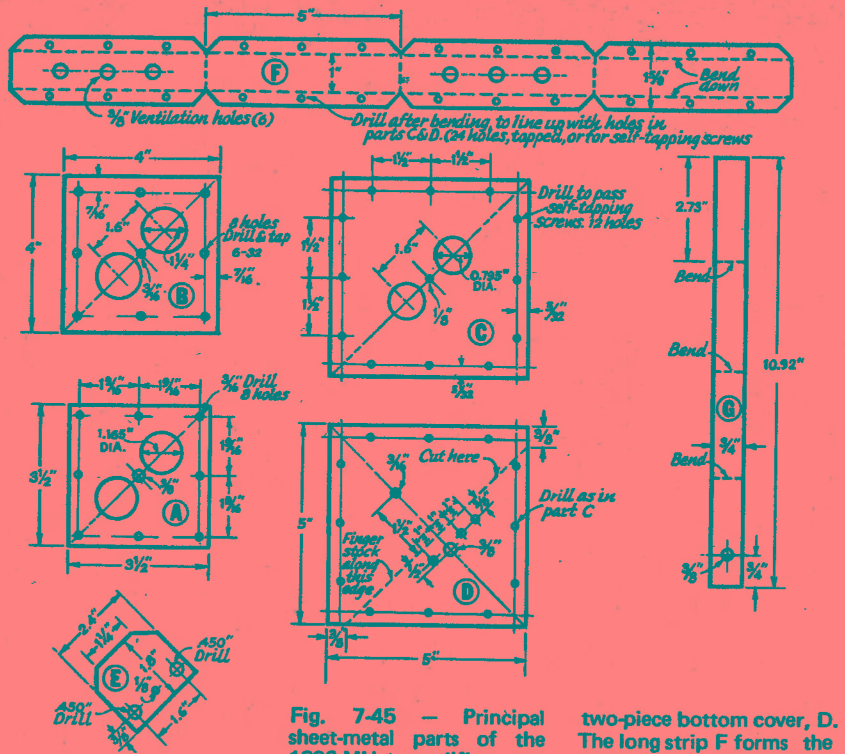


Fig. 7-45 — Principal sheet-metal parts of the 1296-MHz amplifier: top plate of the bypass capacitor, A; its bottom plate and top cover of the plete cavity, B; top plate of the cathode assembly, C; and two-piece bottom cover, D. The long strip F forms the side walls of the cathode assembly, and G forms the side walls of the plete cavity, both before bending into their square shape.

Tuning and Operation

When construction is completed and checked out, apply heater power to the tubes. Connect a milliammeter in series with the cathode resistor. Set the input glass trimmer at the middle of its range, and place the cover plate in position, but without putting in the screws as yet. Keep some pressure on it by hand to insure uniform contact. Apply 10 to 20 watts of driving power, tune C2, and observe the cathode current. Open the cathode compartment, move the input trimmer, replace the cover, and observe the current again. Repeat until highest current is achieved, but do not go over 120 mA. Reduce driving power, if necessary, to keep below this level. Fasten the cover plate in place, and recheck cathode current.

Supply cooling air, if this has not already been done. Be sure that adequate air flow is provided, especially if the plate input is to be near maximum ratings. If there is to be no cowling around the tube fins an air stream of some 150 ft^3/min from a low-pressure blower across the area of the tube fins is required. With an enclosure confining the air flow to a path through the fins a 30 ft^3/min high-pressure blower should suffice. In either case it does no harm to have more. If you have a quiet blower it probably is not enough!

Connect a 50-ohm termination to J2 and apply plate power, preferably at a lower voltage than the maximum that will be used eventually. Apply drive, and tune the input circuit for maximum plate current, and the output circuit for maximum output. A suitable indicator is an incandescent lamp connected at the end of a 50-foot length of RG-58 cable. This will be so lossy that it will look like 50 ohms, regardless of the termination, and the lamp will show relative output. Maximum output may not coincide with minimum plate current.

