

Receiver Principles and Design

§ 7-1 Elements of Receiving Systems

Basic requirements — The purpose of a radio receiving system is to abstract energy from passing radio waves and convert it into a form which conveys the intelligence contained in the transmitted signal. The receiver also must be able to select a desired signal and eliminate those not wanted. The fundamental processes involved are those of amplification and detection.

Detection — The high frequencies used for radio signaling are well beyond the audio-frequency range (§ 2-7), and therefore cannot be used to actuate a loudspeaker directly. Neither can they be used to operate other devices, such as relays, by means of which a message might be transmitted. The process of converting a modulated radio-frequency wave to a usable low frequency, called *detection* or *demodulation*, is essentially that of rectification (§ 3-1). The modulated carrier (§ 5-1) is thereby converted to a unidirectional current, the amplitude of which will vary at the same rate as the modulation. These low-frequency variations are readily amplified, and can be applied to the headphones, loudspeaker or other form of electromechanical device.

Code signals — The dots and dashes of code (c.w.) transmissions are rectified as described, but in themselves can produce no audible tone in the headphones or loudspeaker because they are of constant amplitude. For aural reception it is necessary to introduce a second radio frequency, differing from the signal frequency by a suitable audio frequency, into the detector circuit to produce an audible beat (§ 2-13). The frequency difference, and hence the *beat note*, is generally of the order of 500 to 1000 cycles, since these tones are within the range of optimum response of both the ear and the headset. If the source of the second radio frequency is a separate oscillator, the system is known as *heterodyne* reception; if the detector itself is made to oscillate and produce the second frequency, it is known as an *autodyne* detector.

Amplification — To build up weak signals to usable output level, modern receivers employ considerable amplification — often of the order of hundreds of thousands of times. Amplifiers are used at the frequency of the incoming signal (*r.f. amplifiers*), after detection (*a.f. amplifiers*), and, in superheterodyne receivers, at one or more intermediate radio frequencies (*i.f. amplifiers*). R.f. and i.f. amplifiers practically always employ tuned circuits.

Types of receivers — Receivers may vary in complexity from a simple detector with no amplification to multi-tube arrangements having amplification at several different radio frequencies as well as at audio frequency. A regenerative detector (§ 7-4) with or without audio-frequency amplification (§ 7-5) is known as a *regenerative receiver*; if the detector is preceded by one or more tuned r.f. amplifier stages (§ 7-6), the combination is known as a *t.r.f. (tuned radio frequency) receiver*. The *superheterodyne receiver* (§ 7-8) employs r.f. amplification at a fixed intermediate frequency as well as at the frequency of the signal itself, the latter being converted by the heterodyne process to the intermediate frequency.

At very-high frequencies the superregenerative detector (§ 7-4), usually with audio amplification, is used in the *superregenerative receiver* or *superregenerator*, providing large amplification of weak signals with simple circuit arrangements.

§ 7-2 Receiver Characteristics

Sensitivity — Sensitivity is defined as the strength of the signal (usually expressed in microvolts) which must be applied to the input terminals of the receiver to produce a specified audio-frequency power output at the loudspeaker or headphones (§ 7-5). It is a measure of the amplification or gain of the receiver.

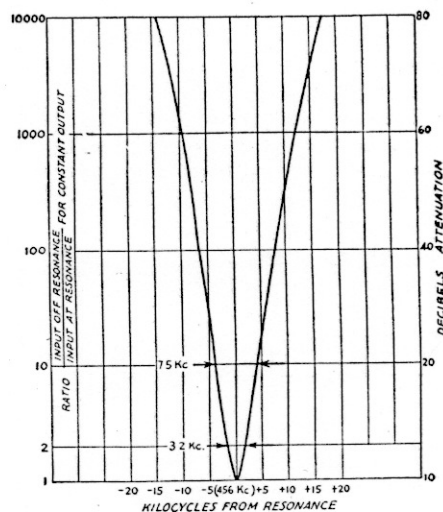


Fig. 701 — Selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios; the corresponding decibel steps are shown at the right.

Signal-to-noise ratio— Every receiver generates some noise of a hiss-like character, and signals weaker than the noise cannot be separated from it no matter how much amplification is used. This relation between noise and a weak signal is expressed by the term *signal-to-noise ratio*. It can be defined in various ways, one simple way being to give it as the ratio of signal power output to noise power output from the receiver at a specified value of modulated carrier voltage applied to the input terminals.

The hiss-like noise mentioned above is inherent in the circuits and tubes of the receiver, and its amplitude depends upon the selectivity of the receiver. The greater the selectivity the smaller the noise, other things being equal (§ 7-6). In addition to inherent receiver noise, atmospheric electricity (natural "static") and electrical devices in the vicinity of the receiver also cause noise which adversely affects the signal-to-noise ratio.

Selectivity— Selectivity is the ability of a receiver to discriminate against signals of frequencies differing from that of the desired signal. The over-all selectivity will depend upon the selectivity of the individual tuned circuits and the number of such circuits.

The selectivity of a receiver is shown graphically by drawing a curve which gives the ratio of signal strength required at various frequencies off resonance to the signal strength at resonance, to give constant output. A *resonance curve* of this type (taken on a typical communications-type superheterodyne receiver) is shown in Fig. 701. The *band-width* is the width of the resonance curve (in cycles or kilocycles) of a receiver at a specified ratio; in Fig. 701, the band-widths are indicated for ratios of response of 2 and 10 ("2 times down" and "10 times down").

Selectivity for signals within a few kilocycles of the desired-signal frequency is called *adjacent-channel selectivity*, to distinguish it from the discrimination against signals considerably removed from the desired frequency.

Stability— The stability of a receiver is its ability to give constant output, over a period of time, from a signal of constant strength and frequency. Primarily, it means the ability to stay tuned to a given signal. However, a receiver which at some settings of its controls has a tendency to break into oscillation, or "howl," also is said to be unstable.

The stability of a receiver is affected principally by temperature variations, supply-voltage changes, and constructional features of a mechanical nature.

Fidelity— Fidelity is the relative ability of the receiver to reproduce in its output the modulation (keying, 'phone, etc.) carried by the incoming signal. For exact reproduction the band-width must be great enough to accommodate the highest modulation frequency transmitted, and the relative amplitudes of the various frequency components within the band must not be changed in the output.

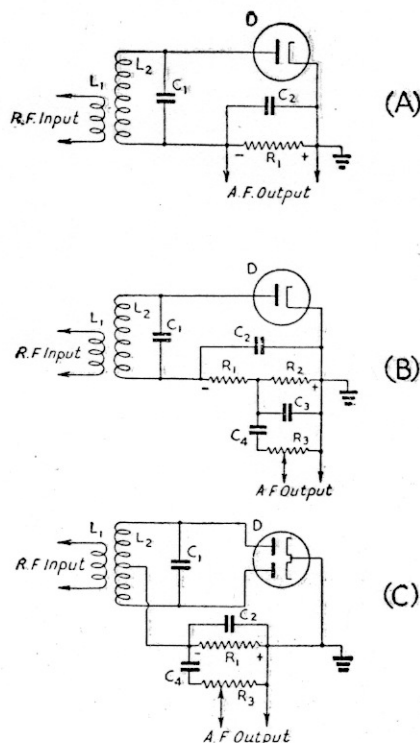


Fig. 702—Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L_2C_1 , is tuned to the signal frequency; typical values for C_2 and R_1 in A and B are 250 $\mu\text{fd.}$ and 250,000 ohms, respectively; in B, C_2 and C_3 are 100 $\mu\text{fd.}$ each; R_1 , 50,000 ohms; and R_2 , 250,000 ohms. C_4 is 0.1 $\mu\text{fd.}$ and R_3 may be 0.5 to 1 megohm.

7-3 Detectors

Characteristics— The important characteristics of a detector are its sensitivity, fidelity or linearity, resistance or impedance, and signal-handling capability.

Detector *sensitivity* is the ratio of audio-frequency output to radio-frequency input. *Linearity* is a measure of the ability of the detector to reproduce, as an audio frequency, the exact form of the modulation on the incoming signal. The *resistance* or *impedance* of the detector is important in circuit design, since a relatively low resistance means that power is consumed in the detector. The *signal-handling capability* means the ability of the detector to accept signals of a specified amplitude without overloading.

Diode detectors— The simplest detector is the diode rectifier. Circuits for both half-wave and full-wave (§ 8-3) diodes are given in Fig. 702. The simplified half-wave circuit at 702-A includes the r.f. tuned circuit, L_2C_1 , a coupling coil, L_1 , from which the r.f. energy is fed to L_2C_1 , and the diode, D , with its load resistance, R_1 , and by-pass condenser, C_2 . The flow of rectified r.f. current through R_1 causes a d.c. voltage to develop across its terminals, and this voltage varies with the modulation on the

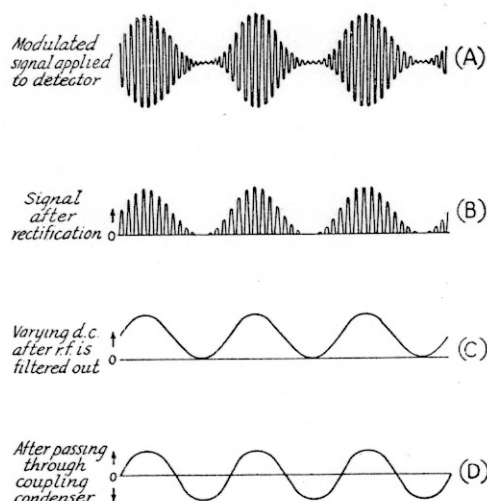


Fig. 703 — Diagrams showing the detection process.

signal. The — and + signs show the polarity of the voltage. The variation in amplitude of the r.f. signal with modulation causes corresponding variations in the value of the d.c. voltage across R_1 . The load resistor, R_1 , usually has a rather high value of resistance, so that a fairly large voltage will develop from a small rectified-current flow.

The progress of the signal through the detector or rectifier is shown in Fig. 703. A typical modulated signal as it exists in the tuned circuit is shown at A. When applied to the rectifier tube, current flows from plate to cathode only during the part of the r.f. cycle when the plate is positive with respect to the cathode, so that the output of the rectifier consists of half-cycles of r.f. still modulated as in the original signal. These current "pulses" flow in the load circuit comprised of R_1 and C_2 , the resistance of R_1 and the capacity of C_2 being so proportioned that C_2 charges to the peak value of the rectified voltage on each pulse and retains enough charge between pulses so that the voltage across R_1 is smoothed out, as shown in C. C_2 thus acts as a filter for the radio-frequency component of the output of the rectifier, leaving a d.c. component which varies in the same way as the modulation on the original signal. When this varying d.c. voltage is applied to a following amplifier through a coupling condenser (C_4 in Fig. 702-B), only the variations in voltage are transferred, so that the final output signal is a.c., as shown in D.

In the circuit at 702-B, R_1 and C_2 have been divided for the purpose of providing a more effective filter for r.f. It is important to prevent the appearance of any r.f. voltage in the output of the detector, because it may cause overloading of a succeeding amplifier tube. The audio-frequency variations can be transferred to another circuit through a coupling condenser, C_4 in Fig. 702, to a load resistor, R_3 , which usually is a "potentiometer" (§ 8-10) so that the volume can be adjusted to a desired level.

The full-wave diode circuit at 702-C is practically identical in operation to the half-wave circuit, except that both halves of the r.f. cycle are utilized. The full-wave circuit has the advantage that very little r.f. voltage appears across the load resistor, R_1 , because the midpoint of L_2 is at the same potential as the cathode, or "ground" for r.f.

The reactance of C_2 must be small compared to the resistance of R_1 at the radio frequency being rectified, but at audio frequencies must be relatively large compared to R_1 (§ 2-8, 2-13). This condition is satisfied by the values shown. If the capacity of C_2 is too large, response at the higher audio frequencies will be lowered.

Compared with other detectors, the sensitivity of the diode is low. Since the diode consumes power, the Q of the tuned circuit is reduced, bringing about a reduction in selectivity (§ 2-10). The linearity is good, however, and the signal-handling capability is high.

Grid-leak detectors — The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuit of Fig. 704-A, the grid corresponds to the diode

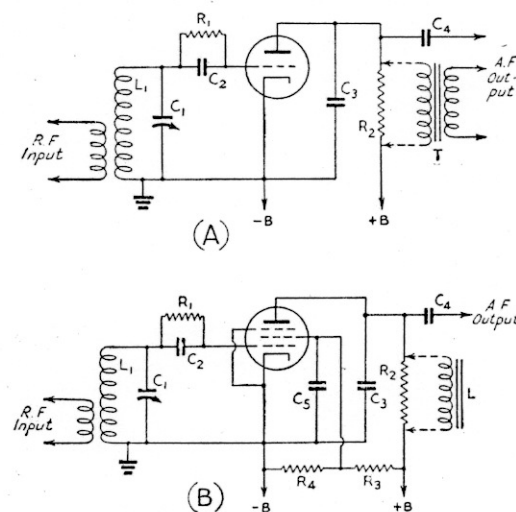


Fig. 704 — Grid-leak detector circuits, A, triode; B, pentode. A tetrode may be used in the circuit of B by neglecting the suppressor-grid connection. Transformer coupling may be substituted for resistance coupling in A, or a high-inductance choke may replace the plate resistor in B. L_1C_1 is a circuit tuned to the signal frequency. The grid leak, R_1 , may be connected directly from grid to cathode instead of across the grid condenser as shown. The operation with either connection will be the same. Representative values for components are:

Component	Circuit A	Circuit B
C_2	100 to 250 $\mu\text{fd.}$	100 to 250 $\mu\text{fd.}$
C_3	0.001 to 0.002 $\mu\text{fd.}$	250 to 500 $\mu\text{fd.}$
C_4	0.1 $\mu\text{fd.}$	0.1 $\mu\text{fd.}$
C_5		0.5 $\mu\text{fd.}$ or larger.
R_1	1 to 2 megohms.	1 to 5 megohms.
R_2	50,000 ohms.	100,000 to 250,000 ohms.
R_3		50,000 ohms.
R_4		20,000 ohms.
T	Audio transformer.	
L		500-henry choke.

The plate voltage in A should be about 50 volts for best sensitivity. In B, the screen voltage should be about 30 volts and the plate voltage from 100 to 250.

plate and the rectifying action is exactly the same as just described. The d.c. voltage from rectified-current flow through the grid leak, R_1 , biases the grid negatively with respect to cathode, and the audio-frequency variations in voltage across R_1 are amplified through the tube just as in a normal a.f. amplifier. In the plate circuit, R_2 is the plate load resistance (§ 3-3) and C_3 is a by-pass condenser to eliminate r.f. in the output circuit. C_4 is the output coupling condenser. With a triode, the load resistor, R_2 , may be replaced by an audio transformer, T , in which case C_4 is not used.

Since audio amplification is added to rectification, the grid-leak detector has considerably greater sensitivity than the diode. The sensitivity can be further increased by using a screen-grid tube instead of a triode, as at 704-B. The operation is equivalent to that of the triode circuit. The screen by-pass condenser, C_5 , should have low reactance (§ 2-8, 2-13) for both radio and audio frequencies. R_3 and R_4 constitute a voltage divider (§ 8-10) from the plate supply to furnish the proper d.c. voltage to the screen. In both circuits, C_2 must have low r.f. reactance and high a.f. reactance compared to the resistance of R_1 ; the same applies to C_3 with respect to R_2 .

Because of the high plate resistance of the screen-grid tube (§ 3-5), transformer coupling from the plate circuit of a screen-grid detector is not satisfactory. An impedance (L in Fig. 704-B) can be used in place of a resistor, with a gain in sensitivity because a high value of load impedance can be developed with little loss of plate voltage as compared to the voltage drop through a resistor. The coupling coil, L_2 , for a screen-grid detector should have an inductance of the order of 300 to 500 henrys.

The sensitivity of the grid-leak detector is higher than that of any other type. Like the diode, it "loads" the tuned circuit and reduces its selectivity. The linearity is rather poor, and the signal-handling capability is limited.

Plate detectors—The plate detector is arranged so that rectification of the r.f. signal takes place in the plate circuit of the tube, as contrasted to the grid rectification just described. Sufficient negative bias is applied to the grid to bring the plate current nearly to the cut-off point, so that the application of a signal to the grid circuit causes an increase in average plate current. The average plate current follows the changes in signal amplitude in a fashion similar to the rectified current in a diode detector.

Circuits for triodes and pentodes are given in Fig. 705. C_3 is the plate by-pass condenser, R_1 is the cathode resistor which provides the operating grid bias (§ 3-6), and C_2 is a by-pass for both radio and audio frequencies across R_1 (§ 2-13). R_2 is the plate load resistance (§ 3-3), across which a voltage appears as a result of the rectifying action described above. C_4 is the output coupling condenser. In the pentode circuit at B, R_3 and R_4 form a voltage divider

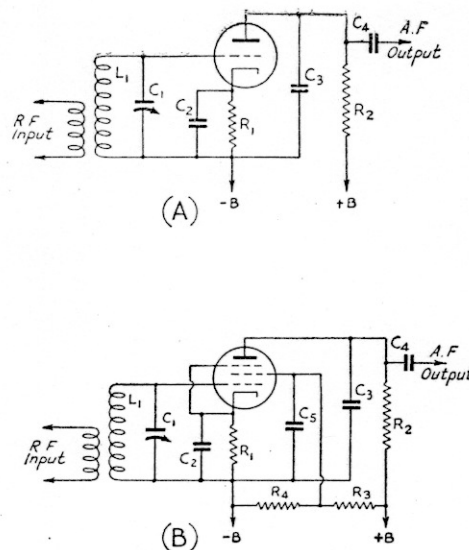


Fig. 705 — Circuits for plate detection. A, triode; B, pentode. The input circuit, $L_1 C_1$, is tuned to the signal frequency. Typical values for the other constants are:

Component	Circuit A	Circuit B
C_2	0.5 μ fd. or larger.	0.5 μ fd. or larger.
C_3	0.001 to 0.002 μ fd.	250 to 500 μ fd.
C_4	0.1 μ fd.	0.1 μ fd.
C_5		0.5 μ fd. or larger.
R_1	25,000 to 150,000 ohms.	10,000 to 20,000 ohms.
R_2	50,000 to 100,000 ohms.	100,000 to 250,000 ohms.
R_3		50,000 ohms.
R_4		20,000 ohms.

Plate voltages from 100 to 250 volts may be used. Effective screen voltage in B should be about 30 volts.

to supply the proper potential (about 30 volts) to the screen, and C_5 is a by-pass condenser between screen and cathode. C_5 must have low reactance for both radio and audio frequencies.

In general, transformer coupling from the plate circuit of a plate detector is not satisfactory, because the plate impedance even of a triode is very high when the bias is set near the plate-current cut-off point (§ 3-2, 3-3). Impedance coupling may be used in place of the resistance coupling shown in Fig. 705. The same order of inductance is required as with the screen-grid detector described previously.

The plate detector is more sensitive than the diode since there is some amplifying action in the tube, but less so than the grid-leak detector. It will handle considerably larger signals than the grid-leak detector, but is not quite so tolerant in this respect as the diode. Linearity, with the self-biased circuits shown, is good. Up to the overload point the detector takes no power from the tuned circuit, and so does not affect its Q and selectivity (§ 2-10).

