another effective method is to insert resistance-capacitance decoupling filters in the plate-supply lead of the first stage and, if necessary, in the second stage as well. These filters are inserted in both stages in Fig. 30.

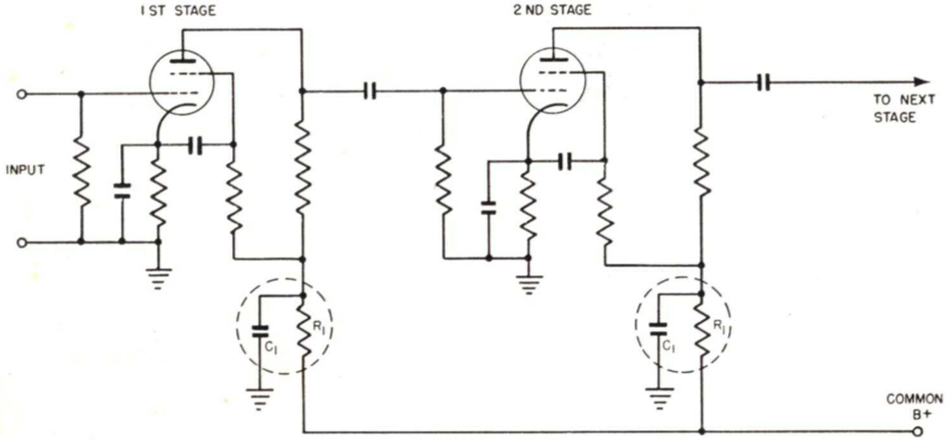

FIG. 30. Use of decoupling filters to prevent "motorboating."

The resistors R_1 are usually about one-tenth the resistance of the plate load resistors, and are placed in series with the plate supply leads. The by-pass condensers C_1 may be of any value from about 0.01 mfd. up to 1 or 2 mfd., or higher to prevent oscillation.

The action of these filters is simple. The condenser C_1 normally is charged to very nearly full plate voltage. Any disturbance of the main supply voltage caused by current drain of the final stage will not be felt immediately through the filters, for it takes a definite time for the by-pass condensers to charge or discharge to the new voltage value. If this period of charge or discharge is made long enough, the output stage is prevented from causing rapid voltage fluctuations in the earlier stages, and the amplifier is stabilized.

Lesson Questions

Be sure to number your Answer Sheet 21RC.

Place your Student Number on *every* Answer Sheet.

Most students want to know their grade as soon as possible, so they mail their set of answers immediately. Others, knowing they will finish the next Lesson within a few days, send in two sets of answers at a time. Either practice is acceptable to us. However, don't hold your answers too long; you may lose them. Don't hold answers to send in more than two sets at a time or you may run out of Lessons before new ones can reach you.

1. What class of amplifier is used in the r.f. amplifier stages following the modulator, when low-level modulation is used?
2. What makes it possible to use a single tube in a class B linear r.f. amplifier, when two tubes in push-pull are necessary for a class B *audio* amplifier?
3. What is the usual *peak,* and also the average power efficiency of a single-tube class B linear stage?
4. List 3 advantages obtained by the use of tubes in push-pull, over single tubes, in a class B linear amplifier.
5. Why is a grid phasing network used in the Doherty amplifier?
6. If sufficient inverse feedback is put into an amplifier to drop the gain 75% (to ¼ its former value), to what per cent would an original output distortion of 16% be reduced?
7. What is the greatest cause of trouble when attempting to bridge a number of stages with a feedback loop?
8. Why is current-controlled feedback never used in a driver stage which feeds a class B audio amplifier?
9. If a coil or a condenser is inserted in the feedback loop, will the feedback be the same at all frequencies?
10. In transmitter over-all r.f. to a.f. feedback systems, what benefits are possible in addition to correction of noise and distortion in the modulator?

THOROUGHNESS

Whatever you do, do well if you would stay on the straight road to success. The habits of carelessness and slipshod work are all too easy to acquire; beware of them as you would the plague. Men who are thorough in their work cannot remain undiscovered for long, because the demand for such men is greater than the supply.

Thoroughness is just as important in study as it is in work; what you get out of a lesson depends upon how completely you master the material presented in it. Some books, as fiction, are read hurriedly and only once, then cast away; the enduring works of literature are carefully read and reread many times but always essentially for the pleasure they give; textbooks, however, must be read quickly, to get the basic ideas, then carefully many times until every important principle has been mastered. Textbooks are always saved for future reference; tomorrow you may have urgent need for the information given in a paragraph which today seems so insignificant.

Thoroughness in study habits leads to thoroughness in work habits, and eventually to a thorough success.

J. E. Smith