Once the amplifier appears to be working normally, plate voltage may be increased. Recheck the tuning adjustments for each change in plate voltage. Use a value of cathode resistance that will result in about 50 mA plate current with no drive. With 1000 volts on the plates do not operate the amplifier for more than a few seconds at a time under key-down conditions. With a normal cw keying duty cycle you can run up to 400 mA plate current. With ssb you may run up to 600 mA peak current, or a 300-mA indicated meter reading during normal voice operation. With the expected 100 watts output, with 300 to 400 in, the RG-58 cable should melt in a few minutes. This is not a very satisfactory method of measuring output, and some reliable power-indicating meter should be used for at least an intermittent check, if at all possible.

500-WATT AMPLIFIER FOR 432 MHz

One of the best tuned circuits, other than a resonant cavity, for an amplifier in the 420-MHz band is a coaxial line. To build a good one requires some metal work, but the assembly described here should not be difficult for the advanced worker. Amplifiers of this type have been built and used by W1QWJ and W1RVW, with excellent results. They run up to 500 watts input on fm and cw, and the amplifiers operate very much as they would on much lower frequencies.

Input circuit details are given for both 144 and 432 MHz, permitting the stage to be set up for tripling or straight-through operation. An inexpensive 4X150A running as a tripler will drive any of the 250-series tubes with ease.

Construction

The basic design should be clear from the photographs, Figs. 7-46, 47 and 49. Structural details may be obtained from Fig. 7-50. The straight-through amplifier and the 144-MHz grid circuit are shown schematically in Fig. 7-48. In the amplifier photographed, W1RVW used two separate 8 by 12-inch chassis, with their 8-inch surfaces fastened to a standard rack panel, 1 inch apart. They are held in firm alignment by an aluminum plate fastened at the back. One chassis carries the amplifier, the other a regulated screen supply.

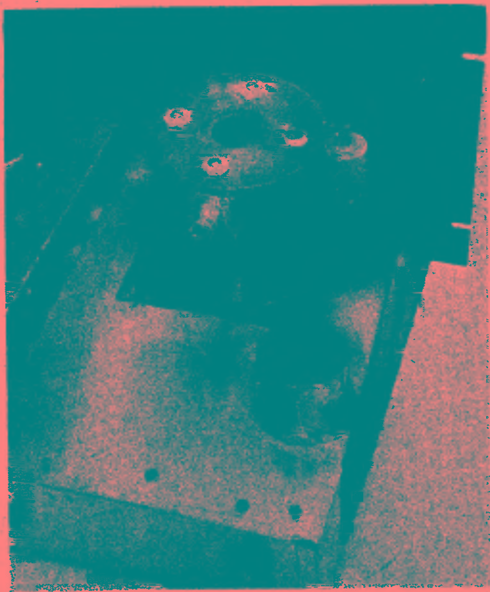


Fig. 7-46 — Looking down at the coaxial plete circuit of the 500-watt uhf amplifier. Air fed into the screened intake, lower edge of the picture, flows through the enclosed chassis below, up through the tube socket end out through the hole at the end of the plate line.

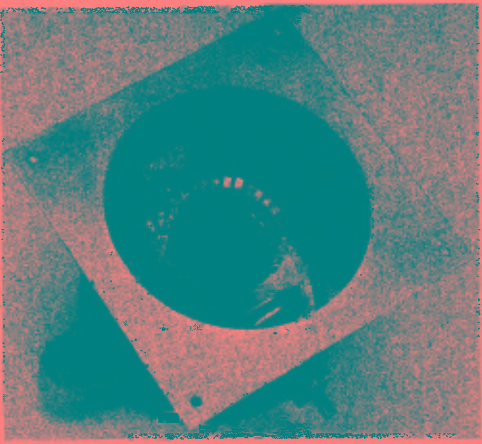


Fig. 7-47 — Interior of the plate circuit assembly, showing tha center conductor with its ring of finger stock, the output-coupling loop, left, and the disk-type tuning capacitor, right.

