HOW IT WORKS

Special Section of Volume XI

RIDER'S MANUAL

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HOW IT WORKS

LOOPS

One feature which is evident in the design of many of the newer receivers is the provision for loop operation to eliminate the need for an external antenna. Loops are now being used not only in small midgets but also in the larger multi-band receivers. In portables, a built-in loop is provided in practically all models.

In most instances, the loop is fixed in position within the receiver cabinet, but can be oriented by turning the receiver in the proper direction. An exception is the rotating loop employed in the RCA 46X1. In some, all the inductance in concentrated in the loop itself, while in others a loading coil is placed in series with the loop to build up the inductance to the required value for the range to be covered.

For operation on more than one band, the loop is either tapped, as exemplified by the Philco Model 40-510 shown in Fig. 1, or additional coils or loops are shunted across the main tuning loop to lower the inductance. Provision for using an external antenna and ground is usually made, either by tapping a portion of the loop inductance for connection of the antenna or by an additional winding inductively coupled to the main loop.

Philco 40-510 Loop

An example of a simple loop, tapped for use with an external antenna when such is used, is shown in Fig. 1. Note that the ground connection is made to the outer end of the loop. This is done in order to

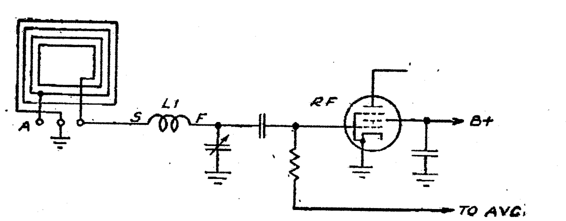


Fig. 1.—The loop used in the Philco Model 40-510 receiver. A tap is provided for use with an external antenna.

minimize variations in capacity to ground which otherwise might be appreciable if the grid were coupled to this point. In operation, the receiver may be moved close to grounded objects and thus additional capacity would be effectively shunted across the tuned circuit. When the grid is connected to the inner end of the loop, this capacity variation is minimized.

The loading coil *L1* is connected in series aiding with the loop so as to add sufficient inductance to enable full coverage of the broadcast range. The designations *S* and *F* refer to the start and finish of the loading coil winding. This connection must be observed when servicing if the tuning range is to be covered.

Belmont 411

Loop operation on both long wave and broadcast bands is secured in the Belmont Model 411, shown in schematic form in Fig. 2, by a combination of loop and series inductances.

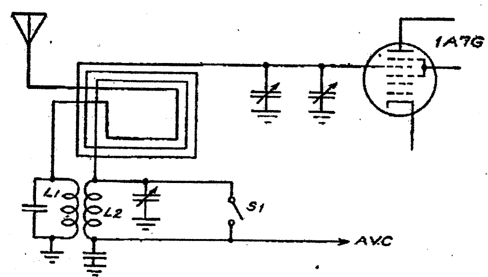


Fig. 2.—The loop employed in the Belmont 411 receiver covers the standard broadcast band and, when S1 is opened, the inductance of L2 is added so that long-wave operation is secured.

As shown, an antenna transformer is used in conjunction with the loop, the primary coil *L1* of the antenna transformer being connected in series with the antenna winding of the loop. The secondary winding of the loop is correspondingly connected to *L2*. When operation on the standard broadcast band is desired, closing *S1* serves to short out the secondary winding *L2* and the main tuning condenser *C1* then tunes the loop alone. The primary coil *L1* remains in the circuit at all times.

Montgomery Ward 93WG-382

In the schematic shown in Fig. 3, two loops are employed in conjunction with a separate tuned coil to cover a frequency range from 528 to 22000 kc in three bands.

For band B, the larger loop is coupled through the band switch to the r-f amplifier grid and is shunted by the tuning condenser *C*. The auxiliary condenser directly across the loop is used for trimming. When the band switch is shifted to band C, the tuned *L1C1* is placed in parallel with the loop circuit, reducing its inductance and enabling coverage of the 2200-7000-kc band. *C1* is used for trimming on this band. When the band switch is set for operation on band D, the small "D" loop and its associated tuned circuit *L2C2* are placed in parallel with *L1C1* and with the larger loop, reduc-

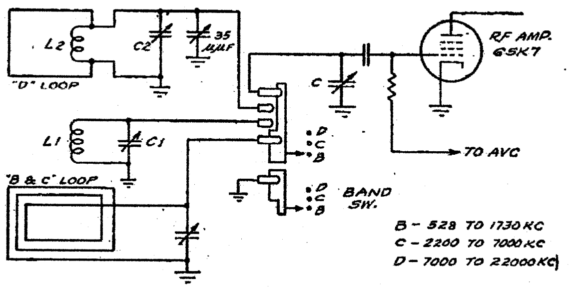


FIG. 3.—In the Montgomery Ward 93WG-382 receiver, shunting inductance across the main loop enables three bands to be covered.

ing the inductance of the circuit to the point where the highest frequency range may be covered.

The G-E Super Beam-a-scope

In the G-E Models H-77, H-78 and H-79 receivers, an electrostatically shielded loop is employed, as shown in Fig. 4. In this schematic, the loop is represented as *L1* and operates over the broadcast band only, while *L* is an antenna choke which functions as a parallel feed to the antenna transformers used on short-wave bands.

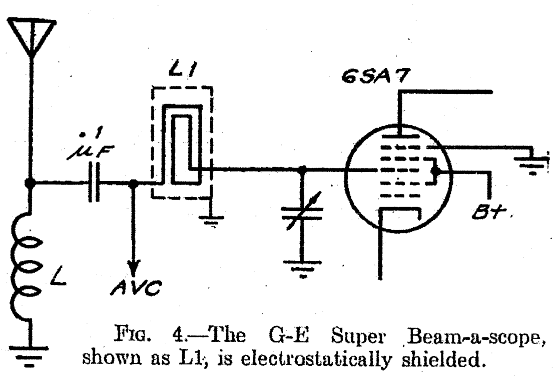


FIG. 4.—The G-E Super Beam-a-scope, shown as L1, is electrostatically shielded.

The Faraday shield around the loop acts as a screen against electrostatic disturbances and serves to improve the signal-to-noise ratio. Since noise voltages are composed of electrostatic and electromagnetic components, and the former predominates, by revolving the loop it is often possible to find a point where the voltages induced by these two components will cancel out. Under such conditions, the desired signal may be received without interference. This adjustment is based on the assumption that the noise arrives from a single source and that the desired signal is not in direct line with the noise source.

The Faraday shield itself consists of a number of closely spaced parallel wires which are joined together at one end only, so that no closed loop results.

Zenith Wavemagnet

In many Zenith receivers, an electrostatically shielded loop, called a Wavemagnet, is used. When the wavemagnet is employed with receivers covering more than one band, a switching arrangement as shown in Fig. 5, is used.

The loop is designated in the schematic as *L1*, and is surrounded by two shields. When the switch *S1* is thrown to the wavemagnet position, these shields are connected together and are returned to ground through the low-impedance short-wave primary winding and a blocking condenser of .05 mf.

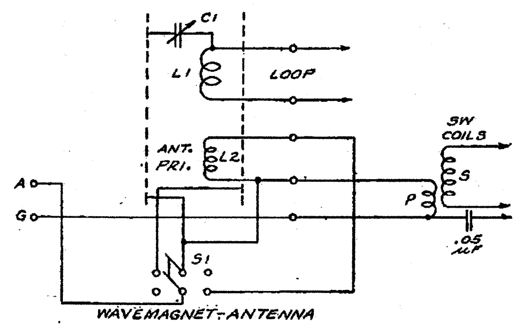


FIG. 5.—Another electrostatically shielded loop is the Zenith Wavemagnet, shown as L1 in the above diagram.

The trimmer condenser *C1* is used to compensate for variations in the distributed capacity between the shield and the loop winding and thus enables better tracking over the high-frequency portion of the standard broadcast band.

The loop is tuned only over the standard broadcast band; on short-wave bands, the loop assembly acts as a small antenna coupled to the short-wave input transformer primary winding.

When *S1* is thrown to the Antenna position, the connection between the two shields is opened and the outer shield serves as a small antenna which is capacity coupled to the loop by *C1*. When an external antenna is employed, the coil *L2* serves as a coupling means to the loop circuit. This coil functions only when an antenna is used.

A feature of special interest concerning the Wavemagnet loop is that it may be removed from the cabinet of portable receivers and can be fastened by means of vacuum cups to windows, sides of buildings, etc. This makes it possible to operate the receiver in shielded buildings, trains and other places where the signal pickup at the receiver itself is low, due to shielding.

