VIDEO AMPLIFIERS AND D.C. RESTORERS

finish

54RH-3

NATIONAL RADIO INSTITUTE
WASHINGTON, D. C.
ESTABLISHED 1914

STUDY SCHEDULE No. 54

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

☐ 1. **Introduction** . . . . **Pages 1-6**
The video amplifier is compared to the audio amplifier in a sound receiver, then signal characteristics are discussed.

☐ 2. **Video Signal Distortions** . . . . **Pages 7-10**
This section covers amplitude distortion, frequency distortion, phase distortion, and time delay.

☐ 3. **Fundamental A.C. Video Stages** . . . . **Pages 11-20**
Low-frequency response, high-frequency response, series compensation, and shunt compensation are covered here.

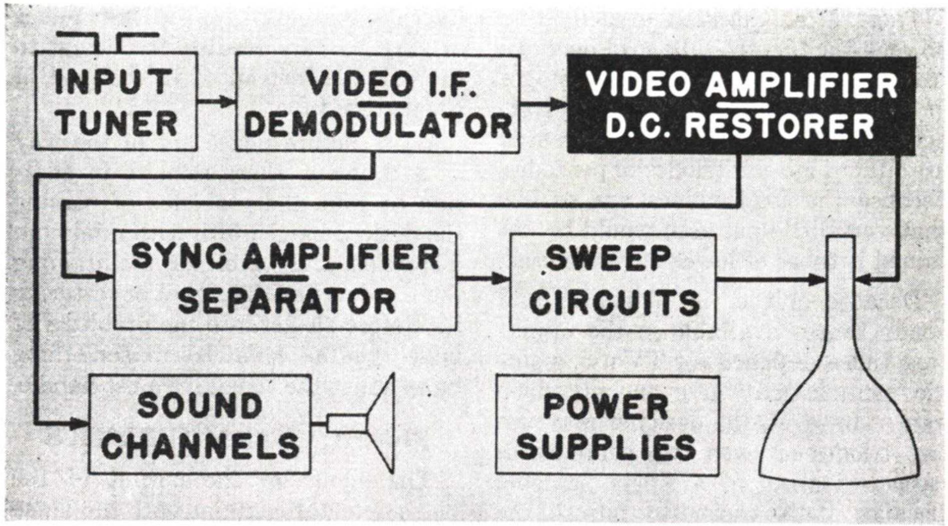
☐ 4. **Fundamental D.C. Amplifier** . . . . **Pages 21-30**
You study the power supply, a typical circuit, and signal operation of a two-stage d.c. amplifier, then learn about brilliancy control and contrast control.

☐ 5. **D.C. Restorers** . . . . **Pages 30-36**
This discusses the basic d.c. restorer, then describes a practical diode restorer, restoration in the last video stage, and in the picture-tube circuit.

☐ 6. **Answer Lesson Questions and Mail your Answers to NRI for Grading.**

☐ 7. **Start Studying the Next Lesson.**

(An Affiliate of the National Radio Institute)
JD6M250 1950 Edition Printed in U.S.A.



NOW that you have followed the video signals through the r.f. section (the input tuner and i.f. amplifier) and have studied the video detector, you are ready to make a detailed study of the video amplifier. This video amplifier can be compared directly to the audio amplifier of a sound receiver; it is the "low-frequency" section of a television set—the section that takes the signal from the detector and delivers it to the picture tube. It is necessary because the signal output of the detector is insufficient to operate the picture tube.

The early history of the development of video amplifiers is much like that of audio amplifiers. Originally, the gain obtainable in the r.f. section was low, so a number of video stages had to be used, even though fidelity was sacrificed thereby. Then, as developments in tubes and circuits permitted more amplification at high frequencies, the number of video stages was decreased, thus permitting wider frequency response. Today, many TV sets have only one video stage, although most use two, and a very few use three.

Basically, the video amplifier stage is either a resistance-coupled or a direct-coupled stage. The transformer coupling commonly used in audio amplifiers is completely impossible in video amplifiers because the very wide frequency range present in video signals simply cannot be passed through audio transformers. Hence, the maximum gain can be no more than the mu of the tubes; actually it will be far less than this because the load values must be kept small.

To get a general idea of the gain needed, we can assume that the output of the video detector will be somewhere around 2 volts in a set installed within a few miles of a television station. The amount of signal required to drive the picture tube may be anywhere between 40 and 100 volts, depending on the tube type. This means that the video gain necessary will be between 20 and 50.

There is no question of delivering power—the picture-tube grid operates from a voltage, so all the video stages in a TV set are essentially just voltage-amplifying stages. The tubes used are either high-mu triodes or pentodes, because the load values are so low that very little net gain would be obtained if tubes of lower mu were used.

Because of the far higher mutual conductances available in the miniature tubes designed for TV use, a single pentode may give enough video gain. However, the designer may use two triodes or even two pentodes in cascade instead of a single pentode, because it is easier to adjust the two load circuits to give the required band width than it is to adjust the single load circuit of a single stage.

Picture Phase. When the signal is fed to the grid of the picture tube, it must have a positive picture phase—that is, the signal voltage must go increasingly positive as the scene increases in brightness. Each video stage reverses the phase 180°, so when an odd number of video stages (one or three) is used, and the signal is fed to the picture-tube grid, the detector must deliver a negative picture phase. On the other hand, when an even number of stages (two) is used, and the signal goes to the picture-tube grid, the detector output must have a positive picture phase. At one time this had to be taken into consideration; today, however, the detector output can be of either phase because it has been found possible to secure a positive picture by feeding a signal of negative picture phase to the cathode of the picture tube. Therefore, today, regardless of the number of stages used in the video amplifier, the picture phase is adjusted properly either by setting the detector to deliver the correct phase for the number of stages or by feeding the signal to the cathode instead of to the grid of the picture tube.

With picture phase out of the way as a problem, there remain, in addition to gain, the problems of passing the desired band width with minimum distortion and of either maintaining the d.c. level of the signal or restoring it. Before we get into the problems of band width, let's learn something more about the signal we must handle.

SIGNAL CHARACTERISTICS

The signal at the output of the video detector contains both the video and synchronizing signals. The complete video signal, including the sync pulses, is fed to the picture tube. It is not desirable to separate the synchronizing signals from the picture signal in the video amplifier, because the sync pulses help to blank out the tube during sweep retraces, as we will show elsewhere.

At some point between the detector and the picture tube, a copy of the entire signal is taken off to be fed through the stages engaged in controlling the sweep circuits. The exact point depends on the sync level and phase, as another Lesson will show.

Frequencies. The video amplifier must pass signals ranging from about 10 cycles to as much as 4.25 megacycles in high-definition systems. The upper limit is subject to some variation; when small picture tubes are used, it is impossible to see fine details at normal viewing distances, and furthermore, such sets are frequently inexpensive types. In such cases, the upper frequency limit may be cut to be only about 3 mc. Receivers with medium definition may cut off around 3.5 to 3.75 mc.

A good low-frequency response is necessary so that the system will reproduce slow changes in brilliancy or gradual changes in shading from light to dark. On the other hand, a good high-frequency response is quite necessary so that the set will be able to reproduce satisfactorily the sudden and sharp changes that occur when the scanning spot moves from a light to a dark object in a scene.

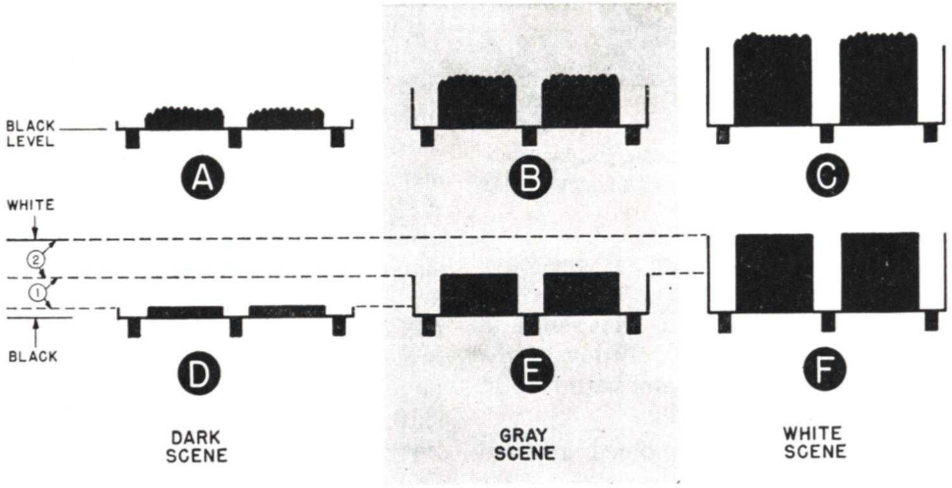


FIG. 1. How the d.c. component affects the video signal.

D.C. Component. In addition to having an extremely wide frequency range, the video signal differs in one other important respect from the sound signal with which you are familiar. The video signal has a d.c. component; it is actually a pulsating d.c. signal rather than a true a.c. one. The effect of the d.c. component is shown in Fig. 1. In A, B, and C of this figure we have the same a.c. signal, except that in each instance the level about which it varies is different. In A we have the signal for a dark scene, in B that for a gray scene, and in C that for a brightly lighted scene. All three a.c. signals may represent two scanning lines crossing the same object: for example, they may be scans of a scene that shows a house in darkness, in moonlight, and in sunlight.

For simplicity, let's wipe out the variations in the a.c., thereby getting the signals shown in Figs. 1D, 1E, and 1F. (These could well represent scenes having a solid over-all dark, gray, or white tone respectively.)

Examining these three signals, you will see that the only basic difference between them is in the amplitude of the video portion, which is shaded in this figure. Thus, as we move from the dark scene to the gray scene, the amplitude represented by the shaded area increases. It increases more as we move from the gray to the white scene. However, this increase is all on one side of the black-level line. That is, the synchronizing pulses are not changed in amplitude at all; we have merely changed the amount of video signal by adding a d.c. to our a.c. signal.

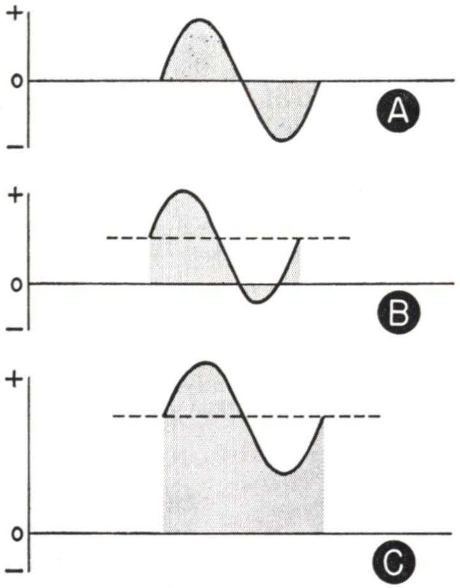


FIG. 2. What happens when a d.c. component is added to an a.c. sine-wave signal.

If we add a d.c. to an a.c. sine-wave signal, as shown in Fig. 2, the effect produced is the same as that shown in Fig. 1. That is, the addition of the d.c. moves the a.c. signal with respect to the zero reference line.

If we pass the combined a.c. and d.c. signal of Fig. 2C through an a.c. amplifier, the d.c. component will be wiped out, and we will have only the a.c. component of Fig. 2A. There is no way for us to regain the d.c. component here, because the signal, after a.c. amplification, contains no component that indicates what its previous d.c. level was.

In a TV signal, however, the presence of the pedestal and sync pulse does let us regain the d.c. component after it has been wiped out by a.c. amplification. Let's see why it is important to regain this d.c. level and learn how it may be done.

Figs. 3A, B, and C show the pulsating d.c. forms of the TV signals we had in Fig. 1. In Figs. 3D, E, and F are their respective a.c. equivalents. Notice that the pedestals of the a.c. signals do not line up, although they do in the d.c. signals. This change has occurred because the areas of an a.c. signal on either side of the zero line are always equal (or, in other words, the average of an a.c. signal is always zero). The pedestals of signals having large d.c. components must therefore shift considerably below the reference line when the signals are converted to a.c.

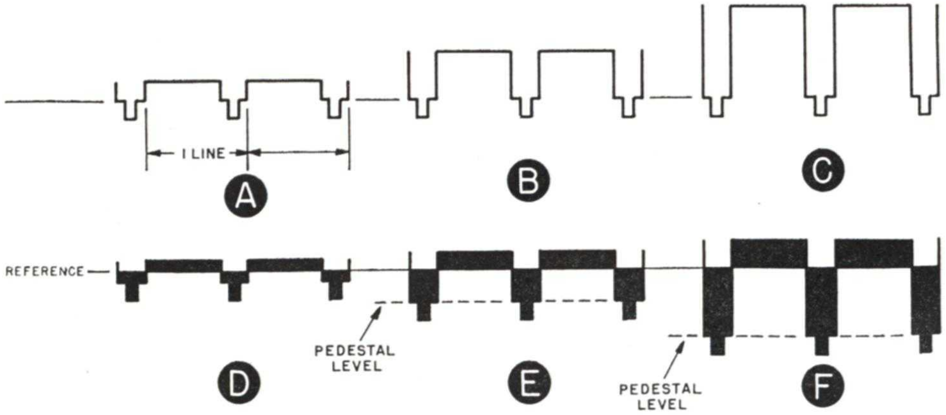


FIG. 3. Pulsating d.c. signals (top) and their a.c. equivalents (bottom).