The amplifier plate circuit is built in a 3 3/4-inch section of 4-inch copper tubing. This is mounted on a 5-inch square brass base plate. The top is a copper disk with a 1 3/8-inch air hole at the center. Inside the cover is a teflon-insulated capacitor plate, soldered to the inner conductor of the plate circuit, L3 in Fig. 7-48. The latter is 1 1/2-inch copper tubing, 2 3/16 inch long. A ring of finger stock extends 5/8 inch below the end of L3, for making contact to the 7203/4CX250B anode. Eimac CF-300 Finger Stock, 31/32 inch wide, is used.

The line is tuned by means of a brass-disk capacitor, C3, details of which are shown in Fig. 7-50. The method of keeping tension on the lead-screw may be of interest, since this is often a problem with this type of tuning device. Two methods have been used by the builders. The amplifier shown has a piece of brass 1/2 inch square and 3/4 inch long fastened to the outer wall. The screw passes through this, and the lower part of the block is slotted, up to the 1/4-inch hole. A tension screw threaded into the block makes it possible to pull the sides together slightly, as required. The other tension system is shown in Fig. 7-50. Here a springy piece of metal is threaded onto the lead-screw, and then put under tension slightly by screws at either end.

The capacitor plate, C5, at the top of the line is insulated from the cover with teflon sheet, the thickness of which is determined by the type of operation intended. If the amplifier is to be plate-modulated this sheet should be 1/32 inch. For cw or fm 0.01 inch is satisfactory. Four ceramic buttons insulate the screws that hold the capacitor together. Dimensions are not given for the holes required, as they will depend on the insulators available.

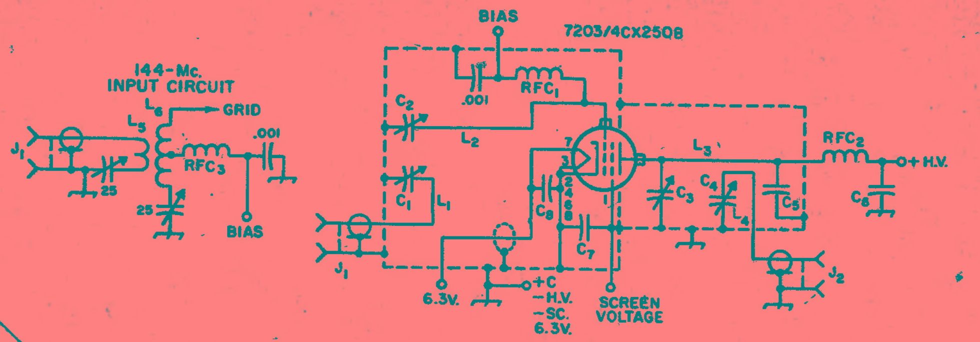


Fig. 7-48 – Schematic diagram of the 432-MHz amplifier, as set up for straight-through operation. An alternate 144-MHz input circuit for tripling is shown at the left.

C1, C2, C4 – 9-pF miniature trimmer (Johnson 160-104 or 9M11).
C3 – Disk-type tuning capacitor, 1 1/2-inch dia brass.
C5 – Teflon-insulated high-voltage bypass. See text.
C6 – 500-pF 20-kV TV-type capacitor.
C7, C8 – Built into socket.
J1, J2 – Coaxial fitting.
L1 – No. 12 wire loop, 6 inches overall. See Fig. 7-49-7-50.
L2 – 1/16-inch brass, 1 1/4 by 3 7/8 inches. See Fig. 7-49.
L3 – 1 1/2-inch copper tubing, with finger stock. See Fig. 7-50.
L4 – No. 16 wire loop, 1/4 inch wide. Top is 1/4 inch from C5.
L5 – 2 turns No. 16 enam., 1/2-inch dia, coupled to L6.
L6 – 4 turns No. 14 enam., 1/2-inch dia, 1 inch long, ct.
RFC1 – 8 turns No. 16 enam., 1/4-inch ID, 7/8 inch long.
RFC2 – 8 turns No. 20 enam., on 1-watt 1-meg-ohm resistor.
RFC3 – 1.4-μH rf choke.

