## RIDER'S VOLUME XVIII

# HOW IT WORKS

JOHN F. RIDER PUBLISHER, INC.

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for their preparation of the technical material contained herein.



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# DETECTOR CIRCUITS IN AM-FM RECEIVERS

In the production of many combination am-fm receivers, one of the primary problems is the design of those circuits that are intended to perform a dual function — that is, circuits that will operate on f.m. as well as on a.m. Such design problems are encountered in every part of the receiver up to the audio system. R-f, converter, i-f, and detector circuits all have their own individual design problems in combination am-fm receivers.

In some receivers a single tube is capable of performing the function of a-m and f-m detection. Other receivers have separate tubes for a-m and f-m detection, but in some instances one of these detector tubes has another function besides detection. Whatever the type of circuit designed, the important prerequisite is that it function only on a.m. when the receiver is tuned to the a-m broadcast band and only on f.m. when it is tuned to the f-m band. In this section we will discuss some of the different types of detector circuits that appear in combination am-fm receivers.

### Admiral Model 9B14——9B16

The Admiral models 9B14, 9B15, and 9B16 employ a separate tube for both a-m and f-m detection, but the tube employed for a-m detection is also used as the f-m second i-f amplifier. The complete service data for these models appears on *Admiral pages 18-33, 34 through 18-38 in Rider's Vol. XVIII.* A duo-diode 6AL5 tube is used on the f-m band in a conventional ratio detector circuit. This tube performs only one function — namely f-m detection. A pentode tube, the 6BA6, serves the dual purpose of f-m second i-f amplifier and a-m detector. The schematic diagram of

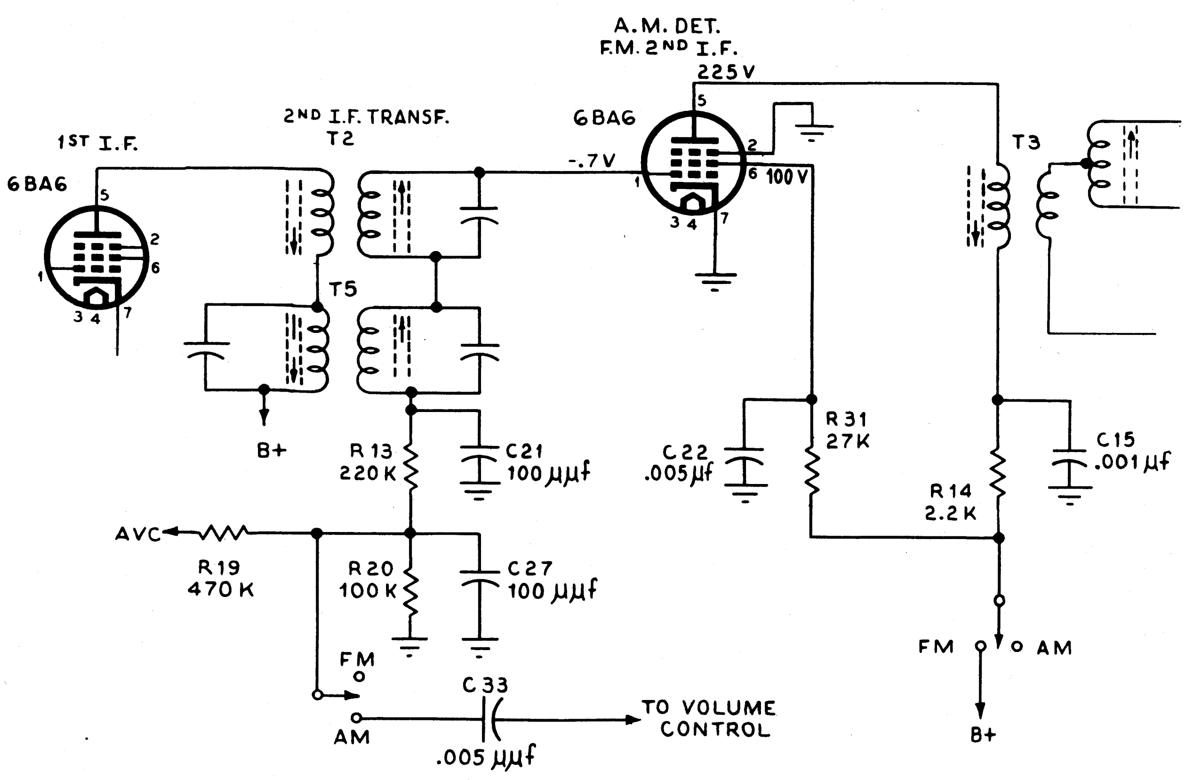



FIG. 1. — The detector and i-f stages of the Admiral models 9B14, 9B15, and 9B16. The second 6BA6 serves as both an fm-if amplifier and an a-m detector.

this circuit is illustrated in Fig. 1. Let us examine this circuit and see how it works when the set is tuned to the f-m band and then to the a-m band.

Although not shown in the drawing, the first a-m and fm-if transformers are in series with each other and precede the first i-f amplifier tube. When the receiver is switched to the f-m band, the primary of the first am-if transformer is shorted and the circuit becomes selective only to the fm-if signal and this signal is fed to the grid circuit of the first i-f amplifier. The output from this tube is coupled to the following stage through the second i-f transformer arrangement as shown in Fig. 1. In this coupling unit *T2* is the fm-if transformer and *T5* the am-if transformer. Since only an fm-if signal is present, transformer *T2* is the active coupling unit due to its being pretuned to the f-m intermediate frequency.

In the f-m position of this set, a switching arrangement in the plate and screen circuits of the second 6BA6 tube enables these electrodes to receive B supply voltage and thus the tube acts as a pentode amplifier to the fm-if signal. The necessary bias for the proper operation of this tube as an amplifier is obtained by the d-c voltage drop across the two series grid resistors *R*13 and *R*20. This voltage is about –0.7 volt. The total value of these resistors is high enough to establish this bias with a very small value of grid current.

Although this d-c bias voltage is shown as –0.7 volt, this is only a typical measuring indication. The exact value of the bias depends upon the average signal strength of the incoming signal and thus may be different from the value mentioned. Due to this phenomenon, this bias is a convenient source for avc voltage on the f-m band. The total bias voltage is not used for avc purposes, only part of it, as can be seen in Fig. 1 since the avc lead is connected to the junction of the two grid resistors, *R*13 and *R*20. In other f-m receivers that employ a ratio detector, avc voltage is often taken from the negative side of the electrolytic capacitor in the output circuit of the detector.

Now let us examine the circuit when the receiver is tuned to the a-m band. In this position the primary of the first fm-if transformer, although not shown in Fig. 1, is shorted and only the first am-if transformer is selective and passes on the am-if signal to the 6BA6 first i-f amplifier tube. The output from this tube is coupled to the grid of the following 6BA6 tube via the second am-if transformer, *T*5, as shown in Fig. 1. The inductance of the coils of the fm-if transformer *T2* is so low as to have negligible effect on the a-m intermediate frequency, and it is, therefore, not necessary to short the coils of *T2*.

At a quick glance at this circuit one might be puzzled as to how the 6BA6 pentode tube, which previously acted as an fm-if amplifier, now acts as an a-m detector. Upon a more thorough investigation of this 6BA6 circuit, its ability to act as a detector will become readily apparent. First of all, note the plate and screen circuits of this tube. When the set is switched to a.m., the supply voltages are removed from these electrodes, and these sections of the tube are, therefore, inoperative. With these electrodes inoperative, the suppressor grid does not have any effect. Thus the only two electrodes left, the control grid and the cathode, must act as the a-m detector — which they do. They function as a diode detector, with the grid acting as the plate of this diode. The action takes place as in the casual type of diode detector. Avc is also established in the normal manner with capacitors *C*21 and *C27* and resistor *R*13 acting as a conventional r-f and i-f filter. The audio signal is developed at the junction of *R*13 and *R*20. By tapping off at this point and feeding the signal through the proper filters, avc voltage is delivered to the necessary tubes of the receiver.

With the receiver in the a-m position, a switch in this part of the circuit allows the audio voltage to be fed to the volume control of the set via the 0.005 $\mu f$ coupling capacitor *C*33, as seen in Fig. 1.

### Farnsworth GK-085 and Firestone 4-A-12

At the beginning of this section it was mentioned that in many am-fm receiver combinations a single tube was used to perform the function of detection of both a.m. and f.m. In glancing through the am-fm receivers in *Rider's Volume XVIII,* you will find many such type circuits. In the two receivers to be discussed here such a multi-purpose tube is employed; and, in addition to serving as an a-m and f-m detector, it also has additional electrodes which serve as the first audio amplifier of the unit. The combination receivers that we have in mind are the Farnsworth models GK-084, 085, 086, 087 and Firestone model 4-A-12. The complete service data for the Farnsworth models appear on *pages 18-6 through 18-12 of Rider's Volume* XVIII and the service data for the Firestone model appear on *pages 18-7 through 18-10* of the same volume.

The schematic representation for each of these special tubes is illustrated in Fig. 2. Each tube is a triple-diode triode but they have somewhat different constructions even though their functions are the same. The 6S8GT tube, shown in part (A) of Fig. 2, is used in the Farnsworth models and the 19T8 tube

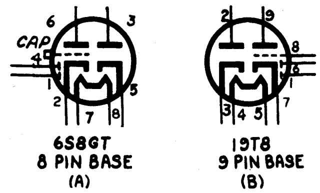



FIG. 2 (A) — Schematic representation of the 6S8GT triple-diode triode tube. (B) The schematic representation of the 19T8 miniature-type tube. Both of these tubes are used for combined a-m and f-m detection.

shown in part (B) is used in the Firestone model. The 6S8GT is a regular sized tube with an octal (8-pin) base. Since there are nine electrodes, including the two for the heaters, and only an eight-pin base, connections to the other electrode, the grid of the triode section, is made through the top of the tube by means of a grid cap. The 19T8 tube is a miniature type that employs a nine-pin base and hence there is no need for a grid cap. For this latter tube a special nine-pin socket has to be employed.

The 6S8GT tube requires 6.3 volts operation for its filaments whereas the filaments of the 19T8 tube requires 18.9 volts. The latter tube is more readily usable in ac-dc receivers than the other type. The 6S8GT tube is usually employed in a-c receivers which use a power transformer with a 6.3-volt filament winding. The circuit connections of both tubes are quite similar so we will show only the hookup for the 6S8GT tube in the Farnsworth models. This circuit is illustrated in Fig. 3. Let us now examine this circuit and see how it functions on the a-m and f-m bands relative to its use as an a-m and f-m detector, avc voltage supplier, and audio amplifier.

The schematic representation of the electrodes of the 6S8GT tube are rearranged from that shown in Fig. 2 (A) solely for the purpose of simplifying the drawing. When the receiver is tuned to the f-m band, the fm-if signal is fed through a series of three i-f amplifiers and then coupled to the 6S8GT tube through the ratio detector transformer. The plates and cathodes of pins 2, 3, 4, and 5 are connected in a conventional type of ratio detector circuit as seen in Fig. 3.* Avs for f.m. is taken from the negative side of the 4 $\mu f$ electrolytic capacitor in the output circuit of the ratio detector. This point is used because the output voltage there changes in accordance with the varying average strength of the incoming signal.

The 22,000-ohm resistor and 0.003-$\mu$f capacitor in the tertiary winding of the ratio detector transformer forms the de-emphasis network of the receiver. The audio signal output from this circuit is fed to a volume and tone control network through the f-m section of the switch as shown in Fig. 3. The audio signal is finally coupled to the grid of the triode section of the tube through a 0.01-$\mu$f capacitor. The plate, pin 6, and cathode, pin 2, are the other electrodes of the amplifier. Bias for this amplifier is obtained by the d-c voltage drop across the 10-megohm resistor in the grid circuit. The signal output from the plate of this first audio amplifier is then coupled to the grid of the audio power output tube.

Up to this point, the cathode, pin 2, has been used

* For a complete discussion of the ratio detector see pages 313-321 in the text "FM Transmission and Reception" by Rider and Uslan, published by John F. Rider, Publisher, Inc.; 480 Canal St., N. Y. 13, N. Y.

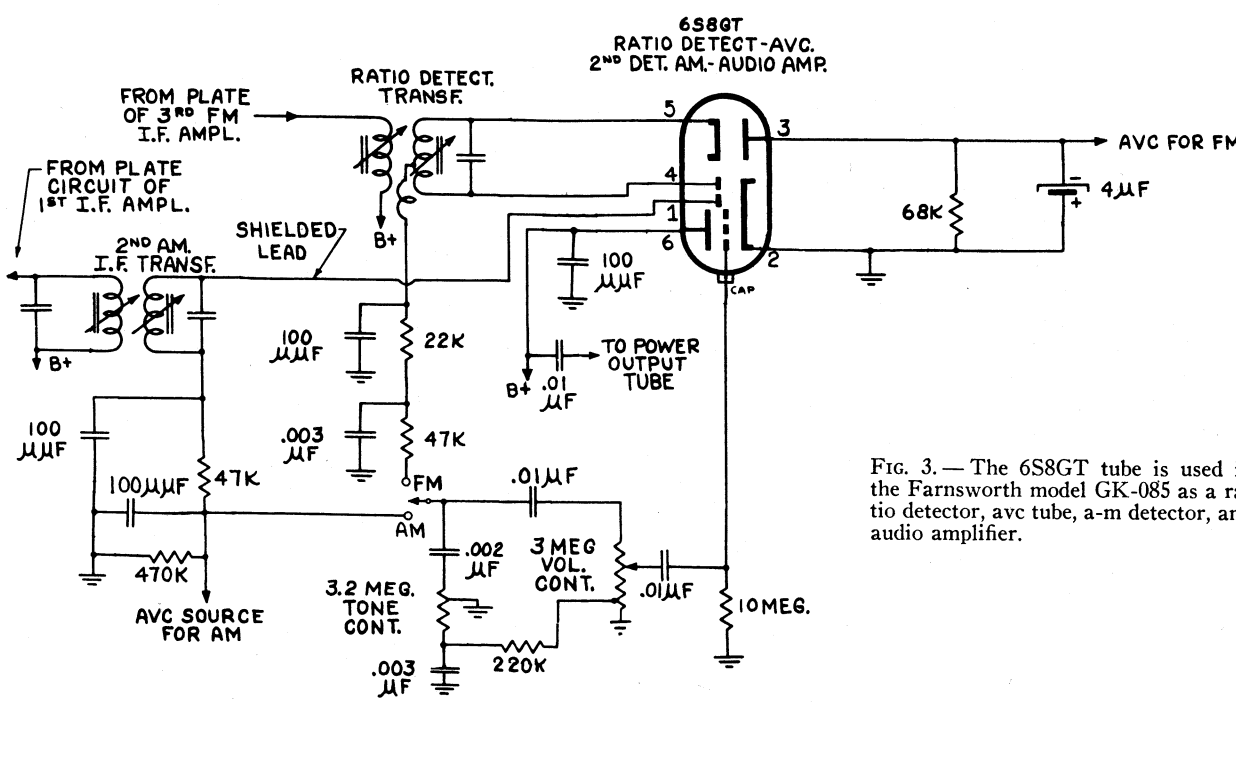



FIG. 3. — The 6S8GT tube is used in the Farnsworth model GK-085 as a ratio detector, avc tube, a-m detector, and audio amplifier.

for two separate applications; one for the ratio detector circuit and the other for the audio amplifier. It also has a third application — that being part of the a-m detector circuit. Let us now examine this circuit when the receiver is tuned to the a-m band. In this position of the im-if signal passes through only one stage of i-f amplification and is then coupled to the plate, pin 1, of the 6S8GT tube through the second am-if transformer. This electrode together with the grounded cathode, pin 2, functions as a conventional diode detector for a-m signals. Since no fm-if signal is present at the ratio detector transformer, the other electrodes, pins 3, 4, and 5 of the 6S8GT tube, are not operative.