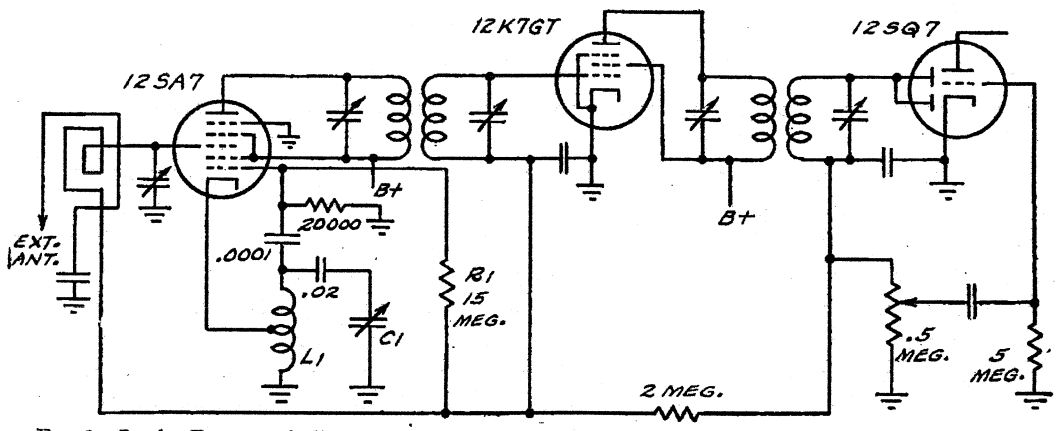


FIG. 6.—In the Farnsworth C1-1 receiver, a portion of the d-c voltage across the oscillator grid leak is used to provide delayed avc action.

OSCILLATOR VOLTAGE FOR BIAS CONTROL

A new method of obtaining a limiting bias on amplifier tubes is incorporated in the Farnsworth C1-1 receiver as shown in the schematic, Fig. 6. This is accomplished by feeding a portion of the rectified d-c voltage across the oscillator grid leak into the avc network. In this manner, sufficient control grid bias is applied to amplifier tubes so that the plate current does not become excessive when the signal level is so low that avc action is not present. This feature eliminates the need for cathode bias and provides an automotac control of sensitivity for variations in oscillator output over the operating range.

In the circuit shown, a negative voltage is normally developed across the 20,000-ohm oscillator grid resistor which varies with the strength of oscillation. This voltage is of the order of —10 to —30 volts in different receivers. As you will note, a portion of this voltage is applied to the avc bus through the resistor *R1*. The portion of this voltage so applied is approximately determined by the ratio of the resistance in the avc network to the total circuit resistance which includes *R1*. In this circuit, the avc network is composed of a 2-meg resistor and the diode load, a 0.5-meg volume control. The fraction of the total d-c voltage across the oscillator grid resistor which is present at the junction of the 2-meg and 15-meg resistors is therefore substantially equal to 2.5/15 + 2.5 or 1/7. If the voltage across the oscillator grid leak is —15 volts, then the actual voltage applied in this manner to the avc system is 1/7 of —15 or about —2 volts. Actually, the voltage under these conditions will be slightly less than this due to the loading effect of the diode across the 0.5-meg volume control.

Since the oscillator voltage varies to some extent over the operating range, the voltage fed into the avc system in this manner will also vary accordingly. Since the conversion gain of the converter tube decreases if the oscillator voltage drops below a certain value, an automatic compensation for this decrease in gain is obtained. For, when the voltage

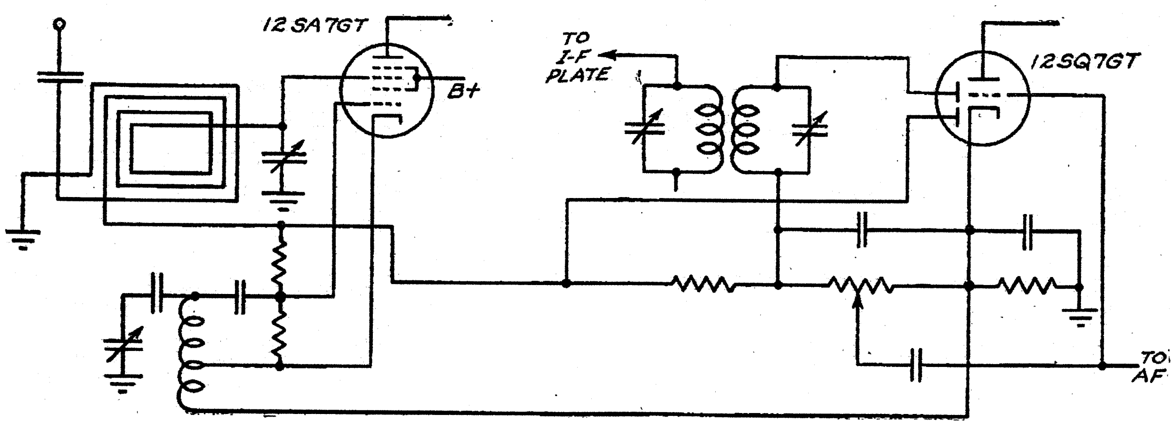


FIG. 7.—The circuit used in the Emerson CV264. This is another example of the application of a portion of d-c voltage across the oscillator grid leak for bias control.

across the oscillator grid leak decreases, the voltage fed into the avc system also decreases, thus increasing the gain of the controlled stages.

Delayed avc action also results from the use of this method. The negative voltage thus fed into the avc network is applied to the diode through the 2-meg resistor, preventing rectification until the signal voltage is greater than the negative voltage thus applied. The voltage which reaches the diode is lower than that applied to the avc bus due to the drop across the 2-meg resistor.

This feature is also incorporated in receivers of other manufacture, such as the Emerson Model CV264, Fig. 7.

POWER-SUPPLY FILTER SYSTEMS

In the past, the field coil of the usual electrodynamic speaker has often served as an inductance in the power-supply filter system. Now that the permanent-magnet dynamic speaker is becoming so widely used, in which no inductance is available for filtering purposes, resistance-capacity filters are becoming more common.

Since the current drain of all the tubes in a radio receiver amounts to an appreciable total, the values of the resistances used in power-supply systems cannot be high or the output voltage will drop. Consequently, to obtain sufficient filtration, the values of capacity are increased above those usually employed in choke-and-condenser filter systems. In addition, the heavy plate current drain of output power tubes is usually drawn from some section of the power-supply filter system where hum filtration is less, since more hum can be tolerated in the amplifier output circuit than in a preceding stage because the hum will not be further amplified.

Montgomery Ward 93BR Series

A resistance-capacity filter in one of its simplest forms is shown in Fig 8. This is the circuit employed in the Montgomery Ward 93BR Series. Note the high values of capacities employed in this two-section filter. The resistance values are kept low and do not exceed materially the d-c resistance of many speaker field coils.

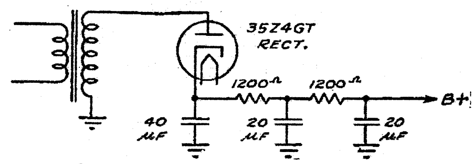


Fig. 8.—The power supply filter system of the Montgomery Ward 93BR series. Note that no filter chokes are employed.

General Electric HM-136

A more elaborate resistance-capacity power-supply filter system is shown in Fig. 9. In this General Electric receiver, the high plate current drain of the two 6L6G output tubes is taken off the power supply ahead of the resistance elements of the circuit. While a high hum voltage is thus impressed on the output tube plates, the hum voltage on each plate of the push-pull output is equal and opposite in phase, therefore it is canceled out and does not appear in the speaker output.

Note that the screen supply is taken off the first section of the filter, thus providing a somewhat lower voltage and sufficient filtration. The second

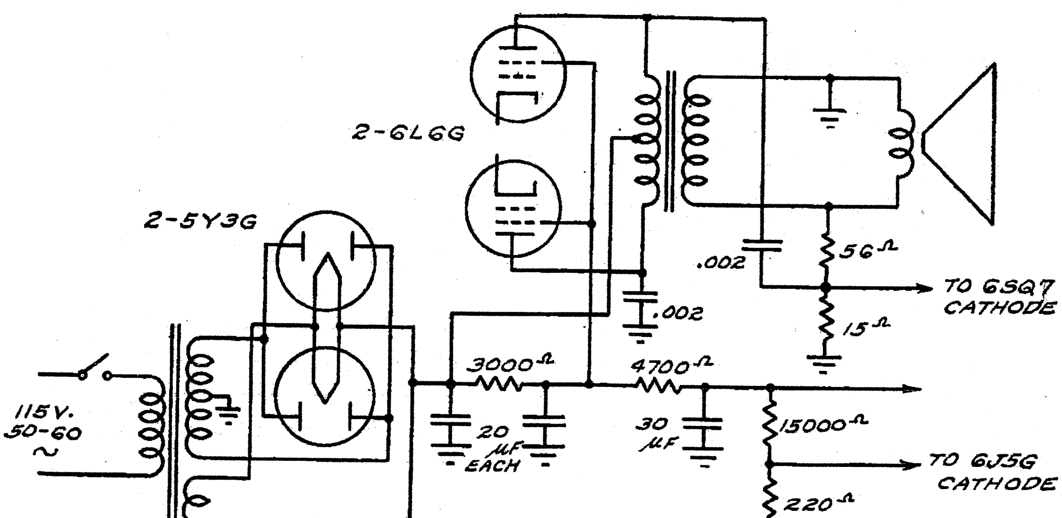


Fig. 9.—The resistance-capacity power-supply filter system employed in the General Electric HM-136 receiver.

section of the resistance-capacity filter employs somewhat higher values of resistance and capacity, since the filtration required for the amplifier tubes is greater. The 15,000 and 220-ohm resistors form a bleeder which prevents excessive voltage surges when the set is first turned on. Due to the fact that filament-type rectifier tubes are used in conjunction with heater-type amplifier tubes, the rectifier tubes will supply a high voltage before the amplifier tubes reach normal operating temperature and consequently before their plate current becomes normal.