Note that the high voltage is on these screws when the amplifier is in operation. It is fed into one of them through a small rf choke, RFC2, the outer end of which is supported on a TV-type 500-pF high-voltage capacitor, C6. The lower end of C6 is supported on a brass angle bracket fastened to the side of the line assembly.

Output coupling from the line is by means of a small loop of wire, L4, mounted in a vertical position near the top of the line. It is series-tuned by C4, directly below it.

Details of the 432-MHz grid circuit and its input coupling are given in Fig. 7-50. The input capacitance of these tubes is high, so a half-wave line must be used. Even with this type of grid circuit, the inductance must be very low to tune to 432 MHz. Note that L1 is less than 4 inches long, despite its 1 1/4-inch width.

Operation

Because of the high-efficiency coaxial plate circuit, the amplifier operates almost as it would on lower frequencies. The manufacturer's ratings may be followed, using the maximum figures if desired. It is usually desirable to make provision for lowering the plate voltage in some way, however, as the difference between the maximum rating and something perhaps 25 to 50 percent lower will make only a trifling difference in results, except where contact is being maintained under marginal conditions.

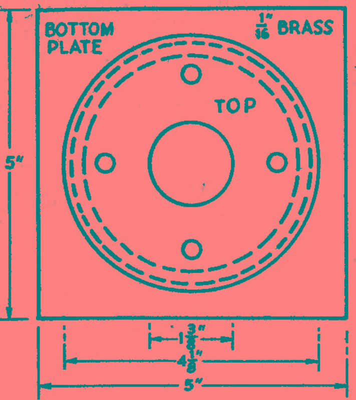
About the only variation from lower-frequency practice is the need for keeping the heater voltage low. The rated voltage for these tubes is 6.0, not 6.3, and at frequencies above 300 MHz it should be reduced. At 432 MHz the voltage should be 5.5. With higher voltages the back-bombardment that the cathode is subjected to raises the overall tube temperature and shortened tube life results. The drifting of operating conditions often observed in vhf and uhf amplifiers is likely to be traceable to excessive heater voltage.

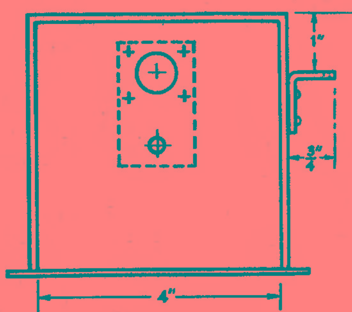
Be sure to use plenty of air flow through the socket and tube anode. In the amplifier shown, air

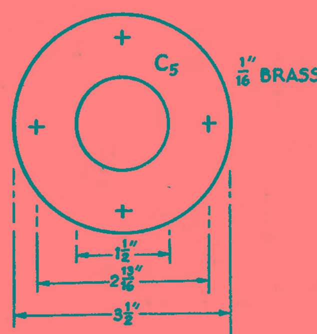
Fig. 7-49 – Bottom view of the amplifier, showing the strip of brass used for the grid circuit inductance, L2.

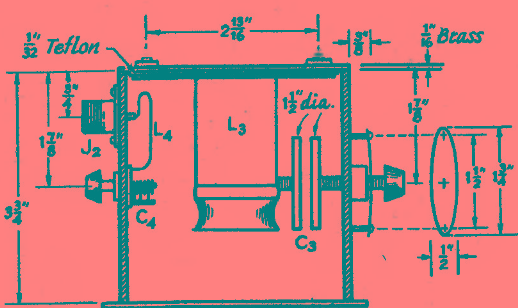
is fed into an opening in the top of the chassis. The bottom has a tight-fitting cover, so that the only air route open is through the socket and out through anode and L3.