Infinite-impedance detector—The circuit of Fig. 706 combines the high signal-handling capabilities of the diode detector with low distortion (good linearity), and, like the plate detector, does not load the tuned circuit to which it is connected. The circuit resembles that of the plate detector, except that the load

resistance, R_1 , is connected between cathode and ground and thus is common to both grid and plate circuits, giving negative feed-back for the audio frequencies. The cathode resistor is by-passed for r.f. (C_1) but not for audio (§ 2-13), while the plate circuit is by-passed

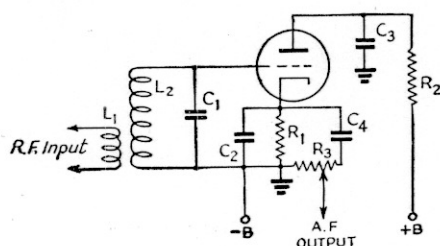


Fig. 706 — The infinite-impedance or linear detector. The input circuit, L_2C_1 , is tuned to the signal frequency. Typical values for the other constants are:

C_2 — 250 μ fd. R_1 — 0.15 megohm.
 C_3 — 0.5 μ fd. R_2 — 25,000 ohms.
 C_4 — 0.1 μ fd. R_3 — 0.25-megohm volume control.

A tube having a medium amplification factor (about 20) should be used. Plate voltage should be 250 volts.

to ground for both audio and radio frequencies. R_2 forms, with C_2 , an RC filter (§ 2-11) to isolate the plate from the "B" supply at a.f.

The plate current is very low at no signal, increasing with signal as in the case of the plate detector. The voltage drop across R_1 similarly increases with signal, because of the increased plate current. Because of this and the fact that the initial drop across R_1 is large, the grid cannot be driven positive with respect to the cathode by the signal, hence no grid current can be drawn.

§ 7-4 Regenerative Detectors

Circuits — By providing controllable r.f. feed-back or regeneration (§ 3-3) in a triode or pentode detector circuit, the incoming signal can be amplified many times, thereby greatly increasing the sensitivity of the detector. Regeneration also increases the effective Q of the circuit, and hence increases the selectivity (§ 2-10) by virtue of the fact that the maximum regenerative amplification takes place only at the frequency to which the circuit is tuned. The grid-leak type of detector is most suitable for the purpose. Except for the regenerative connection, the circuit values are identical with those previously described for this type of detector, and the same considerations apply. The amount of regeneration must be controllable, because maximum regenerative amplification is secured at the critical point where the circuit is just about to oscillate (§ 3-7) and the critical point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned.

Fig. 707 shows the circuits of regenerative detectors of various types. The circuit of A is for a triode tube, with a variable by-pass condenser, C_3 , in the plate circuit to control regeneration. When the capacity is small the

tube does not regenerate, but as it increases toward maximum its reactance (§ 2-8) becomes smaller until a critical value is reached where there is sufficient feed-back to cause oscillation. If L_2 and L_3 are wound end-to-end in the same direction, the plate connection is to the outside of the plate or "tickler" coil, L_3 , when the grid connection is to the outside of L_2 .

The circuit of B is for a screen-grid tube, regeneration being controlled by adjustment of the screen-grid voltage. The tickler, L_3 , is in the plate circuit. The portion of the control resistor between the rotating contact and ground is by-passed by a large condenser (0.5 μ fd. or more) to filter out scratching noise when the arm is rotated (§ 2-11). The feed-back is adjusted by varying the number of turns on L_3 or the coupling (§ 2-11) between L_2 and L_3 , until the tube just goes into oscillation at a screen voltage of approximately 30 volts.

Circuit C is identical with B in principle of operation, except that the oscillating circuit is of the Hartley type (§ 3-7). Since the screen and plate are in parallel for r.f. in this circuit, only a small amount of "tickler" — that is, relatively few turns between the cathode tap and ground — is required for oscillation.

Adjustment for smooth regeneration — The ideal regeneration control would permit the detector to go into and out of oscillation smoothly, would have no effect on the frequency of oscillation, and would give the same value of regeneration regardless of frequency and the loading on the circuit. In practice, the effects of loading, particularly the loading that occurs when the detector circuit is coupled to an antenna, are difficult to overcome. Likewise, the regeneration is affected by the frequency to which the grid circuit is tuned.

In all circuits it is best to wind the tickler at the ground or cathode end of the grid coil, and to use as few turns on the tickler as will allow the detector to oscillate easily over the whole tuning range at the plate (and screen, if a pentode) voltage which gives maximum sensitivity. Should the tube break into oscillation suddenly as the regeneration control is advanced, making a click, the operation often can be made smoother by changing the grid-leak resistance to a higher or lower value. The wrong grid leak plus too-high plate and screen voltage are the most frequent causes of lack of smoothness in going into oscillation.

Antenna coupling — If the detector is coupled to an antenna, slight changes in the antenna constants (as when the wire swings in a breeze) affect the frequency of the oscillations generated, and thereby the beat frequency when c.w. signals are being received. The tighter the antenna coupling is made, the greater will be the feed-back required or the higher will be the voltage necessary to make the detector oscillate. The antenna coupling should be the maximum that will allow the detector to go into oscillation smoothly with the correct voltages on the tube. If capacity

coupling (§ 2-11) to the grid end of the coil is used, only a very small amount of capacity will be needed to couple to the antenna. Increasing the capacity increases the coupling.

At frequencies where the antenna system is resonant the absorption of energy from the oscillating detector circuit will be greater, with the consequence that more regeneration is needed. In extreme cases it may not be possible to make the detector oscillate with normal voltages, causing so-called "dead spots." The remedy for this is to loosen the antenna coupling to the point which permits normal oscillation and smooth regeneration control.

Body capacity—A regenerative detector occasionally shows a tendency to change frequency slightly as the hand is moved near the dial. This condition (*body capacity*) can be caused by poor design of the receiver, or by the antenna if the detector is coupled directly to it. If body capacity is present when the antenna is disconnected, it can be eliminated by better shielding, and sometimes by r.f. filtering of the 'phone leads. Body capacity which is present only when the antenna is connected is caused by resonance effects in the antenna, which tend to cause part of a standing wave (§ 2-12) of r.f. voltage to appear on the ground lead and thus raise the whole detector circuit above ground potential. A good, short ground connection should be made to the receiver and the length of the antenna varied electrically (by adding a small coil or variable condenser in the antenna lead) until the effect is minimized. Loosening the coupling to the antenna circuit also will help.

Hum—Hum at the power-supply frequency may be present in a regenerative detector, especially when it is used in an oscillating condition for c.w. reception, even though the plate supply itself is free from ripple (§ 8-4). The hum may result from the use of a.c. on the tube heater, but effects of this type normally are troublesome only when the circuit of Fig. 707-C is used, and then only at 14 Mc. and higher frequencies. Connecting one side of the heater supply to ground, or grounding the center-tap of the heater transformer winding, is good practice to reduce hum, and the heater wiring should be kept as far as possible from the r.f. circuits.

House wiring, if of the "open" type, will have a rather extensive electrostatic field which may cause hum if the detector tube, grid lead, and grid condenser and leak are not electrostatically shielded. This type of hum is easily recognizable because of its rather high pitch, a result of harmonics (§ 2-7) in the power-supply system. The hum is caused by a species of grid modulation (§ 5-4).

Antenna resonance effects frequently cause a hum of the same nature as that just described which is most intense at the various resonance points, and hence varies with tuning. For this reason it is called tunable hum. It is prone to occur with a rectified a.c. plate supply

(§ 8-1) when a standing wave effect of the type described in the preceding paragraph occurs, and is associated with the non-linear-ity of the rectifier tube in the plate supply.

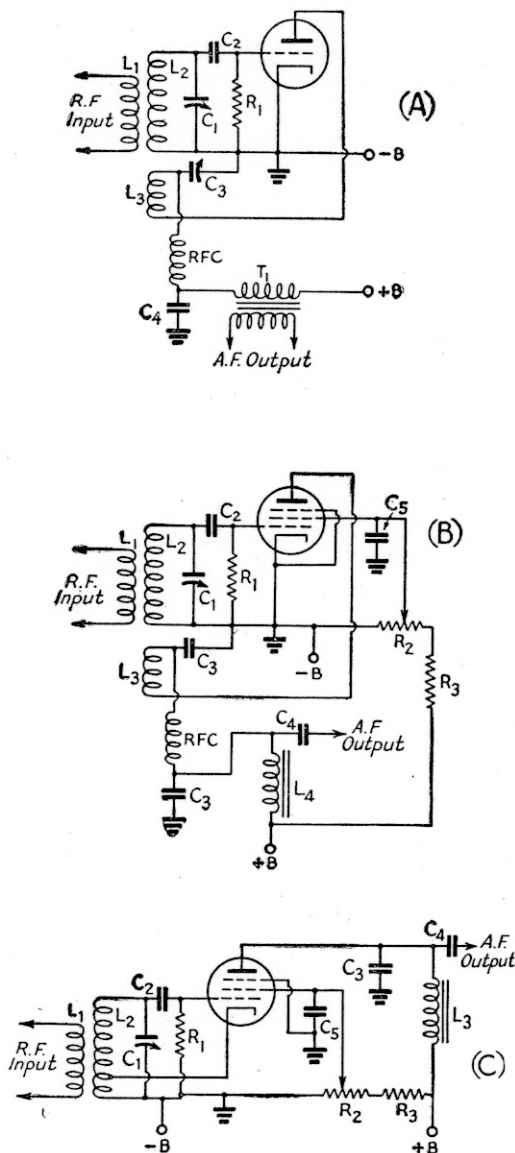


Fig. 707—Triode and pentode regenerative detector circuits. The input circuit, L_2C_1 , is tuned to the signal frequency. The grid condenser, C_2 , should have a value of about $100 \mu\text{fd.}$ in all circuits; the grid leak, R_1 , may range in value from 1 to 5 megohms. The tickler coil, L_3 , ordinarily will have from 10 to 25 per cent of the number of turns on L_2 ; in C, the cathode tap is about 10 per cent of the number of turns on L_2 above ground. Regeneration control condenser C_3 in A should have a maximum capacity of $100 \mu\text{fd.}$ or more; by-pass condensers C_3 in B and C are likewise $100 \mu\text{fd.}$ C_5 is ordinarily $1 \mu\text{fd.}$ or more; R_2 , a 50,000-ohm potentiometer; R_3 , 50,000 to 100,000 ohms. L_4 in B (L_3 in C) is a 500-henry inductance, C_4 is $0.1 \mu\text{fd.}$ in both circuits. T_1 in A is a conventional audio transformer for coupling from the plate of a tube to a following grid. RFC is 2.5 mh. In A, the plate voltage should be about 50 volts for best sensitivity. Pentode circuits require about 30 volts on the screen; plate voltage may be 100 to 250 volts.

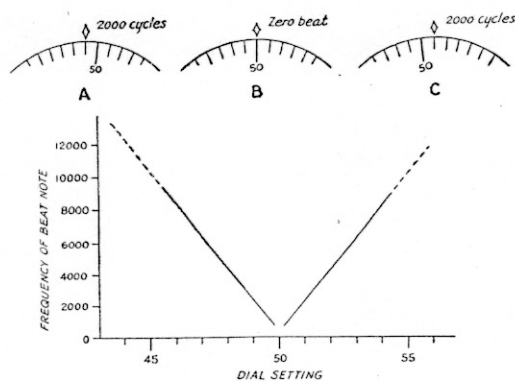


Fig. 708 — As the tuning dial of a receiver is turned past a c.w. signal, the beat note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000 or 10,000 cycles but usually is not heard because of the limitations of the audio system.

Elimination of antenna resonance effects as described and by-passing the rectifier plates to cathode (using by-pass condensers of the order of 0.001 μ fd.) usually will cure it.

Tuning — For c.w. reception, the regeneration control is advanced until the detector breaks into a "hiss," which indicates that the detector is oscillating. Further advancing the regeneration control after the detector starts oscillating will result in a slight decrease in the strength of the hiss, indicating that the sensitivity is decreasing.

The proper adjustment of the regeneration control for best reception of c.w. signals is where the detector just starts to oscillate, when it will be found that c.w. signals can be tuned in and will give a tone with each signal depending on the setting of the tuning control. As the receiver is tuned through a signal the tone first will be heard as a very high pitch, then will go down through "zero beat" (the region where the frequencies of the incoming signal and the oscillating detector are so nearly alike that the difference or beat is less than the lowest audible tone) and rise again on the other side, finally disappearing at a very high pitch. This behavior is shown in Fig. 708. It will be found that a low-pitched beat-note cannot be obtained from a strong signal because the detector "pulls in" or "blocks"; that is, the signal tends to control the detector in such a way that the latter oscillates at the signal frequency, despite the fact that the circuit may not be tuned exactly to resonance. This phenomenon, commonly observed when an oscillator is coupled to a source of a.c. voltage of approximately the frequency at which the oscillator is operating, is called "locking-in"; the more stable of the two frequencies assumes control over the other. "Blocking" usually can be corrected by advancing the regeneration control until the beat-note occurs again. If the regenerative detector is preceded by an r.f. amplifier stage, the blocking can be eliminated by reducing the

gain of the r.f. stage. If the detector is coupled to an antenna, the blocking condition can be eliminated by advancing the regeneration control or loosening the antenna coupling.

The point just after the receiver starts oscillating is the most sensitive condition for c.w. reception. Further advancing the regeneration control makes the receiver less prone to blocking by strong signals, but also less capable of receiving weak signals.

If the receiver is in the oscillating condition and a 'phone signal is tuned in, a steady audible beat-note will result. While it is possible to listen to 'phone if the receiver can be tuned to exact zero beat, it is more satisfactory to reduce the regeneration to the point just before the receiver goes into oscillation. This is also the most sensitive operating point.

Superregeneration — The limit to which ordinary regenerative amplification can be carried is the point at which oscillations commence, since at that point further amplification ceases. The *superregenerative* detector overcomes this limitation by introducing into the detector circuit an alternating voltage of a frequency somewhat above the audible range (of the order of 20 to 100 kilocycles), in such a way as to vary the detector's operating point (§ 3-3). As a consequence of the introduction of this *quench* or *interruption* frequency, the detector can oscillate only when the varying operating point is in a region suitable for the production of oscillations. Because the oscillations are constantly being interrupted, the regeneration can be greatly increased, and the signal will build up to relatively tremendous proportions. The superregenerative circuit is

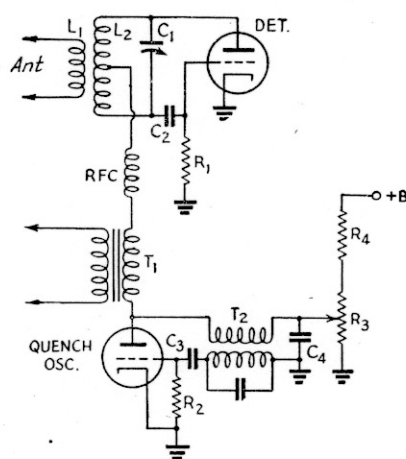


Fig. 709 — Superregenerative detector circuit using a separate quench oscillator. L_2C_1 is tuned to the signal frequency. Typical values for other components are:
 C_2 — 100 μ fd. T_1 — Audio transformer, plate-to-grid type.
 C_3 — 500 μ fd. RFC — R.f. choke, value depending upon the frequency in use. Special low-capacity chokes are required for the very-high frequencies.
 C_4 — 0.1 μ fd.
 R_1 — 5 megohms.
 R_2 — 50,000 ohms.
 R_3 — 50,000-ohm potentiometer.
 R_4 — 50,000 ohms.

suitable only for the reception of modulated signals, and operates best on the very-high frequencies where it has found considerable application in simple receivers.

A typical superregenerative circuit for very-high frequencies is shown in Fig. 709. The regenerative detector circuit is an ultraudion oscillator (§ 3-7). The quench frequency, obtained from a separate quench oscillator, is introduced in the plate circuit. Many other circuit arrangements are possible.

If regeneration in an ordinary regenerative circuit is carried sufficiently far, the circuit will break into a low-frequency oscillation simultaneously with that at the operating radio frequency. This low-frequency oscillation has much the same quenching effect as that from a separate oscillator, hence a circuit so operated is called a *self-quenching* superregenerative detector. This type of circuit is more successful at very-high than at ordinary communication frequencies. The frequency of the quench oscillation depends upon the feed-back and upon the time constant of the grid leak and condenser, the oscillation being a form of "blocking" or "squegging" in which the grid accumulates a strong negative charge which does leak off rapidly enough through the grid leak to prevent a relatively slow variation of the operating point.

The superregenerative detector has relatively little selectivity as compared to a regular regenerative detector, but discriminates against noise such as that from automobile ignition systems. It also has marked automatic volume control action, since strong signals are amplified much less than weak signals.

Adjustment of superregenerative detectors — Because of the greater amplification, the hiss when the superregenerative detector goes into oscillation is much stronger than with the ordinary regenerative detector. The most sensitive condition is at the point where the hiss first becomes marked. When a signal is tuned in, the hiss will disappear to a degree which depends upon the signal strength.

Lack of hiss indicates insufficient feed-back at the signal frequency, or inadequate quench voltage. Antenna loading effects will cause dead spots similar to those with regenerative detectors, and these can be overcome by the same methods. The self-quenching detector may require critical adjustment of the grid leak and grid condenser values for smooth operation, since these determine the frequency and amplitude of the quench voltage.

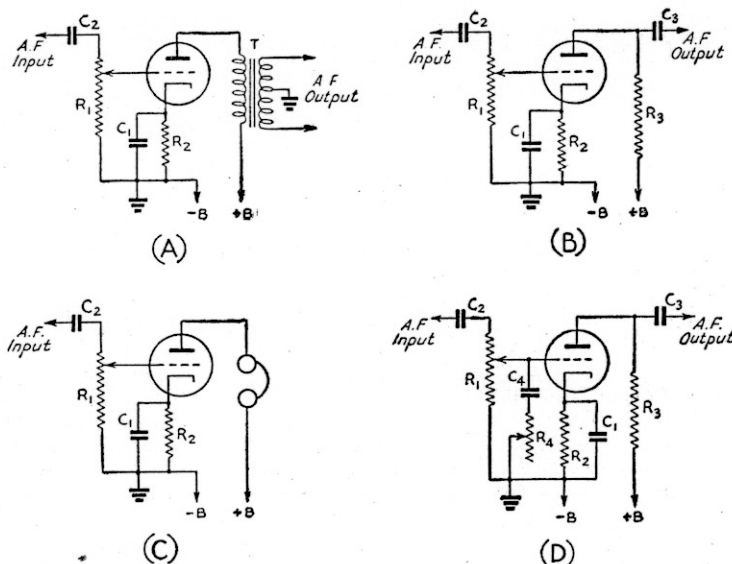


Fig. 710 — Audio amplifier circuits used for voltage amplification and to provide power for headphone output. The tubes are operated as Class-A amplifiers (§ 3-4).

7-5 Audio-Frequency Amplifiers

General — The ordinary detector does not produce very much audio-frequency power output — usually not enough to give satisfactory sound volume, even in headphone reception. Consequently, audio-frequency amplifiers are used after the detector to increase the power level. One amplifier usually is sufficient for headphones, but two stages generally are used where the receiver is to operate a loudspeaker. A few milliwatts of a.f. power is sufficient for headphones, but a loudspeaker requires a watt or more for good room volume.