RCA Model 46X21

An unusual power-supply filter system is employed in the RCA Model 46X21 receiver, a partial schematic of which is shown in Fig. 10. Note that

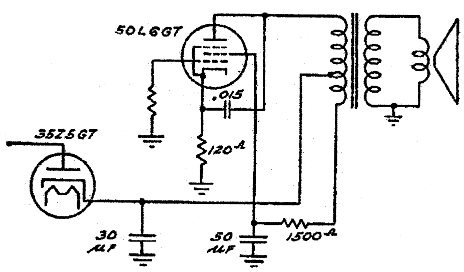


FIG. 10.—In the RCA 46X21 receiver, a portion of the output transformer primary is tapped to form a part of the power-supply filter system.

a portion of the output transformer primary winding is tapped to form a part of the filter network, in combination with a 1500-ohm resistor. The partly filtered power-supply voltage is fed to the tap on the output transformer primary and thence to the plate of the 50L6GT. The hum voltage present at the plate is reduced by the degenerative action of the unbypassed cathode resistor. The screen supply is more completely filtered by the 1500-ohm series resistor and its associated 50 mf filter condenser.

Farnsworth C4-1 Series

In the diagram shown in Fig. 11, the power-supply filter system is somewhat conventional, except

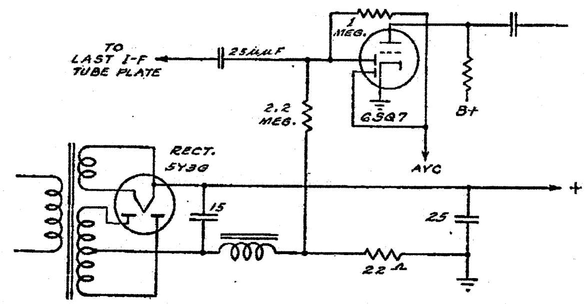


FIG. 11.—The avc delay voltage in the Farnsworth C4-1 series is applied through the 2.2-meg resistor.

that the choke and resistor are in the negative leg of the filter system. The feature here is that avc delay bias is taken off the junction of the filter choke and the 22-ohm resistor. This negative voltage is applied to the avc diode through the 2.2-meg series resistor. This high resistance is necessary in order to prevent any undue loading of the diode circuit.

NEW NEGATIVE FEEDBACK CIRCUITS

Simplified inverse feedback circuits are employed in a number of receiver models of different manufacture. In its simplest form, negative feedback is obtained by omitting the usual cathode bypass condenser across the power tube cathode resistor, as shown in Fig. 12. This circuit is employed in many of the smaller receivers, and by employing a somewhat higher plate load impedance increased power output is obtained without excessive distortion. The customary value of cathode resistor remains unchanged.

The effect of this omission of the cathode bypass condenser is to raise slightly the effective plate resistance of the tube, whereas in other methods the action is to cause the tube to function as if the plate resistance were lower.

In operation, the effective signal input to the tube is composed of the a-f signal applied from the preceding stage and that portion of the output signal developed across the cathode resistor. Since the latter voltage is out of phase with that of the incoming signal, it tends to oppose, and therefore reduce, the total signal voltage from grid to cathode.

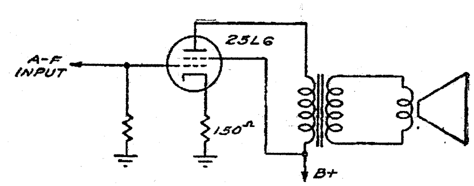


FIG. 12.—In the above circuit, negative feedback is obtained by simply omitting the cathode bypass condenser.

Since the harmonics thus fed back are also in phase opposition to those present in the input signal and are proportionately greater in magnitude, the action is to reduce the harmonic components in the output signal and consequently the distortion.

Another simple method is employed in the Stromberg-Carlson 400-H series, shown in Fig. 13. The 1-meg resistor, joining the plate of the output tube to that of the preceding stage, serves to feed a portion of the output signal voltage into the plate circuit of the preceding tube. Since the output signal voltage is positive at the instant when the signal at the 6SQ7 plate is negative, the feedback is in opposite phase.

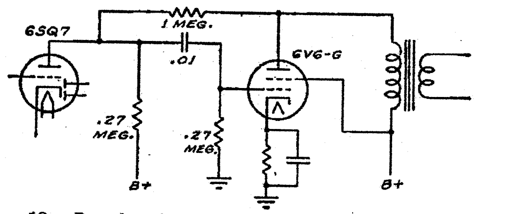


FIG. 13.—In the Stromberg-Carlson 400-H series, the feedback voltage is applied through the 1-meg resistor.

The percentage feedback introduced in this manner amounts to about 12 percent. This is determined by the ratio of the value of the effective input load resistance of the 6V6-G to the sum of the input load resistance and the feedback resistance. The effective input load is composed of the .27-meg plate load resistor and the grid resistor of similar value, which may be considered to be in parallel. The parallel load thus becomes 135,000 ohms. The ratio of 135,000 to 135,000 + 1,000,000 is about $\frac{1}{8}$; and $\frac{1}{8} \times 100\%$ equals about 12%.

In many of the newer receivers, negative feedback is introduced in the volume-control return circuit. When used in this manner, the percentage feedback is determined by the volume-control setting and is greater when the control is set near minimum.

One method of doing this is shown in Fig. 14 which is a partial schematic of the circuit employed in the Gamble 867A receiver. The volume control returns to ground through the voltage divider, *R1* and *R2*, shunted across the output transformer secondary. Since *R2* is 25 ohms and *R1* is 100 ohms, the percentage feedback is equal to 25/100 + 25 times 100% or 20%. This value represents the maximum percentage feedback.

The percentage feedback introduced into the first a-f grid circuit is dependent upon the volume-control setting, as illustrated in Fig. 15. In this diagram, you will note that the feedback voltage is divided by *R3* and *R4*, representing the volume control and the first a-f grid resistor. When the volume control is near maximum, as shown, the feedback voltage is reduced by the voltage drop across *R3*. Under such conditions, since the feedback is less, the gain of the amplifier is greater. When the volume control is set near minimum, the amount of resistance in series with *R4* is decreased

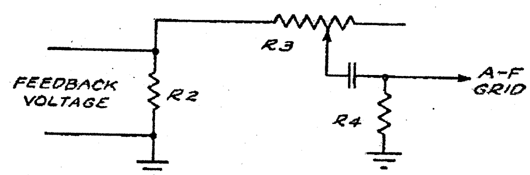


FIG. 15.—The amount of feedback voltage is dependent upon the volume control setting, as shown above.

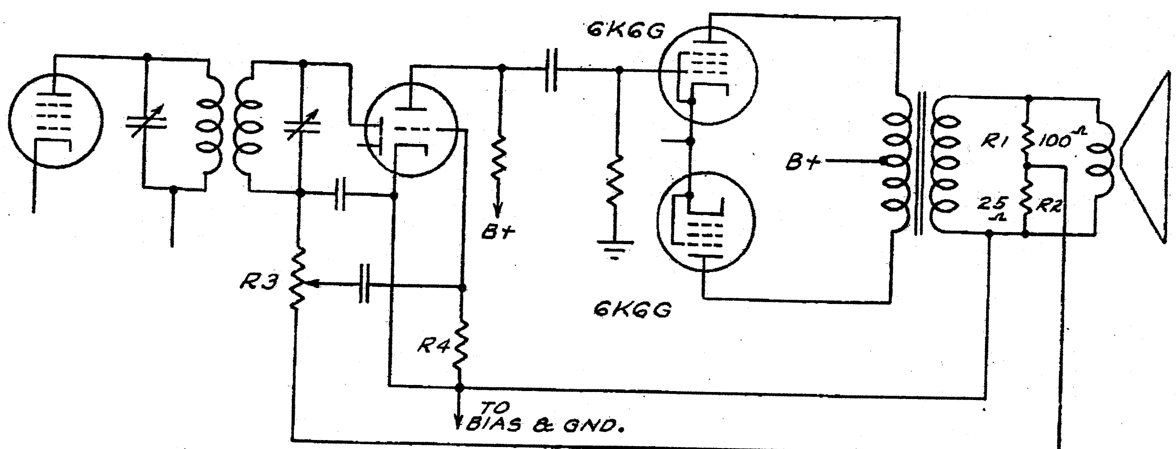


FIG. 14.—In the Gamble 867A receiver, negative feedback is introduced in the volume control circuit by returning the volume control to ground through a voltage divider network across the speaker voice coil.