The two 100-$\mu\mu f$ capacitors and the 47,000-ohm resistor in the secondary circuit of the second am-if transformer serve as an i-f filter. The output signal from this filter is audio and is applied to the volume and tone control circuits of the receiver through the a-m section of the switch, as seen in Fig. 3. This output signal from the i-f filter is used as an avc source for a.m. The avc voltage for the necessary circuits is obtained after the audio signal is properly filtered in the customary manner.

# THE LOCKED-IN OSCILLATOR DETECTOR

Practically all of the a-m broadcast superheterodyne receivers today employ the simple method of diode detection. It has been the accepted method of detection for a-m signals for a long time. In f.m. the situation is much different. Today there are four methods of f-m detection employed: the limiter-discriminator method, the ratio-detector method, the Fremodyne circuit, and the locked-in oscillator detector. The first three types of f-m detector circuits have been discussed in previous *"How It Works"* books (see *Volume XV and XVII*) and the latter type of circuit will be discussed now.

### Philco 48-482

To date only one manufacturer, Philco Corporation, has employed the locked-in oscillator detector. This circuit has all the necessary features for proper f-m detection, such as negligible response to a-m signals and maximum response and linearity to f-m signals. The locked-in oscillator detector circuit, as its name implies, employs the principle of the locked-in oscillator. A locked-in oscillator utilizing a special tube and circuit construction comprises this f-m detector network. It is used in a number of Philco models but the one to be studied here is model 48-482. The complete service data for this model can be found on *pages 18-91 through 18-107 of Rider's Volume XVIII.* While this f-m detector circuit is more complex than the other three types, the action occurring in the circuit can be understood by careful study of the following analysis.

### Circuit Construction

We will first study the individual functions of the associated circuits and then consider the over-all action of the detector under operating conditions. The schematic arrangement for the circuit under consideration is illustrated in Fig. 1. Similar to the ratio detector, a single tube is employed for the process of f-m detection and a-m rejection. In Fig. 1 all the lettered and numbered designations are the manufacturer's except the symbols *L*1, *L*2, *L*3, and *L*4 which we have inserted for ease of discussion.

There are three tuned circuits in this network that have to be considered, namely, *L*1-*C*403B, *L*3-*C*300A, and *L*2-*C*300B. The resonant frequency of operation of the latter tuned circuit is also determined by ca-

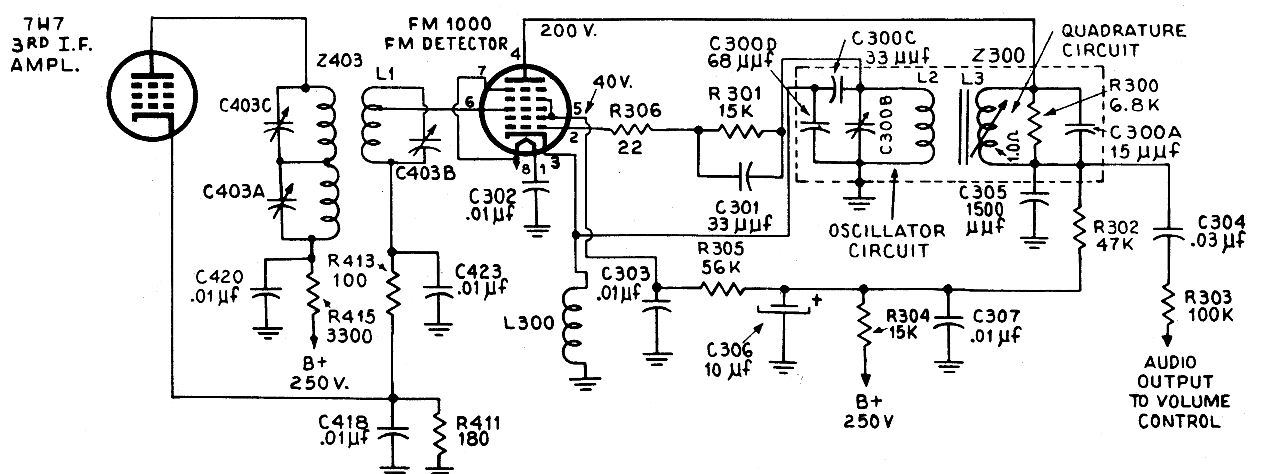



FIG. 1.—The locked-in oscillator detector circuit used in the Philco model 48-482. The FM 1000 tube used in this circuit is a special pentagrid tube.

pacitors *C*300C and *C*300D, as well as trimmer *C*300B. These three tuned circuits are all resonant to the same frequency, which is 9.1 mc, the i.f. of the set. The 7H7 tube is the last i-f amplifier of the receiver. The *Z*403 designation is for the last i-f transformer of the unit and it contains both a-m and f-m sections. However, only those components that are part of the circuit are shown, consequently the secondary of the a-m section of *Z*403 is omitted. The primary of this a-m section, which consists of capacitor *C*403A in conjunction with the coil across it, is illustrated because it is in series with the primary of the fm-if section and constitutes part of the completed f-m signal and d-c path.

The tube used as an f-m detector is designated by Philco as FM 1000 and is of special pentagrid construction, with the second and fourth grids tied together inside the tube and the fifth grid or suppressor connected externally to the filament, pin 8. The first two grids and the cathode of the unit, in conjunction with the associated circuit form a Colpitts oscillator. Components *L*2 and *C*300B in connection with *C*300D comprise the oscillator tank circuit. The parallel circuit arrangement of *R*301 and *C*301 forms the grid-leak bias network of the oscillator. The d-c return path from the oscillator grid is through coil *L*2 to ground. The second grid acts as the oscillator anode. The oscillator signal, which is 9.1 mc, is electron coupled to the plate, pin 4, of the FM 1000 tube. The oscillator anode and the fourth grid, which acts as the screen of the tube, both receive the same supply voltage (because of their internal connections) and they are both at i-f ground potential through the 0.01-$\mu$f capacitor *C*303.

Coupled to the oscillator tank circuit is the tuned plate circuit composed of capacitor *C*300A in parallel with coil *L*3. This circuit, called a *quadrature circuit,* is also resonant to 9.1 mc, the fm-if of the receiver. This tuned plate circuit is called a *quadrature circuit* because it reflects a voltage into the oscillator tank that is 90 degrees out of phase with the oscillator tank voltage. However, the bandwidth characteristics of this circuit are much wider than those of the oscillator tuned circuit because of the parallel 6800-ohm resistor *R*300. The use of this resistor decreases the Q of the circuit and hence increases its bandwidth. This increase in bandwidth is a desirable factor because for proper operation of the detector the impedance of this plate circuit must not change appreciably over the frequency range of the incoming signal. It must be remembered that the incoming signal is frequency modulated and hence is varying in frequency about a mean. A greater bandwidth can be obtained by using a smaller value of *R*300 and hence reducing the Q of the circuit. This further increased bandwidth, although it may be desirable, might result in a complete damping out of the oscillations of the tank circuit. The value of *R*300 that is used is low enough, nevertheless, to cause the bandwidth of the quadrature circuit to be over five times as great as the width of the f-m signal.

**Operating Conditions**

Let us now see how the tube functions with the circuit in operation. The oscillator section operates class C and is so designed that its grid, pin 2, is driven positive over a small part of its positive half cycle of signal; hence the r-f current flow in the tube due to the oscillator is in pulses of short time duration.

When there is no f-m input signal to pin 6, these pulses will continue to flow unchanged to the plate circuit of the tube due to the electron coupling between this electrode and the oscillator circuit. Therefore, the same pulses will flow in the quadrature circuit. Due to the transformer coupling between the quadrature and oscillator circuits, some voltage is fed back to the oscillator tank circuit. Since this quadrature network is always in the circuit the voltage fed back in conjunction with the oscillator voltage that would exist without feedback establishes the operating frequency of the oscillator. This feedback voltage, which in reality is a reflected voltage from *L*3 to *L*2, is also a factor in establishing the relative phase and magnitude of the plate current pulses.

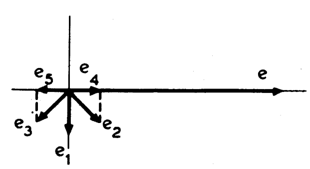



FIG. 2.—The vector diagram indicating the voltage relationships between the signals at the input and oscillator grids of the locked-in oscillator circuit of the Philco model 48-482.

Without any f-m signal applied to the grid, pin 6, but with the proper d-c potentials applied to the FM 1000 tube, the magnitude of these plate-current pulses are constant. However, when an f-m signal is applied to pin 6, the magnitude of these current pulses will vary according to the polarity, or phase, of this incoming signal. If this signal grid swings positive, with respect to its potential before a signal is applied, the magnitude of the current will be increased: that is, there will be an increase in the current flow. On the other hand, if the grid swings negative with respect to its initial potential, there will be a decrease in the amount of current flow. Consequently we see that this signal grid is a controlling factor in the magnitude of the current flow in the FM 1000 tube.

In order to better understand the relationship between the signal at pin 6 and the oscillator voltage with regard to the magnitude of the current, let us study the vector diagram of Fig. 2.* In this drawing vector $e$ represents the oscillator voltage that exists on the oscillator grid, pin 2, and vector $e_1$ represents the signal voltage that exists on the third grid, pin 6. Since the input to the third grid is an fm-if signal, vector $e_1$ represents the mean or center frequency of this f-m signal. In this particular case the frequency is 9.1 mc.

You will notice that vector $e_1$ is shown in quadrature, 90 degrees out-of-phase, with vector $e$ and may wonder why it is drawn in this manner. It is a known fact in the operation of this circuit that when the incoming f-m signal has an instantaneous frequency of 9.1 mc, the center i.f., the pulses of current in the tube remain unchanged. This is the same condition that exists when there is no input signal. Consequently vector $e_1$ can have no component that is in-phase or 180 degrees out-of-phase with vector $e$ because either component would change the magnitude of vector $e$. If vector $e$ is changed, the current flow in the tube will vary and we know this is not the case when the incoming signal is exactly equal to the center i.f. Since vector $e_1$ cannot have any in-phase or 180-degree out-of-phase components with vector $e$, it is drawn in quadrature.

When the instantaneous frequency of the f-m signal is equal to 9.1 mc, this signal must pass through a zero value when the pulses of plate current (due to the oscillator alone) are at a maximum. This is primarily so because the free frequency of the oscillator, that is, when there is no input to the signal grid, is 9.1 mc, the same as the center i.f.

## Incoming Signal Changing in Frequency

Let us now see what happens to the pulse of plate current when the incoming signal is changing in frequency. If there is a phase change between the incoming signal voltage and the oscillator voltage due to a change in input signal frequency, then the two vectors $e$ and $e_1$ will no longer be in quadrature with each other. This means that the signal voltage vector will either have an in-phase or 180-degree out-of-phase component with vector $e$. This in turn will mean a change in the magnitude of oscillator pulse current and hence a change in the amount of plate current. But how can the input f-m signal control the amount of plate current if the magnitude of the f-m signal is constant?

An f-m signal, even though it is undergoing a variation in frequency, also indirectly changes in phase. In Fig. 2 this phase change is indicated by vectors $e_2$ and $e_3$. Vector $e_2$ indicates the input voltage at some instantaneous point where the frequency has decreased from the center i.f. of 9.1 mc and vector $e_3$ represents the signal voltage when the frequency has increased. The phase difference between the signal and oscillator voltage has decreased for the case of vector $e_2$ and has increased for the case of vector $e_3$. Resolving both instantaneous signal vectors $e_2$ and $e_3$, into their horizontal and vertical components, we find that voltage vector $e_2$ has a horizontal component, $e_4$, in phase with vector $e$; and voltage vector $e_3$ has horizontal component, $e_5$, 180 degrees out-of-phase with vector $e$. The vertical components of vectors $e_2$ and $e_3$ are in quadrature with vector $e$ and thus have no effect upon the magnitude of the oscillator voltage. Since vectors $e_4$ and $e$ are in-phase with each other they are additive and the magnitude of plate current is said to increase. Conversely, component vector $e_5$ has to be subtracted from vector $e$, thereby reducing the plate current flow in the tube.

*Therefore, it is seen that when the instantaneous frequency of the incoming fm-if signal is above that of the center i.f. of 9.1 mc, then the magnitude of current pulses decreases; and when the instantaneous frequency is below 9.1 mc, the magnitude of the current pulses increased.*

## Lock-In Action

Let us now consider the feedback action between the quadrature tank circuit and oscillator tank circuit of Fig. 1. It is known that the feedback voltage is proportional in amplitude to the pulse of plate current and will vary in accordance with the change in plate current. The feedback voltage has a phase lead of approximately 90 degrees with respect to the oscillator voltage that exists without feedback. A change in feedback voltage, which is dependent upon the instantaneous frequency of the incoming fm-if signal, effectively changes the frequency of the oscillator. This frequency change is such that the oscillator will *lock-in* at the same frequency as that of the incoming signal. As the input signal changes in frequency, the frequency of the oscillator will follow accordingly, due to the lock-in effect.

For a more complete understanding of why the frequency of the oscillator changes in accordance with

*For a complete analysis of vector diagrams, what they mean and how to use them, see the text "Understanding Vectors and Phase" by Rider and Uslan, published by John F. Rider Publisher, Inc.

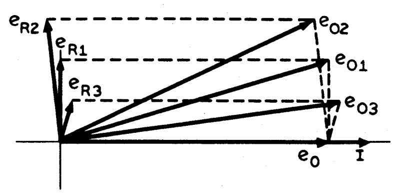



FIG. 3. — The vector diagram indicating the relationships between the reflected voltage from the quadrature circuit and the oscillator voltage in the locked-in oscillator detector circuit.

the incoming signal, let us refer to the vector diagram of Fig. 3. In this diagram vector $e_0$ represents the oscillator voltage that would exist in the absence of feedback and vector $I$ indicates the current flowing through the oscillator tank circuit. Since the oscillator tank is a resonant circuit, the current flowing through it will be in-phase with the voltage across the circuit at resonance. This is illustrated in the vector diagram of Fig. 3. As mentioned previously, with the circuit operating a feedback voltage exists and the total oscillator voltage is equal to the feedback, or reflected, voltage from the quadrature circuit plus the voltage existing without feedback.