In all except battery-operated receivers, the negative grid bias of audio amplifiers usually is secured from the voltage drop in a cathode resistor (§ 3-6). The cathode resistor must be bypassed by a condenser having low reactance at the lowest audio frequency to be amplified, compared to the resistance of the cathode resistor (10 per cent or less) (§ 2-8, 2-13). In battery-operated receivers, a separate grid-bias battery generally is used.

Headset and voltage amplifiers — The circuits shown in Fig. 710 are typical of those used for voltage amplification and for providing sufficient power for operation of headphones (§ 3-3). Triodes usually are preferred to pentodes because they are better suited to working into an audio transformer or headset, the input impedances of which are of the order of 20,000 ohms.

In these circuits, R_2 is the cathode bias resistor and C_1 the cathode by-pass condenser. The grid resistor, R_1 , gives volume control action (§ 5-9). Its value ordinarily is from 0.25 to 1 megohm. C_2 is the input coupling condenser, already discussed under detectors; it is, in fact, identical to C_4 in Figs. 704 and 705, if the amplifier is coupled to a detector.

Power amplifiers—A popular type of power amplifier is the single pentode, operated Class A or AB; the circuit diagram is given in Fig. 711-A. The grid resistor, R_1 , may be a potentiometer for volume control, as shown at R_1 in Fig. 710. The output transformer, T , should have a turns ratio (§ 2-9) suitable for the loudspeaker used; many of the small loudspeakers now available are furnished complete with output transformer.

When greater volume is needed, a pair of pentodes or tetrodes may be connected in push-pull (§ 3-3), as shown in Fig. 711-B. Transformer coupling to the voltage-amplifier stage is the simplest method of obtaining push-pull input for the amplifier grids. The inter-stage transformer, T_1 , has a center-tapped secondary with a secondary-to-primary turns ratio of about 2 to 1. An output transformer, T_2 , with a center-tapped primary must be used. No by-pass condenser is needed across the cathode resistor, R , since the a.f. current does not flow through the resistor as it does in single-tube circuits (§ 3-3).

Tone control—A tone control is a device for changing the frequency response (§ 3-3) of an audio amplifier; usually it is simply a method for reducing high-frequency response. This is helpful in reducing hissing and crackling noises without disturbing the intelligibility of the signal. R_4 and C_4 in Fig. 710-D, together form an effective tone control of this type. The maximum effect is secured when the resistance of R_4 is entirely out of the circuit, leaving C_4 connected directly between grid and ground. R_4 should be large compared to the reactance of C_4 (§ 2-8) so that when its resistance is all in circuit the effect of C_4 on the frequency response is negligible.

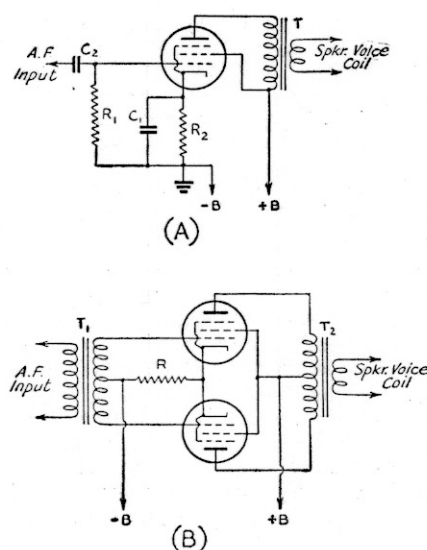


Fig. 711 — Power-output audio amplifier circuits. Either Class A or AB amplification (§ 3-4) may be used.

Headphones and loudspeakers—Two types of headphones are in general use, the *magnetic* and *crystal* types. They are shown in cross-section in Fig. 712. In the magnetic type the signal is applied to a coil or pair of coils having a great many turns of fine wire wound on a permanent magnet. (Headphones having one coil are known as the "single-pole" type, while those having two coils, as shown in Fig. 712, are called "double-pole.") A thin circular diaphragm of iron is placed close to the open ends of the magnet. It is tightly clamped by the earpiece assembly around its circumference, and the center is drawn toward the permanent magnet under some tension. When an alternating current flows through the windings the field set up by the current alternately aids and opposes the steady field of the permanent magnet, so that the diaphragm alternately is drawn nearer to and allowed to spring farther away from the magnet. Its motion sets the air into corresponding vibration. Although the d.c. resistance of the coils may be of the order of 2000 ohms, the a.c. impedance of a magnetic type headset will be of the order of 20,000 ohms at 1000 cycles.

In the crystal headphone, two piezoelectric crystals (§ 2-10) of Rochelle salts are cemented together in such a way that the pair tends to be bent in one direction when a voltage of a certain polarity is applied and to bend in the other direction when the polarity is reversed. The crystal unit is rigidly mounted to the earpiece, with the free end coupled to a diaphragm. When an alternating voltage is applied, the alternate bending as the polarity of the applied voltage reverses makes the diaphragm vibrate back and forth. The impedance is several times that of the magnetic type.

Magnetic-type headsets tend to give maximum response at frequencies of the order of 500 to 1000 cycles, with a considerable reduction of response (for constant applied voltage) at frequencies both above and below this region. The crystal type has a "flatter" frequency-response curve, and is particularly good at reproducing the higher audio frequencies. The peaked response curve of the magnetic type is advantageous in code reception, since it tends to reduce interference from signals having beat tones lying outside the region of maximum response, while the crystal type is better for the reception of voice and music. Magnetic headsets can be used in circuits in which d.c. is flowing, such as the plate circuit of a vacuum tube, providing the current is not too large to be carried safely by the wire in the coils; the limit is a few milliamperes. Crystal headsets must be used only on a.c. (since a steady d.c. voltage will damage the crystal unit), and consequently must be coupled to the tube through a device, such as a condenser, which isolates the d.c. voltage but permits the passage of an alternating current.

The most common type of loudspeaker is the *dynamic* type, shown in cross-section in Fig.

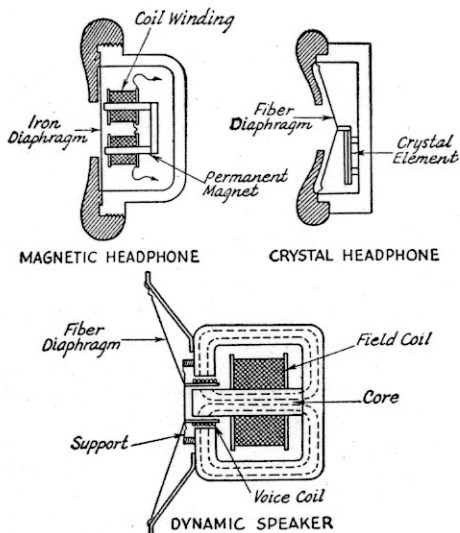


Fig. 712 — Headphone and loudspeaker construction.

712. The signal is applied to a small coil (the *voice coil*) which is free to move in the gap between the ends of a magnet. The magnet is made in the form of a cylindrical coil slightly smaller than the form on which the voice coil is wound, with the magnetic circuit completed through a pole piece which fits around the outside of the voice coil leaving just enough clearance for free movement of the coil. The path of the flux through the magnet is as shown by the dotted lines in the figure. The voice coil is supported so that it is free to move along its axis but not in other directions, and is fastened to a fiber or paper conical diaphragm. When current is sent through the coil it moves in a direction determined by the polarity of the current (§ 2-5), and thus moves back and forth when an alternating voltage is applied. The motion is transmitted by the diaphragm to the air, setting up sound waves.

The type of speaker shown in Fig. 712 obtains its fixed magnetic field by electromagnetic means, direct current being sent through the *field coil* for this purpose. Other types use permanent magnets to replace the electromagnet, and hence do not require a source of d.c. power. The voice coils of dynamic speakers have few turns and therefore low impedance, values of 3 to 15 ohms being representative.

7-6 Radio-Frequency Amplifiers

Circuits— Although there may be variations in detail, practically all r.f. amplifiers conform to the basic circuit shown in Fig. 713. A screen-grid tube, usually a pentode, is used, since a triode will oscillate when its grid and plate circuits are tuned to the same frequency (§ 3-5). The amplifier operates Class A, without grid current (§ 3-4). The tuned grid circuit, L_1C_1 , is coupled through L_2 to the antenna (or, in some cases, to a preceding stage). R_1 and C_2 are the

cathode bias resistor and by-pass condenser, C_3 is the screen by-pass condenser, and R_2 is the screen dropping resistor. L_3 is the primary of the output transformer (§ 2-11), tightly coupled to L_4 , which, with C_5 , constitutes the tuned circuit feeding the detector or following amplifier. The input and output circuits, L_1C_1 and L_4C_5 , are both tuned to the signal frequency.

Shielding— The screen-grid construction of the amplifier tube prevents feed-back (§ 3-3) from plate to grid inside the tube, but in addition it is necessary to prevent transfer of energy from the plate circuit to the grid circuit external to the tube. This is accomplished by enclosing the coils in grounded shielding containers and by keeping the plate and grid leads well separated. With "single-ended" tubes, care in laying out the wiring to obtain the maximum possible physical separation between plate and grid leads is necessary to prevent capacity coupling.

The shield around a coil will reduce the inductance and Q of the coil (§ 2-11) to an extent which depends upon the shielding material and its distance from the coil (see page 420). Adjustments therefore must be made with the shield in place.

By-passing— In addition to shielding, good by-passing (§ 2-13) is imperative. This is not simply a matter of choosing the proper type and capacity of by-pass condenser. Short *separate* leads from C_3 and C_4 to cathode or ground are a prime necessity. At the higher radio frequencies even an inch of wire will have enough inductance to provide feed-back coupling, and hence cause oscillation, if the wire happens to be common to both the plate and grid circuits.

Gain control— The gain of an r.f. amplifier usually is varied by varying the grid bias. This method works best with variable- μ type tubes (§ 3-5), hence this type usually is found in r.f. amplifiers. In Fig. 713, R_3 and R_4 comprise the gain-control circuit. R_3 is the control resistor (§ 3-6) and R_4 a dropping resistor of such value as to make the voltage across the outside terminals of R_3 about 50 volts (§ 8-10). The gain is maximum with the variable arm on R_3 all the way to the left (grounded), and minimum at the right. R_3 could simply be placed in series with R_1 , omitting R_4 entirely, but the range of control with this connection is limited because it depends on the cathode current alone.

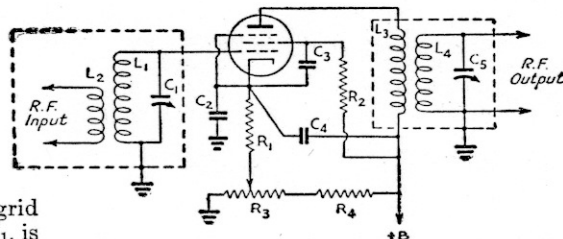


Fig. 713 — Basic circuit of a tuned radio-frequency amplifier. Component values are discussed in the text.

In a multi-tube receiver the gain of several stages may be varied simultaneously, a single control sufficing for all. The lower ends of the several cathode resistors (R_1) are then connected together and to the movable contact on R_3 in Fig. 713.

Circuit values — The value of the cathode resistor, R_1 , should be calculated for the minimum recommended bias for the tube used. The capacities of C_2 , C_3 and C_4 must be such that the reactance is low at radio frequencies; this condition is easily met by using 0.01- μ fd. condensers at communication frequencies, or 0.001 to 0.002 mica units at very-high frequencies up to 112 Mc. R_2 is found by taking the difference between the recommended plate and screen voltages, then substituting this and the rated screen current in Ohm's Law (§ 2-6). R_3 must be selected on the basis of the number of tubes to be controlled; a resistor must be chosen which is capable of carrying, at its low-resistance end, the sum of all the tube currents plus the bleeder current. A resistor of suitable current-carrying capacity being found, the bleeder current necessary to produce a drop through it of about 50 volts can be calculated by Ohm's Law. The same formula will give R_4 , using the plate voltage less 50 volts for E and the bleeder current previously found for I .

The constants of the tuned circuits will depend upon the frequency range, or band, to be covered. A fairly high L/C ratio (§ 2-10) should be used on each band; this is limited, however, by the irreducible minimum capacities. To an allowance of 10 to 20 μ fd. for tube and stray capacities should be added the minimum capacity of the tuning condenser.

If the input circuit of the amplifier is connected to an antenna, the coupling coil, L_2 , should be adjusted to provide critical coupling (§ 2-11) between the antenna and grid circuit. This will give maximum energy transfer. The turns ratio of L_1/L_2 will depend upon the frequency, the type of tube used, the Q of the tuned circuit and the constants of the antenna system, and in general is best determined experimentally. The selectivity will increase as the coupling is reduced below this "optimum" value, a consideration which it is well to keep in mind if selectivity is of more importance than maximum gain.

The output-circuit coupling depends upon the plate resistance (§ 3-2) of the tube, the input resistance of the succeeding stage, and the Q of the tuned circuit, L_4C_5 . L_3 usually is coupled as closely as possible to L_4 (avoiding the necessity for an additional tuning condenser across L_3) and the energy transfer is maximum when L_3 has $\frac{2}{3}$ to $\frac{4}{5}$ as many turns as L_4 , with ordinary receiving pentodes.

Tube and circuit noise — In any conductor electrons will be moving in random directions simultaneously and, as a result, small irregular voltages are developed across the conductor terminals. The voltage is larger the greater the resistance of the conductor and the higher its

temperature. This is known as the *thermal-agitation* effect, and it produces a hiss-like noise voltage distributed uniformly throughout the radio-frequency spectrum. The thermal-agitation noise voltage appearing across the terminals of a tuned circuit will be the same as in a resistor of a value equal to the parallel impedance (§ 2-10) of the tuned circuit, even though the actual circuit resistance is low. Hence, the higher the Q of the circuit, the greater the thermal agitation noise.

Another component of hiss noise is developed in the tube because the rain of electrons on the plate is not entirely uniform. Small irregularities caused by gas in the tube also contribute to the effect. Tube noise varies with the type of tube; in general, the higher the cathode current and the lower the mutual conductance of the tube, the more internal noise it will generate.

To obtain the best signal-to-noise ratio, the signal must be made as large as possible at the grid of the tube, which means that the antenna coupling must be adjusted to that end and also that the Q of the grid tuned circuit must be high. A tube with low inherent noise obviously should be chosen. In an amplifier having good signal-to-noise ratio, the thermal-agitation noise will be greater than the tube noise. This can easily be checked by disconnecting the antenna so that no outside noise is being introduced into the receiver, then grounding the grid through a 0.01- μ fd. condenser and observing whether there is a decrease in noise. If there is no change the tube noise is greatly predominant, indicating a poor signal-to-noise ratio in the stage. The test is valid only if there is no regeneration in the amplifier. The signal-to-noise ratio will decrease as the frequency is raised, because it becomes increasingly difficult to obtain a tuned circuit of high effective Q (§ 7-7).

The first stage of the receiver is the important one from the standpoint of signal-to-noise ratio. Noise generated in the second and subsequent stages, while comparable in magnitude to that generated in the first, is masked by the amplified noise and signal from the first stage. After the second stage, further contributions by tubes and circuits to the total noise are inconsequential in any normal receiver.

Tube input resistance — At high radio frequencies the tube may consume power from the tuned grid circuit, even though the grid is not driven positive by the signal. Above 7 Mc. all tubes "load" the tuned circuit to some extent, the amount of loading varying with the type of tube. This effect comes about because the time necessary for electrons to travel from the cathode to the grid becomes comparable to the time of one r.f. cycle, and because of the degenerative effect (§ 3-3) of the cathode lead inductance. It becomes more pronounced as the frequency is increased. Certain types of tubes may have an input resistance of only a few thousand ohms at 28 Mc. and as little as

a few hundred ohms at very-high frequencies. The input resistance of the same tubes at 7 Mc. and lower frequencies may be so high as to be considered infinite, from a practical standpoint.

This *input-loading* effect is in addition to the normal decrease in the Q of the tuned circuit alone, because of increased losses in the coil and condenser at the higher frequencies. Thus the selectivity and gain of the circuit both are affected adversely by increasing frequency.

Comparison of tubes — At 7 Mc. and lower frequencies, the signal-to-noise ratio, gain, and selectivity of an r.f.-amplifier stage are sufficiently high with any of the standard receiving tubes. At 14 Mc. and higher, however, this is no longer true, and the choice of a tube must be based on several conflicting considerations.

Gain is highest with high mutual-conductance pentodes, the 1851 and 1852 being examples of this type. These tubes also develop less noise than any of the others. The input-loading effect is greatest with them, however, so that selectivity is decreased and the tuned-circuit gain is lowered.

Pentodes, such as the 6K7, 6J7 and corresponding types in glass, have lesser input-loading effects at high frequencies, moderate gain, and relatively high inherent noise.

"Acorn" and equivalent miniature pentodes are excellent from the input-loading standpoint; gain is about the same as with standard types, and the inherent noise is somewhat lower.

Where selectivity is paramount the acorns are best, the standard pentodes second, and the 1851-52 types worst. On signal-to-noise ratio the 1851-52 tubes are first, acorns second and standard pentodes third. The same order of precedence holds for over-all gain.

At 56 Mc. the standard types are usable, but acorns are capable of better performance because of lesser loading. The 954 and 956 and the corresponding types, 9001 and 9003, are practically the only usable types for r.f. amplification at 112 Mc. and higher.

7-7 Tuning and Band-Changing Methods

Band-changing — The resonant circuits which are tuned to the frequency of the incoming signal constitute a special problem in the design of amateur receivers, since the amateur frequency assignments consist of groups or bands of frequencies at widely spaced intervals. The same LC combination cannot be used for, say, 14 Mc. to 3.5 Mc., because of the impracticable maximum-minimum capacity ratio required, and also because the tuning would be excessively critical with such a large frequency range. It is necessary, therefore, to provide a means for changing the circuit constants for various frequency bands. As a matter of convenience the same tuning condenser usually is retained, but new coils are inserted in the circuit for each band.

There are two favorite methods of changing inductances. One is to use a switch having an

appropriate number of contacts, which connects the desired coil and disconnects the others. The second is to use coils wound on forms with contacts (usually pins) which can be plugged in and removed from a socket.

Bandspreading — The tuning range of a given coil and variable condenser will depend upon the inductance of the coil and the change in tuning capacity. For ease of tuning, it is desirable to adjust the tuning range so that practically the whole dial scale is occupied by the band in use. This is called *bandspreading*. Because of the varying widths of the bands, special tuning methods must be devised to give the correct maximum-minimum capacity ratio on each band. Several of these methods are shown in Fig. 714.

In A, a small *bandspread condenser*, C_1 (15 to 25 μfd . maximum capacity), is used in parallel with a condenser, C_2 , which is usually large enough (140 to 175 μfd .) to cover a 2-to-1 frequency range. The setting of C_2 will determine the minimum capacity of the circuit, and the maximum capacity for bandspread tuning will be the maximum capacity of C_1 plus the setting of C_2 . The inductance of the coil can be adjusted so that the maximum-minimum ratio will give adequate bandspread. In practicable circuits it is almost impossible, because of the non-harmonic relation of the various bands, to get full bandspread on all bands with the same pair of condensers, especially when the coils are wound to give continuous frequency coverage on C_2 , which is variously called the *band-setting* or *main-tuning* condenser. C_2 must be reset each time the band is changed.