It is possible to apply these a.c. signals to the grid of the picture tube without first lining up their pedestals. If this is done, however, improper operation will be produced, as you can see from Fig. 4. Here, a dark scene (Fig. 4A) has been lined up so that its pedestal coincides with the operating point X on the tube curve. Video signals now swing to the right to give increases in brilliancy, and sync pulses go to the left into the blacker-than-black region to cut off the tube during retraces. A bright scene, however, as shown in Fig. 4B, would give but little increase in over-all illumination, because the effect of any swing in the direction to the left of the operating point would be lost. On the other hand, if we arrange the circuits so that the pedestal of a bright scene (Fig. 4C) lines up with the operating point of the tube, even the sync pulses of a dark scene (Fig. 4D) will be well to the right of the operating point; the retraces of the dark scene will therefore be visible in the picture, and the scene would be very much brighter than it should be.

All this wouldn't matter if all scenes had the same background brilliancy. Since they don't, however, the pedestals of the signals applied to the picture tube must be lined up, as shown in Fig. 5, so that changes in background illumination will be reproduced accurately and so that the retrace lines will not be visible. In other words, the signals applied to the picture tube must contain a d.c. component that will permit alignment of the pedestals.

Since the signal is produced in the proper form (with the d.c. component so that the pedestals are aligned) at the video detector, we must somehow get this signal to the picture tube unchanged.

One way of avoiding a.c. amplification is to use a d.c. amplifier (one that does not have coupling condensers). A single-stage d.c. amplifier is relatively simple—the grid of the tube is fed directly from the video detector load, and its plate circuit is directly

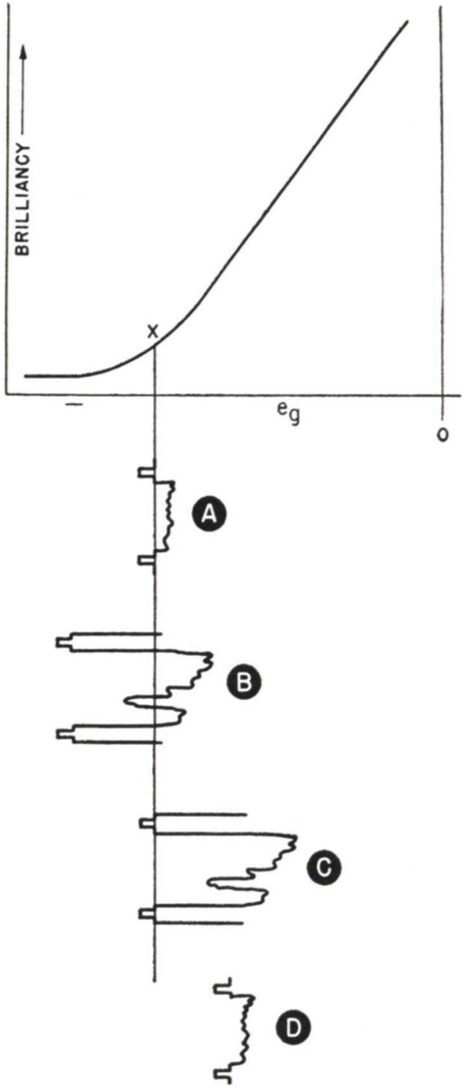


FIG. 4. This illustration shows what happens if the pedestals of the a.c. signals are not lined up before they are applied to the grid of the picture tube.

coupled to the grid circuit of the picture tube.

When we get into more stages, however, a number of major problems arise. A high-voltage power supply is required; if it is not furnished, the tubes must operate at reduced voltages, because the total supply voltage needed is the sum of the maximum voltages required for each stage. In the early days of television, this was a real stumbling block because of the tube types available. However, the tubes now used operate at relatively low voltages and do not draw excessive plate currents, so two-stage direct-coupled amplifiers are practical. They are used to a considerable extent, as we shall see later in this Lesson.

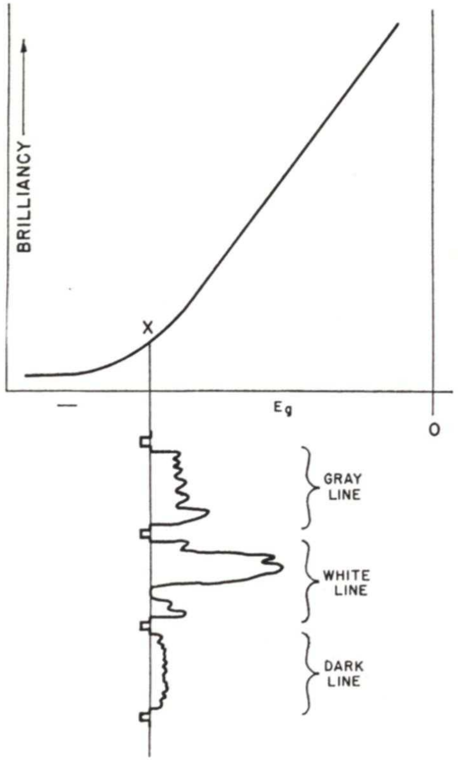


FIG. 5. How the pedestals of a.c. signals should be lined up to produce the proper variation in the relative brightness of the scene reproduced on a picture tube.

With the d.c. amplifier, special precautions must be taken to prevent low-frequency oscillation (corresponding to motorboating in a sound receiver). Also, there is a high cathode-to-heater voltage in the final stage of the d.c. amplifier, which may lead to tube leakage difficulties.

These problems, particularly the one of instability, make a.c. amplifiers quite desirable. Fortunately, we can use a.c. amplifiers even though they do wipe out the d.c. component of the signal, because it is possible to restore the d.c. component, using a circuit known as the d.c. restorer, so that the signal fed to the picture tube grid or cathode will have the pedestals lined up again properly. When this arrangement is used, the video amplifier does not require such extremely high voltages and is in general much more stable. For these reasons, a.c. amplification with d.c. restoration is more common than d.c. amplification. We'll study both types, but first let's go into the distortions that are troublesome in video amplifiers.

Video Signal Distortions

Video signals can be considerably upset by improper phase relationships as well as by frequency and amplitude distortion. Let's review these distortions and learn of the new requirements.

FREQUENCY DISTORTION

As you will recall, frequency distortion occurs whenever an amplifier does not amplify equally all the frequencies in the desired pass band. An amplifier that is defective in this respect may amplify some frequencies more than is desired and some less, or may not pass some frequencies at all.

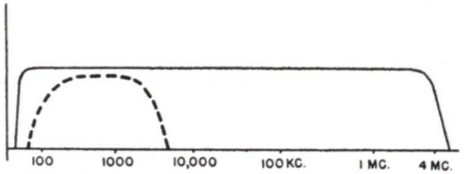


FIG. 6. The solid line shows the frequency response that the video section of a TV set should have; the dotted line shows the response of the average audio amplifier.

Fig. 6 shows the frequency response that the video section of a TV set should have. (Compare this with the frequency response of the average audio amplifier, shown by the dotted line.) It is not absolutely necessary that the response of the video amplifier alone be as flat as this. If there is a loss of high frequencies in the video i.f., for example, it can be compensated for by having a peaked response at these frequencies in the video amplifier. Similarly, any peaking in the r.f. or i.f. sections can be compensated for by a reduced response in the video amplifier at the appropriate frequencies. Thus, the response of the video amplifier can be adjusted to make up for deficiencies in the responses of preceding sections, thereby creating an approximately flat over-all characteristic for the entire set.

AMPLITUDE DISTORTION

When the wave shape of the output of a radio circuit or device is not exactly proportional to the wave shape of the incoming signal, we have what is known as amplitude distortion. This distortion results in the production of harmonics that were not present in the original signal. Such changes in the shape of the signal are commonly produced by operating an amplifier too near one of the bends of its characteristic; they may also be produced by overloading. Overloading is commonly avoided in television through the use of a gain control in the i.f. amplifier or in one of the video stages. Set designers are always careful to choose operating voltages and loads such that amplitude distortion is not appreciable when a set is operating as it should.

While we are on the subject of amplitude distortion, we might mention intermodulation distortion as well. Intermodulation distortion occurs when two signals are mixed and their beat (sum and difference) frequencies are produced. These sum and difference frequencies are not harmonics, but they are produced by operation over a bent characteristic just as amplitude distortion is. Therefore, intermodulation distortion is low whenever the amplitude distortion is low, and vice versa.

PHASE DISTORTION

Phase distortion exists in all amplifiers—sound as well as video. It is of no particular importance in sound amplifiers, but it is extremely important in video circuits.

Phase shifts occur because there is some reactive component in the circuit. Even in a direct-coupled amplifier, the tube capacities shunting the load introduce a certain amount of phase shift. This comes about because the impedance of the load circuit changes as the frequency of the signal changes. Therefore, in most cases, the amount of phase shift changes as the frequency of the input signal changes.

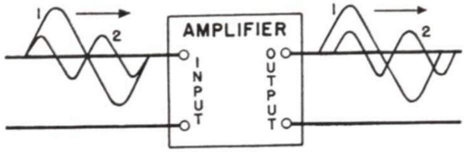


FIG. 7. The phase relationship of the components of a complex wave can be altered by a video amplifier. This causes "phase distortion."

A phase shift causes the wave form of the signal to be changed because the relative positions of the various components in the signal are changed. For example, in Fig. 7, we are applying to the amplifier an input signal consisting of two components, 1 and 2, with the second having twice the frequency of the first. If there were no phase shift, the components in the output would be in the same phase as are the input components.

If, on the other hand, one frequency gets more of a shift than another, the resultant output wave form will be quite different from the input. In Fig. 7, the output signals show that the higher frequency has been shifted 90° in phase with respect to the lower one.

The resultant of the two input signals is shown in Fig. 8A; the 90° phase shift produces the resultant output shown in Fig. 8B. By comparing the two resultant waves, you will see that the over-all wave form of the output is quite different from that of the input, which means that distortion has been introduced.

One effect of this distortion is shown by the change it causes in the displacement from the zero line of points on the resultant wave. As you can see, point X in Fig. 8A is at a considerably greater distance from the zero line than is point X in Fig. 8B, although the two occur at the same time in the cycle. Since its displacement from the reference line determines the shade (relative grayness) of a point in a signal, phase distortion is thus capable of changing the shade of various parts of a picture.

At first glance, it might appear that we could avoid difficulty if we could arrange for the same phase delay at all frequencies. However, another factor is involved here—the fact that a phase delay results in a delay in time. Any time delay in the application of a signal to the picture tube will, of course, cause the part of the scene represented by that signal to appear in the wrong place. Hence, a phase shift may not only change the shade of parts of the picture, but may also produce an actual physical distortion by causing some portions of the image to be out of place. The image blurring caused by time delay is even less desirable than changes in shade are, so television circuits are arranged to produce a constant *time* delay rather than a constant phase shift. If all components of the signal are *time* delayed the same amount, the entire picture is shifted slightly to the right or left, which doesn't matter. We get into trouble only when some elements of the picture are shifted more than others.

TIME DELAY

To get an idea of how the phase shift and time delay are related, refer to Fig. 9. In Fig. 9A, we have assumed that we have a 100 cycle-per-second sine-wave signal. If this signal is shifted 90° (one-quarter cycle), as shown by the dotted-line wave, there will be a certain definite delay between the time that the original signal performs some action (such as going through zero) and the time that the phase-shifted signal performs the same action. After it has passed through zero, for example, the original signal will go through a quarter of a cycle before the phase-shifted signal goes through zero. The time delay suffered by the shifted signal is therefore equal to the time duration of a quarter-cycle. Since the time for one cycle of a 100-cycle signal is 1/100 of a second, the time delay caused by a phase shift of one-quarter cycle is 1/400 of a second.