The feedback voltage which is in quadrature leading the oscillator voltage $e_0$ has the effect of introducing an effective inductance in series with the oscillator tank inductance. This increase in inductance, although small, establishes the operating frequency of the oscillator. If the instantaneous frequency of the incoming signal is equal to the 9.1 mc i.f. of the receiver, or if there is no signal input, a certain amount of voltage is, nevertheless, reflected into the oscillator circuit from the quadrature circuit. This voltage represented by $e_{R1}$ is in quadrature leading the oscillator vector $e_0$ as shown in Fig. 3. The effective oscillator voltage that exists under these conditions can be found by vectorially adding vectors $e_0$ and $e_{R1}$. The resultant oscillator voltage in this case is designated as vector $e_{01}$.

**Incoming Signal Lower Than Center I.F.**

Let us now see what happens to the effective oscillator-voltage vector when the instantaneous frequency of the incoming signal is different from the center i.f. If we assume that the incoming signal decreases in frequency, the pulses of plate current will increase as we have indicated in Fig. 2. Since this current also flows through the quadrature circuit, the increase in current flow will in turn increase the amount of reflected voltage into the oscillator tank. Since the amount of this reflected voltage is increased, then the inductance that is introduced in series with the oscillator coil (due to this reflected voltage) is increased from what it was previously. This means that the total effective oscillator inductance is increased and the frequency of the oscillator is decreased accordingly and is said to lock-in with the frequency of the incoming signal. This reflected voltage is illustrated as vector $e_{R2}$ in Fig. 3. It is drawn somewhat longer than vector $e_{R1}$ because of its increase in magnitude. Due to the decrease in frequency of the input signal, the phase lead of vector $e_{R2}$ is slightly greater than that of $e_{R1}$. When vectors $e_{R2}$ and $e_0$ are added together, a different resultant voltage, designated as $e_{02}$ in Fig. 3, appears across the oscillator tank circuit.

**Incoming Signal Higher Than Center I.F.**

When the incoming signal increases in frequency, the pulses of plate current will decrease. This means that the amount of voltage reflected into the oscillator-tank circuit will likewise decrease. This new reflected voltage is designated as vector $e_{R3}$ in Fig. 3 and is decreased in amplitude compared to vector $e_{R1}$. Due to the increase in frequency, this reflected voltage vector, $e_{R3}$, has a phase lead that is slightly less than that of $e_{R1}$. Because of its decrease in amplitude, this reflected voltage causes a decrease in the effective inductance that is introduced in series with the oscillator coil. This means a decrease in the over-all inductance of the oscillator tank, thereby increasing its frequency of operation. The increase in frequency is such that the new oscillator frequency follows that of the incoming signal and a *lock-in* effect results. Vectorially adding vectors $e_{R3}$ and $e_0$ gives resultant oscillator voltage $e_{03}$.

Glancing at Fig. 3 once more, it can be seen that when the frequency of the incoming signal varies above and below the center i.f., the resultant oscillator voltage also varies in phase. A change in phase is indirectly followed by a change in frequency. The increase in phase of vector $e_{02}$ over $e_{01}$ represents a decrease in oscillator frequency and the decrease in phase of vector $e_{03}$ over $e_{01}$ represents an increase in oscillator frequency.

The circuit is so designed that the plate current will vary linearly with respect to frequency variations of the input signal above and below the center i.f. The frequency variations of the fm-if input signal are well within the linear limits of the circuit. Although the reflected voltage can be seen to vary in amplitude from Fig. 3, it also varies in phase with respect to the resultant oscillator voltage to maintain the oscillator voltage substantially constant. This is an important point to remember because if the oscillator voltage were not constant, distortion would result in the output audio signal.

### Obtaining the Audio

How does the action of this circuit bring about the detection of the audio modulating component of the input f-m signal? To understand how this occurs is simple, all that has to be remembered is that the *rate* of deviation of the f-m signal is dependent upon the *frequency* of the audio modulating signal and the *amount* of deviation is dependent upon the *amplitude* of the audio modulating signal. Since the rate of change in the plate current is in direct accordance with the rate of deviation, then this rate of plate current variation is in turn dependent upon the frequency of the audio modulating signal. The magnitude of the plate current varies in accordance with the amount of frequency deviation of the f-m signal and hence is indirectly dependent upon the magnitude of the audio modulating signal. Consequently we see that the rate of plate current flow is the same as the frequency of the audio modulating signal and the magnitude of the plate current is proportional to the amplitude of the audio modulating signal.

This plate current flows through its load circuit which consists primarily of the quadrature network, load resistor $R302$ and capacitor $C305$ as seen from Fig. 1. The audio signal represented by the varying plate current of the tube appears across this load. Part of this audio signal appears across the 47,000-ohm load resistor $R302$ and represents an available point from which the audio signal can be taken off. The 1500-$\mu\mu$f capacitor $C305$ serves as a bypass for any i-f currents. The 0.03-$\mu$f capacitor $C304$ and the 100,000-ohm resistor $R303$ are employed to directly couple the audio voltage appearing across load resistor $R302$ to the volume control and hence to the following audio stages.

### Suppression of A-M

In the above analysis we have only indicated how the circuit functions as a detector of f-m signals but nothing has been said about suppression of a-m effects which is equally important to the operation of an f-m detector circuit. A-m effects are suppressed in the following manner. If there is a change in the amplitude of the incoming signal, it will tend to change the magnitude of the current pulses. Any change in the magnitude of the current will cause a change in the voltage reflected from the quadrature circuit into the oscillator circuit. This, of course, will tend to cause a change in the frequency of the oscillator, as we have previously indicated. However, we do know that changes in oscillator frequency are accompanied by phase changes between the pulse current and reflected voltage. This was illustrated by the vector diagram of Fig. 3. Hence, we can conclude that there is a phase change between current pulse and input signal when the input f-m signal undergoes a change in amplitude.

However, the oscillator has only a small frequency change which is considered negligible until it once again locks-in frequency with that of the incoming signal. In other words, it is the lock-in effect of the circuit which is the primary controlling factor in the suppression of a.m. The change in pulse current as caused by a.m. is very small and thus this type of f-m detector circuit is highly insensitive to a.m. The slight sensitivity that it has to a.m. is, like other practical f-m detector systems, far less than the sensitivity it has to f-m signals.

### Linearity and Bandwidth

A linear response characteristic is a requirement for the proper operation of f-m detector circuits because, otherwise, distortion would result in the audio output system. With a *minimum* input signal maintained at all times, the response of the circuit under discussion is quite linear. This linearity is over a total bandwidth of 200 kc, or 100 kc on either side of the center i.f. of 9.1 mc. In other words the input fm-if signal to the detector tube can vary between the limits of 9.0 to 9.2 mc and still fall within the linear response of the circuit.

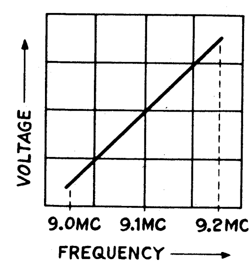



FIG. 4.—Drawing of the f-m detector response of the Philco model 48-482. Note that this curve is linear over a total bandwidth of 200 kc.

This is illustrated by the curve in Fig. 4, which is a drawing of the actual detector response of this model as seen on an oscilloscope. From this drawing it is readily seen that the response characteristic of this circuit is quite linear over a total bandwidth of 200 kc.

If the voltage input to the detector tube is too small, the lock-in effect which is necessary for the proper operation of the detector will not occur; in other words, a certain threshold value of input signal is required at all times. In this receiver the input signal is maintained above its threshold value by providing sufficient i-f amplification preceding the detector.

# UNUSUAL I-F AMPLIFIER CIRCUITS

In a highly competitive field such as radio manufacturing, the producers of the equipment are constantly seeking new ways and methods of improving their products. In order to do this, new developments in engineering design are appearing all the time. In this article we will discuss some methods that are being used in current sets to improve the operation of i-f amplifiers in the simplest and most direct possible fashion.

### Crosley Models 9-119, 9-120W

In the Crosley models 9-119 and 9-120W, appearing on *pages 18-12 and 18-13 of Rider's Volume XVIII,* the tuning capacitor of the first i-f transformer is returned to r-f ground in an unusual manner. This is shown in Fig. 1. It can be seen that the capacitor is returned to the low side of the oscillator transformer primary, which is at r-f ground. Furthermore, this point is physically and electrically near the cathode of the converter, since the oscillator transformer primary, being tuned to a frequency quite different from the i.f., presents a low impedance to i-f currents in the converter plate circuit. Thus, the i-f currents circulating through the converter and the primary tuning capacitor of the i-f transformer are confined to a very short path. If the capacitor were connected to the low end of the i-f transformer primary, the i-f current would have to return to the cathode of the converter through a rather long ground loop which includes i-f currents flowing in the i-f amplifier grid and plate circuits. Such an intermixing of current paths can easily produce instability in high-gain amplifiers. In this circuit such instability is avoided by a very simple circuit arrangement which involves no extra parts.

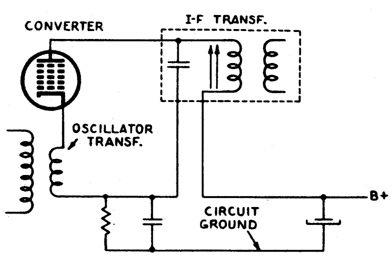



Fig. 1. — Simplified schematic of the converter plate circuit of the Crosley models 9-119 and 9-120W.

### Philco Model 48-300

The i-f amplifier in the Philco model 48-300 portable receiver is conventional in that tuned input and output transformers are used for this stage. As is usual, these transformers are tuned to the same frequency. As a result of this, feedback from the plate to the grid circuit tends to produce oscillation. This feedback is due to several factors, such as the plate-grid capacitance of the i-f amplifier tube, external distributed capacitances between closely placed parts, etc. Another factor is the high gain of the stage, which makes the feedback, small as it is, a potential source of oscillations in the stage.

In order to offset this tendency to oscillate, the stage has been neutralized. Neutralization is a device that is frequently used in transmitter amplifiers, but rarely in receivers. The term "neutralization" is highly descriptive: the tendency to oscillate is produced by positive feedback, so this feedback is cancelled, or neutralized, by negative feedback. For maximum effectiveness, the neutralizing feedback signal should be exactly 180 degrees out of phase with the positive feedback, and of the same amplitude. However, excellent results can be obtained without achieving ideal conditions.

The neutralization of the i-f amplifier in the Philco model 48-300 is shown in simplified form in Fig. 2.

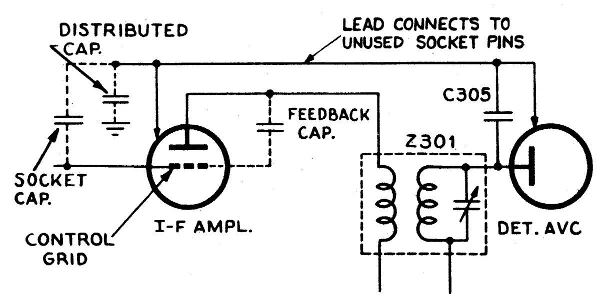



Fig. 2. — Simplified schematic of the neutralized i-f amplifier in the Philco model 48-300.

(The complete data for this model may be found on *pages 18-56 through 18-63 of Rider's Volume XVIII.*) The capacitors shown with dashed leads represent stray capacitances. One of these, labeled "feedback cap," is the source of trouble for it is the path over which positive feedback occurs. It is this capacitance which, if not neutralized, could cause oscillation. Signals traveling over this path arrive at the grid approximately in phase with the plate signal. The i-f transformer, *Z*301, shifts the signals from the plate of the i-f amplifier by 180 degrees. Therefore, signals fed back to the i-f amplifier grid by way of *C*305 are 180 degrees out of phase with the plate signals, *and,* therefore, 180 degrees out of phase with those signals fed back through the dashed-line "feedback cap." *C*305 is considerably larger than the capacitance which it is intended to neutralize, but because the distributed capacitance to ground acts as a voltage divider with *C*305, and also because the socket capacitance of the i-f amplifier is in series with it, the amplitude of the signal *C*305 feeds back to the grid is approximately equal to the signal fed back through the "feedback cap." Thus, two equal signals, of opposite phase, are fed back to the grid. Since they are equal and opposite, they cancel, and thereby provide effective neutralization of the stage.

## Philco Model 48-360

In a battery-operated receiver, particularly a portable one, the power available to operate the tubes is much less than in a set operating from a 110-volt line. As a result, tubes used in portables are incapable of providing the sensitivity obtainable from the 6.3-volt heater-type tubes. In order to overcome this handicap to some extent, positive feedback is used in the i-f amplifier of the Philco model 48-360 to increase the gain of this stage.

At this point the reader may well exclaim, "Design engineers are certainly inconsistent! In one set they go to a lot of trouble to neutralize positive feedback, and in another they take trouble to put it in!" This charge, however, is not well founded; in the model 48-300 the positive feedback is very detrimental to the operation of the set, and *must* be neutralized. In the model 48-360, on the other hand, the positive feedback is controlled and performs a useful and desirable function.

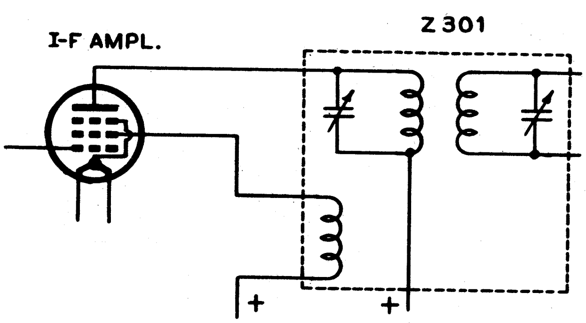



FIG. 3. — Simplified schematic of the i-f amplifier in the Philco model 48-360.

The manner in which positive feedback is obtained in the model 48-360 is shown in Fig. 3. (Complete data for model 48-360 may be found on *pages 18-64 through 18-71 of Rider's Volume XVIII.*) An extra winding, called a tertiary (third) winding, on the second i-f transformer, *Z*301, applies a signal to the screen grid of the i-f amplifier tube. The signal applied to the screen through the transformer is shifted 180 degrees in passing through the transformer, so that it is in phase with the signal at the control grid. In this way, the screen signal is of such a polarity that it increases the effect of the control grid on the tube. Thus, when the control-grid signal is positive, increasing the flow of electrons to the plate, the screen signal is also positive, further increasing the flow. In addition, the screen-grid signal amplitude is proportional to that on the control grid, so that the effect of the control-grid signal is increased without distorting it. Likewise, when the control-grid signal goes negative, the screen-grid signal goes proportionately negative, increasing the effect of the control-grid swing in this direction.