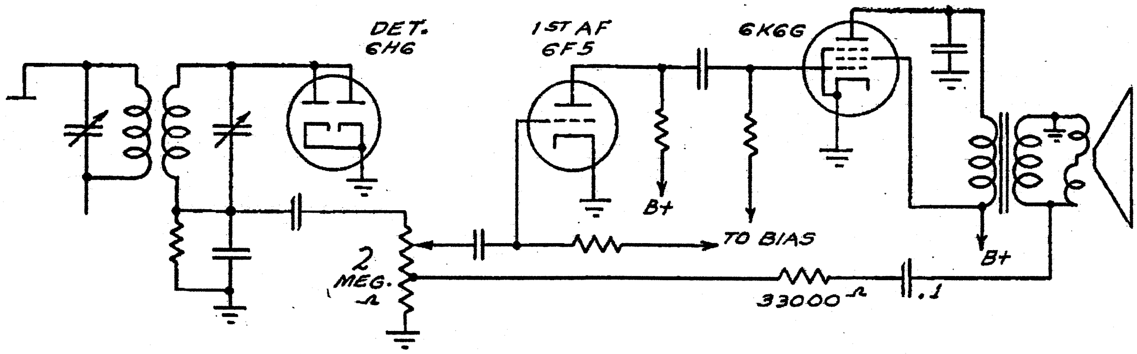


FIG. 16.—Another method of applying negative feedback to the volume control circuit is shown above. This circuit is used in the G-E G-75 receiver.

and consequently the feedback voltage is increased. This causes a decrease in output of the a-f amplifier beyond that contributed by the reduced volume-control setting. Also, since the negative feedback is increased, less distortion is present in the output.

This arrangement has the advantage that maximum feedback is available when most needed—on strong, local signals—while for weak signals maximum gain is obtained, which is the primary consideration under such conditions.

In the General Electric G-75, inverse feedback is obtained in the same manner as described above, though the circuit, as shown in Fig. 16, differs somewhat. The feedback voltage is fed to a tap on the volume control through a 33,000-ohm series resistor and an 0.1-mf condenser. The action, insofar as strong and weak signals are concerned, is similar to the system employed in the Gamble receiver previously described. In addition, the presence of the 0.1-mf condenser introduces another compensating effect. At low frequencies, when the reactance of the condenser is high, the feedback voltage is decreased, while at high frequencies the opposite effect results. Since the gain of the amplifier increases as the feedback decreases, the effect is an increase in gain at the lower frequencies which compensates to some extent for any falling off in the normal response of the amplifier and speaker at low frequencies.

Power Sensitivity

Negative feedback reduces the power sensitivity of any amplifier. Power sensitivity is defined as the ratio of the output power in watts to the square of the input audio signal in volts (rms) as in the equation below:

$$\textit{Power Sensitivity} = \frac{\textit{Watts Output}}{(\textit{Input Signal Volts})^2}$$

For example, if a 1/10 volt audio signal must be applied to the first audio grid of a two-stage audio amplifier to produce one watt output, the power sensitivity, expressed in mhos, is equal to $(1/0.1)^2$, or 1/.01 which equals 100 mhos. Most audio amplifiers in typical commercial receivers have a power sensitivity around this figure.

Since the gain is reduced when negative feedback is employed, the power sensitivity decreases also. In order to make up for this loss, it is customary to provide somewhat greater gain in amplifiers designed to be used with negative feedback. Alternatively, the feedback is so arranged as to have maximum effect over one portion of the a-f range and less over other portions.

NEW RADIO-FREQUENCY COUPLING CIRCUITS

In an number of the new receivers, the familiar tuned r-f transformer is no longer in evidence. Instead we find simpler circuits, circuits which require no tuning or, if tuning is used, the transformer design differs radically from those commonly employed. Resistance coupling, formerly employed only in a-f systems, now finds its way into the r-f stages of broadcast receivers. In some receivers, resistance coupling is combined with shunt chokes to modify the frequency characteristic; in others, simple resistance-capacity coupling is employed.

Untuned transformers, formerly used in the earliest broadcast receivers, again make their appearance in some of the newer receivers, though much improved in performance.

Farnsworth AC-70 Series

As shown in Fig. 17, simple resistance-capacity coupling is employed in the r-f stage of this receiver. Note the low value (1500 ohms) of the plate load resistor. This is characteristic of all applications of resistance coupling in r-f stages, since the bypassing effect of the tube and circuit capacities is so great that there is no advantage in using high values of coupling resistors at high frequencies. In a-f stages, the contrary is true. There, though

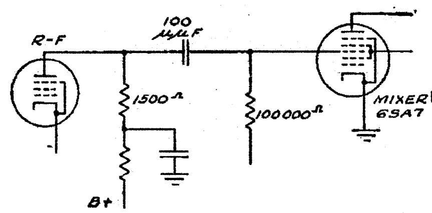


Fig. 17.—Resistance-capacity coupling in the r-f stage of the Farnsworth AC-70 series.

the tube and circuits capacitances are the same or greater, the reactance of these shunt capacities at low frequencies is so high that the signal is not bypassed.

In general, the higher the frequency the lower the value of plate resistance required to give any gain. Since the amount of gain which may be

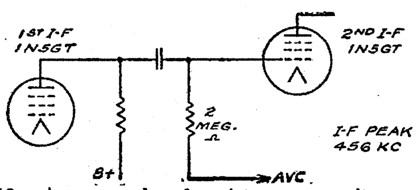


Fig. 18.—An example of resistance-capacity coupling in an i-f stage. The plate load resistor varies, in different models, from 10,000 to 20,000 ohms.

obtained, decreases as the load resistance is decreased it is apparent that the load resistance cannot be made too low without causing a stage loss rather than a stage gain. If the reactance of the shunt capacity of the circuit is lower than that of the load resistor, no increase in gain can result by increasing the plate load resistance. When resistance coupling is employed in all-wave receivers, operating at frequencies up to 18 mc where the reactance of the shunt capacities is extremely low, it is natural to find a very low value of plate load resistor. This is the case in the circuit shown in Fig. 17. Over such ranges, the r-f stage serves as buffer rather than as an amplifier.

Eliminating Oscillator Grid Condenser

To replace the usual mica condenser coupling the oscillator grid to the tuned tank circuit, the capacity coupling between a small dead-end coil and

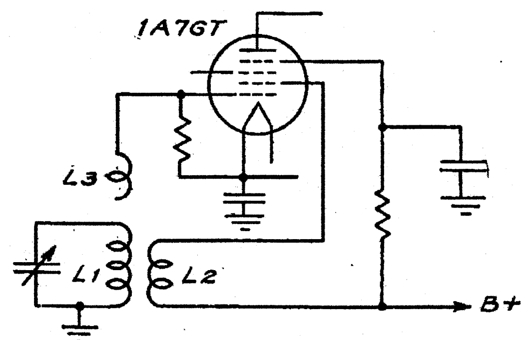


Fig. 19.—In the Emerson DF-302 receiver, the capacity coupling between L3 and L1 acts as a condenser.

the tuned circuit is used in some receivers. This is shown in the diagram, Fig. 19, which represents the oscillator circuit of the Emerson DF-302 receiver.

In the circuit shown, the oscillating circuit is composed of *L1* and *L2*, which operate in a tickler-feedback arrangement. The capacity coupling to

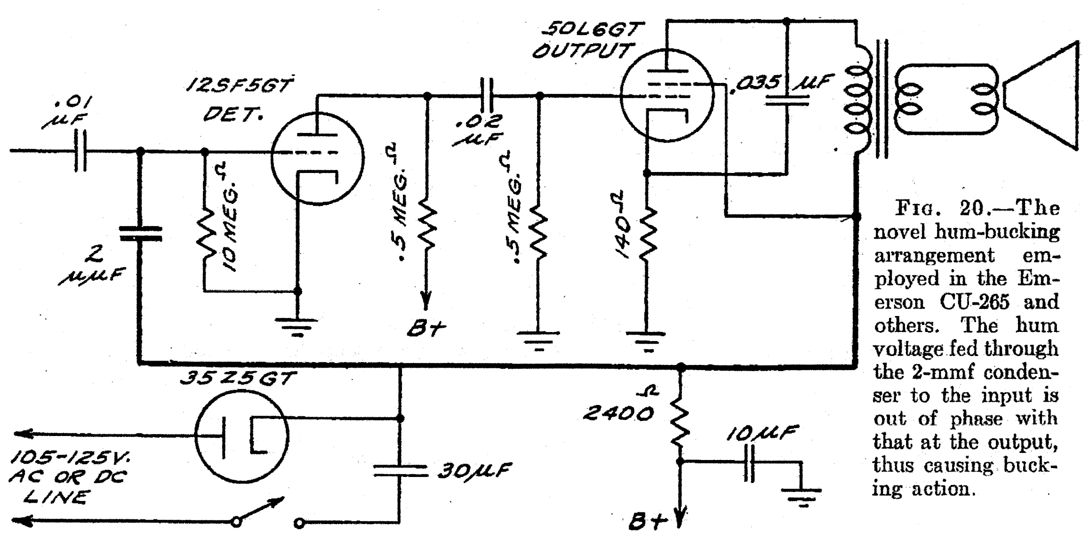


Fig. 20.—The novel hum-bucking arrangement employed in the Emerson CU-265 and others. The hum voltage fed through the 2-mmf condenser to the input is out of phase with that at the output, thus causing bucking action.

the oscillator grid is obtained by close coupling between *L3* and *L1*.

In signal tracing, a negative d-c voltage will be developed across the oscillator grid leak in the same manner as in the more conventional circuit arrangement employing a mica coupling condenser. This method of obtaining capacity coupling in oscillator circuits is also used in some RCA and Farnsworth models.