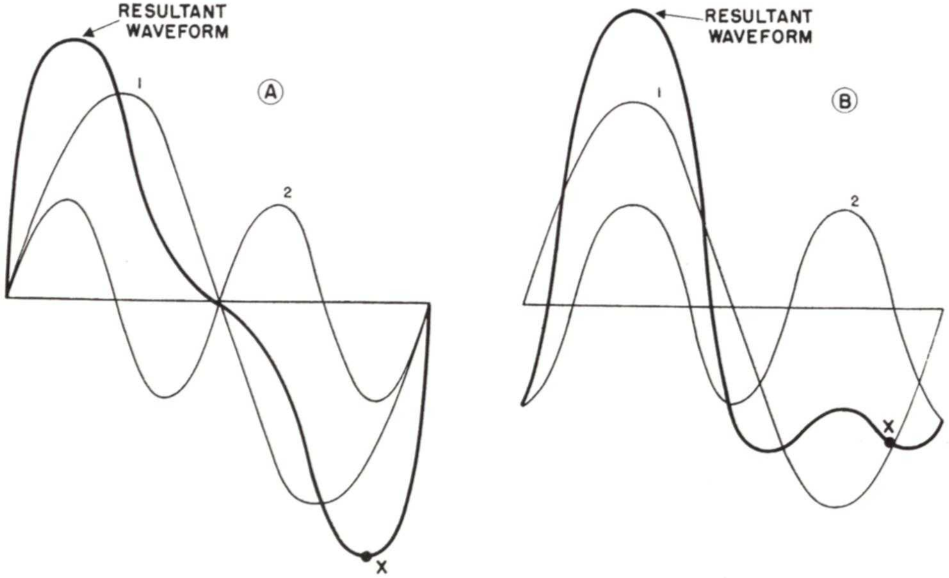


FIG. 8. If two input signals are applied simultaneously to an amplifier, and one is shifted more in phase than the other is, the resultant waveform of the output (B) will be quite different from the resultant waveform of the input (A).

Now let's see what happens with a 200-cycle signal. As shown in Fig. 9B, the time for one cycle is 1/200 of a second, so the time delay for a phase shift of one quarter cycle would be 1/800 of a second. Thus we see that the *time delay* for a phase shift of 90° at 200 cycles is one-half as long as the time delay caused by the same amount of phase shift at 100 cycles. Therefore, if phase-shifted signals of various frequencies are to undergo a *constant time* delay, the amount of phase shift must increase as the frequency increases. In other words, the phase shift must vary linearly with frequency.

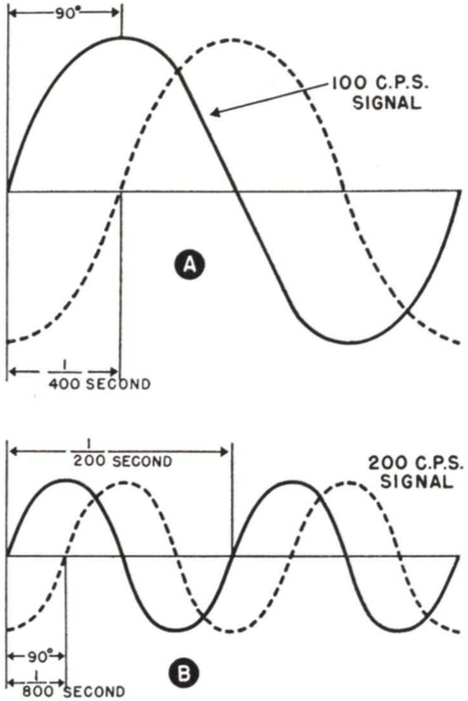


FIG. 9. How phase shift and time delay are related.

As an example, if an amplifier has a phase shift of 6° at 500 cycles, it must have a 60° phase shift at 5000 cycles and only a .6° phase shift at 50 cycles if it is not to produce time-delay distortion of the amplifier signal. The formula that shows the relation between time delay (t) and phase shift (θ) is:

$$t = \frac{\theta}{360\,f}$$

where t is in seconds, f is in cycles per second, and θ is in degrees.

We will give a somewhat better idea of the effects of time delay later, when we consider the effects of such delay at low and at high frequencies. (Incidentally, this really should be called a time shift rather than a time delay, because it is possible for there to be speeding up or advance in time of some components, however, since it is customary to refer to it as "time delay," we shall use that form too.) Let's now turn to a.c. amplifiers to learn of their characteristics, then go on to d.c. amplifiers.

Fundamental A.C. Video Stages

Let's start our study of a.c. video stages by reviewing the resistance-capacitance-coupled amplifier shown in Fig. 10, which is much like the audio amplifier with which you are familiar. Briefly, you will recall that the gain of the VT_1 stage depends on the amplification factor of the tube and on the load used with that tube. At middle frequencies in the pass band, the load is the parallel combination of resistors R_1 and R_2. If R_2 is fairly high in value compared to R_1, the load is approximately the value of R_1 alone.

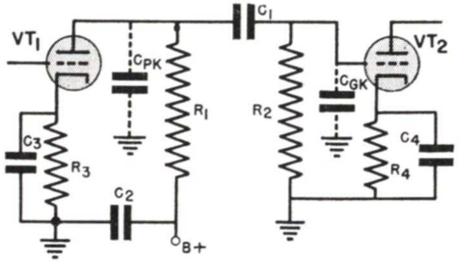


FIG. 10. A fundamental resistance-coupled amplifier.

If such a stage contained no reactive components, its frequency response would be flat from zero to infinity. This is shown by the horizontal line in Fig. 11. As a matter of fact, however, the frequency response of an amplifier of this kind rolls off in both the low- and the high-frequency regions, as indicated by the dotted lines in Fig. 11. The fall-off in the response at the low-frequency end, in the region marked X, is chiefly caused by the coupling condenser C_1 in Fig. 10, which acts as a voltage divider with R_2. As the frequency decreases, the reactance of C_1 increases, and less signal appears across R_2 for application to the next tube.

The low-frequency response is further reduced by the cathode by-pass condensers C_3 and C_4 in Fig. 10, because their reactances rise so much at low frequencies that they are not effective by-pass condensers, with the result that degeneration occurs.

At the high frequencies, the response falls off because the load is shunted by the output capacity of VT_1 (marked C_{PK} in Fig. 10), by the input capacity of VT_2 (marked C_{GK} in Fig. 10), and by such distributed capacities as exist from the plate and grid wires and from C_1 to chassis. The reactances of these capacities fall as the frequency increases, so they shunt the load at high frequencies. This reduction in the net load causes the high-frequency roll-off marked Y in Fig. 11.

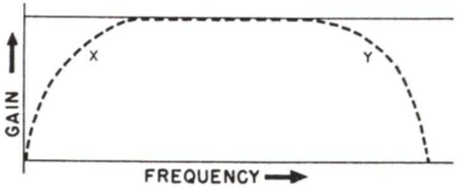


FIG. 11. The frequency response of a simple R-C amplifier rolls off at both the low and the high frequencies.

Thus, we have two different problems—we have to extend the low-frequency roll-off so that the gain will be reasonably high at the lowest frequency we want, which is around 10 cycles in the case of the video amplifier, and we must also keep up the high-frequency response out to about 4 megacycles. Let's investigate both these problems.

LOW-FREQUENCY RESPONSE

As we just mentioned, the primary reason for the roll-off in the low-frequency response is that the coupling condenser and the following grid resistor form a capacity voltage divider. The action is brought out more clearly in Fig. 12. When we apply the voltage e_1 (from the load for VT_1) to the combination shown in Fig. 12A, a current will flow through C_1 and R_2; the current through R_2 produces the voltage drop e_2, which is the grid voltage for the following stage.

When the frequency is so high that the capacitive reactance of C_1 is negligible with respect to the value of R_2, e_1 and e_2 are equal, as shown in Fig. 12B.

As the frequency decreases, however, the reactance of C_1 increases. As it becomes larger, the current flow through the series circuit of C_1 and R_2 causes an increasing voltage drop e_C across the condenser. The same current flows through both C_1 and R_2, but the voltage drop across R_2 is in phase with the current, whereas that across C_1 is 90° out of phase with the current and lags behind it. Hence, as shown in C, D, E, and F of Fig. 12, when the drop e_C increases, the voltages e_1 and e_2 are pulled out of phase, and e_2 decreases. Eventually, at very low frequencies, the reactance of C_1 becomes so high that practically all the voltage e_1 is dropped across the condenser, and there is virtually none left as e_2. Therefore, this voltage divider reduces the output at low frequencies. In addition, as you can see from Fig. 12, the phase difference between e_2 and e_1 increases as the frequency decreases, eventually reaching approximately 90° at the low-frequency extreme of the frequency range.

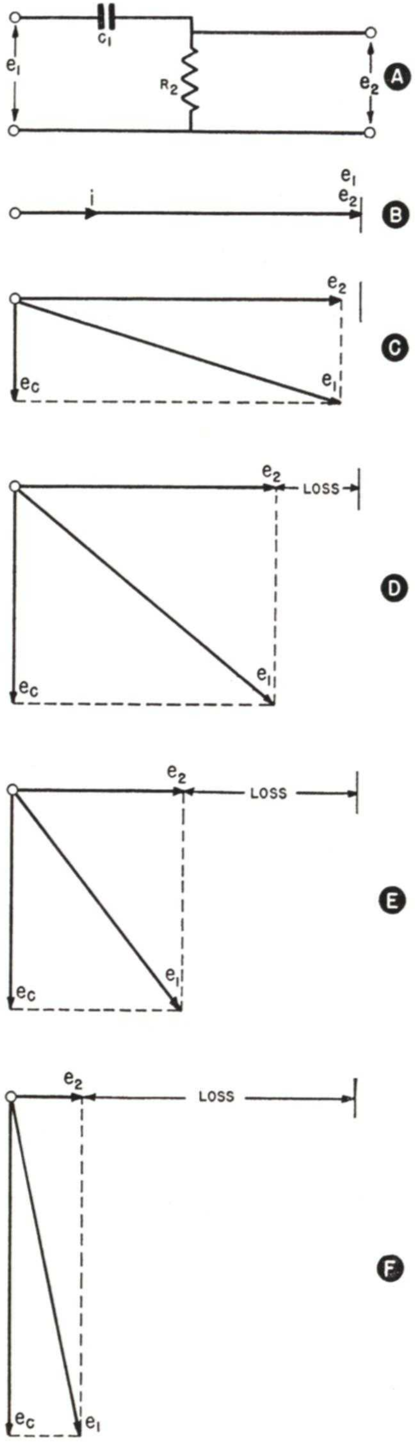


FIG. 12. These vector diagrams show why a roll-off occurs in the low-frequency response of an R-C amplifier.

The amount of this phase shift is very important. If it is as much as 45° or 1/8 of a cycle at 10 cycles, the time delay will be 1/80 of a second or .0125 second. This is 12,500 microseconds, which is almost equal to the time it takes to scan an entire field! Obviously, the phase shift must be kept very small if the amplifier is to handle very low frequencies.

Another way of looking at this action is to consider it as a time-constant problem. Let's suppose we have a scene like that shown in Fig. 13A. At the top of the picture, the signal voltage should rise to the white level. It should remain there until the middle of the picture is reached, when it should instantly shift to the black level, remaining there until the end of the picture. Hence, the signal applied to the grid of the picture should be like that shown in Fig. 13B.

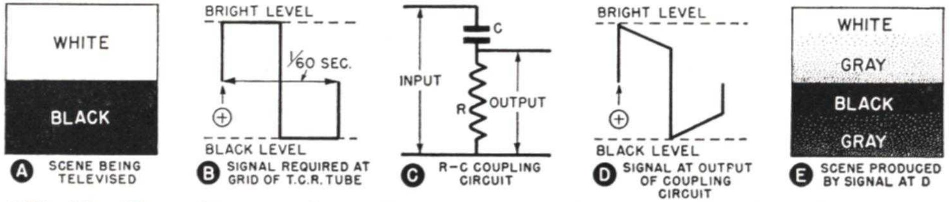


FIG. 13. These diagrams shown how an R-C coupling circuit produces background distortion. A long time constant for the coupling circuit minimizes the distortion.

If this signal is fed through an R-C coupling circuit like that in Fig. 13C, the signal at the output of this circuit will be like that shown in Fig. 13D if the time constant of the circuit is too short. At the instant the voltage changes from zero to the bright level, maximum current flows through R. As the condenser becomes charged, however, the current flow into it decreases, so the voltage across R gradually drops, producing a sloping top on what should be a square wave. When the voltage suddenly changes to the black level, the signal produced by the current flow through R goes to the black level at first, but the charging of the condenser produces a sloping bottom on what should be a square wave. When this signal is applied to the picture tube, the highest swing at the bright level will produce a white background, but since the voltage does not maintain this level but gradually fades off into the gray region, the picture will gradually get gray as the center of the screen is approached. At the center, the signal will suddenly produce a black picture, but this will become gray towards the bottom of the picture.

It would appear that we could prevent this graying of the whites and blacks by increasing the time constant of the coupling network—that is, by using large values of C and R so that it would take longer for the condenser to charge. This would keep the output more nearly constant by reducing the amount of slope in the top of the wave.

By increasing the values of either R_2 or C_1 in Fig. 10, we can keep the output more nearly constant and also reduce the phase distortion that occurs in such a coupling. However, there are practical limits to the values that may be used. If the resistor R_2 has too high a resistance, gas in VT_2 may cause difficulties; this resistor is in the grid circuit of VT_2, and a high resistance develops an unwanted positive "bias" when even small amounts of gas current flow through it. Increasing the value of condenser C_1 may also cause trouble, because this increases the capacity between the condenser and ground and therefore reduces the high-frequency response, as we shall show later. Furthermore, high-capacity condensers are likely to have relatively high leakage, which may upset the C bias in the following stage. Finally, and perhaps most important, making the time constant of the coupling circuit too long will make it impossible for the following stage to recover from transient voltages quickly, with the result that sharp changes in voltage may cause motorboating.