Adjustment of the position of the output coupling loop, L4, with respect to the inner conductor of the line is fairly critical, if maximum efficiency is to be achieved. In one of the amplifiers the coupling loop, the coaxial fitting and the series capacitor were made into a single assembly on a curved plate of copper or brass. This could be removed at will, to permit adjustment of the shape and position of the coupling loop. It is fastened to the outside of the main cylinder with small brass screws, covering a rectangular hole in the cylinder cut for this purpose.









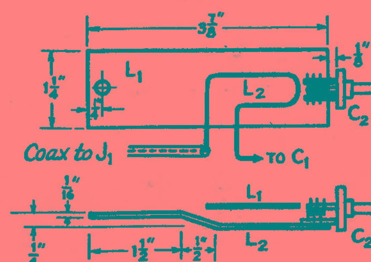


Fig. 7-50 – Principal mechanical details of the 432-MHz amplifier. The coaxial tank circuit is shown in cut-away form at the lower left, and in outline, center. The top view of the assembly and the capacitor plata for C5 are the other views. Details of the strip-line grid circuit are et the lower right.

Receiving Systems

The performance of a communications receiver can be measured by its ability to pick up weak signals and separate them from the noise and interference while at the same time holding them steady at the same dial settings. The difference between a good receiver and a poor one can be the difference between copying a weak signal well, or perhaps not copying it at all.

Whether the receiver is of home-made or commercial origin, its performance can range from excellent to extremely poor, and high cost or circuit complexity cannot assure proper results. Some of the simplest of receivers can provide excellent results if careful attention is given to their design and proper use. Conversely, the most expensive of receivers can provide poor results if not operated in a competent manner. Therefore, the operator's success at sorting the weak signals out of the noise and interference is dependent upon the correct use of a properly designed, correctly operated receiver.

Communications receivers are rated by their **sensitivity** (ability to pick up weak signals), their **selectivity** (the ability to distinguish between signals that are extremely close together in terms of frequency), and by their stability. The latter trait assures that once a stable signal is tuned in it will remain tuned without periodic retuning of the receiver controls (especially the main tuning and BFO controls).

A well-designed modern receiver must be able to receive all of the popular modes of emission if it is to be truly versatile. It should be capable of handling cw, ssb, a-m, fm, and RTTY signals.

The type of detection to be used will depend on the job the receiver is called upon to do. Simple receivers consisting of a single stage of detection (regenerative detector) followed by a one- or two-stage audio amplifier are often adequate for portable and emergency use over short distances. This type of receiver can be quite compact and light weight and can provide many hours of operation from a dry-battery pack if transistorized circuitry is used. Similarly, superregenerative detectors can be used in the same way, but are suitable for copying only a-m and wide-band fm signals. Superheterodyne receivers are the most popular and are capable of better performance than the foregoing types. *Heterodyne* detectors are used for ssb and cw reception in the latter. If a regenerative detector is made to oscillate and provide a steady signal, it is known as an autodyne detector. A beat-frequency oscillator, or BFO, is used to generate a steady signal in the superheterodyne receiver. This signal is applied to the detector stage to permit the reception of ssb and cw signals,

Communications receivers should have a slow tuning rate and a smooth-operating tuning-dial mechanism if any reasonable degree of selectivity is used. Without these features cw and ssb signals are extremely hard to tune in. In fact, one might easily tune past a weak signal without knowing it was there if a fast tuning rate were used.