Hum Bucking In Emerson CU-265, CULW-261, 262, 265, and 274

A novel arrangement to secure hum-bucking action is shown in the schematic, Fig. 20. The hum voltage present at the filter input is fed through a tiny 2-mmf condenser to the grid of the 12SF5GT detector. Since this hum voltage is fed through a capacity into what is essentially a resistance load, its phase is shifted 90 degrees. In the plate circuit of the 12SF5GT, the hum voltage is reversed in phase and is fed to the output tube grid. In the output circuit of the 50L6GT, which is partly inductive, the phase of the hum voltage reaching the plate of the tube is such as to oppose a portion of the hum voltage fed to the plate of this tube from the filter circuit. In this manner an overall reduction in hum is secured.

MODERN PHASE INVERTER CIRCUITS

In more recent receivers, phase inversion circuits are simpler than those heretofore used. Otherwise they do not differ fundamentally from former designs.

RCA 9Q4

The phase inversion circuit used in the RCA 9Q4 is shown in Fig. 21. As illustrated, this is a single tube phase inverter employing the new 6SF5. The output of the triode section of the 6SQ7 first a-f tube is coupled to the lower 6F6-G power pentode and also to a voltage divider composed of *R1* and *R2*. The audio signal voltage developed across *R2* is capacity coupled to the grid of the 6SF5. This audio signal is amplified by the 6SF5 and reversed in phase. The amplified signal voltage across the plate load of the 6SF5 is then coupled to the upper 6F6G power pentode control grid. When this circuit is functioning properly, the signal voltages delivered to each of the output tube grids should be equal, though opposite in phase.

The average gain which the phase inverter tube should normally give is readily determined from the constants of the voltage divider *R1* and *R2* and is represented by the ratio of $R1 + R2$ to *R2* or more simply, because *R1* is much larger than *R2*, by $R1/R2$. Since *R1* is 1 meg and *R2* is 15,000 ohms, the rated gain of the 6SF5 tube and circuit is 1,000,000/15,000 or 67. Thus, when 1/67th of the signal voltage fed to the lower 6F6G is applied to the phase inverter grid and is amplified 67 times, the resulting output signal voltage which is applied to the upper 6F6G grid is the same as that fed to the lower 6F6G grid.

The modification of this circuit which makes for simplicity and greater freedom from hum is the omission of the cathode biasing resistor in the phase inverter tube circuit. By grounding the cathode

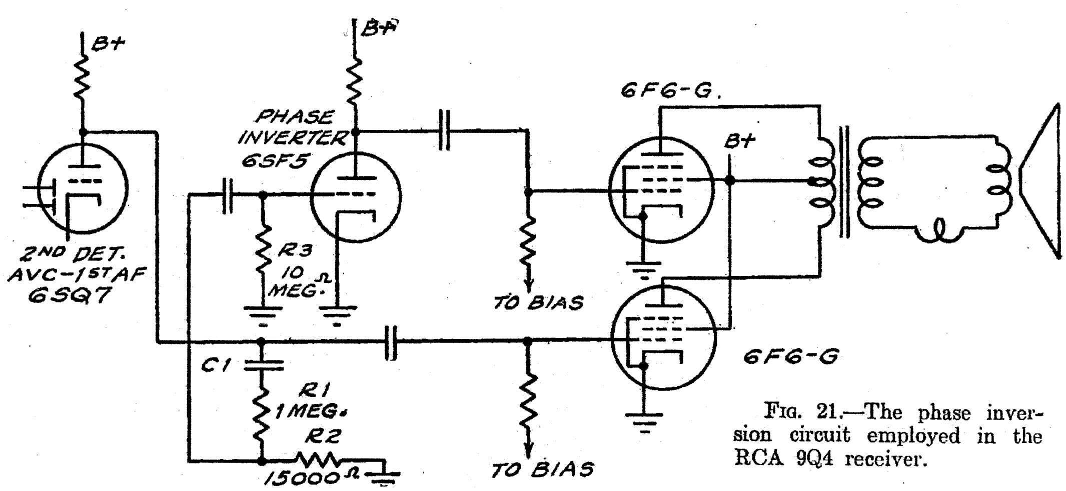


Fig. 21.—The phase inversion circuit employed in the RCA 9Q4 receiver.

directly, one resistor is eliminated and the cathode, being at ground potential, will not pick up hum to the extent as is usually experienced in former circuits.

This arrangement is made possible by the use of a high value of grid resistor. Though no grid bias is applied to the 6SF5, the high value of *R3* limits the amount of grid current which can flow. Thus, though the grid draws current over a large portion of the cycle, little distortion is introduced.

RCA 8Q2 Phase Inverter

An example of a double triode, used as an a-f amplifier and a phase inverter, is shown in Fig. 22. The audio signal is amplified first by one section of the 6SC7 and fed to the upper 6F6 output tube grid. The two grid resistors for this output tube form a voltage divider, as in the previous circuit described above. The audio signal voltage developed across the 10,000-ohm section of this voltage divider is fed to the phase inverter section of the 6SC7. The amplified signal voltage developed across the plate load resistor of the phase-inverter tube is then coupled to the lower 6F6G output tube grid.

The gain required in the phase inverter so that the output signal voltage applied to the lower 6F6 grid equals that fed to the upper output tube grid, is calculated by the same method as that previously described. Since the voltage divider is composed of a 470,000 and a 10,000-ohm resistor in series, 1/47 of the signal voltage fed to the upper 6F6 is applied to the phase-inverter grid. In order that the signal voltages applied to both output tube grids be made equal, then the phase inverter tube and circuit must amplify the signal 47 times.

As in the previous circuit, the cathode of the 6SC7 is directly grounded. The same advantages obtained in the case of this form of connection with a single triode are likewise secured with the double triode.

Note the low value of capacity used in the coupling condensers to the 6SC7 triode sections. Values larger than .0025 mf should not be used for replacement purposes, otherwise blocking may result due to the high value of the grid leaks used in the input circuits of this tube. The time required for a larger condenser to become discharged in a high resistance circuit is appreciable. Thus, if a large capacity is substituted, when the set is switched on, the charge which the coupling condenser accumulates due to the presence of the B-supply voltage at the plates of each tube section is, in part, discharged across the 10-meg grid resistors. Since this resistance is high, the time required for the charge to be dissipated is greater than when the resistance is lower in value. During this period the grids are temporarily at a high negative potential, causing blocking until the charge has leaked off.

Pushpull Parallel Operation

In circuits employing output tubes in pushpull-parallel operation, trouble due to parasitic oscillation often occurs unless special precautions are taken. In the diagram, Fig. 23, which shows the output circuit of the RCA Models K130, U-46, a single grid suppressor in each paralleled pair of output tubes serves to obviate this trouble.

These resistors are 1000 ohms each and are desig-

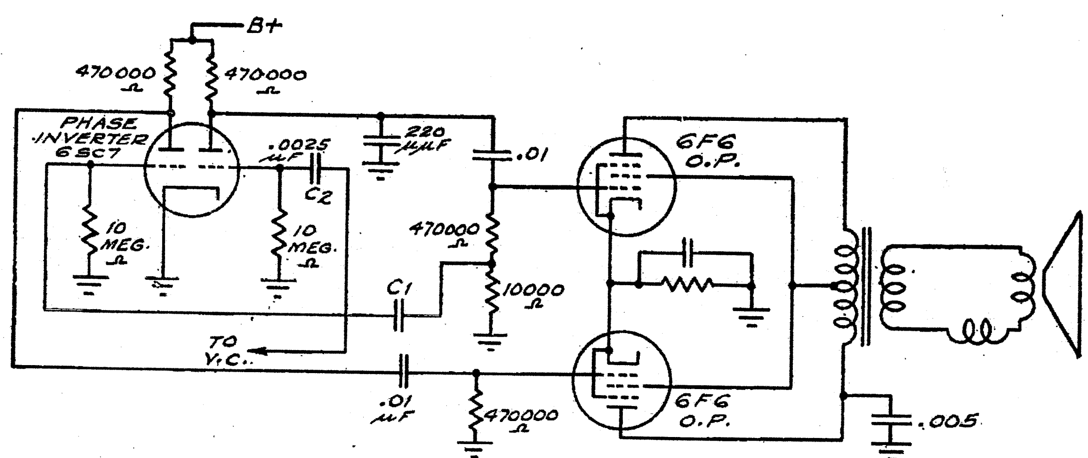


FIG. 22.—In the RCA 8Q2 receiver, the phase inverter is the 6SC7 double triode. Note that no cathode bias is employed in the phase inverter tube circuit.