Because the use of a long time constant has so many drawbacks, designers use other methods to compensate for low- and high-frequency roll-off.

Low-Frequency Compensation. Fig. 14 shows the low-frequency compensating arrangement that is normally used in video amplifiers. This consists essentially of the by-pass condenser C_2 and the resistor R_5. At first glance, this looks like an ordinary decoupling network, and in fact it acts like one at the medium and high frequencies. By-pass condenser C_2 offers negligible impedance at high frequencies, so it prevents signals from getting either into or out of the VT_1 stage through the B+ lead.

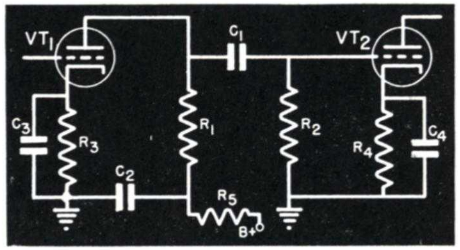


FIG. 14. A circuit that is often used for low-frequency compensation.

The value of C_2 is chosen, however, so that it is somewhat smaller than it would be in a sound receiver, with the result that it becomes part of the load at medium-low frequencies. In other words, at these frequencies, the load for VT_1 is made up of R_1 and C_2, since this condenser completes the a.c. plate circuit back to the cathode. The voltage developed across both R_1 and C_2 is applied to C_1-R_2. Effectively, therefore, we have a circuit in which the load increases in impedance as the frequency decreases. As a result, the voltage applied to C_1-R_2 increases as the frequency decreases, which makes up for the increased voltage drop across C_1.

Fortunately, this also compensates for the phase shift that we described earlier. The capacitive reactance of C_2 causes a phase shift in the source voltage for the C_1-R_2 divider of such a nature that the voltage at the grid of VT_2 is kept in nearly the proper phase.

The resistor R_5 is needed to make C_2 become part of the load. The value

of R_5 should be as high as possible; if it is made too high, however, there will be too great a drop in the B supply voltage. Therefore, in most practical circuits of this kind, R_5 has a resistance about 20 times as great as the reactance of C_2 at the lowest frequency the circuit is designed to pass.

If the C_1-R_2 phase shift were the only shift of importance, it could be removed by making the time constant of R_1 and C_2 equal to that of C_1 and R_2 (the time constant $R \times C$). However, you won't always find these two products equal in practical circuits, because there may be other phase shifts that must be compensated for. For example, the cathode by-pass condenser C_3 is supposed to prevent degeneration at all frequencies. However, at very low frequencies, the reactance of even a very high capacity will be appreciable in comparison to the low resistance it by-passes. Hence, there will be a certain amount of degeneration in the VT_1 stage in Fig. 14 at low frequencies, reducing the gain of the stage at those frequencies. Some of this effect can be compensated for by the C_2-R_5 combination, because the increase in the load will counteract the drop in gain.

Although we have shown triodes, pentodes are quite commonly used in video amplifiers because of their greater gain. The screen-grid circuit of a pentode introduces a low-frequency drop, because the screen-grid by-pass condenser is not large enough to be a completely effective by-pass at these low frequencies. This drop must also be compensated for by the C_2-R_5 combination.

The values used for C_2 and R_5 therefore have a considerable effect on the low-frequency response of the video amplifier. In replacing these parts, it is important to use exact duplicates to avoid upsetting the compensation.

More low-frequency compensation can be obtained, if desired, by allowing more of the vestigial side band to pass through the r.f.-i.f. sections. This will cause the output of the video detector to be higher in the low-frequency region.

HIGH-FREQUENCY RESPONSE

At the higher frequencies, ranging from about 15,000 cycles to 4 megacycles, compensation is absolutely necessary if we are to obtain a reasonable response. As a matter of fact, most engineers consider it more important to get the high-frequency response ironed out than to compensate the low-frequency response. Although it is true that the background brilliancy may be disturbed if the low-frequency response is poor, the *detail* in the picture is determined by the high-frequency response.

The low-frequency response is dependent on three items (the coupling condenser, the cathode by-pass condenser, and the screen-grid by-pass condenser), but the response at the high frequencies is dependent primarily only on the capacity that is in shunt with the load. In other words, the parts that cause the low-frequency roll-off play no part in causing the high-frequency roll-off, because the condensers involved have such low reactances at high frequencies that they act as short circuits and introduce no drop.

Therefore, at high frequencies, the typical circuits shown in Figs. 10 and 14 can be resolved into the equivalent circuit shown in Fig. 15. Here, the load resistance R_1, the grid resistance R_2, and the shunting capacities are all in parallel and make up the net load. The capacity we have marked as C_{PK} represents not only the plate-cathode capacity of VT_1 but also the stray capacities to the chassis of the load resistor and of the tube socket. The capacity C_{GK} represents not only the input capacity of the following tube but also the stray capacities to the chassis of the grid resistor and of the coupling condenser.

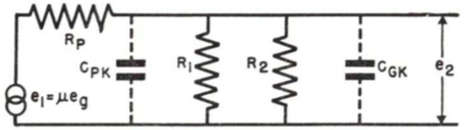


FIG. 15. The equivalent circuit of an R-C amplifier at high video frequencies.

If we combine the parallel resistors and the parallel capacities, we get the equivalent circuit shown in Fig. 16. You can see that, as the reactance of the capacity decreases, the net impedance of the parallel combination of C and R will drop. This means that more and more of the source voltage e_1 will be dropped in the plate resistance R_P of tube VT_1. Hence, the output voltage e_2 will be reduced with increases in frequency.

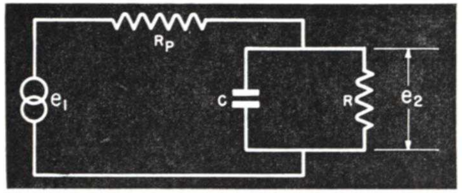


FIG. 16. The circuit in Fig. 15 can be simplified to this form.

This variation in the load impedance with frequency also causes phase-shift difficulties. The total plate current must equal the vector sum of the currents through the resistive and capacitive portions of the load; therefore, as the capacitive current increases because of the reduction in reactance, the total current becomes increasingly out of phase with the output voltage e_2. This is the same as saying that the output voltage is becoming increasingly out of phase with the input voltage, because the input voltage and the total current are in phase. Hence, the variation in the load introduces a phase shift.

Since this difficulty is caused by shunting capacities, the most direct remedy is to reduce these capacities as much as possible. Miniature tubes and sockets are used to reduce the tube capacities, and the parts are mounted well away from the chassis to reduce the capacity of the resistors and of the coupling condenser to the chassis. Even when these capacities have been reduced as much as is practical, they are still too high for the very wide frequency range we need.

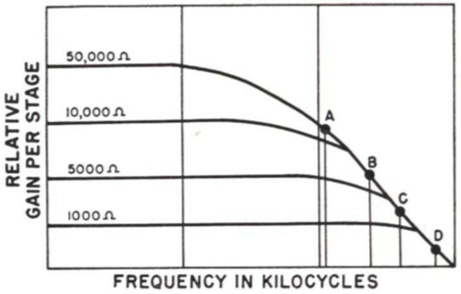


FIG. 17. These curves show that a reduction in the plate-load resistance of an R-C coupled stage improves frequency response at a sacrifice in gain.

As Fig. 17 shows, it is possible to reduce the effects of these shunting capacities further by reducing the plate load resistance. In drawing this figure, we assume that the shunting capacities had the same value in each case. (The actual point at which the curves begin to slope will, of course, depend on what the capacity is. The larger the capacity, the lower the frequency at which the roll-off will occur. In general, however, the curves will have the same shape regardless of capacity.)

You will recall from your studies of wide-band amplifiers that the pass band is considered to extend out to the point where the relative gain is about 70% of the maximum gain. If we follow this principle in examining the curves in Fig. 17, we find that the pass band extends to point A when the load is 50,000 ohms. When the load is reduced to 10,000 ohms, the response extends to point B; reducing it to 5000 ohms extends the response to C; and reducing it to 1000 ohms extends the response to point D. In other words, reducing the load gives an increasingly wider pass band, because the response is extended toward the high-frequency end of the band.

As you can see, however, this extension of the pass band is accompanied by a reduction in gain at all the other frequencies. There is, therefore, a definite limit to how far we can reduce the load impedance. Even though we use tubes of high mutual conductance, the load resistance must be kept at a reasonable value so that normal gain can be obtained. The high-frequency response must therefore be extended in other ways. There are two basic methods of doing this; let's study them now.

SHUNT COMPENSATION

One of the basic ways of extending the high-frequency response is shown in Fig. 18. An inductance coil L_1 is added in series with the plate load resistor. This coil increases the high-frequency response because it forms a parallel-resonant circuit with the shunting capacity. This is called "shunt compensation" because the inductance is in shunt (in parallel) with the capacity (represented as C). The coil is selected so that the circuit is made resonant at a frequency at or above the highest picture frequency to be passed. The resulting over-all response of the stage is then the combination of the resonance curve and the normal response curve.

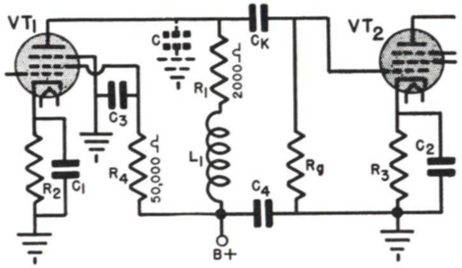


FIG. 18. Coil L_1 in this R-C coupled amplifier provides shunt compensation to offset attenuation of high picture frequencies by C.

If the capacity remains fixed, the resonant peak can be moved to the right or left by choosing different values for the coil L_1, as shown in Fig. 19. The actual height of the peak depends on the Q; if we increase the Q by increasing the inductance without changing the load resistance, we will get a higher peak at a lower frequency. The choice of the coil therefore depends on the frequency range desired and on whether the compensation should raise the high-frequency response above normal.

Since the effect of such a coil is to give a peak in the response, it is known as a "peaking" coil.

The introduction of the coil to cause resonance at the higher-frequency end of the pass band tends to reduce the phase shift. Unfortunately, the values of inductance and load resistance that give the best phase characteristics are not the ones that give the best peak output, so a compromise is necessary.

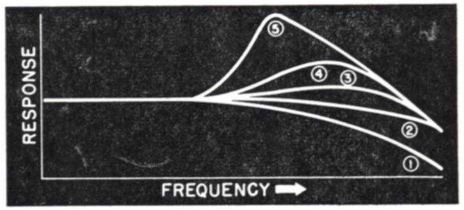


FIG. 19. These curves show how the high-frequency response of the amplifier shown in Fig. 18 depends upon the Q of coil L_1. Curve 1 shows the effect of a low-Q coil. Curve 5 the effect of a high-Q coil; the other curves show the effect of intermediate values of Q.

SERIES COMPENSATION

Another system of compensating for the roll-off in the high-frequency response is to install an inductance in the position occupied by L_1 in Fig. 20. This system is known as "series compensation," because the coil is in series with the signal path.

The effect of this inductance is to split the shunting capacity in half. That is, the output capacity of VT_1, represented as C_A, and the input capacity of VT_2 (C_B) are separated by coil L_1. (Insofar as R_1 is concerned, therefore, it is now shunted only by C_A.) These two shunting capacities and L_1 now form a low-pass filter of such a nature that all frequencies are passed that are below the frequency at which L_1 and C_B become series resonant; above this point, however, the frequency response is cut off sharply.

Since R_1 is now shunted by a much smaller condenser than the one that shunted it before L_1 was installed, the frequency response is extended considerably. As a matter of fact, the characteristics of the filter are such that it is superior to shunt compensation with respect to frequency response, minimum phase shift, and output.

The exact response obtainable with series peaking depends on where coil L_1 is located. As shown in Fig. 21, there are three possible positions for coil L_1. In Fig. 21A, it is located between the load resistor and the coupling condenser, a position that makes the capacity C_B include that between the coupling condenser and ground.

In Fig. 21B, the coil is on the other side of C_1; now the C_1-to-chassis capacity is part of C_A.

In Fig. 21C, the coil is located between the plate of tube VT_1 and the entire coupling network. Now the capacity C_A is just that of the tube VT_1, whereas C_B includes all other shunting capacities.

In each of these cases, the position of the coil determines how much capacity C_A has and how much C_B has. Since the relative capacities of these two have a pronounced effect on the action of the filter, it is possible to change the resonant point by changing the position of the coil in the circuit.