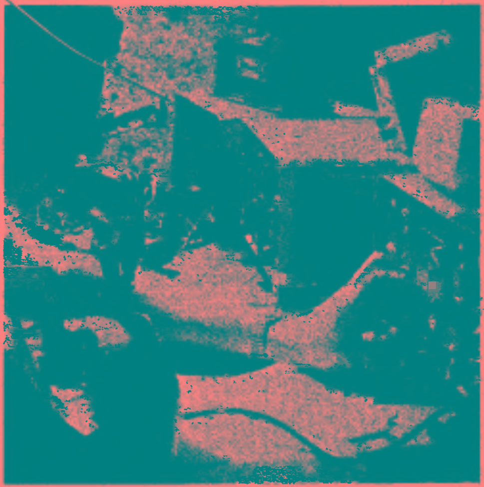


Fig. 8-1 — The success of amateur on-the-air operation is, in a large part, determined by a receiver. A good receiver, mated with a good pair of ears, is an unbeatable combination.

RECEIVER CHARACTERISTICS

Sensitivity

In commercial circles "sensitivity" is defined as the signal at the input of the receiver required to give a signal-plus-noise output some stated ratio (generally 10 dB) above the noise output of the receiver. This is a useful sensitivity measure for the amateur, since it indicates how well a weak signal will be heard. However, it is not an absolute method, because the bandwidth of the receiver plays a large part in the result.

The random motion of the molecules in the antenna and receiver circuits generates small

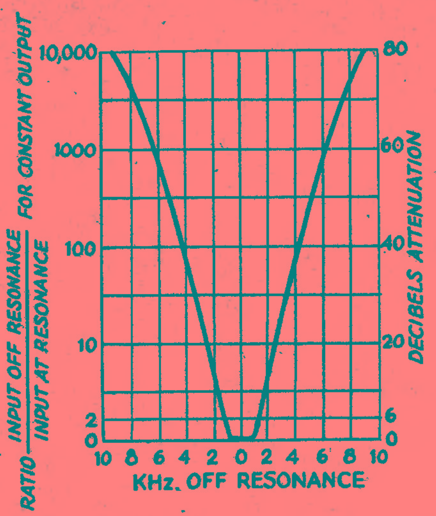


Fig. 8-2 — Typical selectivity curve of e modern superheterodyne receiver. Relative response is plotted against deviations above end below the resonance frequency. The scala et the left is in terms of voltage ratios, the corresponding decibel steps are shown et the right.

voltages called thermal-agitation noise. Thermal-agitation noise is independent of frequency and is proportional to the (absolute) temperature, the resistive component of the impedance across which the thermal agitation is produced, and the bandwidth. Noise is generated in vacuum tubes and semiconductors by random irregularities in the current flow within them; it is convenient to express this shot-effect noise as an equivalent resistance in the grid circuit of a noise-free tube. This equivalent noise resistance is the resistance (at room temperature) that placed in the grid circuit of a noise-free tube will produce plate-circuit noise equal to that of the actual tube. The equivalent noise resistance of a vacuum tube increases with frequency.

An ideal receiver would generate no noise in its tubes or semiconductors and circuits, and the minimum detectable signal would be limited only by the thermal noise in the antenna. In a practical receiver, the limit is determined by how well the amplified antenna noise overrides the other noise of the input stage. (It is assumed that the first stage of any good receiver will be the determining factor; the noise contributions of subsequent stages should be insignificant by comparison.) At frequencies below 20 or 30 MHz the site noise (atmospheric and man-made noise) is generally the limiting factor.

The degree to which a practical receiver approaches the quiet ideal receiver of the same bandwidth is given by the noise figure of the receiver. Noise figure is defined as the ratio of the signal to noise power ratio of the ideal receiver to the signal-to-noise power ratio of the actual receiver output. Since the noise figure is a ratio, it is usually given in decibels; it runs around 5 to 10 dB for a good communications receiver below 30 MHz. Although noise figures of 2 to 4 dB can be obtained, they are of little or no use below 30 MHz except in extremely quiet locations or when a very small antenna is used. The noise figure of a receiver is not modified by changes in bandwidth.