By a proper choice of turns ratio and coupling between the primary and tertiary of *Z*301 the positive feedback is maintained at a level insufficient to produce oscillation. Thus, *controlled* positive feedback is used to attain the desired goal of increased gain.

## Philco Model 48-464

In an ac-dc radio, B— must either be connected directly to the chassis or bypassed to it so that the B— bus and chassis are at the same r-f and i-f potential. This is particularly important when some components, such as the tuning capacitors, are returned directly to the chassis rather than to the B— bus, and also when a high-gain, two-stage i-f amplifier is used, as in the Philco model 48-464 (*see pages 18-72 to 18-79 of Rider's Volume XVIII*). If this is not done, various undesirable effects, such as instability (tendency to oscillation) of the i-f amplifier, will occur.

In many sets, where the B— bus is not tied directly to the chassis, B— is bypassed to the chassis by means of a capacitor alone or a capacitor and choke in series. When a capacitor and choke are used, they are chosen so as to be series resonant at the intermediate frequency. Being resonant at the i. f., they present an even lower impedance to the i. f. than would a capacitor alone.

In order to produce an intermediate frequency, the converter tube is operated on a non-linear part of its characteristic curve. In addition to the production of the i. f. by the mixing action, harmonics of the i. f. will also be generated. Another possible source of harmonic generation is non-linear operation in the i-f amplifiers. Of course, the higher harmonics are usually not very strong, but the second and third harmonics may be strong enough to cause trouble. For this reason, the B— bus in the 48-464 is bypassed to the chassis by the combination of capacitors and chokes shown in Fig. 4.

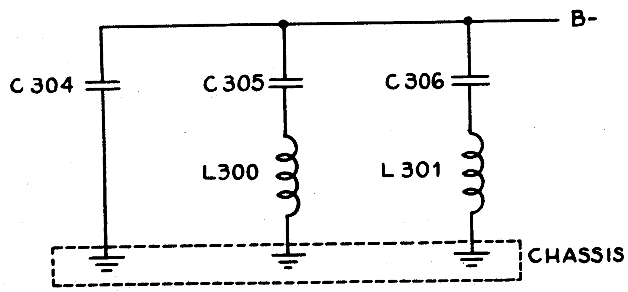



Fig. 4. — Bypassing from B- to ground in the Philco model 48-464 is accomplished by the above circuit.

*C*305 and *L*300 are resonant at 455 kc, which is the i. f. *C*306 and *L*301 resonate at 910 kc, the second harmonic of the i. f., and the two capacitors and two chokes resonate together at 1365 kc, the third harmonic of the i. f. Thus these four components provide a very high degree of bypassing at these three important frequencies. *C*304 takes care of the radio frequencies and the oscillator frequencies. These frequencies, of course, are not fixed, so a fixed-tuned circuit cannot be used to bypass them.

## General Electric Models 210, 211, and 212

Two very interesting features are found in the i-f circuits of GE models 210, 211, and 212 (*see pages 18-21 to 18-25 of Rider's Volume XVIII*). These sets are am-fm receivers; the features that will be discussed here are operative features when the sets are used for f-m reception. Starting at the antenna, we find the first unusual circuit to be a reflex amplifier.

In the reflex circuit a single tube is used to amplify both r-f and i-f signals. The advantages in savings of cost and space by the use of one tube instead of two are obvious. This has long been done in the case of converters (combined mixer and oscillator) and combined detector, avc rectifier, and first audio amplifier; both of these combinations are used in these sets.

The reflex amplifier and its associated converter are shown in Fig. 5. The band switch, *S*1, is in the f-m position. In a good location, f-m signals may be picked

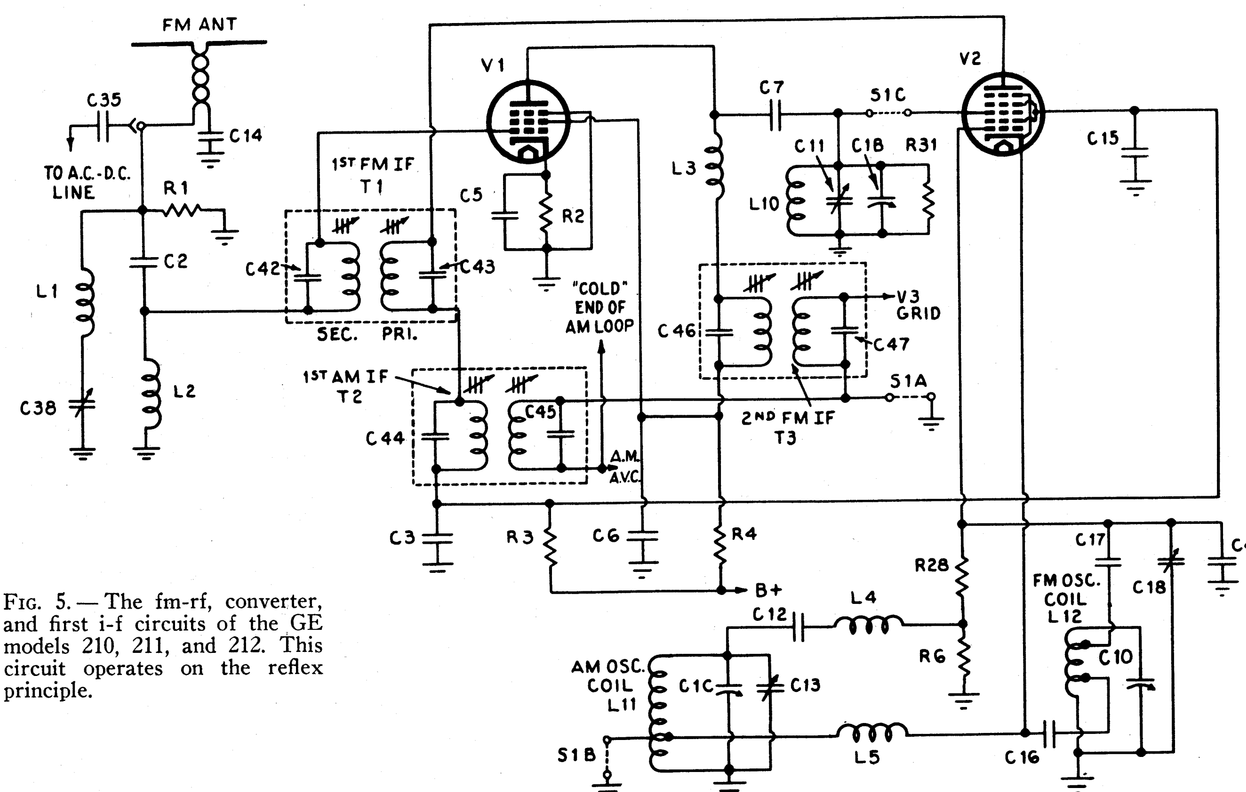



Fig. 5. — The fm-rf, converter, and first i-f circuits of the GE models 210, 211, and 212. This circuit operates on the reflex principle.

off the power line through C35; in a poor location, an outside antenna may be used. In the latter case, C14 in conjunction with the input to the r-f amplifier, which is single-ended, serves to provide a balanced load for the dipole. R1 provides a leakage path to ground for charges which might otherwise accumulate in the antenna circuit. C2 and L2, together with existing stray capacitances (such as distributed capacitance and lead inductance), are resonant at 98 mc, the middle of the f-m band. Since they are series resonant, the voltage across the choke alone is considerably greater than the voltage across the series combination. This increases the sensitivity of the receiver.

The signal is fed to the grid of V1 through the *secondary* of T1. (Because of the unusual circuit that we have here, it is necessary to place the secondary of T1 to the left of its primary for the sake of clearness.) The secondary of T1 is parallel resonant to the i. f., 10.7 mc, so that it appears like a fairly large capacitance to the much higher radio frequencies, which pass through without any difficulty. V1 then amplifies the r-f signals, with L3 functioning as the plate load. The "cold" end of L3 is bypassed to ground through the series circuit consisting of the primary of T3, which is capacitive at the r. f., and C6. R-f signals from the plate of V1 are fed to the signal grid of the converter, V2, through C7 and section S1C of the band switch. The particular r-f signal desired is selected by the parallel tuned circuit consisting of L10, C11, C1B, and R31; C1B is part of the tuning capacitor gang, and R31 is a damping resistor.

The oscillator section of V2 operates in a Hartley circuit for both f.m. and a.m., but different tank circuits must, of course, be used. The a-m tank circuit is decoupled from the f-m tank circuit by L4 and L5, and the cathode tap of the a-m tank circuit is grounded through section S1B of the band switch. A small damping resistor (R28) prevents L4 from affecting resonance conditions in the f-m tank circuit. There are two interesting points to be observed about L12, the f-m oscillator coil; one is that it is made of a short piece of 300-ohm twin lead, shorted at one end and formed into a one-turn loop. The other is that the grid signal is not taken from the top of the tank coil, but is taken from a tap on the coil. The reason for this is to reduce the effect on the tank circuit of variations in the oscillator-grid input capacitance of V2. A fuller explanation of this principle is given in *Rider's Volume XV "How It Works", page 165.*

F-m intermediate frequency signals from the plate of V2 are fed through T1 to the grid of V1. The "cold" end of the primary of T1 is bypassed to ground, first through the primary of T2 and then through C3 and C15. The primary of T2, which is tuned to the a-m i.f. acts like a capacitance at the much higher f-m i.f., just as the f-m i-f transformer primaries appear to be capacitive at the still higher f-m r.f. (These relationships hold true for the transformer secondaries as well.) The "cold" end of the secondary of T1 is returned to ground through L2, which has a very low reactance (about 5 ohms) at the f-m i.f. C2 has a reactance of some 50,000 ohms at the f-m i.f., so that not much i-f signal will leak through it. As an added safeguard, however, a series-tuned trap consisting of L1 and C38 is provided.

V1 now functions as an i-f amplifier. The reactance of C7 is so high at the i. f., and that of L3 so low, that virtually all of the i-f signal available at the plate of V1 appears across the primary of T3. The signal on the primary of T3 is coupled to the secondary in the usual fashion, and is then applied to the control grid of V3. The "cold" end of T3 is grounded through section S1A of the band switch.

It may appear strange that both C3 and C15 are used. The reason is that the frequencies used in f.m. and a.m. differ so widely. C15 is a mica capacitor which acts as an excellent bypass at the f-m r.f. and even the i.f., but is of too small a value to be satisfactory at the a-m frequencies. C3, on the other hand, has sufficient capacitance to be a good bypass at the a-m frequencies. However, it is a paper capacitor, and because of its construction has enough inductance to behave like a choke at the f-m r.f.

When the set is used for a-m reception, V1 is not used. R-f signals are fed directly from the "hot" end of the loop antenna (not shown in illustration) through section S1C of the band switch to the signal grid of V2. In the a-m position of the band switch, section S1B is open, removing the ground from the cathode tap of the oscillator tank coil. L4 and L5 have negligible reactance at the a-m oscillator frequencies, and therefore, do not interfere with its operation. C16, C17, C18, and C41, on the other hand, have very high reactances at these frequencies; as a result, the f-m oscillator tank circuit does not load the a-m oscillator tank circuit.

A-m i-f currents appear in the plate circuit of V2 in the usual fashion. These currents pass easily through the primary of T1, since at the a-m i.f. this winding has a very low reactance. A large i-f voltage is built up across the primary of T2, the first a-m i-f transformer. This produces a voltage across the secondary of T2 which is applied to the grid of V3. In the a-m position of the band switch, section S1A shorts the secondary of T3, so that a direct path is available from the secondary of T2 to the grid of V3. The "cold" end of T2 is connected to the a-m avc bus; the "cold" end of the a-m loop antenna is also connected to the avc at this point.

It may be asked why this reflex principle is not used in a-m reception. The answer lies in the frequencies involved. In the receivers we have just discussed the ratio of the f-m r.f. to the f-m i.f. is over 8 to 1, even at the low end of the f-m band. This makes it relatively easy to keep the r-f and i-f signals separate. In the case of a-m reception, however, the r-f to i-f ratio at the low end of the band is not quite 1.2 to 1. If an attempt were made to operate a reflex circuit under these conditions, the separation of the r-f and i-f signals would be very difficult.

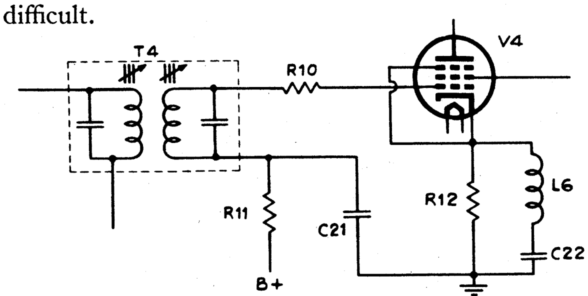



FIG. 6. — Simplified schematic of the limiter in the GE models 210, 211, and 212.

Another interesting aspect of the i-f amplifier in these sets is found in the f-m limiter stage. A simplified schematic of this stage is shown in Fig. 6. This stage not only serves to clip the tops and bottoms of the i-f wave, as is usual in a limiter, but it also introduces a high order of degeneration to any a.m. in the i-f wave. This effect is contributed by the cathode resistor, *R*12, which has the unusually high value of 33,000 ohms. *L6* and *C22* are series tuned to the intermediate frequency, 10.7 mc; they, therefore, have a very effective bypassing action for i-f signals. This permits the stage to operate with the normal gain expected from a limiter.

The grid circuit is not very unconventional, although it appears to be so on first glance. *R*10 is a 10-ohm, anti-parasitic resistor. *C*21 and *R*11 are higher in value than usual, and *R*11 is returned to B+ instead of to ground. The reason for using very high values for these two components is that, speaking colloquially, they do what is usually required of these parts in a limiter, only more so. The usual function of these parts is similar to that of a grid-leak detector, but the parts are ordinarily so chosen that they have a time constant of only a few microseconds.* In the present case, however, the time constant is 50,000 microseconds, a value which is suitable for detection of a-m signals.

With the grid circuit of *V*4 acting as an a-m detector, and the cathode resistor of this stage very large in value and unbypassed (to audio), *V*4 becomes a highly degenerative amplifier for any a-m signals that may reach it. Thus the stage reduces a. m. in two ways, as an ordinary clipping limiter and as a degenerative amplifier.

*R*11 is returned to B+ instead of to ground because of the high value of the cathode resistor. If the normal plate and screen currents for a limiter stage are to flow here, they will produce an unusually high value of cathode voltage, because of the drop across the cathode resistor. In this case the drop amounts to 50 volts. This requires that the grid be almost 50 volts above d-c ground also, so that the correct operating bias can be obtained. This requirement is met by returning *R*11 to B+. In the absence of an i-f signal, this makes the grid slightly positive with respect to the cathode; only slightly, however, because as soon as the grid starts to go positive, current flows in the grid circuit which produces a drop across *R*11 sufficient to keep the grid voltage very close to the required value.