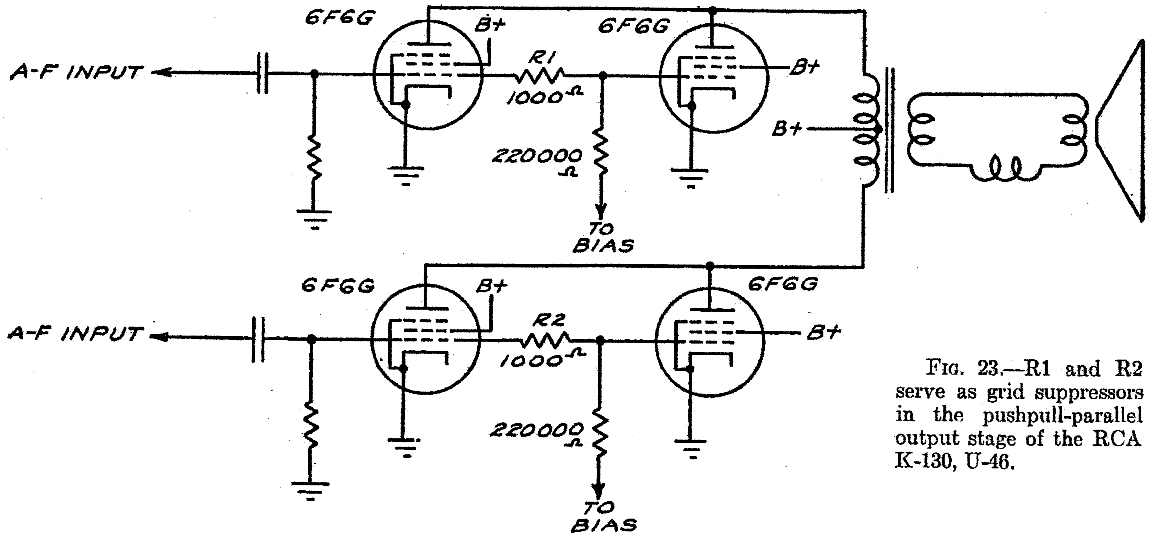


Fig. 23.—R1 and R2 serve as grid suppressors in the pushpull-parallel output stage of the RCA K-130, U-46.

nated on the diagram as *R1* and *R2*. Since the grid resistor in each of the return circuits to which these suppressors connect is much higher in resistance than the suppressors, the signal voltage applied to each grid is substantially the same as that which is fed to the grids of the tubes which do not employ suppressors.

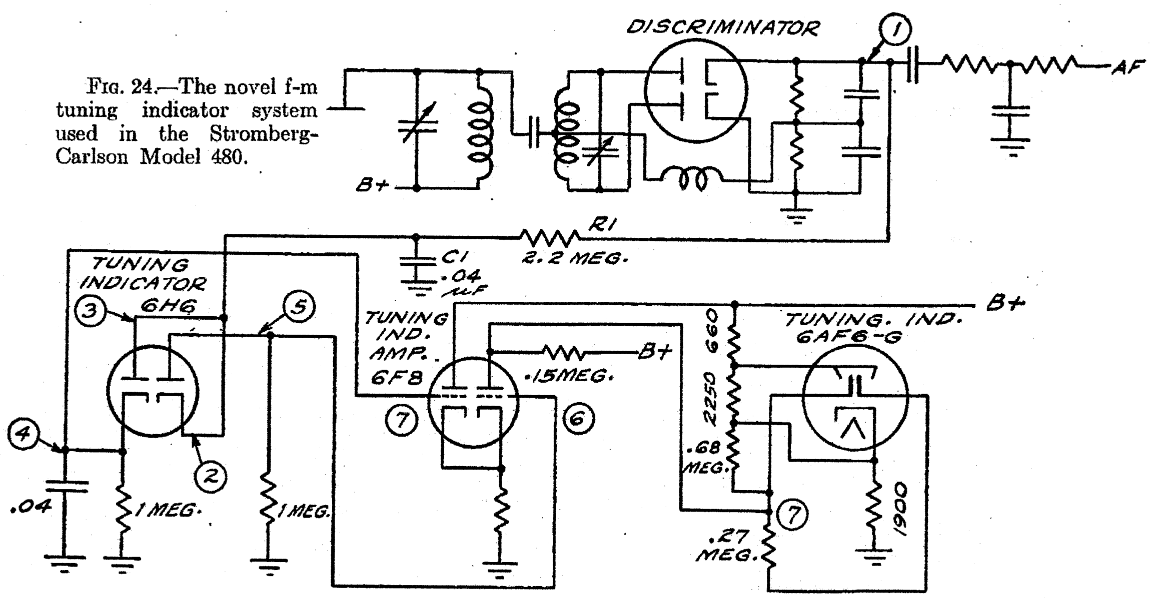
This arrangement is a simplification of former practice, in which suppressors were placed in each grid circuit.

F-M TUNING INDICATOR

A novel tuning indicator circuit, designed to make possible more accurate tuning of f-m receivers, is incorporated in the Stromberg-Carlson Model 480 receiver. This circuit is shown in the partial schematic, Fig. 24.

When the discriminator detector of an f-m receiver is tuned precisely to the intermediate frequency employed in the receiver, the voltage from point *1* to ground is zero. When a frequency-modulated signal is being received, and the receiver is properly tuned, the average voltage over an audio cycle at this point will likewise be zero, since the two halves of the demodulated wave will be equal. On the other hand if the receiver is mistuned, the demodulated output of the discriminator will not be

Fig. 24.—The novel f-m tuning indicator system used in the Stromberg-Carlson Model 480.



symmetrical and consequently point *1* will be either positive or negative with respect to ground. This fact is used in the application of the tuning indicator in this receiver.

In operation, the output of the discriminator is fed through the resistance-capacity filter *R1-C1*, which removes the audio signal variations, so that a d-c voltage is applied to points *2* and *3*. This voltage is zero when the receiver is properly tuned and consequently points *2*, *3*, *4* and *5* are at the same potential. Points *4* and *5* in the double diode circuit are connected to the 6F8 twin triode. This tube serves as a direct-coupled amplifier.

When the set is tuned off resonance, point *1* becomes either positive or negative, depending on whether the receiver is tuned above or below resonance with the incoming signal. When the receiver is tuned above resonance with the signal point *1* becomes negative with respect to ground, this negative voltage being applied to points *2* and *3* in the double diode. When point *3* is negative, the diode plate is negative with respect to cathode and no current flows; consequently no current flows in this diode section and the grid (point *7*) potential of the 6F8 remains unchanged. However, the cathode potential of the other section of the 6H6 becomes more negative with respect to its plate which is equivalent to saying that the plate becomes positive with respect to its cathode, accordingly diode current flows in this section and point *5* becomes negative with respect to ground. This negative potential is applied to point *6* and causes a decrease in plate current so that point *7* becomes more positive, causing the shadow angle of the 6AF6-G to decrease.

When the receiver is tuned below resonance with the incoming signal, point *1* becomes positive as do points *2* and *3* likewise. Point *4* then becomes positive with respect to ground due to conduction in this diode section. This causes an increase in cathode current of the 6F8 due to the positive bias applied to point *7*. Since the potential of point *5* remains unchanged because point *2* is now positive and no current flows in this section of the diode at point *5*, the result is that point *6* effectively becomes negative, the plate current decreases and again point *7* becomes more positive, causing a decreasing shadow angle in the 6AF6-G.

Thus it is seen that tuning either above or below resonance causes a decreasing shadow angle in the indicator tube, while at exact resonance the eye closes.

IMAGE FREQUENCY

In the operation of a superheterodyne receiver, the local oscillator heterodynes, or beats with, the incoming signal and produces in the mixer a signal at the intermediate frequency which has all the characteristics of the desired incoming signal.

Since the selectivity of the mixer input stage is not sufficient to eliminate undesired signals completely, interference may result due to the fact that the local oscillator in the receiver will heterodyne not only the desired signal, but also an undesired one, so that both are fed to the i-f amplifier. One type of interference which may result from this condition is known as image frequency response.

To understand what is meant by image frequency, let us consider the block diagram, Fig. 25, which represents a typical superheterodyne which employs no r-f stage. It is assumed that a 10,000-kc signal is being picked up by the antenna and fed to the mixer input circuit. The receiver is tuned to 10,000 kc; the i-f is 450 kc, hence the local oscillator in the receiver is functioning at 10,450 kc. This is the normal condition of operation.

The i-f amplifier being tuned to 450 kc, any 450-kc signal present in the mixer output circuit will be amplified. When the receiver is tuned to 10,000 kc, a 10-mc signal will present in the mixer circuit. The local oscillator signal at 10,450 kc combines with the incoming signal to produce a new signal which represents the sum and difference of the two frequencies present in the mixer. The difference between 10,450 kc and 10,000 kc is 450 kc and since the i-f amplifier is tuned to this frequency, the signal will be amplified. The sum frequency, which is equal to 10,450 plus 10,000 or 20,450 kc, will also be present but will not be amplified because the i-f amplifier is not tuned to this frequency.

Now let us assume, that in addition to the desired signal of 10,000 kc to which the receiver is tuned, a strong signal of 10,900 kc is present in the mixer. This could be the case, since there is but a single tuned circuit and ordinarily this is not sufficient to cut out completely a strong signal which does not differ by a large percentage in frequency from that of the desired signal.