You will notice that a resistor R_3 is in parallel with the series peaking coil in each case in Fig. 21. In a shunt compensating circuit, the load resistor controls the Q, but the Q for a series peaking coil is not similarly controlled. Therefore, a resistance is added in parallel with the inductance to adjust the Q properly. As a matter of fact, the coil is made with a resistor as an integral part of the assembly: the coil itself is wound right around the resistor so that the two form a single unit. This adjustment of the Q affects the response to frequencies near the resonant cut-off frequency.

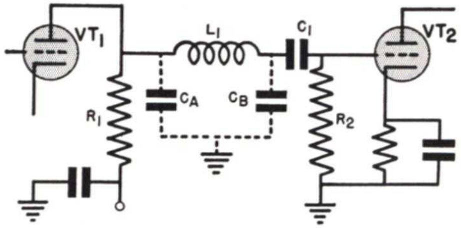


FIG. 20. A series-compensation circuit.

Since a combination of series and shunt peaking gives better response than either alone, it is very common to find both used together in a circuit like that shown in Fig. 22. Here, shunt peaking coil L_2 is shunted by R_4 to give a lower Q than is provided by R_1. (This is necessary in this particular circuit because the value of R_1 is fixed by other requirements at a resistance that is too small to give the Q desired.) The combination C_2-R_5 gives low-frequency compensation in this stage also.

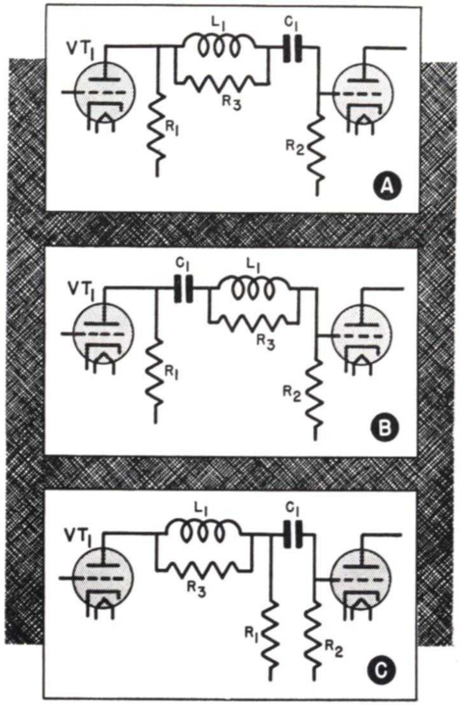


FIG. 21. Three possible positions for the peaking coil in a series compensation circuit.

One stage usually cannot give all the compensation needed, so, to even out the response, it is fairly common to use more than one stage of amplification and to equalize them differently. Thus, you may find two triode stages used to get this equalization, even though the net amplification is no more than could be obtained from a single pentode stage.

Summing up for a.c. amplifiers—one to three stages of resistance-capacitance coupled amplifiers are used in modern sets. Several forms of low- and high-frequency compensation are used with these amplifiers to extend the frequency range and remove frequency and phase distortion as much as possible.

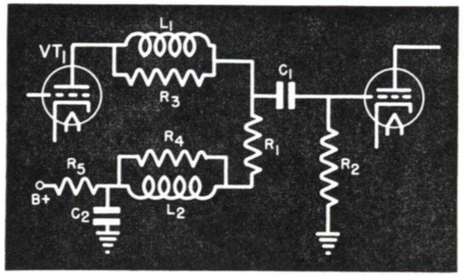


FIG. 22. A combined shunt-peaking and series-peaking coupling circuit. Coil L_1 is the series-peaking coil; coil L_2 is the shunt-peaking coil.

The use of coupling condensers provides an advantage in that it makes each stage independent of all others in its operating potentials. Bias voltages can be fixed in each stage independently. However, coupling condensers affect the low-frequency response and remove the d.c. component of the video signal. Since, as we showed in Figs. 4 and 5, it is necessary to have this d.c. component of the video signal, we must either use d.c. amplification or restore the d.c. component if we are to get the pedestals lined up again so that the circuit will respond properly to any changes that may occur in the background illumination.

Let's first see how the d.c. amplifier differs from the a.c. form, before we study the restorers.

Fundamental D.C. Amplifier

Basically, a d.c. amplifier is one that does not contain blocking condensers or transformers. (The "d.c." in this name can mean either "direct-current" or "direct-coupled"; both names are applied to this amplifier.) This makes it possible to pass along the d.c. component of the signal as well as the a.c. component.

The d.c. component acts throughout a d.c. amplifier as a variable bias that accompanies the signal and arranges the operating points of the amplifier so that the pedestals in the signals will be lined up regardless of the average brightness or darkness of the particular scenes.

Fig. 23 shows a basic circuit of this kind. Here, tube VT_1 is the video detector, which is fed from the i.f. coil L_1. Its load resistor R_1 has the shunt peaking coil L_2 in series with it to increase the high-frequency response. Notice that the control grid of VT_2 is directly connected to this load; no coupling condenser is used. Hence, the d.c. level of the signal is passed right on to tube VT_2. That is, the d.c. drop across R_1 acts as a variable bias that follows the signal brightness level. This bias is applied to VT_2 and causes its operating point to shift as the brightness changes.

You will observe that the plate circuit of tube VT_2 contains the series-peaking coil L_3 shunted by R_6, and that its load resistor R_7 has L_4 as a shunt-peaking coil in series with it. R_8 and C_5 provide a low-frequency compensation as in other circuits we have studied.

The control grid of the picture tube is connected directly to the plate end of the load for VT_2; again, there is no intervening coupling condenser. The picture tube cathode is connected to the chassis by condenser C_6, so the a.c. signal across the L_4-R_7-C_5 load is applied between the grid and cathode of the picture tube. There is also a path from the grid to the cathode (through L_4, R_7, and R_8) across which is developed a d.c. that varies as the

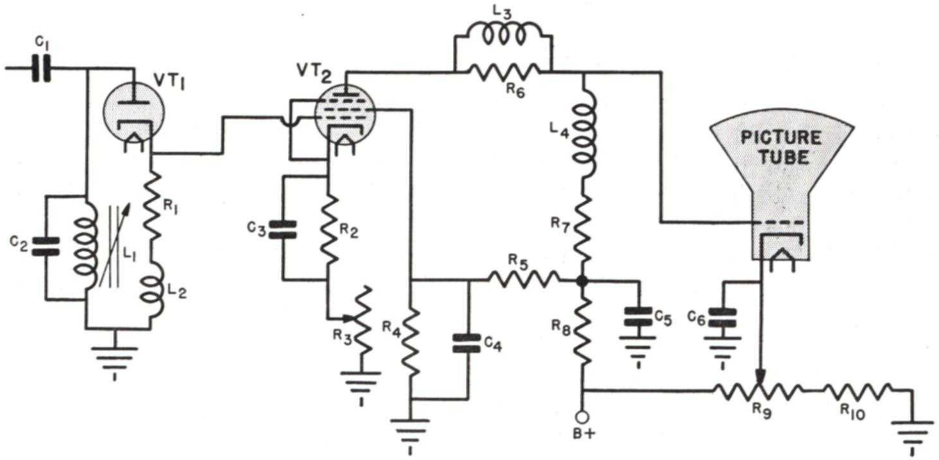


FIG. 23. A basic d.c. amplifier.

brightness changes. Therefore, both the d.c. and a.c. components of the signal are applied to the picture tube, with the result that the pedestals stay lined up.

The absence of coupling condensers removes one of the factors that can cause a low-frequency roll-off, but we still have the cathode circuit of VT_2 to worry about, plus the fact that the over-all design may be such that the low-frequency response is limited. Therefore, although low-frequency compensation is not as necessary in the direct-coupled amplifier as it is in an a.c. amplifier, a certain amount of this compensation may be needed.

TWO-STAGE AMPLIFIER

The single-stage direct-coupled amplifier in Fig. 23 has several factors to recommend it. We do not need voltages as high as are needed in multi-stage d.c. amplifiers, and the likelihood of motorboating is not great in a single stage. Further, there is no need for d.c. restoration, since the d.c. component of the signal is kept throughout.

The only things wrong with the circuit are the facts that there might not be enough gain and that there might not be sufficient high-frequency compensation. For these two reasons, it is sometimes necessary to use a two-stage direct-coupled amplifier.

At once we run into a supply voltage problem. Since the stages are directly coupled, the plate supply of the first stage affects the bias of the second, and the plate supply of the second must be higher than that of the first to get normal operation. Let's run through this problem before studying a typical two-stage amplifier.

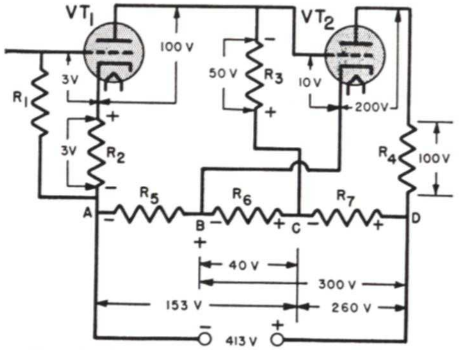


FIG. 24. The power-supply arrangement used for a basic two-stage d.c. amplifier.

Power Supply. Fig. 24 shows an elementary two-stage direct-coupled amplifier. Let's assume that the voltages marked on the diagram are those we require for operation. That is, tube VT_1 requires a 3-volt bias for 100 volts on the plate. The normal VT_1 plate current through load resistor R_3 causes a 50-volt drop. Therefore, the total supply for this tube is 50 + 100 + 3 or 153 volts. This is obtained from the voltage divider between terminals A and C.

For tube VT_2, we need a 10-volt bias and 200 volts on the plate, and the plate current causes a 100-volt drop in the load R_4. First, we have to get the 10-volt bias. We have a 50-volt drop in R_3, which has the right polarity but is 40 volts too much. Therefore, the cathode of VT_2 must be returned to point B on the voltage divider so that it will be 40 volts *negative* with respect to point C. Then, the 40 volts developed across R_6 will be opposed by the 50 volts across R_3 with the result that the difference of 10 volts will appear as the bias for tube VT_2.

Next, we require a total of 300 volts between the cathode and B+ for this tube (200 volts for the plate plus 100 volts that is dropped across R_4). Hence, 300 volts must be developed between point B and point D of the voltage divider. Since there is a 40-volt drop between B and C, a voltage of 260 volts is needed across R_7. The total voltage must therefore be 260 plus 153 or 413 volts. Notice that this direct coupling of two stages makes it necessary to have a high B voltage available even though the plate voltages needed are quite ordinary.

Typical Circuit. A typical two-stage d.c. amplifier circuit is shown in Fig. 25. We can break down the supply circuits of this amplifier into the basic elements shown in Fig. 26. The total supply here is 215 plus 120 or 335 volts. We can analyze this circuit in much the same way as the circuit of Fig. 24 by noticing that the voltages are all measured with respect to ground.

The cathode of VT_2 is tied through R_5 and R_1 to a point that is at —120 volts with respect to ground. There is a 3-volt drop across R_5, which becomes the bias for the grid of VT_2, since the grid returns to this point through R_3.

The plate current of tube VT_2 flows through R_{10} and R_9 to the +215-volt terminal of the power supply. In addition, R_{11} acts as a bleeder to cause additional current to flow through R_9. The total drop across R_9 is therefore more than 215 volts; it is actually 225 volts, which makes the junction of R_9 and R_{11} be at a potential of —10 volts with respect to ground. There is an additional 9 volts across R_{10}, so the plate of VT_2 is —19 volts and its cathode is —117 volts with respect to ground. The plate voltage is the difference between the two, or 98 volts. Although both the plate and the cathode are negative with respect to ground, the cathode is more negative than the plate, which means that the plate is positive with respect to the cathode, as is required.

The bias needed for tube VT_3 is about 4 volts. The grid of this tube is at a potential of —10 volts with respect to ground, because it is connected to the junction of R_9 and R_{11}. To produce the proper bias, the cathode of VT_3 is tied to a terminal that is —6 volts with respect to ground.

The plate voltage needed for VT_3 is about 134 volts, and there is a drop of 81 volts in its load resistor. The total supply needed for VT_3 is therefore 6 + 134 + 81 or 221 volts. The total B supply is 120 + 215 or 335 volts.

As you can see, high power-supply voltages are required for multi-stage direct-coupled amplifiers — so high that practically all circuits use only one or two stages. Occasionally three stages are used, but any more than this would be impractical because of the excessively high voltages that would be needed.