When an i-f signal is applied, a certain amount of rectification (grid-leak detection) takes place. This establishes a charge on *C*21 of such polarity that the grid becomes negative with respect to the cathode, despite the return of the grid resistor to B+.

* For a complete discussion of limiters, see pp. 277-293 of "FM Transmission and Reception", by Rider and Uslan, published by John F. Rider Publisher, Inc., N. Y.

# OSCILLATORS FOR F-M SETS

In an f-m superheterodyne receiver, the function of the local oscillator is naturally the same as that in the more familiar a-m broadcast superhet. However, because the frequencies used in f-m broadcasting are so much higher than those used in a.m., new difficulties are introduced or old ones made worse. One of the major problems encountered is that of frequency drift. This problem has been pretty well solved in a-m sets, but f-m receivers are still sufficiently new so that the solution in their case is not definitely set.

The problem of drift in f-m receiver local oscillators was discussed in the *Volume XV "How It Works"*; however, since the publication of this book enough new tricks have been introduced to make it worthwhile to reopen this subject. Before taking up the technical details, let us briefly review the problem as a whole.

As stated in the *"How It Works" of Volume XV*, oscillator drift is likely to be a more serious matter in f-m than in a-m reception, because of the difference in the nature of the signals and the methods of detection. In a.m. the frequency of the carrier is not made to vary under modulation, so receiver oscillator frequency drift does not produce an effect similar to modulation. In f.m., however, oscillator frequency drift does produce an effect that is similar to modulation. Of course, the rate at which drift takes place is so slow that it does not directly produce an audio output, but if the drift is severe it can cause distortion in the audio.

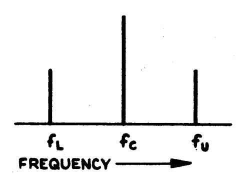
In an a-m receiver, the method of detection is some sort of rectification, which is not in itself frequency sensitive. The rectifying detector is fed from a bandpass i-f amplifier, which is used to avoid reception of any signal but the desired one, and even in this combination severe oscillator drift will cause distortion because of the shape of the i-f pass band. It must be emphasized, however, that the rectifying detector itself will demodulate any a-m wave, regardless of frequency.

In f-m receivers, on the other hand, the detector is a frequency-sensitive device. Although rectification is a part of the process in the operation of both discriminators and ratio detectors, it is not sufficient in itself. In both these and other methods, frequency-sensitive tuned circuits are a necessary part of the detector. Because of this difference, oscillator drift has a direct effect on f-m detectors which it does not have on their a-m counterparts.

It is possible in both a-m and f-m receivers to make the i-f bandwidth sufficiently great so that some oscillator drift is tolerable. It is much easier to do this in a.m. for two reasons; one of which is inherent in the difference between an a-m and f-m wave while the other is in the difference between the broadcast frequencies employed in each.

The inherent difference lies in the effects produced upon a carrier wave by amplitude and frequency modulation. All types of modulation produce sidebands but the factors that determine the characteristics of the sideband distribution for the various types of modulation are not necessarily the same. In a.m.

FIG. 1. — The spectral distribution of an a-m wave. There are always only two sidebands.




only two sidebands are produced (called the upper and lower sidebands) and the frequency separation of the sidebands from the carrier depends directly upon the frequency of the modulating signal. This is readily seen from the spectral distribution of an a-m wave as shown in Fig. 1. In this figure, $f_c$, represents the carrier component of the a-m wave and $f_u$ and $f_l$ represent the upper and lower sidebands respectively. The two sideband components are often referred to as one sideband pair. (The amplitudes of the components are of no interest to us in this discussion.) The frequency of sideband component $f_u$ equals the frequency of the carrier plus that of the audio modulating signal and component $f_l$ equals the frequency of the carrier minus that of the audio.

In other words, a high-frequency audio signal will produce sidebands in a.m. that are relatively distant in frequency from the carrier component and low-frequency audio signals produce sidebands relatively close to the carrier. The distance in separation between $f_l$ and $f_u$ of Fig. 1 determines the operating bandwidth of the a-m wave.

In f.m. the situation is much different. The first and most important factor to consider is that numerous sidebands are possible with f.m. The amount of sidebands is determined by the amplitude and the frequency of the audio modulating signal. The higher the amplitude and the lower the frequency of the audio, the greater the number of sidebands. The spacing of the sidebands are, however, only dependent upon the frequency of the audio modulating signal similar to a.m. The sidebands are distributed equally on either side of the carrier component as illustrated in Fig. 2, the spectral distribution of a typical f-m wave having four sideband pairs (eight sidebands).

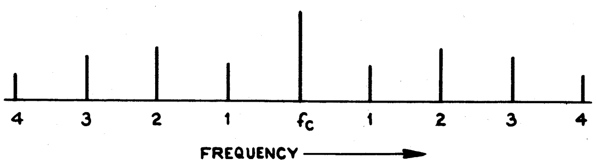



Fig. 2. — The spectral distribution of an f-m wave. There can be numerous sidebands.

In this drawing the sideband pairs are numbered from 1 to 4 with the number 1 components being separated from the carrier component $f_c$ by the frequency of the audio. The frequency separation between each sideband is equal to the audio modulating frequency. In an f-m wave as many as 60 effective sidebands (30 sideband pairs) can exist. The higher order sidebands can be caused by low-frequency as well as high-frequency audio modulating signals. The operating bandwidth of an f-m wave is much greater than an a-m wave even though the audio modulating frequency of the f-m wave is very low as compared to that of the a-m wave.

Consequently in a.m. if sidebands widely spaced from the carrier are lost (which may happen when the oscillator drifts), this will affect only overtones or high fundamental notes, which are not important for intelligibility, and even in musical reproduction are only of secondary importance. In f.m., on the other hand, if sidebands are lost, this may mean distortion of high-amplitude low-frequency (or middle-frequency) modulating signals. Therefore, much less sideband cutting is tolerable in f.m. than in a.m. As a result, all other things being equal, a wider pass band is necessary in f.m. than in a.m. to allow for drift.

The other reason, as we have stated, lies in the difference between the frequencies employed in f-m transmission and reception compared to those used in a.m. In order to illustrate this difference, let us take some definite figures. Assume an a-m receiver tuned to a station operating at 1000 kc; the receiver has an i.f. of 455 kc, so the oscillator frequency should be 1455 kc. If the oscillator is, say, 0.02% high, then its actual frequency is 1455.291 kc, and the actual i.f. becomes 455.291 kc, an error of only 291 cps. Assume now an f-m receiver tuned to a station operating on 100 mc; the receiver has an i.f. of 10.7 mc and an oscillator tuned below the carrier, so that the oscillator frequency should be 89.3 mc. If the oscillator in this case is also 0.02% high, its actual frequency is 89.31786 mc, producing an i.f. of 10.6214 mc. This is an error of 17,860 cps. If the oscillator were set above the carrier, the error would be correspondingly greater.

To view these two errors in their proper perspective, it must be remembered that the i-f center frequencies in a.m. and f.m. are very different, and that the fm-if pass band is also much wider than that in a.m. In the average of production receivers the ratio of fm-if to am-if pass band widths is of the order of 25 to 1. However, in the numerical examples that we chose, the calculated error for the f-m case is slightly more than 60 times that for the a-m case. This indicates that relatively less drift is permissible in the oscillator of an f-m receiver than in that of an a-m set.

Among the receivers shown in *Volume XVIII* are several illustrating a means of reducing drift that may easily become very common, and at least one employing afc (automatic frequency control) to overcome the tendency of the oscillator to drift.

## GE 210, 211, & 212

One of the greatest causes of oscillator drift is changes in the interelectrode capacitances of the oscillator tubes. At the oscillator frequencies used in f-m receivers even the heater-cathode capacitance is important. In addition to its effect on frequency, this capacitance can also be a source of hum leakage from heater to cathode. For these reasons, the cathode is grounded in some oscillators, so that the cathode acts as a shield between the heater and the other elements of the tube, and variations in the heater-cathode capacitance then make no difference. However, it is not always desirable to ground the cathode; this is especially true when a pentagrid converter is used.

One case we have in mind is in the converter section of G. E. models 210, 211, and 212. These models use a miniature type, 12BE6, pentagrid converter

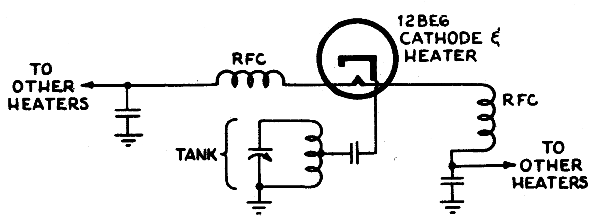



Fig. 3.— Simplified schematic of the converter heater and cathode connections in the GE models 210, 211, and 212. This circuit is designed for f-m oscillator stability.

tube. A simplified schematic of the heater and cathode connections of this converter tube is illustrated in Fig. 3. Complete service data for these models can be found on *pages 18-21 through 18-25 in Rider's Volume XVIII.*

As shown in Fig. 3 the converter heater is isolated from the other heaters in the receivers by r-f chokes and also is isolated from ground by capacitors in addition to the r-f chokes. In addition, the cathode is effectively connected to the heater through the interelectrode capacitance between the cathode and heater of the tube. This internal heater-cathode capacitance of the 12BE6 tube is sufficient to provide adequate coupling between the two tube elements. Regardless of variations in this capacitance, it still presents a sufficiently low reactance compared to that of the r-f chokes, to keep the heater at the same r-f potential as the cathode, and thereby eliminate any effect these internal capacitance changes might have on the oscillator frequency and hum. The capacitors from the heater line to ground helps keep the oscillator voltage from the heater of the remaining tubes in the receiver.

### United Motors Models R-1253, R-1254, and R-1255

An obvious way to overcome the effects of oscillator drift is to retune the oscillator when its frequency changes. It is also obvious that if this chore is forced upon the user of an f-m receiver, he will not take very kindly to the task. However, if the receiver can be made to do the work itself, without any effort on the user's part, a neat solution to the drift problem is obtained. This approach has been taken in a number of f-m sets, including United Motors models R-1253 and R-1254. See *pages 18-11, 12 through page 18-19 of Rider's Volume XVIII* for complete service data on these models.

The circuit that produces automatic retuning of the oscillator when it drifts off frequency is known as an automatic frequency control circuit, usually abbreviated to afc. To operate this circuit it is necessary to obtain a voltage that is proportional to the extent and direction of the drift of the oscillator. Then this voltage must be applied to an electronic tuning control that will return the oscillator toward the correct frequency. Such an arrangement can not make the drift zero, because a control voltage is necessary to operate it, and the control voltage is generated only when the drift is *not* zero. However, it can reduce the drift very considerably, making it very much less than it would be without afc. (By the addition of certain refinements to the basic system just discussed, an afc circuit can be constructed that will reduce the drift to zero, but these refinements are far too complex for a home receiver.)

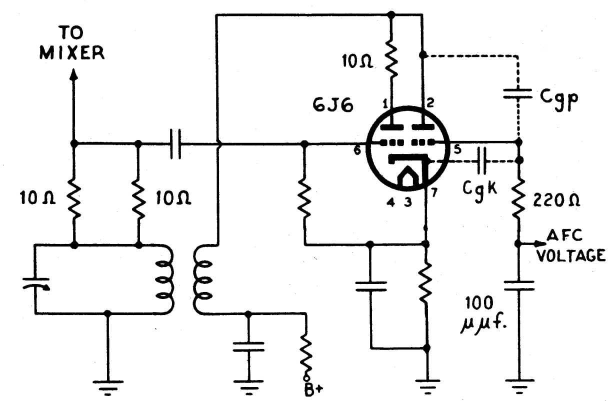



Fig. 4.— Simplified schematic of the oscillator and the reactance-tube circuit used in the United Motors models R-1253, R-1254, and R-1255.

The detector in an f-m receiver must, by the very nature of f.m., respond to frequency changes. Since oscillator drift is a frequency change, the f-m detector can be made to produce an output which indicates this change. This output then serves to control the electronic tuner. In the United Motors models R-1253 and R-1254 an output is obtained from the ratio detector, filtered to remove audio, and then applied to one grid of a 6J6, as shown in Fig. 4.

Fig. 4 shows the oscillator and reactance tube used in these sets. The reactance tube operates as a control for the oscillator and tends to stabilize the oscillator frequency. The diagram has been simplified to show only essentials, and parts values have been shown only for those components that do not perform the ordinary functions found in an oscillator. The 10-ohm resistors are anti-parasitics, to prevent unwanted oscillations. The oscillator is an ordinary tuned-grid feedback type, using one section of a 6J6 duo-triode tube. Since the tube has a single common cathode, and since either positive or negative values of afc voltage may be applied to the grid of the reactance tube section, a cathode resistor is used to produce cathode bias. If this were not done, the reactance-tube grid could go positive with respect to the cathode; this would upset the action of the circuit. The cathode resistor is by-passed, bringing the cathode to r-f ground; for this reason the heater can also be operated at r-f ground.

The grid-plate and grid-cathode capacitances of the reactance tube section of the 6J6 have been shown in Fig. 2 because they are very important in the action of this part of the tube. The grid-plate capacitance is approximately 1.6 $\mu\mu$f, and the grid-cathode capacitance 2.2 $\mu\mu$f. These two capacitances act with the 220-ohm resistance to form a phase-shift network between plate and grid. The bottom end of the 220-ohm resistor is grounded to r.f. by the 100-$\mu\mu$f capacitor, hence that point and the cathode are effectively tied together. For this reason only the grid-plate and grid-cathode capacitances and the 220-ohm resistor are important in the phase-shifting network. This network produces a shift in the neighborhood of 60 degrees in the 100-mc region. Although this falls somewhat short of the 90 degrees desired ideally for a reactance tube, it is sufficient to cause the reactance-tube section of the 6J6 to inject a variable capacitance into the oscillator section of the same tube.* Since the reactance-tube section is controlled by the afc voltage, the variable capacitance injected performs the function of returning the oscillator when its frequency drifts. Thus the oscillator frequency is held within sufficiently close limits so that the slight amount of residual drift does not produce distortion in the audio output of the ratio detector.

An interesting point in connection with this afc circuit is the use of a 6.8-ohm resistor (indicated as *R57* on the schematic in *Volume XVIII*) in series with the heater of the 6AL5 ratio detector. This resistor drops the heater voltage of the 6AL5 and slightly slows down the rate at which the afc output of the ratio detector follows changes in oscillator frequency. This prevents the afc system from over-compensating, or "hunting". Hunting, in this case, would produce an effect on the audio much like motor-boating.

* For a complete discussion of reactance tube operation, see pages 54 to 62 "FM Transmission and Reception," by Rider and Uslan, published by John F. Rider, Publisher, Inc.