The 10,900-kc signal, when mixed with the 10,-450-kc signal supplied by the local oscillator, produces a difference frequency equal to 10,900 — 10,450 or 450 kc. Since this is the frequency to which the i-f amplifier is tuned, the undesired signal will be amplified along with the desired signal, representing the difference between 10,000 and 10,450 kc, and interference will result. The frequency at which this interference results is called the *image frequency.*

Now let us see what relation the image frequency bears to the desired signal frequency. If the receiver is tuned to 10,000 kc and, as we have shown, the image frequency under such conditions is at 10,900 kc, the difference between 10,000 kc and 10,900 kc is 900, which is equal to twice the assumed intermediate frequency of 450 kc. If the intermediate frequency were 465 kc, then the local oscillator would function at 10,465 kc when the receiver was tuned to 10,000 kc. Also, a signal of 10,930 kc would produce an image frequency response. The difference between the desired and undesired signal frequencies would then be 10,930 — 10,000 or 930 kc. Again we see that the image frequency response occurs at a frequency which differs from that of the desired signal by twice the intermediate frequency. And we may set this up as a rule, that the image frequency will always differ from that of the desired signal frequency by twice the intermediate frequency.

In the examples above, we have seen that the image frequency is also higher in frequency than that of the incoming signal. In some receivers, particularly on short-wave bands, the oscillator functions at a frequency which is lower than that of the incoming signal. For instance, if the receiver is tuned to 10,000 kc and the set oscillator operates at 9,550 kc, an i-f signal, representing the difference between 10,000 kc and 9,550 kc, or 450 kc, is produced.

Now, if a 10,900-kc signal were also present in the mixer circuit, the beat between it and the local oscillator would result in the production of a signal frequency of 10,900 — 9,550 or 1350 kc. Since the i-f amplifier is tuned to 450 kc and not 1350 kc, no interference will result. Therefore, 10,900 kc, though it differs by twice the i-f from the desired signal frequency, will not produce interference when the oscillator frequency is lower than that of the signal frequency to which the receiver is tuned.

However, if a signal of 9,100 kc instead of 10,900 kc were present in the mixer circuit, when the set is tuned to 10,000 kc and the oscillator is functioning at a frequency of 9,550 kc, which is 450 kc *lower* than that to which the receiver is tuned, a signal representing the difference between 9,550 kc and 9,100 kc or 450 kc will be formed and will therefore pass through the 450-kc i-f amplifier and cause interference. This 9,100-kc signal, you will note, also differs from the desired signal frequency or 10,000 kc, by 900 kc, an amount which is also equal to twice the intermediate frequency. This then is the image frequency when the oscillator is lower in frequency than that of the incoming signal.

So we may see from the above illustrations that *the image frequency always differs from the desired signal frequency by an amount which is equal to twice the intermediate frequency.* Also, that when the set oscillator operates at a frequency which is higher than that to which the receiver is tuned, the image frequency will always be higher, by twice the intermediate frequency than the desired signal frequency. And, on the other hand, when the set oscillator functions at a frequency which is lower than that of the desired signal frequency, the image frequency will likewise be lower in frequency than that of the desired signal.

One point which deserves particular attention in this analysis is that image frequency has nothing to do with harmonics. While interference can also be produced due to harmonics of the oscillator beating with undesired signals, this type of interference is not due to image frequency response.

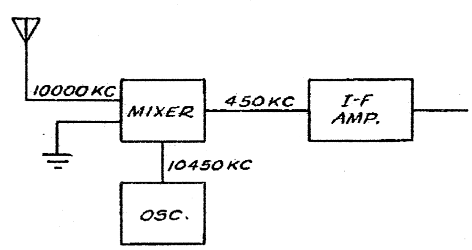


FIG. 25.—Block diagram showing how an i-f signal is formed.

The extent to which interference is produced because of image response will depend upon the strength of the interfering signal, the intermediate frequency employed in the receiver and the percentage difference in frequency between the image and the desired signal. Thus, when the intermediate frequency is 450 kc, the image frequency differs from the desired signal by 900 kc; when the i.f. is 175 kc, the image frequency is only 350 kc removed from the desired signal frequency. When the receiver is tuned to 550 kc, at the low frequency end of the standard broadcast band, and the i.f. is 450 kc, the image frequency occurs at 1450 kc . . .

the high frequency end of the band. The percentage difference in frequency in this instance is large. But, when the receiver is tuned to 20,000 kc under the same conditions, the image frequency at 20,900 kc differs but little in percentage from that of the desired signal. Accordingly, interference due to this cause will be much worse on short-wave bands than on the standard broadcast band.

Alignment Checks of Image Frequency

The fact that the local oscillator on short-wave bands can often be tuned to a frequency which differs from that of the desired signal by the i.f., but in the wrong direction, makes it desirable to check the alignment by making certain that the image response occurs at the proper point.

Thus when the receiver is to be aligned at 18 mc, the oscillator will normally be tuned to 18,450 kc, (if the intermediate frequency is 450 kc) but if the trimmer is screwed down too far, the oscillator frequency may be changed to 17,550 kc, which will likewise produce the required 450-kc i.f. when an 18-mc signal is tuned in.

To make certain the receiver oscillator is properly adjusted, after aligning at 18 mc, tune the test oscillator to 18,900 kc (or whatever frequency which is twice the i.f. higher than that frequency to which the receiver is tuned) and without changing any of the adjustments, note if a signal response is obtained. If the oscillator is adjusted to a frequency which is higher than that to which the set is tuned, the response should be obtained. If the set oscillator is adjusted to a frequency below that of the incoming signal, no response will result.

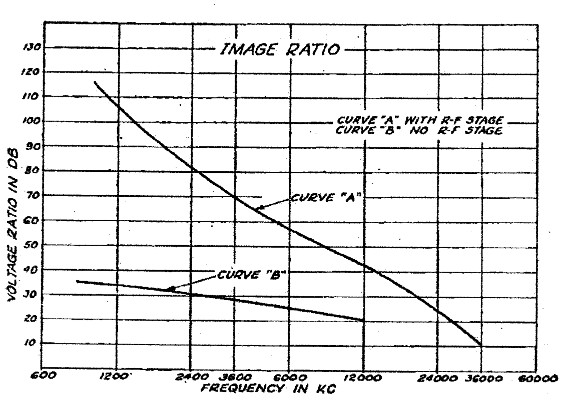


Fig. 26.—Curves showing the improvement in image ratio which results when an r-f stage precedes the mixer.

Image Ratio

When a receiver is tuned to a given frequency, the ratio of the input signal voltage at the image frequency to that required at the frequency to which the receiver is tuned is called the *image ratio.*

For example, if the receiver is tuned to 1000 kc and the i.f. is 450 kc, assuming that a 10-microvolt signal at 1000 kc produces a 50 milliwatt output at the receiver voice coil, then, if a 10,000-microvolt signal is required at the image frequency of 1900 kc to produce the same output while the receiver is tuned to 1000 kc, the image ratio is 10,000/10 or 1000.

Curves showing how the image ratio becomes lower as the frequency to which the receiver is tuned is increased are shown in Fig. 26. In this graph, Curve A is representative of the image ratios secured when an r-f stage is used ahead of the mixer in a high-grade receiver, while Curve B shows the lower image ratios which result when no r-f stage is employed.

In Curve A, the image ratio resulting when the receiver is tuned to 1200 kc is 106 db, corresponding to a voltage ratio of 200,000 to 1. That is, the signal input at the image frequency must be 200,000 times that required at the frequency to which the receiver is tuned, to produce the same output. This ratio decreases on the higher frequency bands, because of the relatively small percentage difference in frequency between the image frequency and the desired signal frequency on such bands. At 12 mc, the image ratio is 43 db, or 140 to 1, while at 36 mc it is only 11 db, or 3.5 to 1.

The improvement resulting from the use of an r-f stage is much more evident at frequencies in the broadcast band than at higher frequencies. As shown in Curve B, the image ratio secured with a representative receiver employing no r-f stage is 34 db, or 3500 to 1, at 1200 kc compared with 200,000 to 1 which is obtained with a receiver employing an r-f stage, over 50 times as great. Yet at 12 mc, where the image ratio on Curve B is 20 db, or 10 to 1, that obtained with the receiver employing the r-f stage is 140 to 1 at this frequency; only 14 times better. The improvement decreases more rapidly at still higher frequencies.

Silvertone Model 6335, 6435, 6490, 6495 Phase Inverter

A phase-inverter circuit employing degeneration is shown in Fig. 27. The audio signal voltage developed across the plate load resistor of the 6Q7G is fed directly to one of the pushpull output 25L6G's. A portion of this signal voltage also appears at point *2* and is fed to the grid of the 6J5G phase-inverter tube where it is amplified and reversed in phase. The output voltage of the phase inverter, appearing across *R4*, is applied through a .01-mf coupling condenser to the grid of the other 25L6G output tube.

This is the fundamental action of this circuit. However, since *R1* forms a portion of the grid return circuits of both output tubes, a portion of the signal voltage at point *1* will also appear at point *3* as well as point *2*. And, by the same token, the out-of-phase voltage developed in the phase-inverter circuit and applied to point *3* will, in part, also be applied to points *1* and *2*.

In action, the signal voltages at points *3* and *1* should be equal but opposite in phase. If the signal voltage at point *3* should tend to become greater than that at point *1*, then a greater out-of-phase voltage would appear at point *2* which would buck the signal voltage at that point derived from point *1*. This would result in lowering the signal voltage at point *3*.

If, on the other hand, we suppose that the gain of the phase inverter falls off due to tube depreciation or for any other reason so that the signal voltage appearing at point *3* tends to become less than that at point *1*, then the degenerative feedback voltage at point *2* also tends to become less and the resulting amplified voltage at point *3* increases.