Signal Operation. Returning now to Fig. 25, we can analyze the circuit in regard to its operation on the signal. Tube VT_1, the video detector,

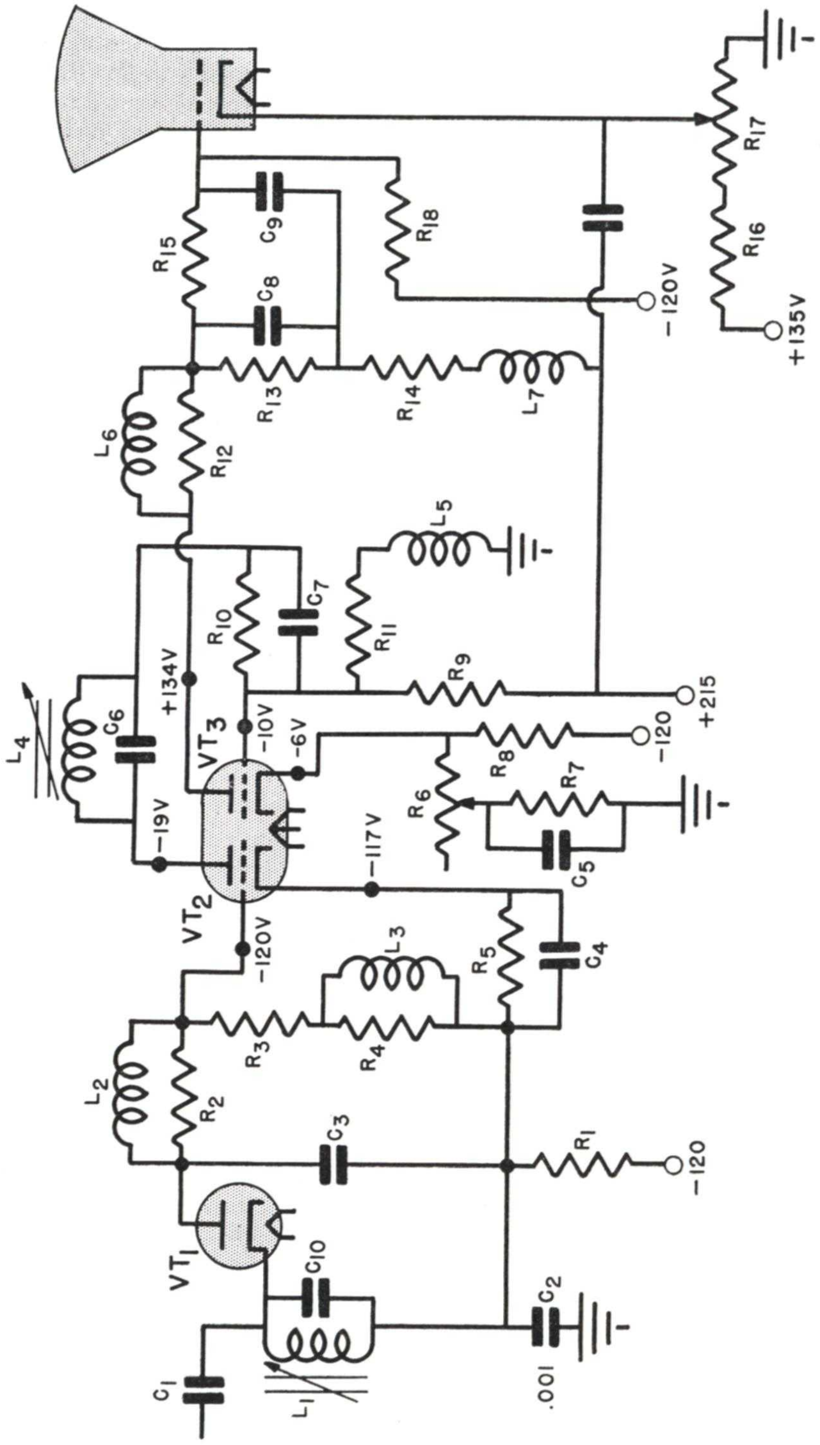


FIG. 25. The schematic diagram of a typical two-stage d.c. amplifier.

feeds into load R_3, shunt compensating coil L_3, and series compensating coil L_2. The signal from this stage is fed directly to the grid of VT_2.

In the plate circuit of VT_2 are a coil L_4 and condenser C_6 that together form a 4.5-megacycle trap. This trap, which is called a "grain" trap, eliminates whatever beat there may be between the sound and video carrier signals; such a beat, if not removed, would cause a fine-grained dot pattern on the picture.

The plate of VT_2 is directly coupled through R_{10} to the grid of VT_3, and the plate of VT_3 is in turn directly connected through its load and a filtering circuit to the grid of the picture tube. Effectively, R_{15}, C_8, and C_9 form a low-pass filter that tends to eliminate components that are higher in frequency than the desired signal.

Although triode tubes are shown in this circuit, pentodes might equally well have been used. Also, separate tubes might have been used instead of the dual triode shown here. Such minor changes could be introduced without making any basic modification of the circuit.

From what we have said, you can see that the direct-coupled amplifier works much like an a.c. amplifier. Its advantages are that the d.c. component of the signal is not removed and that the elimination of coupling condensers helps to maintain a good low-frequency response.

Its disadvantages are that a high operating voltage is needed and that the very good low-frequency response makes motorboating likely to occur. Motorboating may result from the fact that any slow supply-voltage

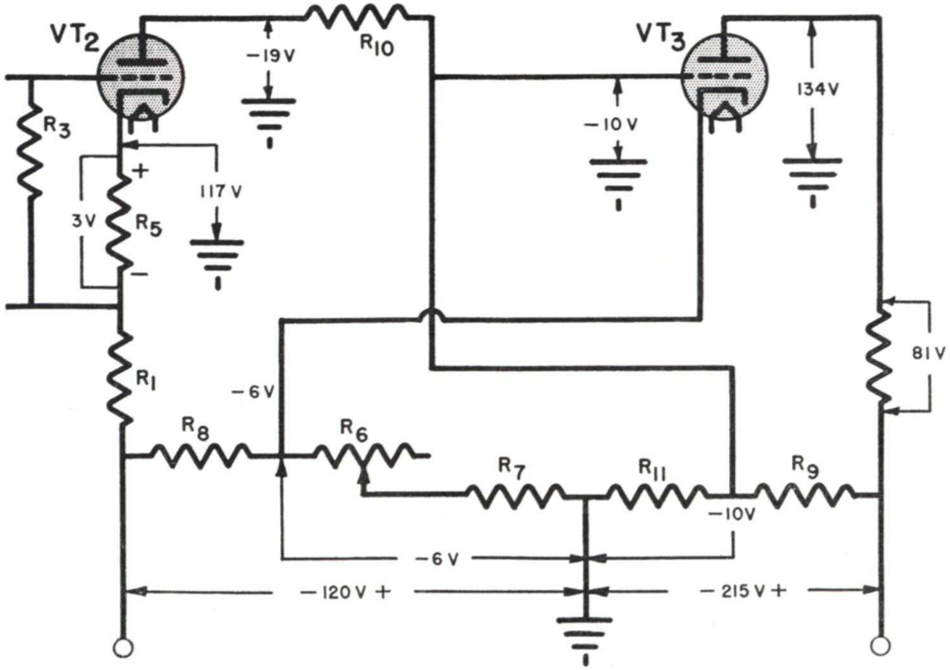


FIG. 26. The power-supply circuits of the amplifier shown in Fig. 25 can be drawn in this form to make it easier to understand the requirements the supply must meet.

change in the first tube will affect the bias on the second, causing a plate current change in the second tube that will aid the change in the first tube. For example, suppose the plate current of the first tube drops. This will reduce the bias on the second tube; its plate current will consequently increase, causing a greater drop in the power supply. This drop will reduce the plate voltage on the first tube decreasing its plate current further and thus continuing the action of decreasing the bias and increasing the current in the second tube. An opposite series of actions will occur if the plate current of the first tube increases. Once this action starts, the circuit is likely to go into oscillation or to motorboat, because the plate current of the second tube can be driven between saturation and zero alternately.

To prevent this from happening, the power supply must be carefully designed to have good regulation. This means it must have low internal resistance and use high-capacity condensers, which makes it more expensive to build.

The a.c. amplifier does not have these two disadvantages, and d.c. restoration will give back the d.c. level. Before we go on to d.c. restorers, however, let's learn more about brilliancy and contrast controls.

BRILLIANCY CONTROL

The picture tube must have an initial bias that will set its operation at the proper point on the brilliancy-grid voltage curve. The operating point is set at the point of greatest curvature, as at point X in Fig. 27A. This point then represents the "black" level at which the pedestals line up; the video signal swings the grid toward zero to give increases in brilliancy, and the sync signals make it more negative so that the screen is blanked completely during the sweep retraces.

Let us momentarily assume that we can couple the signal to the picture tube through a coupling condenser C_1, as shown in Figs. 27B and 27C. The initial brilliancy bias must then be obtained from the power supply. The variable bias control resistor R_2 in both these figures can be used to adjust the bias to cause operation at the proper point on the curve. The fixed resistor R_3 is added to prevent the setting from ever being reduced to zero; this is necessary because the tube might be ruined by the excessive current flow if the bias became zero for very long.

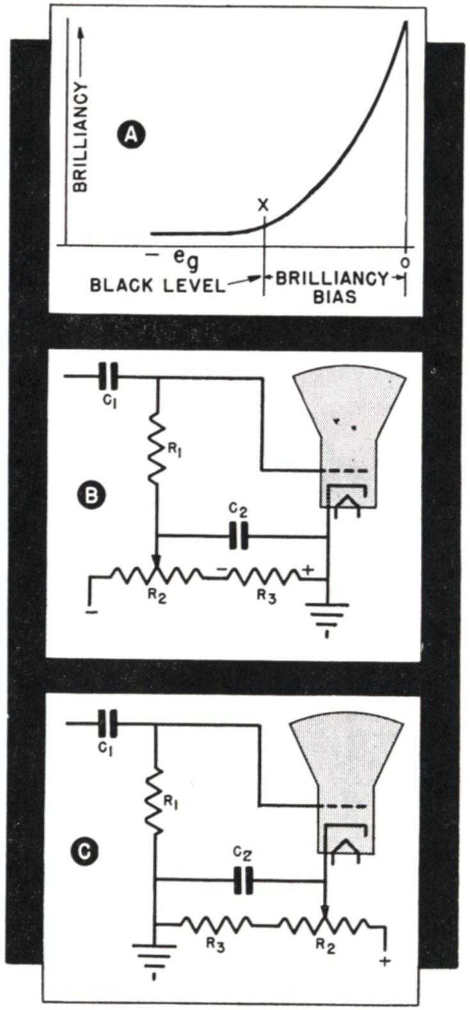


FIG. 27. Basic methods used in a.c. amplifiers for setting the operating point of the picture tube at the desired point on the tube characteristic (A) are shown in (B) and (C).

Since any change in the setting of the variable bias control changes the average brilliancy of the scene by moving the "black" level, it is known as the brilliancy control. If the brilliancy control is set so that the bias is not sufficiently negative, the overall picture will be light; the retrace lines can be seen if the control is far enough from the proper setting. If the control is set so that the operating point is too far in the negative direction, the darker portions of the scene will be telescoped together, with the result that some portions that should be gray will be black. On some sets, the brilliancy control is a screw-driver adjustment; on most, however, it is operated by a front-panel knob so that the set owner may adjust the brightness of the picture to a level that suits him.

Of course, a d.c. restorer (which we will study later) would be needed in circuits like those in Figs. 27B and C. However, these are the basic methods used in a.c. amplifiers for getting the initial bias for the picture tube.

The absence of a coupling condenser in a d.c. amplifier means that the drop in the load of the last video stage will affect the bias of the picture tube. Thus, in the direct-coupled system, the inital bias (the bias when no signal is being received) is obtained from the d.c. voltage drop across R_7, as shown in Fig. 28A. The drop across this resistor is determined by the plate current of VT_2; with usual current and load values, this drop is normally more than is wanted.

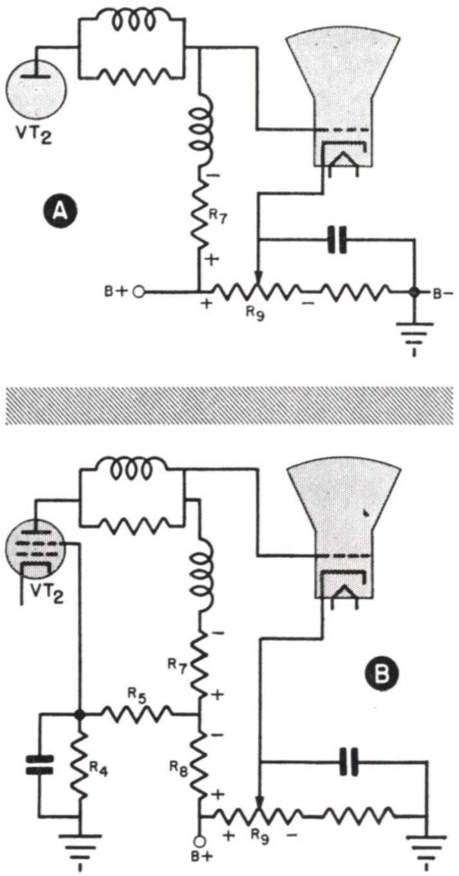


FIG. 28. Two methods of obtaining the initial bias on the picture tube when a d.c. amplifier is used.