The method shown at B makes use of condensers in series. The tuning condenser, C_1 , may have a maximum capacity of 100 μfd . or more. The minimum capacity is determined principally by the setting of C_3 , which usually has low capacity, and the maximum capacity by the setting of C_2 , which is of the order of 25 to 50 μfd . This method is capable of close adjustment to practically any desired degree of bandspread. Either C_2 and C_3 must be adjusted for each band or separate pre-adjusted condensers must be switched in.

The circuit at C also gives complete spread on

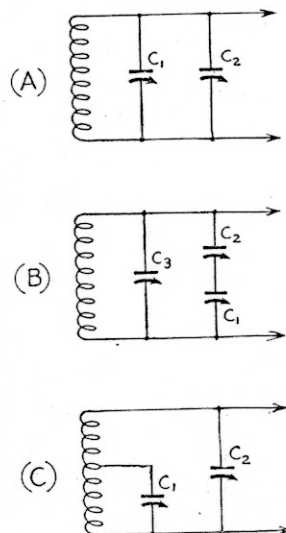


Fig. 714 — Essentials of three bandspread tuning systems.

each band. C_1 , the bandspread condenser, may have any convenient value of capacity; 50 $\mu\text{fd.}$ is satisfactory. C_2 may be used for continuous frequency coverage ("general coverage") and as a band-setting condenser. The effective maximum-minimum capacity ratio depends upon the capacity of C_2 and the point at which C_1 is tapped on the coil. The nearer the tap to the bottom of the coil, the greater the bandspread, and vice versa. For a given coil and tap, the bandspread will be greater if C_2 is set at larger capacity. C_2 may be mounted in the plug-in coil form and pre-set, if desired. This requires a separate condenser for each band, but eliminates the necessity for resetting C_2 each time the band is changed.

Ganged tuning—The tuning condensers of the several r.f. circuits may be coupled together mechanically and operated by a single control. However, this operating convenience involves more complicated construction, both electrically and mechanically. It becomes necessary to make the various circuits *track*—that is, tune to the same frequency at each setting of the tuning control.

True tracking can be obtained only when the inductance, tuning condensers, and circuit minimum and maximum capacities are identical in all "ganged" stages. A small *trimmer* or *padding* condenser may be connected across the coil, so that variations in minimum capacity can be compensated. The fundamental circuit is shown in Fig. 715, where C_1 is the trimmer and C_2 the tuning condenser. The use of the trimmer increases the minimum circuit capacity, but is a necessity for satisfactory tracking. Condensers having maximum capacities of 15 to 30 $\mu\text{fd.}$ are commonly used.

The same methods are applied to bandspread circuits which must be tracked. The circuits are identical with those of Fig. 714. If both general-coverage and bandspread tuning are to be available, an additional trimmer condenser must be connected across the coil in each circuit shown. If only amateur-band tuning is desired, however, then C_3 in Fig. 714-B, and C_2 in Fig. 714-C serve as trimmers.

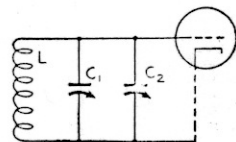


Fig. 715—Showing the use of a trimmer condenser, C_1 , across the tuned circuit to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

The coil inductance can be adjusted by starting with a larger number of turns than necessary and removing a turn or fraction of a turn at a time until the circuits track satisfactorily. An alternative method, provided the inductance is reasonably close to the correct value initially, is to make the coil so that the last turn is variable with respect to the whole coil, or to use a single short-circuited turn the position of which can be varied with respect to the coil. These methods are shown in Fig. 716.

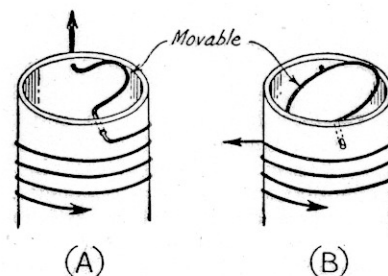


Fig. 716—Methods of adjusting the inductance for ganging. The half turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the plane of the loop is parallel to the axis of the coil, and will give maximum reduction of the coil inductance when perpendicular to the coil axis.

V.h.f. circuits—Interelectrode capacities are practically constant for a given tube regardless of the operating frequency, and the same is approximately true of stray circuit capacities. Hence, at very-high frequencies these capacities become an increasingly larger part of the usable tuning capacity, and reasonably high L/C ratios (§ 2-10) are more difficult to secure as the frequency is raised. Because of this irreducible minimum capacity, standard types of tubes cannot be tuned to frequencies higher than about 200 Mc., even when the inductance in the circuit is simply that of a straight wire between the tube elements.

Along with these capacity effects, the input loading (§ 7-6) increases rapidly at very-high frequencies, so that ordinary tuned circuits have very low effective Q s when connected to the grid circuit of a tube. The effect is still further aggravated by the fact that losses in the tuned circuit itself are higher, causing a still further reduction in Q . For these reasons, the frequency limit at which an r.f. amplifier will give any gain is in the vicinity of 60 Mc. with standard tubes. At higher frequencies there will be a loss, instead of amplification. This condition can be mitigated somewhat by taking steps to improve the effective Q of the circuit, either by tapping the grid down on the coil, as shown in Fig. 717-A, or by using a lower L/C ratio (§ 2-10). The Q of the tuned circuit alone can be greatly improved by using a linear circuit (§ 2-12), which when properly constructed will give Q s much higher than those attainable at lower frequencies with conventional coils and condensers. The concentric type of line, Fig. 717-B, is best both from the standpoint of Q and of adaptability to non-symmetrical circuits such as are used in receivers. Since the capacity and resistance loading effects of the tube are still present, the Q of such a circuit will be destroyed if the grid-cathode circuit of the tube is connected directly across it. Hence, tapping down on the line, as shown, is necessary.

Very-high-frequency amplifiers should employ tubes of the acorn type, which have the

least loading effect as well as low interelectrode capacities. This is because the smaller loading effect means higher input resistance, and, for a given loaded Q of the tuned circuit, a higher voltage developed between the grid and cathode. Thus the amplification of the stage is higher.

A concentric circuit may be tuned by varying the length of the inner conductor (usually by using close-fitting tubes, one sliding inside the other) or by connecting an ordinary tuning condenser across the line. Tapping the condenser down, as shown in Fig. 717-B, gives a bandspread effect, which is advantageous. It also helps to keep the Q of the circuit higher than it would be with the condenser connected directly across the open end of the line, since at very-high frequencies most condensers have losses which cannot be neglected.

V.h.f. oscillators such as those used in the superregenerative detector usually will work well at frequencies where r.f. amplification is impossible with standard tubes (as in the 112-Mc. band), since tube losses are compensated for by energy taken from the power supply. Ordinary coil and condenser circuits are practicable with such tubes at 112 Mc.

7-8 The Superheterodyne

Principles — In the *superheterodyne*, or *superhet*, receiver the frequency of the incoming signal is changed to a new radio frequency, the *intermediate frequency* (i.f.), then amplified, and finally detected. The frequency is

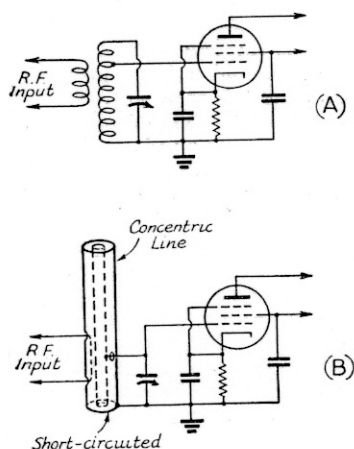


Fig. 717 — Circuits of improved Q for very-high frequencies. A, reducing tube loading by tapping down on the resonant circuit; B, use of a concentric-line circuit, with the tube similarly tapped down. The line should be a quarter-wave long, electrically; because of the additional shunt capacity represented by the tube, the physical length will be somewhat less than given by the formula (§ 10-5). In general, this reduction in length will be greater the higher the grid tap on the inner conductor. The coupling turn should be parallel to the axis of the line and must be insulated from the outer conductor.

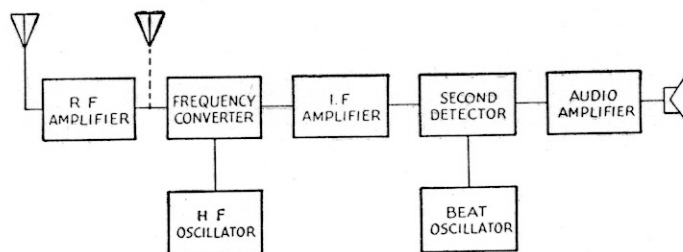


Fig. 718 — Block diagram of the basic elements of the superheterodyne

changed by means of the heterodyne process (§ 7-1), the output of an adjustable *local oscillator* (the *h.f. oscillator*) being combined with the incoming signal in a *mixer* or *converter* stage (*first detector*) to produce a beat frequency equal to the i.f. Fig. 718 gives the essentials of the superheterodyne in block form. C.w. signals are made audible by heterodyning the signal at the second detector by the *beat-frequency oscillator* (b.f.o.) or *beat oscillator*, set to differ from the i.f. by a suitable audio frequency.

As a numerical example, assume that an intermediate frequency of 455 kc. is chosen and that the incoming signal is on 7000 kc. Then the h.f. oscillator frequency may be set to 7455 kc., in order that the beat frequency (7455 minus 7000) will be 455 kc. The h.f. oscillator also could be set to 6545 kc., which will give the same frequency difference. To produce an audible c.w. signal of, say, 1000 cycles at the second detector, the beat oscillator would be set to either 454 kc. or 456 kc.

Characteristics — The frequency-conversion process permits r.f. amplification at a relatively low frequency. Thus high selectivity can be obtained, and this selectivity is constant regardless of the signal frequency. Higher gain also is possible at the lower frequency. The separate oscillators can be designed for stability, and, since the h.f. oscillator is working at a frequency considerably removed from the signal frequency, its stability is practically unaffected by the incoming signal.

Images — Each h.f. oscillator frequency will cause i.f. response at two signal frequencies, one higher and one lower than the oscillator frequency. If the oscillator is set to 7455 kc. to respond to a 7000-kc. signal, for example, it will respond also to a signal on 7910 kc., which likewise gives a 455-kc. beat. The undesired signal of the two is called the *image*.

The radio-frequency circuits of the receiver (those used before the frequency is converted to the i.f.) normally are tuned to the desired signal, so that the selectivity of the circuits reduces the response to the image signal. If the desired signal and image have equal strengths at the input terminals of the receiver, the ratio of the receiver voltage output from the desired signal to that from the image is called the *signal-to-image ratio*, or *image ratio*.

The image ratio depends upon the selectivity of the r.f. tuned circuits preceding the mixer

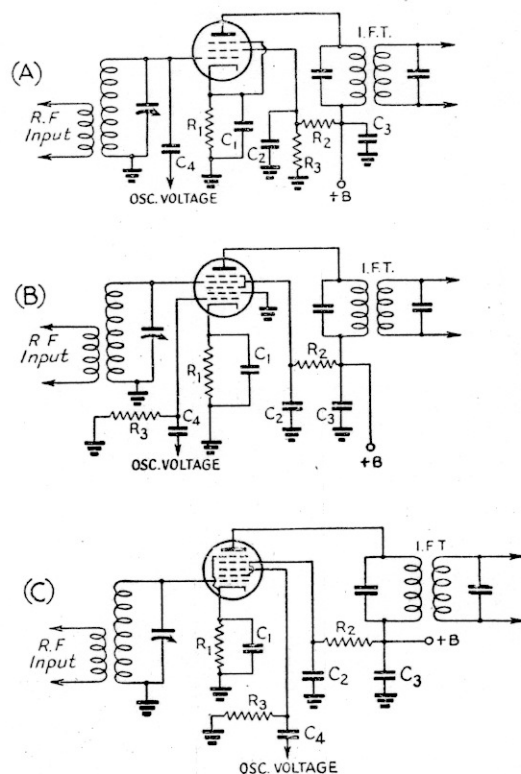


Fig. 719 — Mixer or converter circuits. A, grid injection with a pentode plate detector; B and C, separate injection circuits for converter tubes. Circuit values are:

Component	Circuit A	Circuit B	Circuit C
C ₁ , C ₂ , C ₃	0.01–0.1 μ fd.	0.01–0.1 μ fd.	0.01–0.1 μ fd.
C ₄	Approx. 1 μ fd.	50–100 μ fd.	50–100 μ fd.
R ₁	10,000 ohms.	300 ohms.	500 ohms.
R ₂	0.1 megohm.	50,000 ohms.	15,000 ohms.
R ₃	50,000 ohms.	50,000 ohms.	50,000 ohms.

Plate voltage should be 250 in all circuits. If an 1851 or 1852 tube is used in Circuit A, R₁ should be 500 ohms.

tube. Also, the higher the intermediate frequency, the higher the image ratio, since raising the i.f. increases the frequency separation between signal and image and places the latter farther away from the peak of the resonance curve (§ 2-10) of the signal-frequency circuits.

Other spurious responses — In addition to images, other signals to which the receiver is not ostensibly tuned may be heard. Harmonics of the high-frequency oscillator may beat with signals far removed from the desired frequency to produce output at the intermediate frequency; such spurious responses can be reduced by adequate selectivity before the mixer stage, and by using sufficient shielding to prevent signal pick-up by any means other than the antenna. When a strong signal is received, the harmonics (§ 2-7) generated by rectification in the second detector may, by stray coupling, be introduced into the r.f. or mixer circuit and converted to the intermediate frequency, to go through the receiver in the same way as an ordinary signal. These "birdies" appear as a heterodyne beat on the desired signal, and are

principally bothersome when the frequency of the incoming signal is not greatly different from the intermediate frequency. The cure is proper circuit isolation and shielding.

Harmonics of the beat oscillator also may be converted in similar fashion and amplified through the receiver; these responses can be reduced by shielding the beat oscillator and operating it at low output level.

The double superhet — At high and very-high frequencies it is difficult to secure an adequate image ratio when the intermediate frequency is of the order of 455 kc. To reduce image response the signal frequently is converted first to a rather high (1500, 5000, or even 10,000 kc.), and then — sometimes after further amplification — reconverted to a lower i.f. where higher adjacent-channel selectivity can be obtained. Such a receiver is called a *double superheterodyne*.

7-9. Frequency Converters

Characteristics — The first detector or mixer resembles an ordinary detector. A circuit tuned to the intermediate frequency is placed in the plate circuit of the mixer, so that the highest possible i.f. voltage will be developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are bypassed to ground, since they are not wanted in the output. The i.f. tuned circuit should have low impedance for these frequencies, a condition easily met if they do not approach the intermediate frequency.

The *conversion efficiency* of the mixer is the ratio of i.f. output voltage from the plate circuit to r.f. signal voltage applied to the grid. High conversion efficiency is desirable. The mixer tube noise also should be low if a good signal-to-noise ratio is wanted, particularly if the mixer is the first tube in the receiver.

The mixer should not require too much r.f. power from the h.f. oscillator, since it may be difficult to supply the power and maintain good oscillator stability (§ 3-7). Also, the conversion efficiency should not depend too critically on the oscillator voltage (that is, a small change in oscillator output should not change the gain), since it is difficult to maintain constant output over a wide frequency range.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called *pulling*. If the mixer and oscillator could be completely isolated, mixer tuning would have no effect on the oscillator frequency; but in practice this is a difficult condition to attain. Pulling causes oscillator instability and should be minimized, because the stability of the whole receiver depends critically upon the stability of the h.f. oscillator. Pulling effect decreases with separation of the signal and h.f. oscillator frequencies, hence is less with high intermediate frequencies and greater with low i.f.s.

Circuits — Typical frequency-conversion circuits are given in Fig. 719. The variations are chiefly in the way in which the oscillator volt-

age is introduced. In Fig. 719-A, the screen-grid pentode functions as a plate detector; the oscillator is capacity-coupled to the grid of the tube, in parallel with the tuned input circuit. Inductive coupling may be used instead. The conversion gain and input selectivity generally are good, so long as the sum of the two voltages (signal and oscillator) impressed on the mixer grid does not exceed the grid bias. It is desirable to make the oscillator voltage as high as possible without exceeding this limitation. The oscillator power required is negligible.

A pentagrid-converter tube is used in the circuit at B. Although intended for combination oscillator-mixer use, this type of tube usually will give more satisfactory performance when used in conjunction with a separate oscillator, the output of which is coupled in as shown. The circuit gives good conversion efficiency, and, because of the electron coupling, affords desirable isolation between the mixer and oscillator circuits. A small amount of power is required from the oscillator.

Circuit C is for the 6L7 mixer tube. The oscillator voltage can vary over a considerable range without affecting the conversion gain. There are no critical adjustments, and the oscillator-mixer isolation is good. The oscillator must supply somewhat more power than in B.

A more stable receiver generally results, particularly at the higher frequencies, when separate tubes are used for the mixer and oscillator. Practically the same number of circuit components is required whether or not a combination tube is used, so that there is little difference from the cost standpoint.

Tubes for frequency conversion — Any sharp cut-off pentode may be used in the circuit of Fig. 719-A. The 1851 and 1852 give high conversion gain and excellent signal-to-noise ratio — comparable, in fact, to the gain and signal-to-noise ratio obtainable with r.f. amplifiers — and in these respects are far superior to any other tubes used as mixers. However, this type of tube loads the circuit more (§ 7-6) and thus decreases the selectivity.

The 6K8 is a good tube for the circuit at B; its oscillator plate connection may be ignored. The 6SA7 also is excellent in this circuit, although it has no anode grid (No. 2 grid, in the diagram). In addition to these two types, any pentagrid converter tube may be used.

7-10 The High-Frequency Oscillator

Design considerations — Stability of the receiver (§ 7-2) is dependent chiefly upon the stability of the h.f. oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to changes in voltage, loading, and mechanical shock. Thermal effects (slow change in frequency because of tube or circuit heating) should be minimized. These ends can be attained by the use of good insulating materials and circuit components, suitable electrical design, and careful mechanical construction.

In addition, the oscillator must be capable of furnishing sufficient r.f. voltage and power for the particular mixer circuit chosen, at all frequencies within the range of the receiver, and its harmonic output should be as low as possible to reduce spurious response (§ 7-8).

It is desirable to make the L/C ratio in the oscillator tuned circuit low (high- C), since this results in increased stability (§ 3-7). Particular care should be taken to insure that no part of the oscillator circuit can vibrate mechanically. This calls for short leads and "solid" mechanical construction. The chassis and panel material should be heavy, and rigid enough so that pressure on the tuning dial will not cause torsion and a shift in the frequency. Care in mechanical construction is well repaid by increased frequency stability.

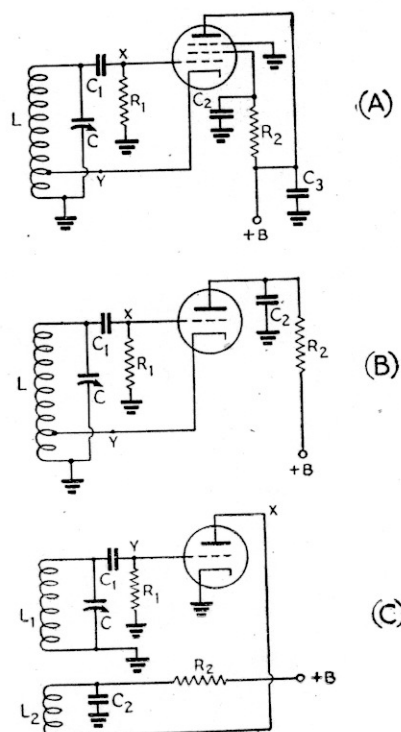


Fig. 720 — High-frequency oscillator circuits. A, screen-grid pentode oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points X and Y. In A and B, coupling from Y will reduce pulling effects, but gives less voltage than from X; this type is best adapted to mixer circuits with small oscillator-voltage requirements. Typical values for components are as follows:

	Circuit A	Circuit B	Circuit C
C ₁ —	100 μ fd.	100 μ fd.	100 μ fd.
C ₂ —	0.1 μ fd.	0.1 μ fd.	0.1 μ fd.
C ₃ —	0.1 μ fd.		
R ₁ —	50,000 ohms.	50,000 ohms.	50,000 ohms.
R ₂ —	50,000 ohms.	10,000 to 25,000 ohms.	10,000 to 25,000 ohms.