This bucking action takes place at point *1* as well as at point *2*, insofar as the out-of-phase signal

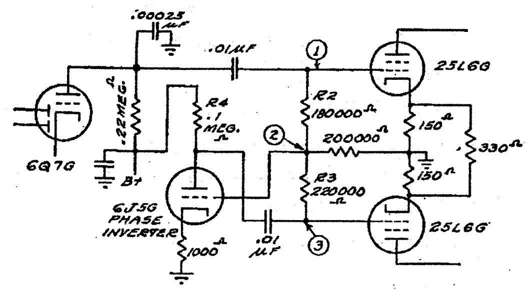


FIG. 27.—The degenerative phase inverter employed in the Silvertone Model 6335.

voltage is concerned. Also, the same action results from the presence of the in-phase voltage at points *2* and *3* in that it tends to oppose any increase in the out-of-phase voltages developed at points *2* and *3*. The net result is a high degree of stability of output.

Degeneration is also used in the phase inverter itself, by eliminating the cathode bypass condenser. This method is also employed in the 25L6G cathode circuits and by coupling the cathodes by means of a 330-ohm resistor.

RADIO-PIANO AMPLIFIER

A combination amplifier, designed for radio, record and piano amplification is incorporated in the Ansley Dynatone, shown on Ansley pages 11-1 in Riders Manual, Volume XI. The use of this amplifier in conjunction with an ordinary piano converts it into what is essentially an electronic piano. A five-stage audio amplifier is used in this instrument, though only four stages are utilized when the amplifier is used for radio or phonograph amplification.

The input stage for piano amplification consists of two triode-connected type 1223 tubes. The input grids of these tubes are coupled to adjustable condenser microphones which act as piano string pickups. The output is transformer-coupled to another amplifier stage composed of two triode-connected 1223's, the output circuits of which are in parallel, rather than pushpull.

One of the input grids of the second stage connects to a volume control which connects, in series with a 100,000-ohm resistor, to the secondary of the first stage transformer; the other grid is fed by the radio-record volume control.

The output of the second stage is resistance-coupled to one section of a 6F8 double triode, the second section of which serves as a phase inverter. The output of these two sections of the first 6F8 feeds to a similar double triode which is pushpull connected and drives a pair of 2A3 output triodes. The pushpull 2A3's operate with fixed grid bias, which is supplied by a 45 triode connected as a diode and operating as a half-wave rectifier. The bias voltage is taken off the junction of the 7,000 and 4000-ohm resistors which act as a voltage divider and is filtered by a 16-mf electrolytic condenser.

The radio section of this instrument utilizes but three tubes. The pre-selector tuned circuit ahead of the 6A8 converter serves to increase the selectivity. AVC is applied to the converter and i-f stages and is obtained from the section of the 6H6 which also acts as the second detector.

It is possible to operate the instrument as a piano amplifier in conjunction with record reproduction, thus enabling the pianist to accompany the record, if desired. When so operated, the second stage amplifier serves also as a mixer. When operated solely as a piano amplifier, the instrument is played in the usual manner and the action of the keyboard and pedals is the same as if the amplifier were not connected.

BATTERY-OPERATED PORTABLE RECEIVERS

Modern portable receivers are smaller, lighter and more convenient to use than those which were heretofore available. Better performance is also obtained, due to an increase in sensitivity which better tubes and improved circuit design make possible. Greater economy in operation is secured by the use of specially designed batteries in conjunction with modern low-current, multi-purpose tubes. In many models, also, provision is made for operating the receiver from either a-c or d-c power lines when battery operation is not required.

Combination "3-in-1" Receivers

Receivers which are designed to operate either from an a-c or d-c line supply or from self-contained batteries, represent the latest development in portable models. In all of these receivers, loop operation is used. Usually the loop is built within the cabinet, though one of the Motorola models has the loop woven into a shoulder strap. In the receivers employing a built-in loop, directional operation is secured simply by turning the receiver until the desired station is received with greatest volume.

The number of tubes required is kept at a minimum by using multi-purpose types, such as the new 3A8GT. This tube is a diode-triode-pentode type which performs three functions: the pentode section serves as an i-f amplifier, the diode as the second detector and the triode, which is a high-mu type, as the first a-f amplifier. This tube requires but 50 ma filament current at 2.8 volts, thus reducing the battery consumption as well as the space requirements over those which would be needed if individual tubes for each purpose were employed.

In most receivers, conventional coupling methods are used in r-f, mixer and i-f circuits. An exception is the use of resistance coupling in one i-f stage of the Emerson DF-302, in which some additional gain is obtained without appreciably increasing the space requirements which would be necessitated if an i-f transformer were used. Further, the additional tuning adjustments ordinarily needed with transformer coupling are eliminated.

In many models, the sensitivity to weak signals is increased by using delayed avc action, which is often obtained by applying a bucking voltage obtained from some portion of the filament circuit.

Operation from a-c power lines is usually obtained by using the new 117L7GT combination half-wave rectifier and output pentode. Since this tube has a 117-volt heater, it may be connected directly across the power line, thus eliminating the need for any voltage-dropping resistor.

When the receiver is line-operated, the pentode section of the 117L7GT, the input of which is connected in parallel with the battery-operated output pentode, functions alone, the switching arrangement being so devised that the filament circuit of the battery-operated output tube is opened when line operation is employed. The filaments of the remaining battery-operated tubes are connected in series with the cathode of the 117L7GT pentode so that the cathode current of this tube serves to operate the filaments of the other tubes.

Resistance-capacity filter systems are employed in the power-supply circuits when line operation is being used.

The maximum power output is greatly increased in line operation, being of the order of two watts, while with battery operation the usual output rating is approximately 275 milliwatts. Since the battery-operated output tube requires a higher plate load for efficient operation than the pentode section of the 117L7GT, the entire output transformer primary winding is employed in its plate circuit. For the 117L7GT pentode, the output transformer primary is tapped to provide the proper load.

The A-battery employed is normally a 6-volt flat-type unit, considerably larger than that formerly employed in older receivers. In some receivers, two 4.5-volt units, similar in design are used when the total voltage required for all the series filaments is 9 volts. The B batteries are 45-volt units, designed to fit in a small space. No binding posts are used; pin receptacles, so arranged that the batteries cannot be improperly connected, are employed in all current receiver designs.

In some receivers, a female receptacle is provided for the line plug so that the battery circuit remains open until this plug is inserted. For line operation, the plug must be removed and inserted in the power line receptacle. This complete protection against turning on the battery circuit when line operation is being employed is obtained. In other models, the same result is accomplished by the switch design.

The rated life of the batteries, in a representative type of portable, is 250 hours when the set is being used 4 hours a day.

REMOTE CONTROL

A variation of the wireless record player idea is employed in the RCA Model 5X5 receiver to enable remote control of another receiver. Instead of using record modulation, the broadcast program picked up by the 5X5 is used to modulate a self-contained oscillator which sends out the modulated signal to another receiver, which receiver is tuned to the frequency of the transmitting oscillator in the remote control unit. Means are provided so that the volume can also be controlled by the control unit.

RCA Model 5X5

A schematic diagram of this receiver is shown in Riders Manual, Volume X. The 12SA7 is employed as a conventional converter, the output of which is fed to the pentode section of the 12C8. The output of the pentode section is coupled by the second i-f transformer to the diode section of this tube, which also acts as the second detector and avc tube.

The 12SC7 serves as the first a-f amplifier and remote control oscillator. The a-f output of the 12SA7 diode appears across the volume control and 47,000-ohm series resistor and is fed to the first section of the 12SC7. The output of this section is resistance-capacity coupled to the 35L6GT output tube; thence to the speaker in the usual manner when the receiver is operated as a conventional set.

For remote control operation, the radio-remote switch in the output circuit of the 35L6GT is thrown to "Remote" position. When this is done, the primary of the output transformer is placed in series with the plate circuit of the remote control oscillator plate coil as well as the plate of the 35L6GT output tube. The 33,000-ohm series resistor and its associated 0.1-mf bypass condenser serve to drop the B voltage present at the output tube plate to a value suitable for the second section of the 12SC7. The 0.1-mf bypass condenser is used to allow the full audio voltage present at the output tube plate to be used to modulate the plate current of the remote control oscillator. In this way high percentage modulation can be obtained.

A plate-tuned oscillator circuit is employed in the remote control oscillator and, as has been shown, the primary of the 35L6GT output transformer serves as a modulation choke. The oscillator frequency is usually adjusted to 540 kc, but by readjusting C7 it may be tuned to any frequency between 540 and 800 kc.

The modulated output signal is fed to the power line by means of the coupling coil adjacent to the grid coil of the remote control oscillator. When remote control operation is employed, the voice coil circuit of the 5X5 speaker is opened and a 5-ohm resistor is automatically shunted across the output transformer secondary, thus providing the proper operating load. The speaker is then inoperative.

Since the degree of audio modulation applied to the remote control oscillator is dependent on the strength of the audio signal present across the primary of the output transformer which is in turn controllable by means of the receiver volume control, the output volume of receiver used to pick up the signal sent out by the remote control oscillator can also be controlled by the 5X5.

Phonograph records may be reproduced and transmitted in like manner by using a turntable and pickup connected to the phono jack provided in the receiver.