In the circuits in Figs. 23 and 28, therefore, there is a control in the cathode circuit of the picture tube that makes it possible to apply a bucking voltage that will reduce the bias. When the slider on R_9 is all the way to the left in Fig. 28A, the bias is that across R_7 alone. However, when the slider is moved toward the right, a voltage of opposite polarity is added between the cathode and grid of the picture tube, thus reducing the bias. Resistor R_9 thus acts as the brilliancy control.

The arrangement shown in Fig. 28A has one serious defect. If tube VT_2 ever became defective so that its plate current was cut off, there would be no voltage across R_7, in which case there would be a positive voltage applied to the grid of the picture tube from the drop across R_9. Since a positive voltage on the picture tube grid could ruin the picture tube, the circuit is sometimes modified as shown in Fig. 28B to prevent such an occurrence. In this latter circuit, a bleeder resistor R_4 draws an additional current through resistors R_8 and R_5. The drop across R_8 is dependent on both the plate current of VT_2 and the size of the bleeder resistor R_4. When the set is operating normally, this drop increases the bias on the picture tube, an effect that is compensated for by adjusting R_9 so that it provides a greater bucking voltage. If the tube VT_2 becomes defective, there will still be a drop across R_8 because of the bleeder current. The values of the various resistors can be so chosen that this drop will keep the grid of the picture tube at a safe voltage.

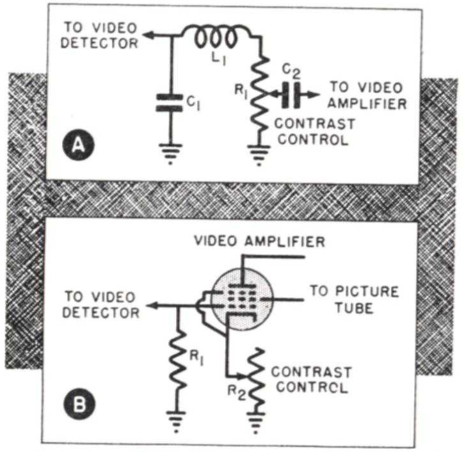


FIG. 29. Two kinds of contrast controls.

A basically similar arrangement was shown earlier in Fig. 25. Here, the voltage fixing the initial operating point of the picture tube is determined by the d.c. drop in R_{14} and R_{13} caused by the plate current of VT_3 and also by the voltage division along the path R_{18}, R_{15}, R_{13}, R_{14}, L_7 from the —120-volt to the +215-volt terminal of the power supply. If tube VT_3 should fail for any reason, the drop across R_{13} and R_{14} caused by the flow of its plate current through them would disappear, but the voltage resulting from current flow through the other path would keep the tube safely biased.

In normal operation of the circuit in Fig. 25, the picture tube grid is slightly positive with respect to ground, but its cathode is considerably more positive. As a matter of fact, under normal operating conditions, the grid-to-ground potential is about +14 volts and the cathode-to-ground potential is about +42 volts. The difference represents a bias of approximately —30 volts on the grid. The brilliancy control R_{17} permits adjustment of the cathode potential, so this bias level can be varied.

Service Hint. When d.c. coupling is used, any defect that reduces the plate current of the output video amplifier, or that increases the plate current of the first video stage, will cause a bright screen. Conversely, either an increased plate current in the output video stage or a decreased plate current in the first video stage will reduce the screen brilliancy.

Notice that this applies ONLY to d.c.-coupled video stages. If a.c. coupling is used, the coupling condensers block the d.c. path, so each stage is independent as long as there is no leakage in the coupling condensers. Hence, in a set using a.c. video coupling, changes in video amplifier plate current cannot normally affect the initial brilliancy setting of the picture tube.

CONTRAST CONTROL

The contrast control is basically the "volume" control for the picture signal. When the set does not have a.g.c., the contrast control is always arranged so that the gain of the i.f. and r.f. stages can be varied with it. When a.g.c. is used to control these stages and prevent overloading, the contrast control is in either the a.g.c. circuit or the video amplifier.

In any case, the contrast control is used to increase gain on weak signals and to reduce gain on signals strong enough to overload the set. When used in the video amplifier, it may be either a control used to vary the signal itself (see Fig. 29A) or one used to adjust the gain by varying the operating bias (Fig. 29B). The latter control is favored because it introduces fewer difficulties with the shunting capacities across the control and its terminals.

The contrast control gets its name from its effect on the signal. When the gain of the receiver is increased, the signal is "stretched" so that the contrast range from white to dark is increased. The contrast control is therefore used to adjust the peak voltage value S of the signal, as shown in Figs. 30B and 30C. Thus, with the signal shown in Fig. 30B, we have a certain range from light to dark. If the gain of the set is increased, the value S is increased (Fig. 30C). As you can see, we have the same basic signal form, but its amplitude has increased. If the black level remains the same, the peak value will show up as a whiter signal.

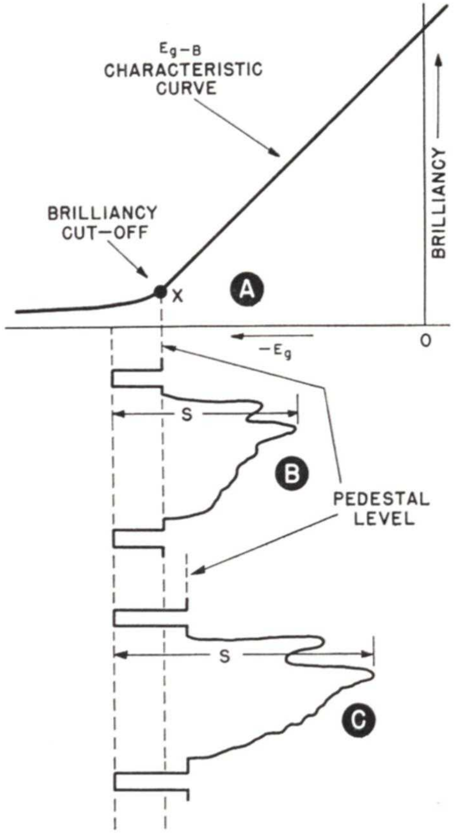


FIG. 30. What the contrast control does to the video signal (B) is shown in (C).

Notice, however, that the stretching of the signal has also increased the amplitude of the sync pulse, with the result that the pedestal level has moved. In other words, the pedestal level in Fig. 30C no longer lines up with the operating point and with the pedestal level in Fig. 30B. Therefore, if it is necessary to increase the contrast control setting, it will also be necessary to vary the brilliancy control to bring the pedestal levels back into line with the brilliancy cut-off X on the curve. If this is not done, dim retrace lines may be visible in the picture.

In general, this "interlock" between the brilliancy and contrast controls occurs regardless of where the contrast control may be.

However, when the contrast control is in a direct-coupled video amplifier like that in Fig. 23, adjustment of the control automatically tends to reset the brilliancy properly. In this circuit, adjustment of resistor R_3 varies the contrast by changing the bias on tube VT_2 and hence changing the gain of this stage. At the same time, this adjustment changes the plate current through R_7 and thus changes the bias applied to the picture tube, thereby resetting the brilliancy level. The circuit arrangement is such that increasing the contrast also tends to move the brilliancy cut-off point in a more negative direction, which automatically tends to line up the pedestals. This is a desirable feature if the parts values used in the circuit can be selected so that the compensation is exact.

D.C. Restorers

To review briefly, Fig. 31 shows the difference between the a.c. and d.c. types of video signals. You will recall that in the a.c. versions shown in D, E, and F of this figure, the areas on either side of the reference line are equal (because the average of an a.c. signal is always zero). As a result, the pedestals and sync pulses of the a.c. signals do not line up if they are associated with lines of different brightness: those representing the brightest lines go farthest below the zero reference line. The displacement of the peak of each sync pulse from the zero line is proportional to the brightness of the scanning line with which that pulse is associated.

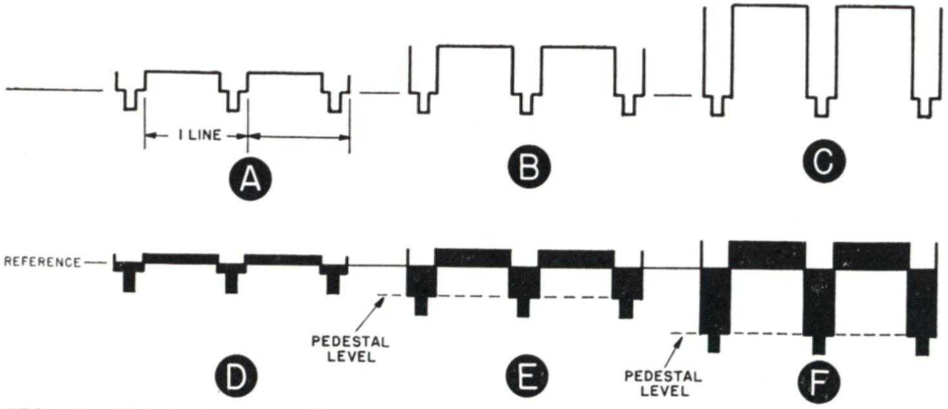


FIG. 31. This is the same illustration that was shown earlier in Fig. 3; it is repeated here for your convenience.

This last fact makes d.c. restoration possible. Fundamentally, we can secure d.c. restoration by applying the a.c. signal to a rectifier circuit that can develop a d.c. voltage that is proportional to the peak values reached by the sync pulses. This d.c. voltage can then be added directly to the a.c. signal to produce the original pulsating d.c. signal form, examples of which are shown in A, B, and C of Fig. 31.

BASIC D.C. RESTORER

An elementary d.c. restoration circuit is shown in Fig. 32. The a.c. signal voltage is applied to the input terminals of this circuit from the last video amplifier. This a.c. signal passes through coupling condenser C_1 and develops across resistor R_1 a duplicate of itself for application to the grid of the picture tube.

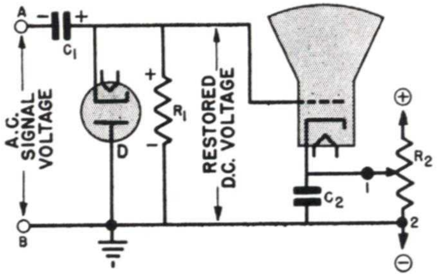


FIG. 32. Simple d.c. restoring circuit in which D is a diode tube. In some television receivers, a germanium crystal rectifier of small area may be used in place of the tube.

The diode rectifier tube D is connected so that it will conduct only when terminal A is negative with respect to terminal B. Therefore, when the a.c. signal is applied, that portion of the a.c. signal that makes terminal A positive with respect to terminal B is merely applied across R_1 to the picture tube. On the negative alternations, however, when the sync pulses make terminal A negative with respect to terminal B, diode D conducts heavily and puts a charge on condenser C_1 that is proportional to the peak value of the negative swing of the a.c. signal.

On the following positive swing condenser C_1 discharges as well as it can through R_1; however, the time constant of C_1 and R_1 is chosen so that it would take several lines and hence several sync pulses of time before C_1 could discharge completely. Usually, the time constant of C_1 and R_1 is made approximately equal to the time required to scan about ten to twenty lines, because this has been found sufficient for the normal changes in brilliancy that occur in the average scene. The time constant cannot be much longer than this, because then it would tend to hold over from one brightness level to the next. If it were made too short, changes in brightness level along a scanning line would begin to affect the background brightness.

When the time constant is correct, conduction of the diode puts a charge on C_1 that is proportional to the average scene brightness. A d.c. voltage resulting from this charge is across R_1 when the diode is cut off; it therefore varies the operating point of the picture tube, thus moving the a.c. signal so that the pedestals line up with the cut-off point of the tube. The brighter the line, the greater the d.c. voltage developed across C_1, and hence the greater the sum of the d.c. and a.c. voltages. That is, the d.c. voltage added to the a.c. signal of a bright scene is much higher than is the d.c. added to the signal of a gray or a black scene. This means that brighter scenes will drive the grid of the picture tube harder in the positive direction, as they should.

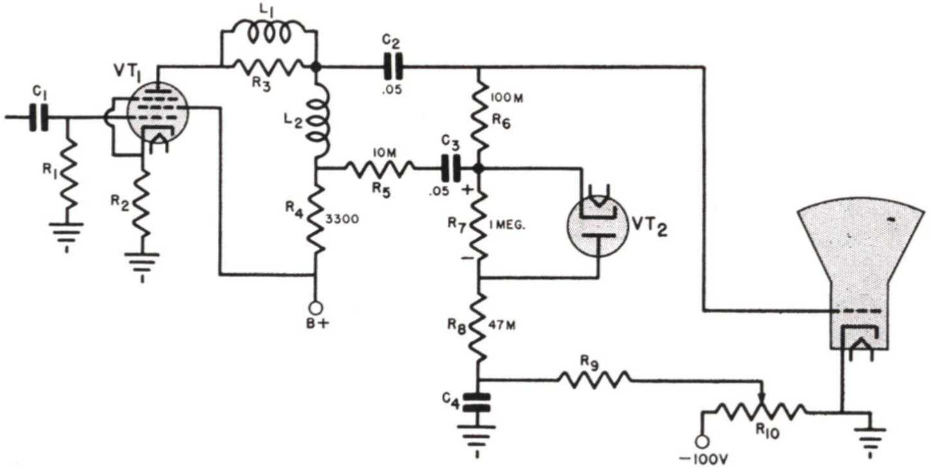


FIG. 33. Schematic of a typical practical d.c. restorer.