The plate-supply voltage should be 250 volts. In circuits B and C, R₂ is used to drop the supply voltage to 100–150 volts; it may be omitted if voltage is obtained from a voltage divider in the power supply (§ 8-10).

Circuits—Several oscillator circuits are shown in Fig. 720. The point at which output voltage is taken for the mixer is indicated in each case by X or Y. Circuits A and B will give about the same results, and require only one coil. However, in these two circuits the cathode is above ground potential for r.f., which often is a cause of hum modulation of the oscillator output at 14 Mc. and higher frequencies when 6.3-volt heater tubes are used. Hum usually is not bothersome with 2.5-volt tubes, nor, of course, with tubes which are heated by direct current. The circuit of Fig. 720-C overcomes hum, since the cathode is grounded. The two-coil arrangement is advan-

tageous in construction, since the feed-back adjustment (altering the number of turns on L_2 or the coupling between L_1 and L_2) is simple mechanically.

Besides the use of a fairly high C/L ratio in the tuned circuit, it is necessary to adjust the feed-back to obtain optimum results. Too much feed-back will cause the oscillator to "squeg," or operate at several frequencies simultaneously (§ 7-4); too little feed-back will cause the output to be low. In the tapped-coil circuits (A, B), the feed-back is increased by moving the tap toward the grid end of the coil; in C, by increasing the number of turns on L_2 or by moving L_2 closer to L_1 .

The oscillator plate voltage should be as low as is consistent with adequate output. Low plate voltage will cause reduced tube heating and thereby reduce frequency drift. The oscillator and mixer circuits should be well isolated, preferably by shielding, since coupling other than by the means intended may result in pulling.

To avoid plate-voltage changes which may cause the oscillator frequency to change, it is good practice to use a regulated plate supply employing a gaseous voltage-regulator tube (§ 8-8).

Tracking—For ganged tuning, there must be a constant difference in frequency between the oscillator and mixer circuits. This difference must be exactly equal to the intermediate frequency (§ 7-8).

Tracking methods for covering a wide frequency range, suitable for general-coverage receivers, are shown in Fig. 721. The tracking capacity, C_5 , commonly consists of two condensers in parallel, a fixed one of somewhat less capacity than the value needed and a smaller variable in parallel to allow for adjustment to the exact proper value. In practice, the trimmer, C_4 , is first set for the high-frequency end of the tuning range, and then the tracking condenser is set for the low-frequency end. The tracking capacity becomes larger as the percentage difference between the oscillator and signal frequencies becomes smaller (that is, as the signal frequency becomes higher). Typical circuit values are given in the tables under Fig. 721.

In amateur-band receivers, tracking is simplified by choosing a bandspread circuit which gives practically straight-line-frequency tuning (equal frequency change for each dial division), and then adjusting the oscillator and mixer tuned circuits so that both cover the same total number of kilocycles. For example, if the i.f. is 455 kc. and the mixer circuit tunes from 7000 to 7300 kc. between two given points on the dial, then the oscillator must tune from 7455 to 7755 kc. between the same two dial readings. With the bandspread arrangement of Fig. 714-C, the tuning will be practically straight-line-frequency if the capacity actually in use at C_2 is not too small; the same is true of 714-A if C_1 is small compared to C_2 .

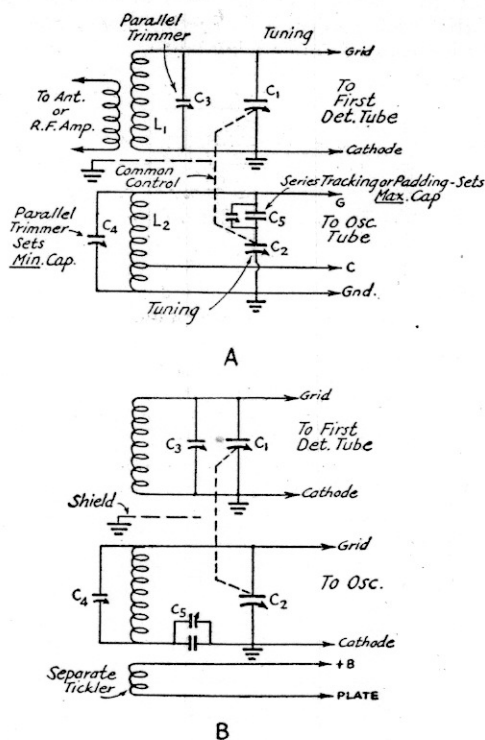


Fig. 721—Converter-circuit tracking methods. Following are approximate circuit values for 450- to 465-kc. i.f.s, with tuning ranges of approximately 2.15-to-1 and C_2 having 140 $\mu\text{fd.}$ maximum, and the total minimum capacitance, including C_3 or C_4 , being 30 to 35 $\mu\text{fd.}$

Tuning Range	L_1	L_2	C_5
1.7-4 Mc.	50 $\mu\text{h.}$	40 $\mu\text{h.}$	0.0013 $\mu\text{fd.}$
3.7-7.5 Mc.	14 $\mu\text{h.}$	12.2 $\mu\text{h.}$	0.0022 $\mu\text{fd.}$
7-15 Mc.	3.5 $\mu\text{h.}$	3 $\mu\text{h.}$	0.0045 $\mu\text{fd.}$
14-30 Mc.	0.8 $\mu\text{h.}$	0.78 $\mu\text{h.}$	None used

Approximate values for 450- to 465-kc. i.f.s with a 2.5-to-1 tuning range, C_1 and C_2 being 350- $\mu\text{fd.}$ maximum, minimum including C_3 and C_4 being 40 to 50 $\mu\text{fd.}$

Tuning Range	L_1	L_3	C_5
0.5-1.5 Mc.	240 $\mu\text{h.}$	130 $\mu\text{h.}$	425 $\mu\text{fd.}$
1.5-4 Mc.	32 $\mu\text{h.}$	25 $\mu\text{h.}$	0.00115 $\mu\text{fd.}$
4-10 Mc.	4.5 $\mu\text{h.}$	4 $\mu\text{h.}$	0.0028 $\mu\text{fd.}$
10-25 Mc.	0.8 $\mu\text{h.}$	0.75 $\mu\text{h.}$	None used

7-11 The Intermediate-Frequency Amplifier

Choice of frequency — The selection of an intermediate frequency is a compromise between various conflicting factors. The lower the i.f. the higher the selectivity and gain, but a low i.f. brings the image nearer the desired signal and hence decreases the image ratio (§ 7-8). A low i.f. also increases pulling of the oscillator frequency (§ 7-9). On the other hand, a high i.f. is beneficial to both image ratio and pulling, but the selectivity and gain are lowered. The difference in gain is least important.

An i.f. of the order of 455 kc. gives good selectivity and is satisfactory from the standpoint of image ratio and oscillator pulling at frequencies up to 7 Mc. The image ratio is poor at 14 Mc. when the mixer is connected to the antenna, but adequate when there is a tuned r.f. amplifier between antenna and mixer. At 28 Mc. and on the very-high frequencies, the image ratio is very poor unless several r.f. stages are used. Above 14 Mc., pulling is likely to be bad unless very loose coupling can be used between mixer and oscillator.

With an i.f. of about 1600 kc., satisfactory image ratios can be secured on 14, 28 and 56 Mc., and pulling can be reduced to negligible proportions. However, the i.f. selectivity is considerably lower, so that more tuned circuits must be used to increase the selectivity. For very-high frequencies, including 28 Mc., the best solution is to use a double superheterodyne (§ 7-8), choosing one high i.f. for image reduction (5 and 10 Mc. are frequently used) and a lower one for gain and selectivity.

In choosing an i.f. it is wise to avoid frequencies on which there is considerable activity by the various radio services, since such signals may be picked up directly on the i.f. wiring. The frequencies mentioned are fairly free of such interference.

Fidelity, sideband cutting — As described in § 5-2, modulation of a carrier causes the generation of sideband frequencies numerically equal to the carrier frequency plus and minus the highest modulation frequency present. If the receiver is to give a faithful reproduction of modulation which contains, for instance, audio frequencies up to 5000 cycles, it must be capable of amplifying equally all frequencies contained in a band extending from 5000 cycles above to 5000 cycles below the carrier frequency. In a superheterodyne, where all carrier

frequencies are changed to the fixed intermediate frequency, this means that the i.f. amplifier should amplify equally well all frequencies within that band. In other words, the amplification must be uniform over a band 10 kc. wide, with the i.f. at its center. The signal-frequency circuits usually do not have enough over-all selectivity to affect materially the "adjacent channel" selectivity (§ 7-2), so that only the i.f. amplifier selectivity need be considered.

A 10-kc. band is considered sufficient for reasonably faithful reproduction of music, but much narrower band-widths can be used for communication work where intelligibility rather than fidelity is the primary objective. If the selectivity is too great to permit uniform amplification over the band of frequencies occupied by the modulated signal, the higher modulating frequencies are attenuated as compared to the lower frequencies; that is, the upper-frequency sidebands are "cut." While sideband cutting reduces fidelity, it is frequently preferable to sacrifice naturalness of reproduction in favor of greater selectivity.

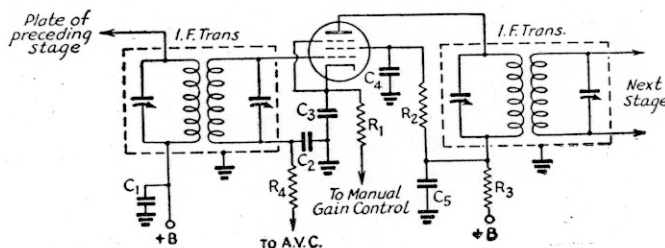
The selectivity of an i.f. amplifier, and hence the tendency to cut sidebands, increases with the number of amplifier stages and also is greater the lower the intermediate frequency. From the standpoint of communication, sideband cutting is not serious with two-stage amplifiers at frequencies as low as 455 kc.

Circuits — I.f. amplifiers usually consist of one or two stages. Two stages at 455 kc. give all the gain usable, in view of the minimum receiver noise level, and also give suitable selectivity for good-quality 'phone reception.

A typical circuit arrangement is shown in Fig. 722. A second stage would simply duplicate the circuit of the first. In principle, the i.f. amplifier is the same as the tuned r.f. amplifier (§ 7-6). However, since a fixed frequency is used, the primary as well as the secondary of the coupling transformer is tuned, giving higher selectivity than is obtainable with a closely coupled untuned primary. The cathode resistor, R_1 , is connected to a gain control circuit of the type previously described (§ 7-6); usually both stages, if two are used, are controlled by a single variable resistor. The decoupling resistor, R_3 (§ 2-11), helps isolate the amplifier, and thus prevents stray feed-back. C_2 and R_4 are part of the automatic volume-control circuit (§ 7-13); if no a.v.c. is used, the lower end of the i.f. transformer secondary is simply connected to ground.

Fig. 722 — Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

C_1 — 0.1 μ fd. at 455 kc.; 0.01 μ fd. at 1600 kc. and higher.
 C_2 — 0.01 μ fd.
 C_3, C_4, C_5 — 0.1 μ fd. at 455 kc.; 0.01 μ fd. above 1600 kc.
 R_1 — 300 ohms. R_3 — 2000 ohms.
 R_2 — 0.1 megohm. R_4 — 0.25 megohm.



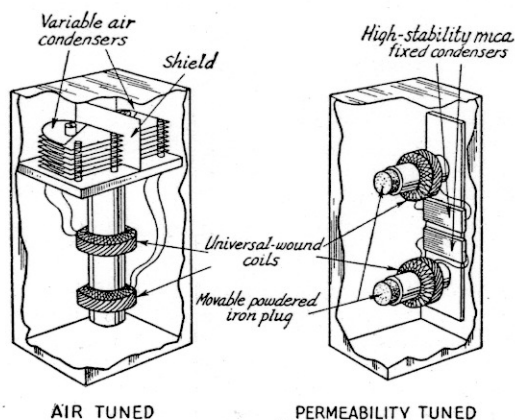


Fig. 723 — Representative i.f. transformer construction. Coils are supported on insulating tubing or (in the air-tuned type) on wax-impregnated wooden dowels. The shield in the air-tuned transformer prevents capacity coupling between the tuning condensers. In the permeability-tuned transformer the cores consist of finely divided iron particles supported in an insulating binder, formed into cylindrical "plugs." The tuning capacity is fixed, and the inductances of the coils are varied by moving the iron plugs in and out.

In a two-stage amplifier the screen grids of both stages may be fed from a common supply, either through a resistor (R_2) as shown, the screens being connected in parallel, or from a voltage divider (§ 8-10) across the plate supply. Separate screen voltage-dropping resistors are preferable for preventing undesired coupling between stages.

When two stages are used the high gain will tend to cause instability and oscillation, so that good shielding, by-passing, and careful circuit arrangement to prevent stray coupling, with exposed r.f. leads well separated, is necessary.

I.f. transformers — The tuned circuits of i.f. amplifiers are built up as transformer units consisting of a metal-shield container in which the coils and tuning condensers are mounted. Both air-core and powdered-iron-core universal-wound coils are used, the latter having somewhat higher Q s and, hence, greater selectivity and gain per unit. In universal windings the coil is wound in layers but with each turn traversing the length of the coil, back and forth, rather than being wound in a plane perpendicular to the axis as it is in ordinary single-layer coils. With straight layer winding, the turns on adjacent layers at the edges of the coil have a rather large potential difference between them as compared to the difference between any two adjacent turns in the same layer; hence a fairly large capacity current can flow between layers. Universal winding, with its "criss-crossed" turns, tends to avoid building up such potential differences, and hence reduces distributed-capacity effects (§ 2-8).

Variable tuning condensers are of the midge type, air-dielectric condensers being preferable because their capacity is practically unaffected by changes in temperature and humidity. Iron-core transformers may be tuned by varying the

inductance (permeability tuning), in which case stability comparable to that of variable air-condenser tuning can be obtained by use of high-stability fixed mica condensers. Such stability is of great importance, since a circuit whose frequency "drifts" with time eventually will be tuned to a different frequency than the other circuits, thereby reducing the gain and selectivity of the amplifier. Typical i.f. transformer construction is shown in Fig. 723.

Besides the type of i.f. transformer shown in Fig. 723, special units to give desired selectivity characteristics are available. For higher than ordinary adjacent-channel selectivity (§ 7-2) *triple-tuned* transformers, with a third tuned circuit inserted between the input and output windings, are used. The energy is transferred from the input to the output windings via this *tertiary* winding, thus adding its selectivity to the over-all selectivity of the transformer. *Variable-selectivity* transformers also can be obtained. These usually are provided with a third (untuned) winding which can be connected to a resistor, thereby loading the tuned circuits and decreasing the Q and selectivity (§ 2-10) to broaden the selectivity curve. The variation in selectivity is brought about by switching the resistor in and out of the circuit. Another method is to vary the coupling between primary and secondary, overcoupling being used to broaden the selectivity curve and undercoupling to sharpen it (§ 2-11).

Selectivity — The over-all selectivity of the i.f. amplifier will depend on the frequency and the number of stages. The following figures are indicative of the band-widths (§ 7-2) to be expected with good-quality transformers in amplifiers so constructed as to keep regeneration to a minimum:

Intermediate frequency	Band-width in kilocycles		
	2 times down	10 times down	100 times down
One stage, 455 kc. (air core) . . .	8.7	17.8	32.3
One stage, 455 kc. (iron core) . .	4.3	10.3	20.4
Two stages, 455 kc. (iron core) . .	2.9	6.4	10.8
Two stages, 1600 kc.	11.0	16.6	27.4
Two stages, 5000 kc.	25.8	46.0	100.0

Tubes for i.f. amplifiers — Variable- μ pentodes (§ 3-5) are almost invariably used in i.f. amplifier stages, since grid-bias gain control (§ 7-6) is practically always applied to the i.f. amplifier. Tubes with high plate resistance will have least effect on the selectivity of the amplifier, and those with high mutual conductance will give greatest gain. The choice of i.f. tubes has practically no effect on the signal-to-noise ratio, since this is determined by the preceding mixer and r.f. amplifier (if the latter is used).

When single-ended tubes (§ 3-5) are used, care should be taken to keep the plate and grid leads well separated. With these tubes it is advisable to mount the screen by-pass condenser directly on the bottom of the socket, cross-wise between the plate and grid pins, to provide additional shielding. The outside foil of the condenser should be connected to ground.

Single-signal effect — In heterodyne c.w. reception with a superheterodyne receiver, the beat oscillator is set to give a suitable audio-frequency beat note when the incoming signal is converted to the intermediate frequency. For example, the beat oscillator may be set to 456 kc. (the i.f. being 455 kc.) to give a 1000-cycle beat note. Now, if an interfering signal appears at 457 kc., it will also be heterodyned by the beat oscillator to produce a 1000-cycle beat. This *audio-frequency image* corresponds to the high-frequency images already discussed (§ 7-8). It can be reduced by providing enough i.f. selectivity, since the image signal is off the peak of the i.f. resonance curve.

When this is done, tuning through a given signal will show a strong response at the desired beat note on one side of zero beat only, instead of the two beat notes on either side of zero beat characteristic of less-selective reception; hence the name, "single-signal" reception.

The necessary selectivity is difficult to obtain with non-regenerative amplifiers using ordinary tuned circuits unless a very low intermediate frequency or a large number of circuits is used. In practice it is secured either by regenerative amplification or by a crystal filter.

Regeneration — Regeneration can be used to give a pronounced single-signal effect, particularly when the i.f. is 455 kc. or lower. The resonance curve of an i.f. stage at critical regeneration (just below the oscillating point) is extremely sharp, a band-width of 1 kc. at 10 times down and 5 kc. at 100 times down being obtainable in one stage. The audio-frequency image of a given signal thus can be reduced by a factor of nearly 100 for a 1000-cycle beat note (image 2000 cycles from resonance).

Regeneration is easily introduced into an i.f. amplifier by providing a small amount of capacity coupling between grid and plate. Bringing a short length of wire, connected to the grid, into the vicinity of the plate lead usually will suffice. The feed-back may be controlled by the regular cathode-resistor gain control. When the i.f. is regenerative, it is usually preferable to operate the tube at reduced gain (high bias) and depend upon the regeneration to bring the signal strength back to normal. This prevents overloading on strong signals and increases the effective selectivity.

The higher selectivity with regeneration reduces the over-all response to noise generated in the earlier stages of the receiver, just as does high selectivity produced by other means, and therefore improves the signal-to-noise ratio. The disadvantage is that the regenerative gain varies with signal strength, being less on strong signals, and the selectivity varies accordingly.