# GROUNDED-GRID INPUT CIRCUITS

Today there is really not much of a design problem regarding the antenna and input r-f section of a-m broadcast receivers. Straight pieces of wire and indoor antennas serve adequately well because of the nature of a-m broadcast signals. Impedance matching between the antenna and receiver is not required in such receivers. However, with f-m receivers the situation is quite different. The nature of f-m signals almost invariably requires a special type of antenna construction, whether indoor or outdoor, to be used with f-m receivers. This is necessary in order to achieve the maximum signal pickup by the antenna.

In order to supply the maximum signal from the antenna to the input section of the f-m receiver, these units must be impedance matched to each other. The usual type of f-m antennas have a very low input impedance and for maximum transfer of energy from the antenna to the receiver, the input impedance of the receiver should equal or closely approximate that of the antenna. The correct type of transmission line should be employed to complete the circuit between antenna and receiver and maintain or provide the impedance match.

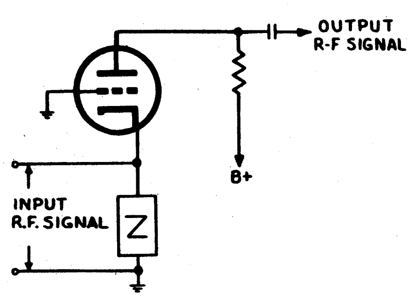



FIG. 1. — A simplified schematic of a grounded-grid amplifier.

This impedance-matching principle creates a design problem especially when the input signal is fed to the control grid of the first r-f tube because the impedance between grid and cathode is comparatively much higher than the low input impedance of f-m antennas. Such connections require specially designed antenna coupling circuits in order to provide the necessary low input impedance for matching purposes. There are a number of f-m receivers on the market today which simplify this impedance-matching problem by employing *grounded-grid* r-f amplifiers in the input circuit. Such arrangements as these feed the input signal to the cathode circuit and the grid is effectively grounded as far as r-f signals are concerned.

### The Grounded-Grid Circuit

A simplified schematic diagram of a grounded-grid amplifier is illustrated in Fig. 1. The input r-f signal, which is usually secured from the antenna circuit, is applied directly across whatever impedance is in the cathode circuit of the tube. This cathode impedance is represented by $Z$ in the drawing and it may be a resistor, capacitor, or inductor, or any possible combination of these components. With such a system as this, it can readily be seen that almost any value of input impedance can easily be secured by the choice of $Z$. In this schematic the grid is shown directly grounded in order to simplify the drawing. The grid may, however, be returned to ground through some circuit component as long as it has a negligible impedance value as far as the fm-rf signals are concerned. The grid must also have a d-c return path to ground. The output signal is taken from the plate circuit of the tube.

The fundamental operation of any amplifier circuit depends upon a difference in potential between the grid and cathode of the tube. In the usual amplifier circuit, the cathode is grounded at signal frequencies and the signal fed to the grid. In the grounded-grid circuit, the reverse is true but a difference of potential still exists between the grid and cathode and the tube can function perfectly well in this manner. With such circuits as that shown in Fig. 1, a low input impedance is possible. This means that a grounded-grid type of input circuit can very easily be made to provide a satisfactory impedance match to an antenna.

Besides providing a low-impedance input source, such a circuit is very stable compared to the conventional type of input circuit, especially at the very high frequencies involved in f.m. This highly stable circuit permits the use of triodes as r-f amplifiers without the worry of regeneration causing oscillation. Triodes are very useful as r-f amplifiers because they have a higher signal-to-noise ratio than pentode amplifiers.

Thus we see that two advantageous features are derived by using a grounded-grid amplifier for the input to f-m receivers; namely, simplification of impedance matching to the antenna and greater stability in the r-f circuit. Let us now study some typical grounded-grid amplifier circuits such as are used in the r-f input circuits of today's f-m receivers.

### Admiral 9B14

In the Admiral models 9B14, 9B15, and 9B16 a 6BA6 miniature type pentode is employed as the grounded-grid input tube for the f-m band. The service data for these models appears *on pages 18-33, 34 through 18-38 in Rider's Volume XVIII.*

The schematic diagram for the circuit under discussion is illustrated in Fig. 2. Note that the first r-f amplifier has its screen connected directly to the plate and thus the tube effectively functions as a triode.

The input f-m signal is fed into the cathode circuit of the first r-f tube. Inductance *L4* is termed the "f-m coupling coil" and inductance *L5* a "cathode choke". These two components together with the 0.001-μf coupling capacitor *C1* and the 100-ohm resistor *R2* comprise the complete load in the cathode circuit of the tube. The r-f voltage across *L4* is the complete input signal and this voltage is also found across the total series combination of *C1*, *R2*, and *L5* because the series arrangement of these components is in parallel with *L4*. Glancing at Fig. 2 again, you will note that the cathode of the first r-f tube is connected to the junction point of *C1* and *R2*. At first it might be thought that only part of this input voltage appears across the cathode because of the voltage divider network of *C1*, *R2*, and *L5*. However, the reactance of the 0.001-*μf* capacitor *C1* is so low at the f-m frequencies (being 1.6 ohms at 100 mc) compared to the total series impedance of *R2* and *L5* that negligible signal voltage drop occurs across *C1*. Thus, the total available signal voltage is considered as being applied to the cathode circuit. The total impedance value of the four components in the cathode circuit, as seen by the antenna and its transmission line, has been so chosen that the impedance match will be correct.

Without any signal applied to the cathode circuit but with B plus and heater voltage supplied, a continuous flow of direct current through the tube will result. A d-c voltage drop occurs across the cathode circuit components *R2* and *L5* to ground.

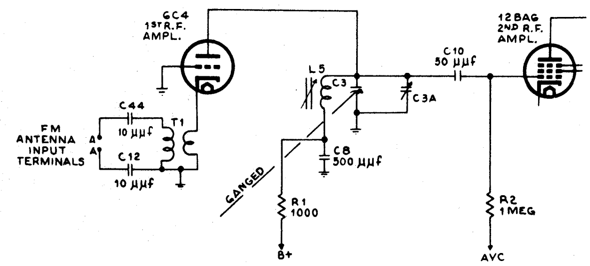



FIG. 3.—A simplified schematic of the r-f section of the Noblitt-Sparks models 280TFM and 281TFM.

Since the grid is directly grounded as seen in Fig. 2, a difference of potential exists between the grid and cathode of this tube, with the cathode being more positive than the grid by the amount of the d-c voltage drop. This establishes the so-called bias of the tube as it does in the ordinary amplifier circuit. When the f-m signal is applied to the cathode circuit, the bias is alternately changed which in turn causes the plate current flowing in the tube to fluctuate in the same manner as this input signal. Consequently the f-m signal appears in the plate circuit of the tube and is coupled to the control grid circuit of the second r-f amplifier via transformer *L6* as seen in Fig. 2. Only the secondary of this transformer is tuned and the tuning is ganged with the other r-f tuning sections of the set.

### Noblitt-Sparks 280 TFM

The f-m section of the Nobbitt Sparkes models 280TFM and 281TFM also employs two r-f ampli-

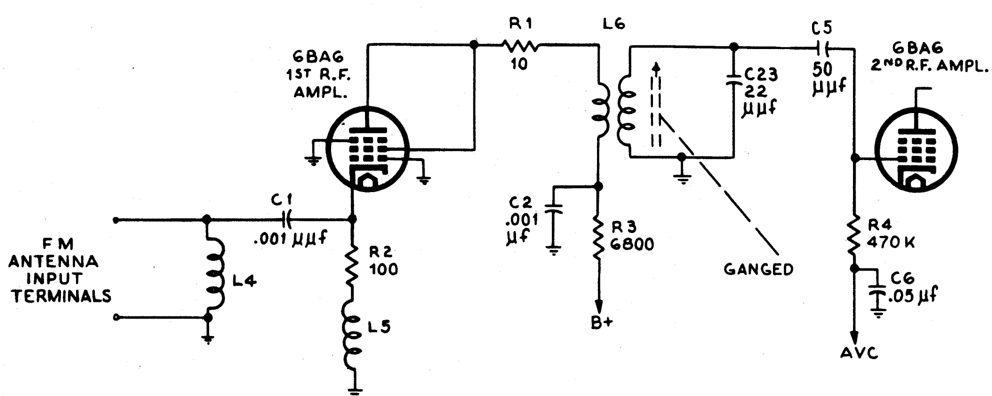



FIG. 2.—The grounded-grid input r-f amplifier of the Admiral models 9B14, 9B15, and 9B16.

fier stages with the first stage being a grounded-grid amplifier. A simplified schematic drawing of this r-f section is shown in Fig. 3. For complete service data for these models see *pages 18-8 through 18-12 in Rider's Volume XVIII.*

A triode tube, the miniature type 6C4, is employed as the first r-f amplifier. In this circuit the grid of the tube is directly grounded, similar to that of the 6BA6 circuit of Fig. 2. The primary difference between the network of Fig. 3 and the previous one is in their cathode circuits. In the circuit under discussion, the input signal is transformer coupled to the cathode via *T*1. Points *A-A* represent the antenna terminal connections. The two 10-$\mu\mu f$ capacitors *C*12 and *C*44 are inserted to balance the input circuit. The actual load in the cathode circuit is the secondary of *T*1 plus whatever impedance is reflected from the primary into the secondary. This complete input circuit is so designed that the impedance seen looking into the antenna input terminals is low enough so that an f-m antenna can easily be matched to this input circuit.

An f-m signal appears at the plate of the 6C4 tube in a similar fashion to that described for the previous model. This signal is then coupled to the grid circuit of the 12BA6 second r-f amplifier via the parallel tuned circuit consisting of coil *L*5 and capacitors *C*3 and *C*3A. Capacitor *C*3 is the tuning unit of the circuit and it is ganged to the other r-f tuning capacitors of the receiver. Avc is applied to the grid of the second r-f amplifier through *R*2, a one-megohm resistor. In this circuit and the one previously discussed the grid of the first r-f tube is grounded directly and hence does not receive any avc voltage.

### Westinghouse H-164

In the Westinghouse models H-164, H-166, H-166A, and H-167 a grounded-grid amplifier is also employed for the input circuit but there are a few interesting details about this circuit arrangement not encountered in the others. See *pages 18-12 through 18-19 in Rider's Volume XVIII* for complete service data of this model. A simplified schematic diagram for the fm-rf section of this receiver appears in Fig. 4.

This r-f section employs the duo-triode 7F8 loctal type tube. Each triode section of this tube is used as an r-f amplifier, the input triode section is a grounded-grid amplifier and the second r-f amplifier is used as a cathode-follower type circuit. Looking at the first r-f amplifier, it is seen that the cathode circuit consists of inductance *L*1 in parallel with a 15-$\mu\mu f$ capacitor *C*30. The high side of the f-m antenna circuit is attached directly to the coil but across only part of the coil windings. In this manner *L*1 acts as an auto-transformer to the incoming signal. Capacitor *C*30 in conjunction with the tapped section of coil *L*1 offers the correct impedance match to the f-m antenna. This *L*1-*C*30 circuit is fairly selective to f-m signals. The simplicity of such a circuit for impedance matching purposes is readily apparent when it is compared to the ordinary amplifier where the signal is fed to the grid circuit of the tube.

Another interesting feature about this first r-f amplifier is the grid circuit. It was mentioned at the beginning of the section on grounded-grid amplifiers that the grid of the tube in question should be effectively grounded as far as r-f signals are concerned This is exactly the case in Fig. 4 although it is not immediately apparent from the arrangement of the input circuit. This grid circuit is not tied directly to ground as in the circuits of Figs. 2 and 3.

As noticed from Fig. 4 a 470-$\mu\mu f$ capacitor, *C*40, and 470,000-ohm resistor *R*17 are attached directly to the grid with the other side of the capacitor grounded. The reactance of capacitor *C*40 at the f-m signal frequencies is so low that the grid is considered grounded

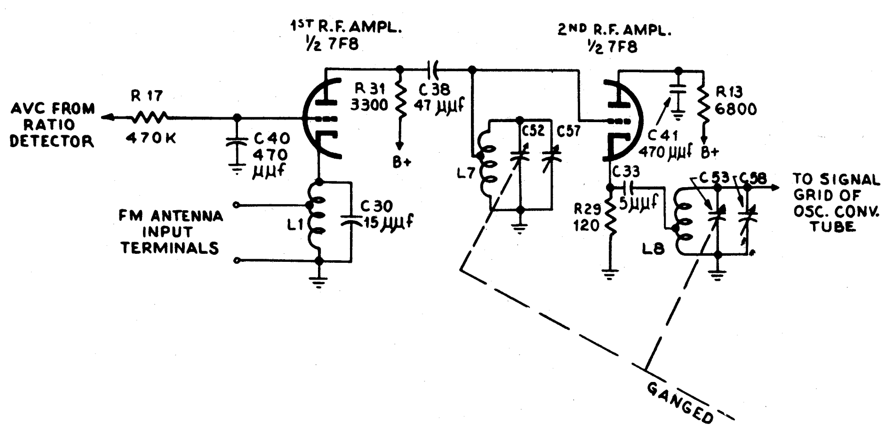



FIG. 4.—A simplified schematic of the fm-rf section of the Westinghouse models H-164, H-166, H-166A, and H-167. The grid of the first r-f amplifier is grounded to r-f signals through the 470-$\mu\mu f$ capacitor *C*40.

as far as r-f is concerned. At 100 mc, which is about the center of the f-m band, the reactance of this 470-$\mu\mu f$ capacitor is 3.4 ohms which is virtually a short circuit. Thus we can see how this triode section of the 7F8 tube functions as a grounded-grid amplifier.

Although not shown in the drawing, the other end of resistor *R17* is connected to the output circuit of the ratio detector of the receiver. In this manner avc voltage is applied to the grid circuit of this triode. In fact no other tube in these Westinghouse models receives avc voltage on the f-m band. (This is in contrast to the previous models discussed wherein avc voltage was not applied to the grounded-grid amplifiers.) The d-c return path for the grid of the input tube is through R17 and the ratio detector load resistors to ground.

The second r-f amplifier of this unit also presents an interesting feature. The output signal from the first r-f amplifier is coupled to the grid circuit of the second triode section via the tuned parallel circuit consisting of *L7*, *C52*, and *C57*. The output signal, instead of being taken from the plate circuit of the tube, is taken from the cathode circuit. Such a circuit is termed a *cathode follower*. The output load on this stage is in the cathode circuit and consists primarily of the cathode resistor *R29*, to which is coupled capacitor *C33* and the tuned circuit comprising *L8*, *C53*, and *C58*. Signal currents exist in the cathode circuit of the tube as well as in the plate circuit. The output signal is taken from the high side of the tuned circuit and fed to the signal grid of the oscillator-converter tube. This tuned circuit which has its tuning capacitor *C53* ganged with the other units of the set, increases the selectivity of the f-m band.