PRACTICAL DIODE RESTORER

In an actual receiver, the diode d.c. restorer would not be connected across the entire grid resistor of the picture tube as shown in Fig. 32, because the diode capacity would then be shunting the amplifier plate load. This would seriously reduce the high-frequency response of the system.

A basic circuit that is typical of those actually found in TV receivers is shown in Fig. 33. Here, VT_1 is the tube in the output video stage. The plate load for this tube consists of the series-peaking coil L_1, the shunt-peaking coil L_2, and the load resistor R_4. Low-frequency compensation is not added in this stage; it probably is used in preceding stages, however.

The a.c. signal that is developed across the load resistor and shunt-peaking coil is applied to the grid of the picture tube through coupling condenser C_2. The a.c. grid circuit for the picture tube consists of the resistances R_6, R_7, and R_8 and the by-pass condenser C_4.

The internal capacity of the diode restorer VT_2 acts as a by-pass across R_7, but the high-frequency components of the signal coming through C_2 are developed across R_6 and R_8, which are sufficiently high in resistance to act as a normal grid resistance.

However, VT_2 must charge a condenser if it is to operate properly as a restorer, and it is isolated from C_2 by R_6. Therefore, the signal is fed to it through condenser C_3. The C_3-to-diode path is isolated from the plate load resistor R_4 by the resistance R_5. However, since most of the high frequencies appear across L_2, there is little other than the middle and low frequencies across R_4; therefore, the by-passing action of the diode on R_4 is small, and R_5 need not be high in resistance.

The d.c. restoration action is much like that in the circuit in Fig. 32. The negative swings of the pedestal and sync pulses are applied to the diode through C_3 (and also through the path C_2-R_6, but this path is not considered to be very effective). When the diode conducts, it charges C_3 (through a path consisting of R_8, R_9, R_{10}, the B supply, R_4, and R_5). When the diode ceases to conduct, R_7 is added to the other resistors in the d.c. path between the terminals of C_3. Since the resistance of R_7 is high in comparison to that of the other resistors in the path, most of the voltage across C_3 is developed across R_7. The time constant of C_3-R_7 is such that C_3 is held at a charge that corresponds to the average brightness of ten to twenty lines. Hence, the d.c. voltage across R_7 corresponds to the average brightness of the scene, and, since it is applied to the grid of the picture tube, it acts to line up the pedestals.

R_9 and C_4 isolate the grid circuit of the picture tube from the brightness control, which is a part of the power pack. The a.c. signal path is through C_4 to the cathode, and R_9 acts as a blocking resistance.

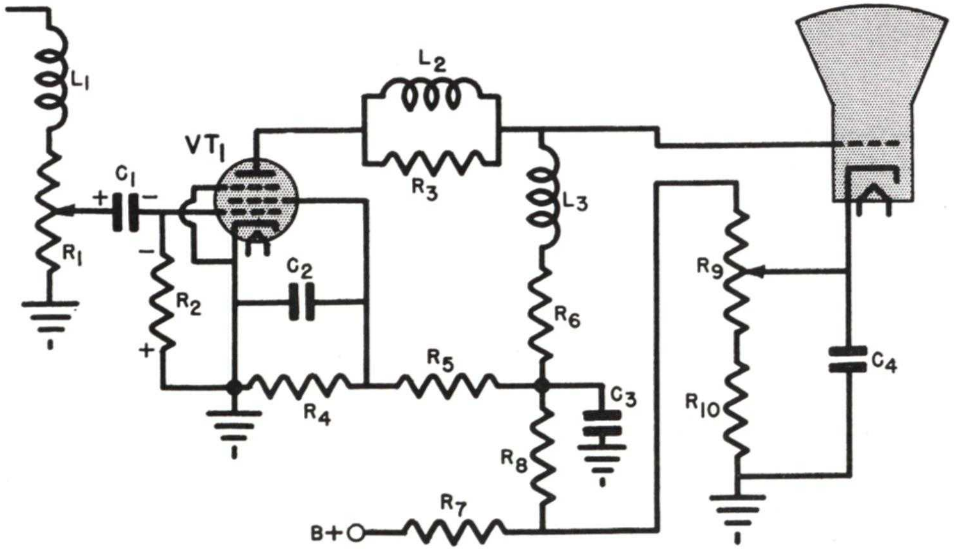


FIG. 34. How d.c. restoration can be produced in the last video stage.

RESTORATION IN LAST VIDEO STAGE

It is possible to make the last video stage do its own restoring as long as it is directly coupled to the picture tube. Fig. 34 shows a typical example of a circuit of this kind. Here, the signal is fed to the grid of VT_1 through the coupling condenser C_1. This is an a.c. coupling, since any d.c. that may be in the signal from the preceding stage will be wiped out by C_1.

The plate load for VT_1 is the series-peaking coil L_2, the shunt-peaking coil L_3, and the load resistor R_6. Resistor R_8 and condenser C_3 provide low-frequency compensation. The signal developed across L_3, R_6, and C_3 is applied directly to the grid of the picture tube.

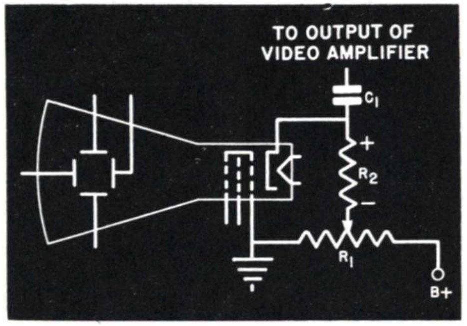


FIG. 35. How d.c. restoration can be produced in the picture-tube circuit.

Restoration. The restoration in this stage occurs in the grid circuit of VT_1. This tube has no initial operating bias; it gets the bias for operation by grid rectification. The action, which is much like that of a grid-leak detector, occurs as follows:

To begin with, you know that the signal applied to the grid of the picture tube must have a positive picture phase—that is, it must swing in the positive direction for increases in brilliancy. Therefore, the signal applied to the grid of VT_1 must have a negative picture phase, because VT_1 inverts the entire signal 180°. (This is not the same kind of phase shift that we studied earlier, because here the action occurs on the entire signal. All frequencies are held in their same relative positions with respect to each other—the entire signal is "flipped over" as a unit.)

Since the signal applied to the grid VT_1 has a negative picture phase, the sync signals drive the grid in the positive direction, and the picture components drive it in the negative direction.

Since VT_1 has no initial grid bias, the sync signals drive the grid positive, and the grid draws current through R_2. This current flow sets up a bias voltage across R_2 that is proportional to the amount the grid is swung positive by the sync signal peaks. This puts a charge on condenser C_1 that creates a d.c. voltage across it having the polarity indicated in Fig. 34. This d.c. voltage acts as a varying bias on VT_1 during the negative part of the signal, moving the operating point of the tube in accordance with the brightness level of the picture to make the pedestals line up.

Basically, the only difference between the grid circuit of the VT_1 stage in Fig. 34 and any similar grid circuit in an a.c. amplifier is the fact that there is no (or little) initial bias in this circuit, which means that the grid of VT_1 can go positive on the signal swings. The time constant of C_1 and R_2 must equal the time of several lines, but this is needed anyway to give reasonable low-frequency response.

Since the pedestals are aligned in the grid circuit, they remain aligned in the plate circuit. The plate of VT_1 is directly coupled to the picture tube grid, so the restored signal is applied to the picture tube much as it is in a d.c. amplifier. In fact, the output coupling is practically identical with that of the d.c. circuits we have studied.

D.C. RESTORATION IN PICTURE-TUBE CIRCUIT

It is also possible to obtain a d.c. restoration action in the picture tube circuit itself when the signal is fed to the cathode. (We cannot obtain restoration by grid rectification in the grid circuit of the picture tube, because we cannot allow the grid to go positive.) The circuit in Fig. 35 is used for this purpose.

In this circuit, the initial operating point is set by the bias developed across R_2. This bias is determined by the beam current of the picture tube, which flows to the cathode through R_2. Resistor R_1 is a vernier brightness control that is used to provide an additional bias to set the final operating point.

It is possible to feed the video signal to the cathode of the picture tube this way as long as it has a negative picture phase. Then the cathode will be swung more negative by brighter elements of the picture; since this is the same as making the grid more positive, we will have normal a.c. signal action.

In addition, d.c. restoration occurs because of a shift in the bias developed across R_2. When the pedestal and sync pulses drive the cathode positive (which is the same as making the grid more negative), the beam current through the picture tube is cut down. This reduces the bias that is produced across R_2. An increase in brightness therefore causes a reduction in the beam current and in the bias applied to the grid of the tube. This allows the grid to go less negative (by making the cathode less positive with respect to the grid) and thus produces a brighter picture.

The coupling condenser C_1 prevents the bias from following changes in the signal too rapidly. The time constant of C_1 and R_2 is such that the bias is determined by the average brilliancy of several scanning lines, as it is in other d.c. restoration circuits.

One manufacturer calls this particular arrangement "automatic brightness control," and another refers to it as "stabilized brightness control." Although it is true that this and all other d.c. restoration methods are effectively variable brightness controls, we usually consider the brightness level to be set by the initial adjustment of the operating point of the picture tube.

CONCLUSION

In the last several Lessons, you have followed the video signal through the r.f.-i.f. section, the video detector, and the video amplifier. This completes the journey of the video signal—it is now applied to the grid of the picture tube and serves to vary the brightness of the spot produced on the face of the tube. However, having a varying spot is not enough; we must move this spot to the proper position to reproduce each element in the scene. Therefore, in addition to varying the content of the electron beam, we must sweep it horizontally and vertically across the face of the tube. Furthermore, it must be kept in step with the transmitter so that each line will start at the proper time.

In the next Lessons, we shall study the circuits that produce sweep voltages and those that synchronize the sweeps with the transmitted signal. These are sections of a TV set that have no counterparts in a sound receiver.

As you will learn, the sweep circuits basically consist of an oscillator, followed by a special wave-shaping circuit or network that is employed to give a voltage of a particular shape. In turn, this voltage is fed through an output or amplifying stage to the deflection plates or deflection coils of the picture tube. One sweep chain operates to sweep the electron beam in the picture tube from left to right in a horizontal direction to form the lines, while another entirely separate sweep chain produces the deflection signal for moving the beam from top to bottom of the picture-tube face.

In the sync-control circuits, the control signal is stripped from the video signal, and then the vertical and horizontal control pulses are separated from each other. These signals are then used to control the frequencies of the oscillators in the sweep circuits so that each line and each frame starts exactly in step with the scanning at the transmitter.

After you have studied sweep circuits and synchronizing control circuits, you will study receiver power supplies, the sound channels, and special systems used in receivers. This will complete your basic theory of television, after which you will go into the study of television antennas, and the installation, adjustment, and servicing of television receivers.

Lesson Questions

Be sure to number your Answer Sheet 54RH-3.

Place your Student Number on every Answer Sheet.

***Send in your set of answers for this Lesson immediately after you finish them,* as instructed in the Study Schedule. This will give you the greatest possible benefit from our speedy personal grading service.**

1. Is the final video amplifier a power or voltage amplifier?
2. If the signal available at the output of the video amplifier has a negative picture phase, how must it be fed to the picture tube?
3. Why is good low-frequency response necessary in a video amplifier?
4. What causes the high-frequency response of a video amplifier to fall off?
5. Is the detail in the picture determined by the high-frequency response or by the low-frequency response?
6. What is done in the video section to eliminate the beat between the sound and video carrier signals?
7. Why does an increase in contrast control setting make it necessary to vary the brilliancy control?
8. Why is it not practical to connect a diode as a d.c. restorer directly across the entire picture tube grid resistor?
9. Which of the following statements is correct: the time constant in the d.c. restoration circuit is such that the condenser is held at a charge that corresponds to the average brightness of: *one line; several lines; several frames?*
10. If a two-stage video amplifier is a.c.-coupled throughout, and the first tube burns out, will the picture tube bias be affected?

Be sure to fill out a Lesson Label and send it along with your answers.

TAKE TIME

Here is a quotation from the *Santa Fe Magazine* which appealed to me as containing much good, common sense. I hope you too will enjoy it—perhaps profit by it:

"Take time to live. That is what time is for. Killing time is suicide.

"Take time to work. It is the price of success.

"Take time to think. It is the source of power.

"Take time to play. It is the fountain of wisdom.

"Take time to be friendly. It is the road to happiness.

"Take time to dream. It is hitching your wagon to a star.

"Take time to look around. It is too short a day to be selfish.

"Take time to laugh. It is the music of the soul.

"Take time to play with children. It is the joy of joys.

"Take time to be courteous. It is the mark of a gentleman."

J. E. Smith