Crystal filters — The most satisfactory method of obtaining high selectivity is by the use of a piezoelectric quartz crystal as a selective filter in the i.f. amplifier (§ 2-10). Compared to a good tuned circuit, the Q of such a crystal is extremely high. The dimensions of the crystal are made such that it is resonant at

the desired intermediate frequency. It is then used as a selective coupler between i.f. stages.

Fig. 724 gives a typical crystal-filter resonance curve. For single-signal reception, the audio-frequency image can be reduced by a factor of 1000 or more. Besides practically eliminating the a.f. image, the high selectivity of the crystal filter provides great discrimination against signals very close to the desired signal in frequency, and, by reducing the bandwidth, reduces the response of the receiver to noise both from sources external to the receiver and in the r.f. stages of the receiver itself.

Crystal filter circuits; phasing — Several crystal filter circuits are shown in Fig. 725. Those at A and B are practically identical in performance, although differing in details. The crystal is connected in a bridge circuit (§ 2-11), with the secondary side of T_1 , the input transformer, balanced to ground either through a pair of condensers, $C-C$, (A) or by a center-tap on the secondary, L_2 (B). The bridge is completed by the crystal, X , and the *phasing condenser*, C_2 , which has a maximum capacity somewhat higher than the capacity of the crystal in its holder. When C_2 is set to balance the crystal-holder capacity, the resonance curve of the crystal circuit is practically symmetrical; the crystal acts as a series-resonant circuit of very high Q and thus allows signals of the desired frequency to be fed through C_3 to L_3L_4 , the output transformer. Without C_2 , the holder capacity (with the crystal acting as a dielectric) would by-pass signals of undesired frequencies to the output circuit.

The phasing control has an additional function besides neutralization of the crystal-holder capacity. The holder capacity becomes a part of the crystal circuit and causes it to act as a parallel-tuned resonant circuit at a frequency

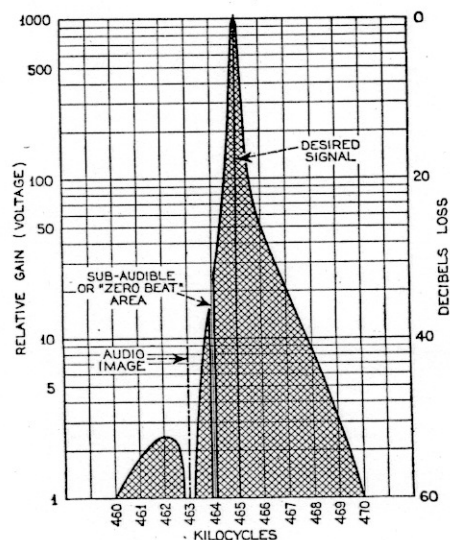


Fig. 724 — Graphical representation of single-signal selectivity. The shaded area indicates the overall band-width, or region in which response is obtainable.

slightly higher than its series-resonant frequency. Signals at the parallel-resonant frequency thus are prevented from reaching the output circuit. The phasing control, by varying the effect of the holder capacity, permits shifting the parallel-resonant frequency over a considerable range, providing adjustable rejection of interfering signals. The effect of rejection is illustrated in Fig. 724, where the audio image is reduced, by proper setting of the phasing control, far below the value that would be expected if the resonance curve were symmetrical.

Variable selectivity — In circuits such as A and B, Fig. 725, variable selectivity is obtained by adjustment of the variable input impedance, which is effectively in series with the crystal resonator. This is accomplished by varying C_1 (the *selectivity control*), which tunes the balanced secondary circuit of T_1 . When the secondary is tuned to i.f. resonance the parallel impedance of the L_2C_1 combination is maximum and is purely resistive (§ 2-10).

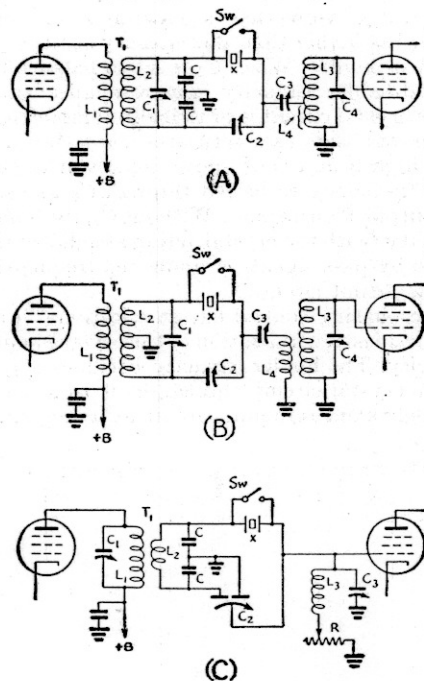


Fig. 725 — Crystal filter circuits of three types. All give variable band-width, with C having the greatest range of selectivity. Their operation is discussed in the text. Suitable circuit values are as follows: Circuit A, T_1 , special i.f. input transformer with high-inductance primary, L_1 , closely coupled to tuned secondary, L_2 ; C_1 , 50- μ fd. variable; C , each 100- μ fd. fixed (mica); C_2 , 10- to 15- μ fd. (max.) variable; C_3 , 50- μ fd. trimmer; L_3C_4 , i.f. tuned circuit, with L_3 tapped to match crystal-circuit impedance. In circuit B, T_1 is the same as in circuit A except that the secondary is center-tapped; C_1 is 100- μ fd. variable; C_2 , C_3 and C_4 , same as for circuit A; L_3L_4 is a transformer with primary, L_4 , corresponding to tap on L_3 in A. In circuit C, T_1 is a special i.f. input transformer with tuned primary and low-impedance secondary; C , each 100- μ fd. fixed (mica); C_2 , opposed stator phasing condenser, approximately 8 μ fd. maximum capacity each side; L_3C_3 , high- Q i.f. tuned circuit; R , 0 to 3000 ohms (selectivity control).

Since the secondary circuit is center-tapped, approximately one-fourth of this resistive impedance is in series with the crystal through C_3 and L_4 . This lowers the Q of the crystal circuit and makes its selectivity minimum. At the same time, the voltage applied to the crystal circuit is maximum.

When the input circuit is detuned from the crystal resonant frequency the resistance component of the input impedance decreases, and so does the total parallel impedance. Accordingly, the selectivity of the crystal circuit becomes higher and the applied voltage falls off. At first the resistance decreases faster than the applied voltage, with the result that the c.w. output from the filter *increases* as the selectivity is increased. The output falls off gradually as the input circuit is detuned further from resonance, and the selectivity becomes still higher.

In the circuits of A and B in Fig. 725, the minimum selectivity is still much greater than that of a normal two-stage 455-kc. amplifier and it is desirable to provide a wider range of selectivity, particularly for 'phone reception. A circuit which does this is shown at Fig. 725-C. The principle of operation is similar, but a much higher value of resistance can be introduced in the crystal circuit to reduce the selectivity. The output tuned circuit, L_3C_3 , must have high Q . A compensated condenser is used at C_2 (phasing) to maintain circuit balance, so that the phasing control does not affect the resonant frequency. The output circuit functions as a voltage divider in such a way that the amplitude of the carrier delivered to the next grid does not vary appreciably with the selectivity setting. The variable resistor, R , may consist of a series of separate fixed resistors selected by a tap switch.

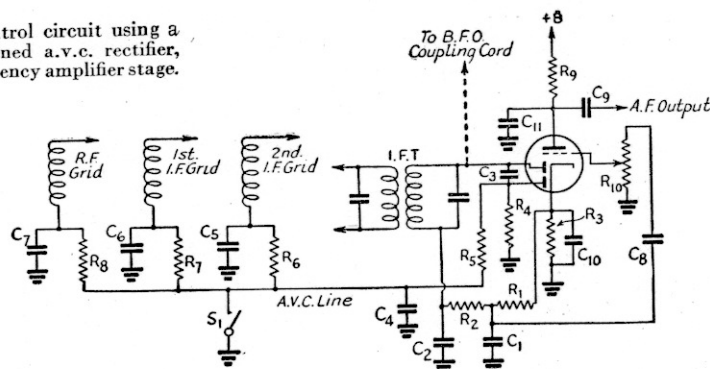
7-12 The Second Detector and Beat Oscillator

Detector circuits — The second detector of a superheterodyne receiver performs the same function as the detector in the simple receiver, but usually operates at a higher input level because of the relatively great r.f. amplification. Therefore, the ability to handle large signals without distortion is preferable to high sensitivity. Plate detection is used to some extent, but the diode detector is most popular. It is especially adapted to furnishing automatic gain or volume control (§ 7-13). The basic circuits are as described in § 7-3, although in many cases the diode elements are incorporated in a multi-purpose tube which also has an amplifier section in addition to the diode unit.

The beat oscillator — Any standard oscillator circuit (§ 3-7) may be used for the beat oscillator. Special beat-oscillator transformers are available, usually consisting of a tapped coil with adjustable tuning; these are most conveniently used with circuits such as those shown at Fig. 720-A and -B, with the output taken from Y . A variable condenser of about 25- μ fd. capacity may be connected between

Fig. 726 — Automatic volume control circuit using a duo-diode-triode tube as a combined a.v.c. rectifier, second detector and first audio-frequency amplifier stage.

- R_1 — 0.25 megohm.
 R_2 — 50,000 to 250,000 ohms.
 R_3 — 2000 ohms.
 R_4 — 2 to 5 megohms.
 R_5 — 0.5 to 1 megohm.
 R_6, R_7, R_8 — 0.25 megohm.
 R_9 — 0.25 megohm.
 R_{10} — 0.5-megohm variable.
 C_1, C_2, C_3 — 100 μ fd.
 C_4 — 0.1 μ fd.
 C_5, C_6, C_7 — 0.01 μ fd.
 C_8, C_9 — 0.01 to 0.1 μ fd.
 C_{10} — 5- to 10- μ fd. electrolytic.
 C_{11} — 250 μ fd.



cathode and ground to provide fine adjustment. The beat oscillator usually is coupled to the second-detector tuned circuit through a fixed condenser of a few μ fd. capacity.

The beat oscillator should be well shielded, to prevent coupling to any part of the circuit except the second detector and to prevent its harmonics from getting into the front end of the receiver and being amplified like regular signals. To this end, the plate voltage should be as low as is consistent with sufficient audio-frequency output. If the beat oscillator output is too low, strong signals will not give a proportionately strong audio response.

An oscillating second detector may be used to give the audio beat note, but, since the detector must be detuned from the i.f., the selectivity and signal strength will be reduced, while blocking (§ 7-4) will be pronounced because of the high signal level at the second detector.

7-13 Automatic Volume Control

Principles — Automatic regulation of the gain of the receiver in inverse proportion to the signal strength is a great advantage, especially in 'phone reception, since it tends to keep the output level of the receiver constant regardless of input signal strength. It is readily accomplished in superheterodyne receivers by using the average rectified d.c. voltage, developed by the received signal across a resistance in a detector circuit (§ 7-3), to vary the bias on the r.f. and i.f. amplifier tubes. Since this voltage is proportional to the average amplitude of the signal, the gain is reduced as the signal strength becomes greater. The control will be more complete as the number of stages to which the a.v.c. bias is applied is increased. Control of at least two stages is advisable.

Circuits — A typical circuit using a diode-triode type tube as a combined a.v.c. rectifier, detector and first audio amplifier is shown in Fig. 726. One plate of the diode section of the tube is used for signal detection and the other for a.v.c. rectification. The a.v.c. diode plate is fed from the detector diode through the small coupling condenser, C_3 . Negative bias resulting from the flow of rectified carrier current is developed across R_4 , the diode load re-

sistor. This negative bias is applied to the grids of the controlled stages through the filtering resistors (§ 2-11), R_5, R_6, R_7 and R_8 . When S_1 is closed the a.v.c. line is grounded, thereby removing a.v.c. bias from the amplifier.

It does not matter which of the two diode plates is selected for audio and which for a.v.c. Frequently the two plates are connected together and used as a combined detector and a.v.c. rectifier. This could be done in Fig. 726. The a.v.c. filter and line would connect to the junction of R_2 and C_2 , while C_3 and R_4 would be omitted from the circuit.

Delayed a.v.c. — In Fig. 726 the audio diode return is made directly to the cathode and the a.v.c. diode return to ground. This places negative bias on the a.v.c. diode equal to the d.c. drop through the cathode resistor (a volt or two) and thus delays the application of a.v.c. voltage to the amplifier grids, since no rectification takes place in the a.v.c. diode circuit until the carrier amplitude is large enough to overcome the bias. Without this delay the a.v.c. would start working even with a very small signal. This is undesirable, because the full amplification of the receiver then could not be realized on weak signals. In the audio diode circuit this fixed bias would cause distortion, and must be avoided; hence, the return is made directly to the cathode.

Time constant — The time constant (§ 2-6) of the resistor-condenser combinations in the a.v.c. circuit is an important part of the system. It must be high enough so that the modulation on the signal is completely filtered from the d.c. output, leaving only an average d.c. component which follows the relatively slow carrier variations with fading. Audio-frequency variations in the a.v.c. voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal, and in practice would cause frequency distortion. On the other hand, the time constant must not be too great or the a.v.c. would be unable to follow rapid fading. The capacity and resistance values indicated in Fig. 726 give a time constant satisfactory for high-frequency reception.

Signal-strength and tuning indicators — A useful accessory to the receiver is an indicator which will show relative signal strength.

Not only is it an aid in giving reports to transmitting stations, but it is helpful also in aligning the receiver circuits, in conjunction with a test oscillator or other steady signal.

Three types of indicators are shown in Fig. 727. That at A uses an electron-ray tube (§ 3-5), several types of which are available. The grid of the triode section usually is connected to the a.v.c. line. The particular type of tube used depends upon the voltage available for its grid; where the a.v.c. voltage is large, a remote cut-off type (6G5 or 6N5) should be used in preference to the sharp cut-off type (6E5).

In B, a milliammeter is connected in series with the d.c. plate lead to one or more r.f. and i.f. tubes, the grids of which are controlled by a.v.c. voltage. Since the plate current of such tubes varies with the strength of the incoming signal, the meter will indicate relative signal intensity and may be calibrated in "S" points. The scale range of the meter should be chosen to fit the number of tubes in use; the maximum plate current of the average remote cut-off r.f. pentode is from 7 to 10 milliamperes. The shunt resistor, R , enables setting the plate current to the full-scale value ("zero adjustment"). With this system the ordinary meter reads downwards from full scale with increasing signal strength, which is the reverse of normal pointer movement (clockwise with increasing reading). Special instruments with the zero-current position of the pointer on the right-hand side of the scale are used in commercial receivers.

The system at C uses a 0-1 ma. milliammeter in a bridge circuit, arranged so that the meter reading and the signal strength increase together. The current through the branch containing R_1 should be approximately equal to the current through that containing R_2 . In some manufactured receivers this is brought about by draining the screen voltage-divider current and the current to the screens of three r.f. pentodes (r.f. and i.f. stages) through R_2 , the sum of these currents being about equal to the maximum plate current of one a.v.c.-controlled tube. Typical values for this type of circuit are given. The sensitivity can be increased by increasing the resistance of R_1 , R_2 and R_3 . The initial setting is made with the manual gain control set near maximum, when R_3 should be adjusted to make the meter read zero with no signal.

§ 7-14 Preselection

Purpose — Preselection is added signal-frequency selectivity incorporated before the mixer stage is reached. An r.f. amplifier preceding the mixer generally is called a *preselector*, its purpose, in part at least, being to discriminate in favor of the signal against the image. The preselector may consist of one or more r.f. amplifier stages. When its tuning control is ganged with those of the mixer and oscillator, its circuits must track with the mixer circuit.

The circuit is the same as discussed earlier (§ 7-6). An external preselector stage may be

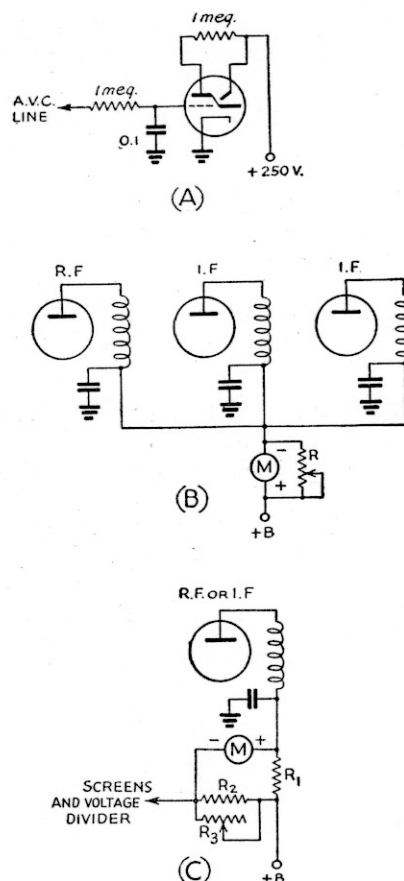


Fig. 727 — Tuning indicator or "S"-meter circuits for superhet receivers. A, electron-ray indicator; B, plate-current meter for tubes on a.v.c.; C, bridge circuit for a.v.c.-controlled tube. In B, resistor R should have a maximum resistance several times that of the milliammeter. In C, representative values for the components are: R_1 , 250 ohms; R_2 , 350 ohms; R_3 , 1000-ohm variable.

used with receivers having inadequate image ratios. In this case it is built as a separate unit, often with a tuned output circuit which gives a further improvement in selectivity. The output circuit usually is link-coupled (§ 2-11) to the receiver.

Signal/noise ratio — An r.f. amplifier will have a better signal-to-noise ratio (§ 7-2) than a mixer because the gain is higher and because the mixer-tube electrode arrangement results in higher internal tube noise than does the ordinary pentode structure. Hence, a preselector is advantageous in increasing the signal-to-noise ratio over that obtainable when the mixer is fed directly from the antenna.

Image suppression — The image ratios (§ 7-8) obtainable at frequencies up to and including 7 Mc. with a single preselector stage are high enough, when the intermediate frequency is 455 kc., so that for all practical purposes there is no appreciable image response. Average image ratios on 14 Mc. and 28 Mc. are 50-75 and 10-15, respectively. This is the overall selectivity of the r.f. and mixer tuned cir-

cuits. A second preselector stage, adding another tuned circuit, will increase the ratios to several hundred at 14 Mc. and to 30-40 at 28 Mc.

On very-high frequencies, it is impracticable to attempt to secure a good image ratio with a 455-kc. i.f. Good performance can be secured only by using a high i.f. or a double superheterodyne (§ 7-8) with a high-frequency first i.f.

Regeneration—Regeneration may be used in a preselector stage to increase both gain and selectivity. Since its use makes tuning more critical and increases ganging problems, regeneration is seldom employed except at 14 Mc. and above, where adequate image suppression is difficult to obtain with non-regenerative circuits. The same disadvantages exist as in the case of a regenerative i.f. amplifier (§ 7-11). The effect of regeneration is roughly equivalent to adding another non-regenerative preselector stage.

Regeneration may be introduced by the same method as used in regenerative i.f. amplifiers (§ 7-11). The manual gain control of the stage will serve as a volume control.

Regeneration in a preselector does not improve the signal-to-noise ratio, since the tube noise is fed back to the grid circuit along with the signal to add to the thermal-agitation noise originally present. This noise also is amplified.