Space does not permit a lengthy theoretical description of a cathode-follower circuit. However such a circuit does offer many advantages even though its gain is less than unity. In other words a loss rather than a gain is the result of such a circuit. One of the chief advantages in such a circuit is that its high-frequency response is excellent which is primarily due to the fact that the circuit has an *equivalent* plate resistance that is very low. The effective input capacitance of a cathode follower tube is less than if the tube were used as an ordinary amplifier. This reduction in input capacitance is advantageous because the frequency response characteristic of the stage preceding the cathode follower is improved. Thus we see that due to the cathode-follower circuit of Fig. 4 the frequency response of both r-f tuned circuits is improved.

# APPLICATION OF THE PRINTED CIRCUIT

In the design of radio receivers and other allied electronic equipment, the trend is toward simplification of circuit construction and betterment of performance. Some examples of this have been the design of indoor loop antennas, the design of miniature and subminiature-type tubes, and also the use of selenium rectifiers. These advancements are well known, with the simplified loop antenna dating back over ten years. There have been many other developments — too numerous to mention here. A development that is comparatively new as far as its use in broadcast receivers is concerned will be discussed here.

### Majestic 6FM714, 6FM773

This new development is used in Majestic models 6FM714 and 6FM773. It is the resistance-capacitance coupling circuit that exits between the first audio amplifier and power output stage. The circuit in question is the same in each of these models and is illustrated in Fig. 1. Complete service data for model 6FM714 will be found on *pages 18-1 and 18-2* and for model 6FM773 on *pages 18-3 and 18-4 of Rider's Volume XVIII.*

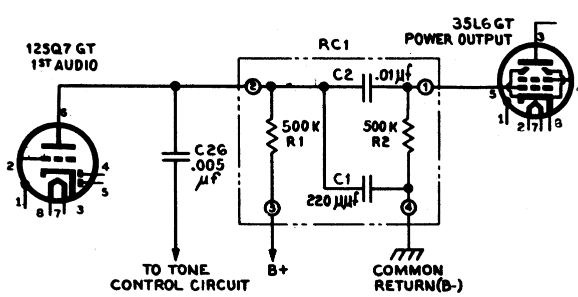



FIG. 1. — The printed circuit that acts as the resistance-capacitance coupling unit between the first audio amplifier and the power output stage is enclosed in the dashed box.

Let us examine the circuit which appears in Fig. 1 and see what the functions of the individual components are. Resistors *R*1 and *R*2 and capacitors *C*1 and *C*2 comprise the so-called coupling circuit between the two tubes. Resistor *R*1 is the plate load for the triode section of the 12SQ7GT tube, capacitor *C*2 is the coupling capacitor which also blocks d.c. from getting to the grid of the power output tube and resistor *R*2 is the grid-leak for the 35L6GT tube. The 220-$\mu\mu$f capacitor *C*1, which is attached directly to the plate of the first audio tube, is used to bypass any r-f or i-f signals that might appear at the plate circuit of this tube.

After looking at this circuit, one is likely to ask what is so special about it that warrants discussion. It appears the same as any ordinary resistance-capacitance coupling circuit which exists between an audio voltage amplifier and power output tube. Although it is true that in schematic arrangement the circuit represents the usual type of R-C coupling, the physical construction of this unit is the interesting thing. The audio coupling circuit composed of components *R*1, *R*2, *C*1, and *C*2 is shown schematically enclosed in a dashed box and labeled *RC*1. The construction of these four circuit elements are in the form of what is known as a *printed circuit*. The manufacturer refers to this *RC*1 as a "printed circuit plaque" (audio coupling). The lettered designations of these units are ours, not the manufacturer's, and were inserted for ease of discussion. The other components in the receiver, such as *C*26 in Fig. 1, have the manufacturer's lettered symbols.

### What Is a Printed Circuit?

Printed circuits are a relatively new phase of electronic engineering and are considered as a wartime development.* Today great advancements have been made in printed circuits; complete transmitters and receivers are now printed in various forms. Before we go too far in telling the advancements made with printed circuits, let us first understand what a printed circuit is and the reasons for its use.

A printed circuit, in brief, is a method whereby wiring and certain circuit components are printed on ceramic or other type surfaces. This printing may be accomplished by stencilling, spraying, or painting on

*For a more detailed discussion of printed circuits see "Printed Electronic Circuits" by C. Brunetti and A. S. Khouri in Electronics, April 1946 and "Printed Electronics Circuits," *NBS Technical News Bulletin*, Feb. 1947.

the surface. Resistors, inductors, and capacitors have been successfully printed onto different types of surfaces. There are cases where a complete circuit is printed around the glass envelope of a subminiature type of tube. **Many different types of circuits, from audio amplifiers to high-frequency receivers and transmitters, have utilized the printed circuit technique.

Reduction in the size of certain electronic devices was the original purpose of printed circuits. The savings in space by the use of such circuits are remarkable. In most cases, even though the reduction in size is great, the over-all electrical performance of the unit is as good, if not even better, than that of ordinary type circuits. The same type of frequency response as that of ordinary audio circuits has been obtained from audio amplifiers using printed circuit technique. In certain cases, such as in high frequency circuits, the use of printed circuits has given better performance primarily because of the smaller size of the component required and the smaller values of stray capacitances. Complete units can be made of a number of different printed circuit sheets, thereby enabling easy servicing by replacing the sheet that contains the defective component. This is considered an advantage in service work where time is of the essence and the expense of the extra printed sheets of secondary nature.

### The Serviceman's Viewpoint

Now let us return to the circuit of Fig. 1. The circled numbers 1, 2, 3, and 4 indicate the four connections that are made from the printed circuit to the rest of the audio circuit. The use of this printed circuit in the receiver enables the unit to be somewhat more compact. As far as factory assembly is concerned, it is much faster to wire in a single unit (the printed circuit) with four connecting points than it is to wire in four individual circuit components comprising a total of eight connecting points. In the production of printed circuits, the stray and wiring capacitances will be practically the same for every printed unit. Consequently, by the use of such circuits in electronic devices, fewer adjustments in the manufacture and service of receivers will be required.

**See "Printed Circuits," a series starting with October 1948 issue of "The APCO Bulletin."

Let us try to analyze the use of such a "printed circuit plaque" from the serviceman's viewpoint. First of all if the insertion of a new printed circuit can be easily and quickly made then such a feature will benefit the serviceman. This is primarily so because time is one of the most valuable assets of the radio technician. Another important consideration is the cost. If only the coupling capacitor in the printed circuit of Fig. 1 becomes defective, the complete printed unit, $RC1$, has to be changed. This, of course, is a disadvantage if one feels that the difference in cost between a 0.01-$\mu$f capacitor and the printed circuit is too great. On the other hand, the repairman may find that he has saved enough time by working with the printed circuit to withstand the additional cost.

Another important point that should be considered is the availability of such replacement parts. Most servicemen have a ready supply of resistors and capacitors available in order to fix such common faults as defective coupling capacitors and load resistors. But what of the printed circuit, such as that being discussed here? Such components may be available through parts jobbers, but probably the only way they can be obtained is through the manufacturer, but here the element of time enters into the picture again. The serviceman has to put the set aside, write away for the unit, and wait for its delivery. Requesting a new printed circuit from the manufacturer may be considered in the same manner as ordering any other item, such as an oscillator coil, which is not available from the parts jobber. Whether the time lost in such a transaction makes the use of the printed circuit less valuable is something every serviceman must decide for himself.

Although very few commercially manufactured broadcast radio receivers employ printed circuits in whole or in part, nevertheless, these circuits are definitely here to stay. Exactly what the future will bring in the design and manufacture of receivers is debatable, but it is believed that the use of printed circuits will play a very important role. It is difficult to predict just when printed circuits will be used extensively. It may be 5 or even 10 years from now. Things are changing so rapidly in this modern world of ours that it is impossible to set any precise time for such developments.

# AUDIO NOISE SUPPRESSION

A very important problem in the recording and reproduction of sound is the elimination of unwanted sounds, or noise. The art of making disc recordings has advanced to the point today where a brand new pressing has a very low noise level. Unfortunately, this high quality is not permanent; as a record is used and re-used the surface noise, or scratch, increases in intensity. At the same time, the very high treble notes originally pressed on the disc by the manufacturer are gradually erased. As a result, *soft* treble passages are masked by scratch after a record has had some use, *loud* treble passages, of course, are not affected this way. This discrepancy is put to good use in the Philco Electronic Scratch Eliminator and the H. H. Scott Dynamic Noise Suppressor.

Before discussing these two devices, let us recall the simplest type of scratch suppressor. This is the treble tone control, or scratch filter; the former name is usually used when the control is variable, the latter when it is fixed. The effect, in either case, is simply to produce a certain amount of treble attenuation. Even in the case of the variable tone control this amount is likely to be fixed for the duration of at least one record; he is a rare listener who is willing constantly to monitor the treble tone control on his set! As a result, there are two possible conditions: the treble may be cut sufficiently to obscure scratch on soft treble passages, in which case loud treble passages lose naturalness because of the poor high-frequency response; or the tone control may be set to give good reproduction on loud treble parts, resulting in excessive scratch being heard when the treble is soft. The remedy for this difficulty, of course, is to provide an automatic treble control, which will reduce the high-frequency response on soft treble passages and give wide-band response when the treble is loud. Unfortunately, this is more cheaply said than done, so that it has only recently become practical to include such a control in a home-type radio receiver.

The problem that exists with the treble tone control exists also with the base control, but to a much lesser extent. For this reason, the Philco Electronic Scratch Eliminator operates only in the treble range; but the Scott Dynamic Noise Suppressor, which is somewhat more complex, regulates the bass response as well.

### Philco Electronic Scratch Eliminator

The Philco Electronic Scratch Eliminator is used in Philco model 48-1286. (This model is shown on *pages 18-165 through 18-179 of Riders Volume XVIII.*) The basic principle is illustrated in Fig. 1. Here is seen a resistive voltage divider, part of which is shunted by a variable capacitor. In the bass and middle ranges the effect of the capacitor, even at its maximum, is slight, so that all frequencies in these ranges are equally attenuated. Because of the values of resistance used, this attenuation is negligible. In the treble register, however, the reactance of the variable capacitor may be considerably less than one megohm, in which case the highs will be very materially

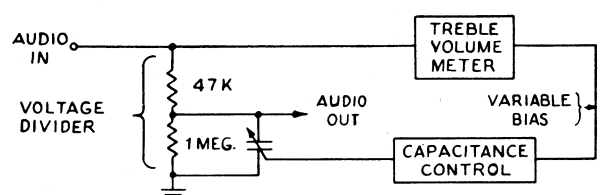



Fig. 1. — The principle of the Philco Electronic Scratch Eliminator is illustrated by this basic circuit.

attenuated by the divider. On the other hand, if the capacitive reactance is made very high, even the highs will suffer no more attenuation than the low and middle tones. In this manner the variable capacitor controls the treble response. The setting of the variable capacitor, in turn, is determined by the level of the treble input to the circuit. The "treble volume meter" shown in Fig. 1 produces a variable bias, rather than an indication on a visual meter, and this bias is applied to a special vacuum-tube circuit which acts as a variable capacitor. This circuit depends for its action upon the Miller Effect, so called in honor of a research worker for the National Bureau of Standards who first described and analyzed this effect.

In Fig. 2 are two diagrams that illustrate in simplified fashion how this effect arises. Part (A) shows a vacuum tube used as an amplifier, and indicates the grid-plate and grid-cathode capacitances of the tube. In part (B) is emphasized the fact that a capacitive feedback current flows from the plate of the amplifier to the grid, through the grid-plate capacitance. Since this current is generated in the plate circuit, its amplitude depends upon the gain of the amplifier. Now, the effective input impedance of the amplifier is the ratio of the voltage in the input to the current. Therefore, the input impedance depends upon the gain of the amplifier. This impedance consists, of course, of resistance and reactance. If the load of the amplifier is a resistance, the reactive component of the input impedance will be capacitive, since it is determined by the feedback, which is capacitive. Neglecting the effect of the screen grid, the input capacitance of the amplifier tube in question can be calculated by the formula:

$$C_{in} = C_{gk} + (1 + A)C_{gp}$$

where A is the gain of the amplifier.

If an external capacitor, C, is added from the plate to the grid, in parallel with the existing grid-plate capacitance of the tube, its capacitance is added to the grid-plate capacitance, as in any case where two capacitors are in parallel. The input capacitance then becomes

$$C_{in} = C_{gk} + (1 + A)(C_{gp} + C)$$

If, as in the case of the circuit actually used in the Philco Electronic Scratch Eliminator, the external capacitor is much larger than the internal grid-cathode and grid-plate capacitances of the tube, the effects of the internal capacitances become negligible, and the formula for the input capacitance can be simplified to

$$C_{in} = (1 + A)C$$

Since the gain, A, is easily varied by means of the bias applied to the control grid of the amplifier tube, it is possible to use an amplifier with an external capacitor from plate to grid as an electronic variable capacitor.

This is what is done in the Philco Electronic Scratch Eliminator. Referring again to the simplified diagram of Fig. 1, we see that a variable capacitor is shunted across part of a voltage divider. The variable capacitor in this case represents a vacuum-tube amplifier and its associated network taking advantage of the Miller Effect.

The complete circuit of the Philco Electronic Scratch Eliminator as used in Model 48-1286 is shown

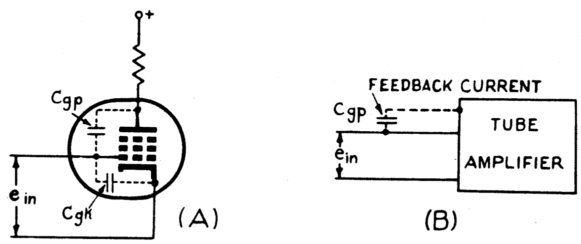



FIG. 2(A).— Simplified amplifier tube circuit showing the interelectrode capacitances which influence the Miller effect. In (B) the feedback circuit which produces the Miller effect is illustrated in block form.