7-15 Noise Reduction

Types of noise—In addition to tube and circuit noise (§ 7-6), much of the noise interference experienced in reception of high-frequency signals is caused by domestic electrical equipment and by automobile ignition systems. The interference is of two types in its effects. The first is the "hiss" type, consisting of overlapping pulses similar in nature to the receiver noise. It is largely reduced by high selectivity in the receiver, especially for code reception. The second is the "pistol-shot" or "machine-gun" type, consisting of separated impulses of high amplitude. The "hiss" type of interference usually is caused by commutator sparking in d.c. and series-wound a.c. motors, while the "shot" type results from separated spark discharges (a.c. power leaks, switch and key clicks, ignition sparks, and the like).

Impulse noise—Impulse noise, because of the extremely short duration of the pulses as compared to the time between them, must have high pulse amplitude to contain much average energy. Hence, noise of this type strong enough to cause much interference generally has an instantaneous amplitude much higher than that of the signal being received. The general

principle of devices intended to reduce such noise is that of allowing the signal amplitude to pass through the receiver unaffected, but making the receiver inoperative for amplitudes greater than that of the signal. The greater the amplitude of the pulse compared to its time of duration, the more successful the noise reduction, since more of the energy can be suppressed.

In passing through selective receiver circuits, the time duration of the impulses is increased, because of the Q or flywheel effect (§ 2-10) of the circuits. Hence, the more selectivity ahead of the noise-reducing device, the more difficult it becomes to secure good noise suppression.

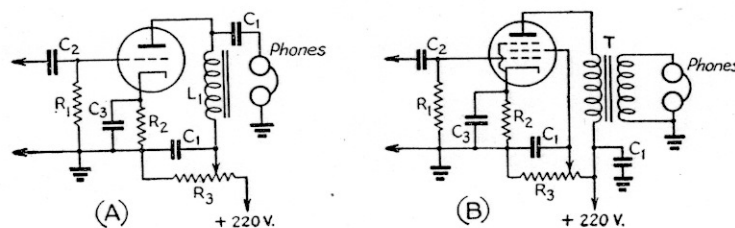
Audio limiting—A considerable degree of noise reduction in code reception can be accomplished by amplitude-limiting arrangements applied to the audio output circuit of a receiver. Such limiters also maintain the signal output nearly constant with fading. Diagrams of typical output-limiter circuits are shown in Fig. 728. Circuit A employs a triode tube operated at reduced plate voltage (approximately 10 volts), so that it saturates at a low signal level. The arrangement of B has better limiting characteristics. A pentode audio tube is operated at reduced screen voltage (35 volts or so), so that the output power remains practically constant over a grid excitation-voltage range of more than 100 to 1. These output-limiter systems are simple, and adaptable to most receivers. However, they cannot prevent noise peaks from overloading previous circuits.

Second-detector circuits—The circuit of Fig. 729 "chops" noise peaks at the second detector of a superhet receiver by means of a biased diode, which becomes non-conducting above a predetermined signal level. The audio output of the detector must pass through the diode to the grid of the amplifier tube. The diode normally would be non-conducting with the connections shown were it not for the fact that it is given positive bias from a 30-volt source through the adjustable potentiometer, R_3 . Resistors R_1 and R_2 must be fairly large in value to prevent loss of audio signal.

The audio signal from the detector can be considered to modulate (§ 5-1) the steady diode current, and conduction will take place so long as the diode plate is positive with respect to the cathode. When the signal is sufficiently large to swing the cathode positive with respect to the plate, however, conduction ceases, and that portion of the signal is cut off from the audio amplifier. The point at which cut-off occurs can be selected by adjustment of

Fig. 728 — Audio output-limiting noise-reducing circuits.

C_1 — 0.25 μ fd.
 C_2 — 0.01 μ fd.
 C_3 — 5 μ fd.
 R_1 — 0.5 megohm.
 R_2 — 2000 ohms.
 R_3 — 50,000-ohm potentiometer.
 T — Output transformer.
 L_1 — 15-henry choke.



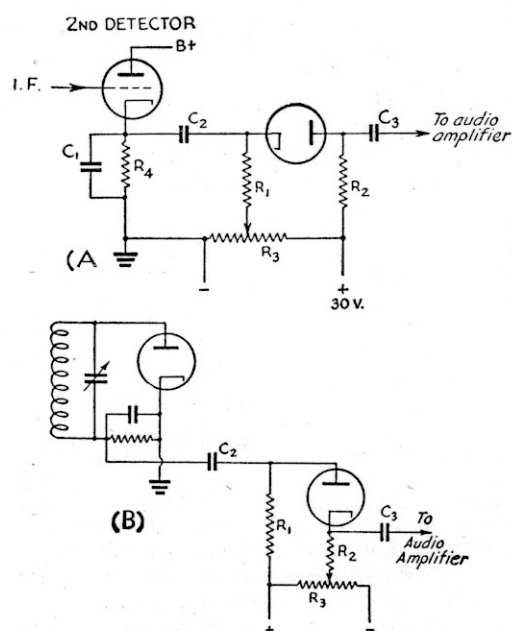


Fig. 729—Series-valve noise-limiter circuits. A, as used with an infinite-impedance detector; B, with a diode detector. Typical values for components are as follows:

R_1 — 0.25 megohm.	R_4 — 20,000 to 50,000 ohms.
R_2 — 50,000 ohms.	C_1 — 250 μ fd.
R_3 — 10,000-ohms.	C_2, C_3 — 0.1 μ fd.

All other diode-circuit constants in B are conventional.

R_3 . By setting R_3 so that the signal just passes through the "valve," noise pulses higher in amplitude than the signal will be cut off. The circuit of Fig 729-A, using an infinite-impedance detector (§ 7-3), gives a positive voltage on rectification. When the rectified voltage is negative, as it is from the usual diode detector (§ 7-3), the circuit arrangement shown in Fig. 729-B must be used.

An audio signal of about ten volts is required for good limiting action. When a beat oscillator is used for c.w. reception the b.f.o. voltage should be small, so that incoming noise will not have a strong carrier to beat against and so produce large audio output.

A second-detector noise-limiting circuit which automatically adjusts itself to the received carrier level is shown in Fig. 730. The diode load circuit (§ 7-3) consists of R_6 , R_7 , R_8 (shunted by the high-resistance audio volume control, R_4) and R_5 in series. The cathode of the 6N7 noise limiter is tapped on the load resistor at a point such that the average rectified carrier voltage (negative) at its grid is approximately twice the negative voltage at the cathode, both measured with reference to ground. A filter network, R_1C_1 , is inserted in the grid circuit, so that the audio modulation on the carrier does not reach the grid; hence, the grid potential is maintained at substantially the rectified carrier voltage alone. The cathode, however, is free to follow the modulation, and when the modulation is 100 per cent

the peak cathode voltage will just equal the steady grid voltage.

At all modulation percentages below 100 per cent the grid is negative with respect to cathode, and current cannot flow in the 6N7 plate-cathode circuit. A noise pulse exceeding the peak voltage which represents 100 per cent modulation will, however, make the grid positive with respect to cathode. The relatively low plate-cathode resistance of the 6N7 then shunts the high-resistance audio output circuit, effectively short-circuiting it, so that there is practically no response for the duration of the noise peak over the 100 per cent modulation limit.

R_5 is used to make the noise-limiting tube more sensitive by applying to the plate an audio voltage out of phase with the cathode voltage, so that, at the instant the grid goes positive with respect to cathode, the highest positive potential also is applied to the plate, thus further lowering the effective plate-cathode resistance.

I.f. noise silencer — In the circuit shown in Fig. 731, noise pulses are made to decrease the gain of an i.f. stage momentarily and thus silence the receiver for the duration of the pulse. Any noise voltage in excess of the desired signal's maximum i.f. voltage is taken off at the grid of the i.f. amplifier, amplified by the noise amplifier stage, and rectified by the full-wave diode noise rectifier. The noise circuits are tuned to the i.f. The rectified noise voltage is applied as a pulse of negative bias to the No. 3 grid of the 6L7 i.f. amplifier, wholly or partially disabling this stage for the duration of the individual noise pulse, depending on the amplitude of the noise voltage. The noise amplifier-rectifier circuit is biased by means of the "threshold control," R_2 , so that rectification will not start until the noise voltage exceeds the desired-signal amplitude. For reception with automatic volume control, the a.v.c. voltage can be applied to the grid of the noise amplifier, to augment this threshold bias. This system of noise silencing gives a signal-to-noise ratio improvement of the order of 30 db. (power ratio of 1000) with heavy ignition interference, raising the signal-to-noise ratio from -10 db. without the silencer to +20 db. with the silencer in a typical instance.

Circuit values are normal for i.f. amplifiers (§ 7-11), except as indicated. The noise-rectifier transformer, T_1 , has an untuned secondary closely coupled to the primary and center-tapped for full-wave rectification. The center-tap rectifier (§ 8-3) is used to reduce the possibility of r.f. feed-back into the i.f. amplifier (noise-silencer) stage. The time constant (§ 2-6) of the noise-rectifier load circuit, $R_1C_1C_2$, must be small, to prevent disabling the noise-silencer stage for a longer period than the duration of the noise pulse. The r.f. choke, RFC , must be effective at the intermediate frequency.

Adequate shielding and isolation of the noise-amplifier and rectifier circuits from the noise-silencer stage must be provided to prevent

possible self-oscillation and instability. This circuit should be applied to the first i.f. stage of the receiver, before the high-selectivity circuits are reached, and is most effective when the signal and noise levels are fairly high (one or two r.f. stages before the mixer) since several volts must be obtained from the noise rectifier for good silencing.

7-16 Operating Superheterodyne Receivers

C.w. reception — For making code signals audible, the beat oscillator should be set to a frequency slightly different from the intermediate frequency (§ 7-8). To adjust the beat-oscillator frequency, first tune in a moderately weak but steady carrier with the beat oscillator turned off. Adjust the receiver tuning for maximum signal strength, as indicated by maximum hiss. Then turn on the beat oscillator and adjust its frequency (leaving the receiver tuning unchanged) to give a suitable beat note. The beat oscillator need not subsequently be touched, except for occasional checking to make certain the frequency has not drifted from the initial setting. The b.f.o. may be set on either the high- or low-frequency side of zero beat.

The use of a.v.c. (§ 7-13) is not generally satisfactory in c.w. reception because the receiver gain rises in the spaces between the dots and dashes, giving an increase in noise in the same intervals, and because the rectified beat-oscillator voltage in the second detector circuit also operates the a.v.c. circuit. This gives a constant reduction in gain and prevents utilization of the full sensitivity of the receiver.

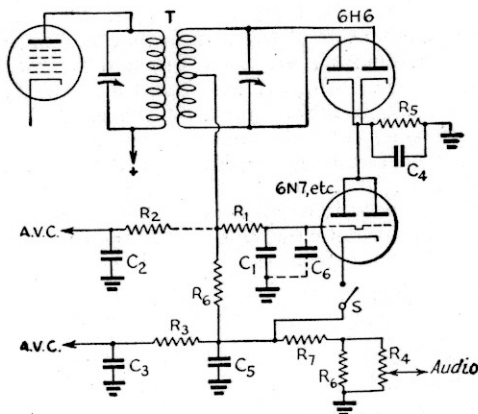


Fig. 730 — Automatic noise-limiter for superheterodynes.

T — I.f. transformer with a balanced secondary for working into a diode rectifier.

R₁, R₂, R₃ — 1 megohm. C₁ — 0.1-μfd. paper.
R₄ — 1-megohm variable. C₂, C₃ — 0.05-μfd. paper.
R₅ — 250,000 ohms. C₄, C₅ — 50-μfd. mica.
R₆, R₈ — 100,000 ohms. C₆ — 0.001-μfd. mica (for
R₇ — 25,000 ohms. r.f. filtering, if
Sw — S.p.s.t. toggle (on-off switch). needed).

The switch should be mounted close to the circuit elements and controlled by an extension shaft if necessary.

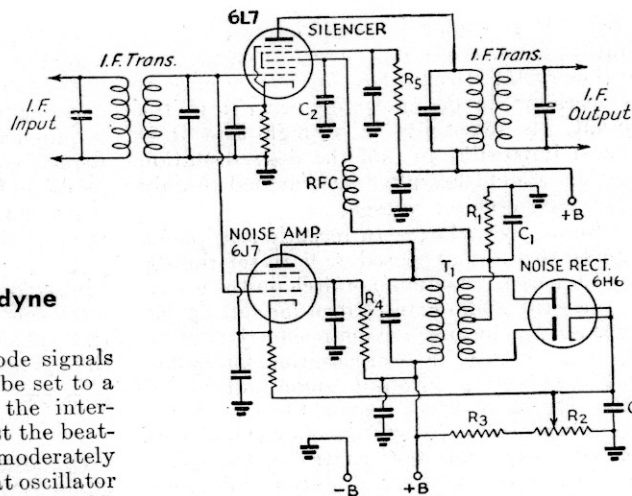


Fig. 731 — I.f. noise-silencing circuit. The plate supply should be 250 volts. Typical values for components are:
C₁ — 50–250 μμfd. (use smallest value possible without r.f. feedback).

C₂ — 50 μμfd. R₂ — 5000-ohm variable.
C₃ — 0.1 μfd. R₃ — 20,000 ohms.
R₁ — 0.1 megohm. R₄, R₅ — 0.1 megohm.
T₁ — Special i.f. transformer for noise rectifier.

Hence, the gain preferably should be manually adjusted to give suitable audio-frequency output.

To avoid overloading in the i.f. circuits, it is usually better to control the i.f. and r.f. gain and keep the audio gain at a fixed value than to use the a.f. gain control as a volume control and leave the r.f. gain fixed at its highest level.

Tuning with the crystal filter — If the receiver is equipped with a crystal filter the tuning instructions in the preceding paragraph still apply, but more care must be used both in the initial adjustment of the beat oscillator and in tuning. The beat oscillator is set as described above, but with the crystal filter in operation and adjusted to its sharpest position, if variable selectivity is available. The initial adjustment should be made with the phasing control (§ 7-11) in the intermediate position. After it is completed, the beat oscillator should be left set and the receiver tuned to the other side of zero beat (audio-frequency image) on the same carrier to give a beat note of the same tone. This beat will be considerably weaker than the first, and may be "phased out" almost completely by careful adjustment of the phasing control. This is the adjustment for normal operation; it will be found that one side of zero beat has practically disappeared, leaving maximum response on the desired side.

An interfering signal having a beat note differing from that of the a.f. image can be similarly phased out, provided its carrier frequency is not too near the desired carrier.

Depending upon the filter design, maximum selectivity may cause the dots and dashes to lengthen out so that they seem to "run together." This, plus the fact that tuning is quite critical with extremely high selectivity, may make it desirable to use somewhat less

selectivity in ordinary operation. However, it must be emphasized that, to realize the benefits of the crystal filter in reducing interference, it is necessary to do *all* tuning with it in the circuit. Its selectivity is so high that it is almost impossible to find the desired station quickly, should the filter be switched in only when interference is present.

'Phone reception — In reception of 'phone signals, the normal procedure is to set the r.f. and i.f. gain at maximum, switch on the a.v.c., and use the audio gain control for setting the volume. This insures maximum effectiveness of the a.v.c. system in compensating for fading and maintaining constant audio output on either strong or weak signals. On occasion a strong signal close to the frequency of a weaker desired station may take control of the a.v.c., in which case the weaker station will practically disappear because of the reduced gain. In this case better reception may result if the a.v.c. is switched off, using the manual r.f. gain control to set the gain at a point which prevents "blocking" by the stronger signal.

A crystal filter will do much toward reducing interference in 'phone reception. Although the high selectivity cuts sidebands (§ 7-11) and thereby reduces the audio output, especially at the higher audio frequencies, it is possible to use quite high selectivity without destroying intelligibility even though the "quality" of the transmission may suffer. As in the case of c.w. reception, it is advisable to do all tuning with the filter in the circuit. Variable-selectivity filters permit a choice of selectivity to suit interference conditions.

An undesired carrier close in frequency to a desired carrier will heterodyne with it to produce a beat note equal to the frequency difference. Such a heterodyne can be reduced by adjustment of the phasing control in the crystal filter. It cannot be prevented in a "straight" superheterodyne having no crystal filter.

A tone control often will be of help in reducing the effects of high-pitched heterodynes, sideband splatter (§ 5-2) and noise, by cutting off the higher audio frequencies. This, like sideband cutting with high selectivity, causes some reduction in naturalness.

Spurious responses — Spurious responses can be recognized without a great deal of difficulty. Often it is possible to identify an image by the nature of the transmitting station, if the frequency assignments applying to the frequency to which the receiver is tuned are known. However, an image also can be recognized by its behavior with tuning. If the signal causes a heterodyne beat note with the desired signal and is actually on the same frequency, the beat note will not change as the receiver is tuned through the signal; but if the interfering signal is an image, the beat will vary in pitch as the receiver is tuned. The beat oscillator in the receiver must be turned off for this test. Using a crystal filter with the beat oscillator on, an image will peak on the side of

zero beat opposite that on which the desired signal peaks.

Harmonic response can be recognized by the "tuning rate," or movement of the tuning dial required to give a specified change in beat note. Signals getting into the i.f. via high-frequency oscillator harmonics tune more rapidly (less dial movement) through a given change in beat note than do signals received by normal means.

Harmonics of the beat oscillator can be recognized by the tuning rate of the beat-oscillator pitch control. A smaller movement of the control will suffice for a given change in beat note than is necessary with legitimate signals.

7-17 Servicing Superheterodyne Receivers

I.f. alignment — A calibrated signal generator or test oscillator is a practical necessity for initial alignment of an i.f. amplifier. Some means for measuring the output of the receiver also is needed. If the receiver has a tuning meter, its indications will serve for this purpose. Alternatively, if the signal generator is of the modulated type, an a.c. output meter (high-resistance voltmeter with copper-oxide rectifier) can be connected across the primary of the output transformer, or from the plate of the last audio amplifier through a 0.1- μ fd. blocking condenser (§ 2-13) to the receiver chassis. The intensity of sound from the loudspeaker can be judged by ear, if no output meter is available, but this method is not as accurate as those using instruments.

The procedure is as follows: The test oscillator is adjusted to the desired intermediate frequency, and the "hot" or ungrounded output lead is clipped on the grid terminal of the last i.f. amplifier tube. The grounded lead is connected to the receiver chassis. The trimmer condensers of the transformer feeding the second detector are then adjusted for maximum signal output. The hot lead from the generator is next clipped on the grid of the next-to-last i.f. tube, and the second from last i.f. transformer is brought into alignment by adjusting its trimmers for maximum output. This process is continued, working back from the second detector, until all of the i.f. transformers have been aligned. It will be necessary to reduce the output of the signal generator as more of the i.f. amplifier is brought into use, because the increased gain otherwise may cause overloading and consequent inaccurate results. It is desirable always to use the minimum signal strength which gives useful output readings.

The i.f. transformer in the plate circuit of the mixer is aligned with the signal-generator output lead connected to the mixer grid. Since the tuned circuit feeding the mixer grid is tuned to a considerably higher frequency, it can effectively short-circuit the signal-generator output, and therefore it may be necessary to disconnect this circuit. With tubes having a top grid-cap connection, this can be done by simply removing the grid clip from the tube cap.

If the tuning indicator is used as an output meter the a.v.c. should be on; if the audio-output method is used, the a.v.c. should be off. The beat oscillator should be off in either case.

If the i.f. amplifier has a crystal filter, the filter should be switched out. Alignment is then carried out as described above, setting the signal generator as closely as possible to the frequency of the crystal. After alignment, the crystal should be switched in and the oscillator frequency varied back and forth over a small range either side of the crystal frequency to find its exact frequency, which will be indicated by a sharp rise in output. Leaving the signal generator set on the crystal peak, the i.f. trimmers may be realigned for maximum output. The necessary readjustment should be small. The signal generator frequency should be checked frequently, to make sure it has not drifted from the crystal peak.

A modulated signal is not of much value for aligning a crystal-filter i.f. amplifier, since the high selectivity cuts sidebands and the results may be inaccurate if the audio output of the receiver is used as a criterion of alignment. Lacking an a.v.c. tuning meter the transformers may be aligned by ear, using a weak unmodulated signal adjusted to the crystal peak. Switch on the beat oscillator, adjust to a suitable tone, and align the transformers for maximum audio output.

An amplifier which is only slightly out of alignment, as a result of normal drift from temperature, humidity or aging effects, can be realigned by using any steady signal, such as a local broadcasting station, in lieu of a test oscillator. Allow the receiver to warm up thoroughly (an hour or so), tune in the signal as usual, and "touch up" the i.f. trimmers for maximum output.

R.f. alignment — The objective in aligning the r.f. circuits in a gang-tuned receiver is to secure adequate tracking over each tuning range. The adjustment may be carried out with a test oscillator of suitable frequency range, or even on noise or such signals as may be heard. First set the tuning dial at the high-frequency end of the range in use. Then set the test oscillator to the frequency indicated by the receiver dial. The test-oscillator output may be connected to the antenna terminals of the receiver for this test. Adjust the oscillator trimmer condenser in the receiver to give maximum response on the test-oscillator signal, then reset the receiver dial to the low-frequency end of the range. Set the test-oscillator frequency near the frequency indicated by the receiver dial and carefully tune the test oscillator until its signal is heard in the receiver. If the frequency of the signal as indicated by the test-oscillator calibration is higher than that indicated by the receiver dial, more inductance (or more capacity in the tracking condenser) is needed in the receiver oscillator circuit; if the frequency is lower, less inductance (less tracking capacity) is required in the receiver oscil-

lator. Most commercial receivers provide some means for varying the inductances of the coils or the capacity of the tracking condenser, to permit aligning the receiver tuning with the dial calibration. Set the test oscillator to the frequency indicated by the receiver dial, and then adjust the tracking capacity or inductance of the receiver oscillator coil to obtain maximum response. After making this adjustment, recheck the high-frequency end of the scale as previously described. It may be necessary to go back and forth between the ends of the range several times before the proper combination of inductance and capacity is secured. In many cases, better over-all tracking will result if frequencies near but not actually at the ends of the tuning range are selected, instead of taking the extreme dial settings.

After the oscillator range is properly adjusted, set the receiver and test oscillator to the high-frequency end of the range. Adjust the mixer trimmer condenser for maximum hiss or signal, then the r.f. trimmers. Reset the tuning dial and test oscillator to the low-frequency end of the range, and repeat; if the circuits are properly designed, no change in trimmer settings should be necessary. If it is necessary to increase the trimmer capacity in any circuit, more inductance is needed; if less capacity resonates the circuit, less inductance is required.

Tracking seldom is perfect throughout a tuning range, so that a check of alignment at intermediate points in the range may show it to be slightly off. Normally the gain variation from this cause will be small, however, and it will suffice to bring the circuits into line at both ends of the range. If most reception is in a particular part of the range, such as an amateur band, the circuits may be aligned for maximum performance in that region, even though the ends of the frequency range as a whole may be slightly out of alignment.

Oscillation in r.f. or i.f. amplifiers — Oscillation in high-frequency amplifier and mixer circuits may be evidenced by squeals or "birdies" as the tuning is varied, or by complete lack of audible output if the oscillation is strong enough to cause the a.v.c. system to reduce the receiver gain drastically. Oscillation can be caused by poor connections in the common ground circuits, especially to the tuning-condenser rotors. Inadequate or defective bypass condensers in cathode, plate and screen-grid circuits also can cause such oscillation. In some cases it may be advisable to provide a shield between the stators of pre-r.f. amplifier and first-detector ganged tuning condensers, in addition to the usual tube and interstage shielding. A metal tube with an ungrounded shell will cause trouble. Improper screen-grid voltage, resulting from a shorted or too-low screen-grid series resistor, also may be responsible.

Oscillation in the i.f. circuits is independent of high-frequency tuning, and is indicated by a continuous squeal which appears when the gain is advanced with the c.w. beat oscillator

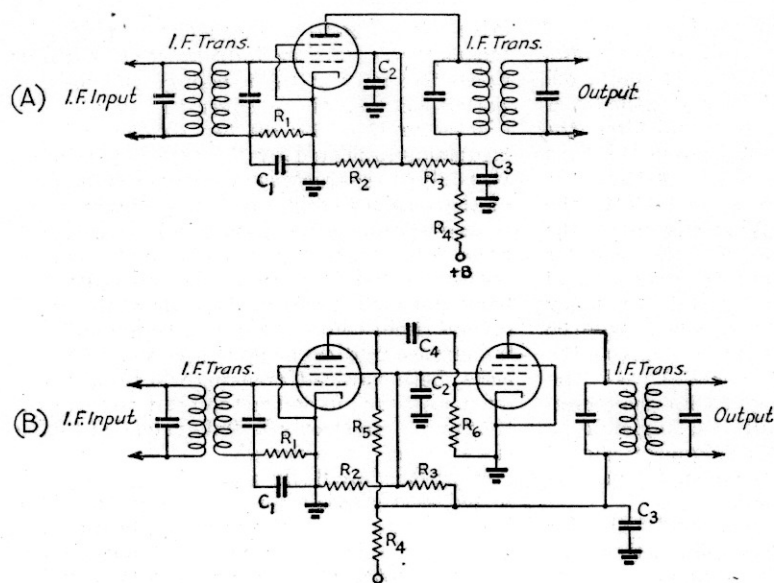


Fig. 732 — F.m. limiter circuits. A, single-tube plate-saturation limiter; B, cascade limiter. Typical values are:

	Circuit A	Circuit B
C ₁ —	100 μ fd.	100 μ fd.
C ₂ , C ₃ —	0.1 μ fd.	0.1 μ fd.
C ₄ —		250 μ fd.
R ₁ —	0.1 megohm.	50,000 ohms.
R ₂ —	2000 ohms.	2000 ohms.
R ₃ —	50,000 ohms.	50,000 ohms.
R ₄ —	0 to 50,000 ohms.	0 to 50,000 ohms.
R ₅ —		4000 ohms.
R ₆ —		0.2 megohm.

Plate-supply voltage should be 250 volts in both circuits.

on. It can result from similar defects in i.f. amplifier circuits. Inadequate cathode by-pass capacitance is a common cause of such oscillation. An additional by-pass condenser of 0.1 to 0.25 μ fd. usually will remedy it. Similar treatment can be applied to the screen-grid and plate by-passes of i.f. tubes.

Instability — “Birdies” or a mushy hiss occurring with tuning of the high-frequency oscillator may indicate that the oscillator is “squegging” or oscillating simultaneously at high and low frequencies (§ 7-4). This may be caused by a defective tube, too-high oscillator plate or screen-grid voltage, excessive feedback, or too-high grid-leak resistance.

A varying beat note in c.w. reception indicates instability in either the h.f. oscillator or beat oscillator, usually the former. The stability of the beat oscillator can be checked by introducing a signal of intermediate frequency (from a test oscillator) into the i.f. amplifier; if the beat note is unstable, the trouble is in the beat oscillator. Poor connections or defective parts are the likely cause. Instability in the high-frequency oscillator may be the result of poor circuit design (§ 7-10), loose connections, defective tubes or circuit components, or poor voltage regulation in the oscillator plate and/or screen supply circuits. Mixer pulling of the oscillator circuit (§ 7-9) also will cause the

beat-note to “chirp” on strong c.w. signals, because the oscillator load changes slightly under these conditions.

In ‘phone reception with a.v.c., a peculiar type of instability (“motorboating”) may appear if the h.f. oscillator frequency is sensitive to changes in plate voltage. As the a.v.c. voltage rises the electrode currents of the controlled tubes decrease, decreasing the load on the power supply and causing its output voltage to rise. Since this increases the voltage applied to the oscillator, its frequency changes correspond-

ingly, throwing the signal off the peak of the i.f. resonance curve and reducing the a.v.c. voltage, thus tending to restore the original conditions. The process then repeats itself, at a rate determined by the signal strength and the time constant of the power-supply circuits. This effect is most pronounced with high i.f. selectivity, as when a crystal filter is used, and can be cured by designing the oscillator circuit to be relatively insensitive to plate voltage changes and by regulating the voltage applied to the oscillator (§ 7-10).

§ 7-18 Reception of Frequency-Modulated Signals

F.m. receivers — A frequency-modulation receiver differs in circuit design from one designed for amplitude modulation chiefly in the arrangement used for detecting the signal. Detectors for amplitude-modulated signals do not respond to frequency modulation. It is also necessary, for full realization of the noise-reducing benefits of the f.m. system, that the signal applied to the detector be completely free from amplitude modulation. In practice, this is attained by preventing the signal from rising above a given amplitude by means of a limiter (§ 7-15). Since the weakest signal must be amplitude-limited, high gain must be provided ahead of the limiter; the superheterodyne type of circuit almost invariably is used to provide the necessary gain.

The r.f. and i.f. stages in a superheterodyne for f.m. reception are practically identical in circuit arrangement with those in an a.m. receiver. Since the use of f.m. is confined to the very-high frequencies (above 28 Mc.) a high intermediate frequency is employed, usually between 4 and 5 Mc. This not only reduces image response but also provides the greater band-width necessary to accommodate wide-band frequency-modulated signals.

Receiver requirements — The primary requirements are sufficient r.f. and i.f. gain to "saturate" the limiter even with a weak signal, sufficient band-width (§ 7-2) to accommodate the full frequency deviation either side of the carrier frequency without undue attenuation at the edges of the band, a limiter circuit which functions properly on both rapid and slow variations in amplitude, and a detector which gives a linear relationship between frequency deviation and amplitude output. The audio circuits are the same as in other receivers (§ 7-5), except that it is desirable to cut off the upper audio range by a low-pass filter (§ 2-11) because higher-frequency noise components have the greatest amplitude in an f.m. receiver.

The limiter — Limiter circuits generally are of the plate-saturation type (§ 7-15), where low plate and screen voltage are used to limit the plate-current flow at high signal amplitudes. Fig. 732-A is a typical circuit. The tube is self-biased (§ 3-6) by a grid leak, R_1 , and condenser, C_1 . R_2 , R_3 and R_4 form a voltage divider (§ 8-10) which puts the desired voltages on the screen and plate. The lower the voltages the lower the signal level at which limiting occurs, but the r.f. output voltage of the limiter also is lower. C_2 and C_3 are the plate and screen by-pass condensers, of conventional value for the intermediate frequency used. The time constant (§ 2-6) of R_1C_1 determines the behavior of the limiter with respect to rapid and slow amplitude variations. For best operation on impulse noise (§ 7-15) the time constant should be small, but a small time constant limits the range of signal strengths the limiter can handle without departing from the constant-output condition. A larger time constant is better in the latter respect but is not so effective for rapid variations, hence compromise constants must be used.

The cascade limiter, Fig. 732-B, overcomes this by making the time constant in the first grid circuit suitable for effective operation on impulse noise, and that in the second grid (C_4R_6) optimum for a wide range of input signal strengths. This results, in addition, in more constant output over a very wide range of input signal amplitudes because the voltage at the grid of the second stage already is partially amplitude-limited, thus giving the second stage less work to do. Resistance coupling ($R_5C_4R_6$) is used in preference to transformer coupling for simplicity and to prevent unwanted regeneration, additional gain at this point being unnecessary.

The rectified voltage developed across R_1 in either circuit may be used for a.v.c. (§ 7-13).

Discriminator circuits and operation — The f.m. detector commonly is called a *discriminator*, because of its ability to discriminate

between frequency deviations above and those below the carrier frequency. The circuit generally used is shown in Fig. 733-A. A special i.f. coupling transformer is used between the limiter and detector. Its secondary, L_1 , is center-tapped and is connected back to the plate side of the primary circuit, which otherwise is conventional. C_4 is the tuning condenser. The load circuits of the two diode rectifiers (R_1C_1 , R_2C_2) are connected in series; the constants are the same as in ordinary diode detector circuits (§ 7-3). The audio output is taken from across the two load resistances.

The primary and secondary circuits are both adjusted to resonance in the center of the i.f. pass-band. The voltage applied to the rectifiers consists of two components, that induced in the secondary by the inductive coupling and that fed to the center of the secondary through C_2 . The phase relations between the two are such that at resonance the rectified load currents are equal in amplitude but flow in opposite directions through R_1 and R_2 , hence the net voltage across the terminals marked "audio output" is zero. When the carrier deviates from resonance the induced secondary current either lags or leads, depending upon whether the deviation is to the high- or low-frequency side, and this phase shift causes the induced current to combine with that fed through C_2 in such a way that one diode gets more voltage than the other when the frequency is below resonance, while the second diode gets the larger voltage when the frequency is higher than resonance. The voltage appearing across the output terminals is the difference between the two diode voltages. Thus a characteristic like that of Fig. 734 results, where the net rectified output voltage has opposite polarity

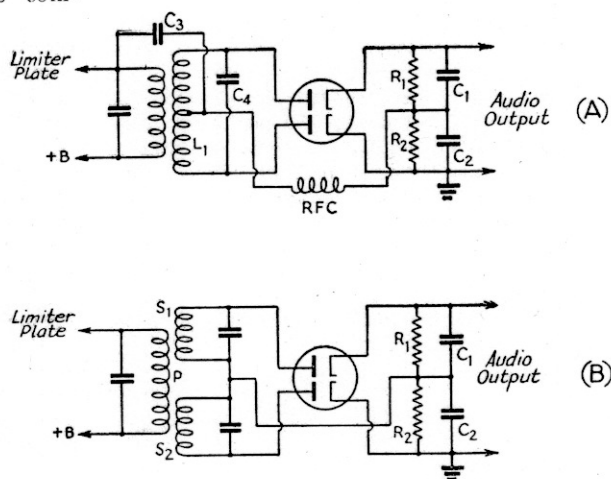


Fig. 733 — F.m. discriminator circuits. In both circuits typical values for C_1 and C_2 are 100 $\mu\text{fd.}$ each; R_1 and R_2 , 0.1 megohm each. C_3 in A is approximately 50 $\mu\text{fd.}$, depending upon the intermediate frequency; RFC should be of a type designed for the i.f. in use (2.5 mb. is satisfactory for i.f.s of 4 to 5 megacycles). The special three-winding transformer in B is described in the text. In either circuit the ground may be moved from the lower end of C_2 to the junction of C_1 and C_2 , for push-pull audio output.

for frequencies on either side of resonance, and up to a certain point becomes greater in amplitude as the frequency deviation is greater. The straight-line portion of the curve is the useful detector characteristic. The separation between the peaks which mark the ends of the linear portion of the curve depends upon the Q s of the primary and secondary circuits and the degree of coupling. The separation becomes greater with low Q s and close coupling. The circuit ordinarily is designed so that the peaks fall just outside the limits of the pass-band, thus utilizing most of the straight portion of the curve. Since the audio output is proportional to the change in d.c. voltage with deviation, it is advantageous from this standpoint to have the peak separation the minimum necessary for a linear characteristic.

A second type of discriminator circuit is shown in Fig. 733-B. Two secondary circuits, S_1 and S_2 , are used, one tuned above the center frequency of the i.f. pass-band and the other below. They are coupled equally to the primary, which is tuned to the center frequency. As the carrier frequency deviates the voltages induced in the secondaries will change in amplitude, the larger voltage appearing across the secondary being nearer resonance with the instantaneous frequency. The detection characteristic is similar to that of the first type of discriminator. The peak separation is determined by the Q s of the circuits, the coefficient of coupling, and the tuning of the two secondaries. High Q s and loose coupling are necessary for close peak separation.

F.m. receiver alignment — Alignment of f.m. receivers up to the limiter is carried out as described in § 7-17. For output measurement, a 0-1 milliammeter or 0-500 microammeter should be connected in series with the limiter grid resistor (R_1 in Fig. 732) at the grounded end; or, if the voltage drop across R_1 is used for a.v.c. and the receiver is provided with a tuning meter (§ 7-13), the tuning meter may be used as an output meter. An accurately calibrated signal generator or test oscillator is desirable, since the i.f. should be aligned to be as symmetrical as possible; that is, the output reading should be the same for any two test oscillator settings the same number of kilocycles above or below resonance. It is not necessary to have uniform response over the whole band to be received, although the output at the edges of the band (limit of deviation (§ 5-11)

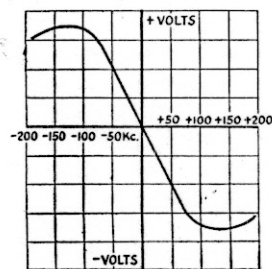


Fig. 734 — Characteristic of a typical f.m. detector. The vertical axis represents the voltage developed across the load resistor as the frequency varies from the exact resonance frequency. This detector would handle f.m. signals up to a band-width of 150 kc. over the linear portion of the curve.

of the transmitted signals) should not be less than 25 per cent of the voltage at resonance. In communications work, a band-width of 30 kc. or less (15 kc. or less deviation) is commonly used.

Output readings should be taken with the test oscillator set at intervals of a few kilocycles either side of resonance until the band limits are reached.

After the i.f. (and front-end) alignment, the limiter operation should be checked. This can be done by temporarily disconnecting C_3 , if the discriminator circuit of Fig. 733-A is used, disconnecting R_1 and C_1 on the cathode side, and inserting the milliammeter or microammeter in series with R_2 at the grounded end. This converts the discriminator to an ordinary diode rectifier. Varying the signal-generator frequency over the channel, with the discriminator transformer adjusted to resonance, should show no change in output (at the band-widths used for communications purposes) as indicated by the rectified current read by the meter. At this point various plate and screen voltages can be tried on the limiter tube or tubes, to determine the set of conditions which gives maximum output with adequate limiting (no change in rectified current).

When the limiter has been checked the discriminator connections can be restored, leaving the meter connected in series with R_1 . Provision should be made for reversing the connections to the meter terminals, to take care of the reversal in polarity of the net rectified current. Set the signal generator to the center frequency of the band and adjust the discriminator transformer trimmer condensers to resonance, which will be indicated by zero rectified current. Then set the test oscillator at the deviation limit (§ 5-11) on one side of the center frequency, and note the meter reading. Reverse the meter terminals and set the test oscillator at the deviation limit on the other side. The two readings should be the same. If they are not, they can be made so by a slight adjustment of the primary trimmer. This will necessitate rechecking the response at resonance to make sure it is still zero. Generally, the secondary trimmer will chiefly affect the zero-response frequency, while the primary trimmer will have most effect on the symmetry of the discriminator peaks. A detector curve having satisfactory linearity can be obtained by cut-and-try adjustment of both trimmers.

Tuning and operation — An f.m. receiver gives greatest noise reduction when the carrier is tuned exactly to the center of the receiver pass-band and to the point of zero response in the discriminator. Because of the decrease in noise, this point is readily recognized.

When an amplitude-modulated signal is tuned in its modulation practically disappears at exact resonance, only those nonsymmetrical modulation components which may be present being detected. If the signal is to one side or the other of resonance, however, it is capable of causing interference to an f.m. signal.