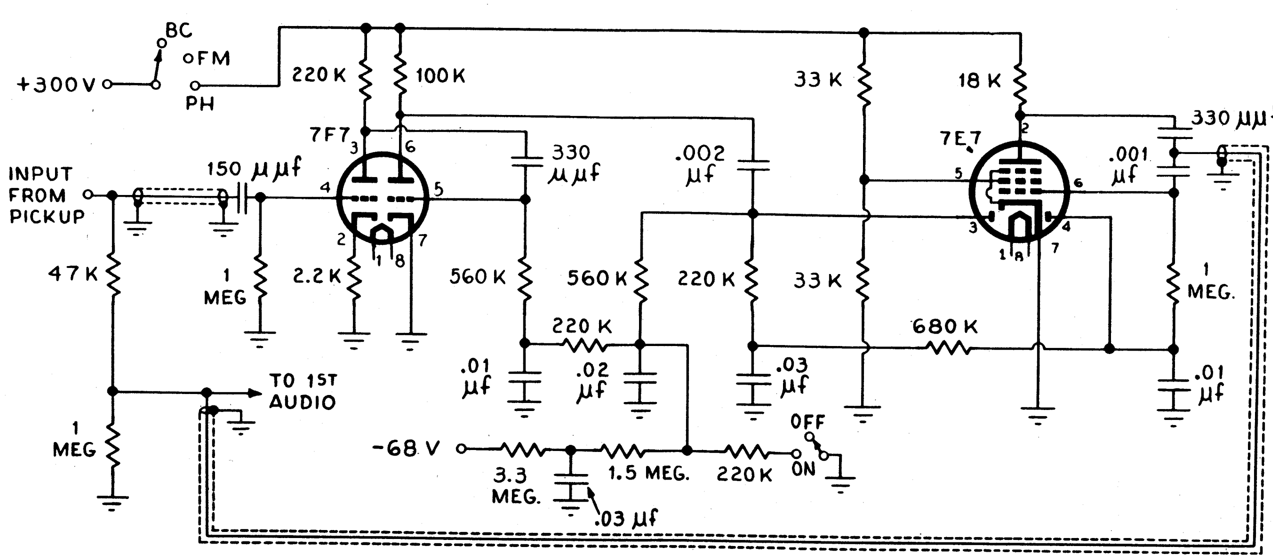



FIG. 3.— The schematic diagram of the Philco Scratch Eliminator as it is used in the Philco model 48-1286.

in Fig. 3. The 7F7 is a twin high-mu triode, providing two stages of amplification of the treble to drive the treble volume meter. These two stages are conventional except in the use of unusually small coupling capacitors. Through the use of these capacitors discrimination of the bass and middle register occurs, but the treble is amplified many times. The output of the treble amplifier feeds one of the diode plates of the 7E7, a duo-diode, remote-cutoff pentode. The rectifying action of this diode provides the variable bias output of the treble volume meter shown in Fig. 1. The variable bias is applied through a decoupling network to the grid of the 7E7, to control the gain of the pentode section of the tube. (The other diode plate, connected to pin 4 of the tube, performs no function in the circuit; this position on the socket of the 7E7 is simply used as a convenient tie point in the wiring of the set.)

A 3300-$\mu\mu$f and a 0.001-$\mu$f capacitor are connected in series between the plate and control grid of the 7E7. These components provide the large external capacitance which minimizes the effects of the internal tube capacitances, as explained above, and gives the tube a high variable input capacitance because of the Miller Effect. Only one capacitor is required for the variable capacitance effect, but two are used to isolate the grid, plate, and audio circuits from one another as regards d.c.

Let us review the action of this scratch eliminator: The treble portion of the audio input is amplified by the 7F7, and rectified by one of the diode sections of the 7E7. The resultant d-c voltage is negative with respect to ground, and its amplitude is proportional to the level of the treble in the audio input. Thus, for low levels of treble this negative voltage, which is applied to the control grid of the 7E7, is small, and the gain of the 7E7 is at its maximum. When the treble level is high, however, this bias voltage is also high, and the gain of the 7E7 is much reduced. From the equation for the input capacitance we see that the input capacitance is directly proportional to the gain of the amplifier, being large when the gain is high, and small when the gain is low. As a result, at low levels of treble input, the input capacitance of the 7E7 is high; but when the treble level is high, the input capacitance is low.

As was pointed out at the beginning of this discussion of the Philco Electronic Scratch Eliminator, a large value of capacitance in the voltage divider circuit produces considerable attenuation of the highs; this is what happens when the treble level, to begin with, is low, and record scratch would be objectionably noticeable. On the other hand, when the capacitance in the divider is small, the attenuation of highs is also small; this is the condition that exists when the treble level in the input is high, and the scratch is masked. Thus an automatic treble tone control action is obtained, which passes the highs when they are desirable, but eliminates them when they are not.

Before leaving this subject, we should consider the actions of the two switches shown in Fig. 3. The three-position switch (a portion of the band switch) is arranged to provide plate voltage to the eliminator only when the phonograph is being used, since it is only then that the eliminator is connected to the audio circuits of the set. The on-off switch provides a means of removing the eliminator from operation, if the user of the set so desires. When this switch is in the "ON" position, grounding the 220K resistor, the voltage at the junction of this resistor and the 1.5 meg. resistor is about −3 volts, providing suitable bias voltages for the second section of the 7F7 and for the 7E7. When the switch is in the "OFF" position, −68 volts is applied through the decoupling networks to these tubes, cutting them off, and preventing the eliminator from operating.

(Note: Shortly before the Second World War the E. H. Scott Radio Laboratories, Inc. used a somewhat similar circuit in several of their receivers. Circuit data and a brief description can be found in *Rider's Volume XIV*.)

### Scott "Metropolitan" Receiver

The "Metropolitan" receiver (see *pages 18-81, 82 through 18-83, 84, Rider's Volume XVIII*) made by the E. H. Scott Radio Laboratories, Inc., employs a version of the Dynamic Noise Suppressor developed by Hermon Hosmer Scott, Inc. The operation of this device is based on the same principle as the Philco Electronic Scratch Eliminator, namely, that when the signal level at the ends of the audio band is high, these signals at the two frequency limits are retained, but when the signal levels are low, they are rejected. The reason for this is that the noise level for a given set of conditions (any one phonograph record, say, or album of records) is fairly constant, but the signal level, particularly at the ends of the audio band, varies greatly. Since most noise is at the ends of the audio band, particularly the high end, it is desirable to reduce the audio bandwidth when the audio signal level is low, and to increase it when the level is high, as this mode of operation will maintain a more nearly constant signal-to-noise ratio. The high-frequency end of the audio band contains by far the greater proportion of the noise under most conditions, and particularly when phonograph records are played. For this reason, a device that controls only the treble response will produce considerable reduction in the apparent noise level.

At the same time, the reduction of treble noise tends to produce an apparent increase in bass noise, such as hum. Another objection to audio bandwidth variation at the high-frequency end only is based upon the characteristics of human hearing. It has been found that when the bandwidth of a reproducing system is varied, a natural effect is best preserved by changing the high- and low-frequency responses simultaneously in such a way that the mid-point of the band remains fixed. (If only one end of the band is cut, the middle of the resultant band is effectively shifted toward the other end.) It is therefore desirable on two counts that, when the ultimate in fidelity is a requirement, a noise suppressor should control *both* ends of the audio band.

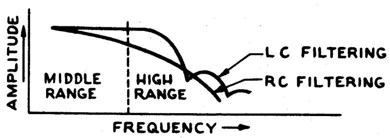
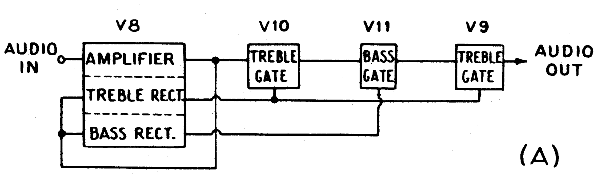



FIG. 4.—The different effects that R-C and L-C low-pass filters have on the frequency range.

A further desirable quality is that the ends of the audio band be more sharply defined than is possible when resistance-capacitance filtering is used. The reason for this is that the gradual change produced by RC filtering must extend into the middle-frequency range if high attenuation is to be obtained at the ends of the audio band. However, when a low- (or high-)



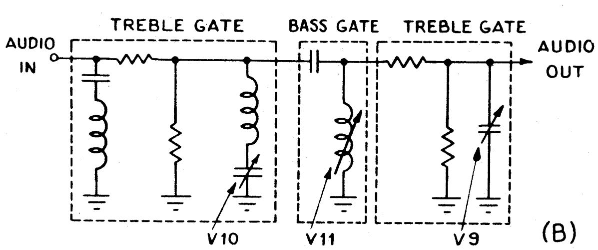




FIG. 5(A).—The block diagram of the Scott Dynamic Noise Suppressor; in (B), the simplified circuit of this noise suppressor is shown.

pass filter employing inductance and capacitance is used, a much sharper cutoff can be obtained. The difference between the RC and LC cases for low-pass filtering is shown (somewhat exaggerated for the sake of clarity) in Fig. 4. The quality of sharp cutoff is particularly desirable at the high end, because the noise level is higher here than at the low end. At the same time, though, this sharpness must not be carried too far, or transient oscillations may occur.

Fig. 5(A) shows in block diagram form the Dynamic Noise Suppressor of the "Metropolitan" receiver. (The blocks are numbered to correspond with the tube numbers on the main schematic.) The amplifier is the pentode section of a 6B8, and employs a conventional circuit. The treble and bass rectifiers are the diodes of the same 6B8; they are fed from the pentode section through high- and low-pass RC filters, respectively. Their outputs are applied as bias voltages to the associated "gates", or electronic filters. In this respect, the Dynamic Noise Suppressor is much like the Electronic Scratch Eliminator. However, where the Philco circuit employs the Miller Effect to obtain variable reactance, the Scott device makes use of reactance tubes such as are used in f-m transmitters and r-f generators. (The principles of reactance tubes have been described in many places; see, for example, "FM Transmission and Reception" by Rider and Uslan, pages 54 to 62.)

The output of the amplifier is fed through three gates, two treble and one bass, which act as an audio band-pass filter of variable width. Two treble gates are used, as against one bass, because the amplitude of treble noise is usually greater than that of bass noise, and because greater treble boost than bass boost is available in the main audio amplifier. The circuits of the gates, highly simplified, are shown in Fig. 5(B). The first treble gate employs two series-resonant LC circuits, one of which is fixed, and the other variable by means of reactance tube *V*10. The two resonant circuits are decoupled by means of the fixed resistors, which prevent interaction. (The grounds shown in Fig. 5(B) are audio grounds, but may not be d-c grounds.) This gate has an amplitude-frequency characteristic similar to that shown for LC filtering in Fig. 4; the position of the first sharp dip is movable, being controlled by *V*10 and having a minimum frequency of 4 kc (determined by a preset control), while the second dip is fixed at 10 kc (also determined by a preset control). The two dips correspond to the resonant frequencies of the series LC circuits.

The bass gate also depends upon series resonance for its action. The variable inductance required here is produced by reactance tube *V*11.

The second treble gate does not make use of LC resonance; instead, it uses resistance and capacitance in much the same way as in the Philco circuit. However, the values of resistance used are such that noticeable attenuation begins at a higher minimum-level frequency. The effect of this gate is to produce a more rapidly-falling high-frequency characteristic than would be possible with one gate, without adding more sharp dips, such as are introduced by the first treble gate. In addition, it possesses a frequency characteristic which approximately complements the frequency boost available from the second stage of the main audio amplifier. That is, its frequency response curve is approximately a mirror image of the curve of the treble boost amplifier; when treble boost is used, the response curves upward in the treble region, but the response of the second gate curves downward. In this fashion, the effect of the treble boost is cancelled to a varying extent when there is little desirable signal in the treble register, but the boost is retained when it is desirable. The cancellation is obtained in a smooth manner because of the similar shapes of the boost and gate curves.

A five-position rotary switch (SW-3) provides manual control over the range through which the noise suppressor acts. In one position, it removes the suppressor entirely from the audio path through the set. In the other positions, one section varies the proportion of the audio signal fed to the bass and treble rectifiers; this controls the extent to which the gates will widen the audio pass band during loud passages. At the same time, another section changes the constants in the second treble gate, so that the rate at which this gate attenuates the highs is affected. These two sections are so arranged that under conditions of high noise level the switch may be set to feed relatively small audio signals to the rectifiers, so that even on loud passages the audio pass band is not very greatly expanded; with this same setting, the second treble gate produces relatively high attenuation of the treble. On the other hand, if the noise level is low, the switch can be set to apply a high audio level to the rectifiers, causing the gates to provide a wide pass band even on passages that are only moderately loud. At this setting the second treble gate constants are such that it produces only moderate attenuation of the highs at most. Thus, the effect of the Dynamic Noise Suppressor can be varied through four settings to compensate for conditions of noise that may be only slight, moderate, or highly objectionable.

## Garod 306

The previous two circuits were rather intricate in design. Of course, such circuits are extremely beneficial to the over-all performance of the unit, but they also, nevertheless, add to the cost of the set. In designing phonograph circuits (with or without receivers) in the lower price range, naturally such types of circuits cannot be included without increasing the cost of the set. In many such units it is desired to eliminate certain types of noise which occur at different frequencies. In the phonograph section of the Garod model 306, a simple noise filter is used in the form of a parallel resistor-capacitor combination. The service data for this model appears on *page 18-8 in Rider's Volume XVIII.*

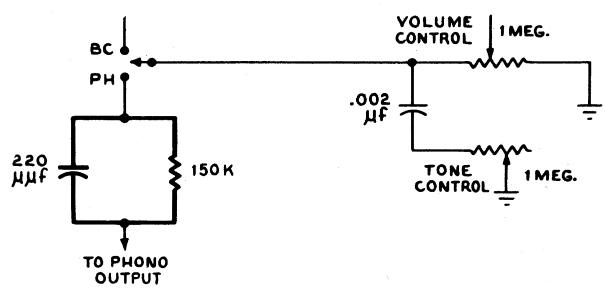



FIG. 6. — The parallel resistor-capacitor combination that is used as the noise filter in the phono section of the Garod Model 306.

A simplified schematic arrangement of this filter is shown in Fig. 6. The filter is located between the output of the phonograph and the phono section of the radio-phono switch. The resistor has a value of 150,000 ohms and the capacitor a value of 220 μμf. The primary reason for this circuit is to eliminate low-frequency noise, such as rumble in the phonograph unit.

The low-frequency attenuation is not at any one sharply defined frequency but occurs rather gradually. At the low audio frequencies, the impedance of the parallel resistance-capacitance unit is considered to be approximately equal to 150,000 ohms, that of the resistor alone. The reason for this is that the reactance of the capacitor becomes so high at the low audio frequencies that it is considered an open circuit. (At 100 cycles the reactance of 220 μμf is approximately 7.3 megohms.) At the high frequencies, however, the reactance of the capacitor decreases, thereby decreasing the over-all impedance of the network — hence providing a ready path for the high audio frequencies.

Many of the small-sized speakers that are used in phonograph circuits or phonograph-receiver combinations have a natural mechanical resonant frequency of a low order — the majority of these frequencies being within the range of 100 to 300 cycles. A low audio-frequency signal equal to or approximating the mechanical resonant frequency of a speaker will cause the speaker to resonate. If the signal is strong enough, the cone of the speaker will undergo a greater degree of displacement than is usual. The displacement is not gradual but occurs rather suddenly. This is undesirable because the abrupt transportation of the speaker cone is accompanied by an annoying "thumping" sound.

Orchestral recordings have many such low audio-frequency tones and often will cause the undesired condition described above. To eliminate this annoying characteristic some receivers are designed with a simple resistance-capacitance circuit the same as, or similar to, that shown in Fig. 6. This circuit will attenuate the low audio-frequency output from the phonograph and thus prevent the speaker from breaking into mechanical resonance. Of course, with such a circuit as this a disadvantage exists in that all the low audio frequencies are attenuated a certain amount and not just those frequencies causing the trouble.