Selectivity

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

The selectivity of a receiver is shown graphically by drawing a curve that 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 is shown in Fig. 8-2. The bandwidth is the width of the resonance curve (in Hz or kHz) of a receiver at a specified ratio; in the typical curve of Fig. 8-2 the bandwidths for response ratios of 2 and 1000 (described as "-6 dB" and "-60 dB") are 2.4 and 12.2 kHz respectively.

The bandwidth at 6 dB down must be sufficient to pass the signal and its sidebands if faithful reproduction of the signal is desired. However, in the crowded amateur bands, it is generally advisable to sacrifice fidelity for intelligibility. The ability to reject adjacent-channel signals depends upon the skirt selectivity of the receiver, which is determined by the bandwidth at high attenuation. In a receiver with excellent skirt selectivity, the ratio of the 6-dB bandwidth to the 60-dB bandwidth will be about 0.2 for code and 0.3 for phone. The minimum usable bandwidth at 6 dB down is approximately 150 Hz for code reception and approximately 2000 Hz for phone.

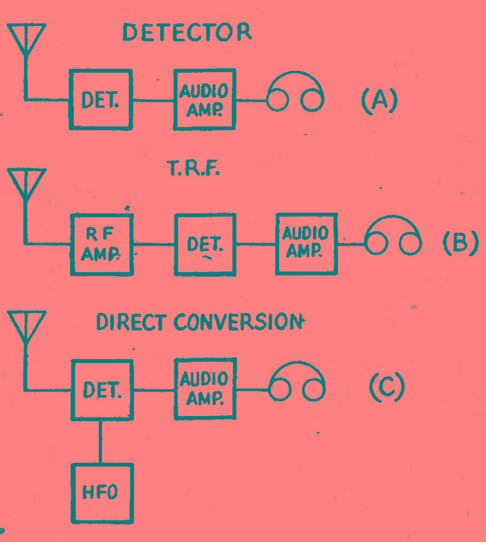


Fig. 8-3 — Block diagrams of three simple receivers.

Stability

The stability of a receiver is its ability to "stay put" on a signal under varying conditions of gain-control setting, temperature, supply-voltage changes and mechanical shock. The term "unstable" is also applied to a receiver that breaks into oscillation or a regenerative condition with some settings of its controls that are not specifically intended to control such a condition.

SIMPLE RECEIVERS

The simplest receiver design consists of a detector followed by an audio amplifier, as shown in Fig. 8-3A. Obviously, the sensitivity of the detector determines how well this receiver will work. Various schemes have been developed to increase detector sensitivity, including the regenerative and superregenerative detectors described later in this chapter. Another way to increase receiver sensitivity is to add one or more rf-amplifier stages before the detector. This approach is called the **tuned-radio-frequency**, or TRF receiver, Fig. 8-3B.

Another design which has become popular for use in battery-powered equipment is the direct-conversion receiver, Fig. 8-3C. Here, a detector is employed along with a variable-frequency oscillator which is tuned just slightly off the frequency of the incoming signal to produce a beat note. A narrow-bandwidth audio filter located between the detector and the aduio amplifier provides selectivity. However, the lack of automatic gain control limits the range over which the receiver can handle strong signals unless a manual rf-gain control is employed. FETs and ICs can be used as detectors to provide up to 90 dB of dynamic range – typically 3 μV to 100 mV of input signal.

DETECTION AND DETECTORS

Detection (demodulation) is the process of extracting the signal information from a modulated carrier wave. When dealing with an a-m signal, detection involves only the rectification of the rf signal. During fm reception, the incoming signal must be converted to an a-m signal for detection. See Chapter 14.

Detector sensitivity is the ratio of desired detector output to the input. Detector linearity is a measure of the ability of the detector to reproduce the exact form of the modulation on the incoming signal. The resistance or impedance of the detector is the resistance or impedance it presents to the circuits it is connected to. The input resistance is important in receiver design, since if it is relatively low it means that the detector will consume power, and this power must be furnished by the preceding stage. The signal-handling capability means the ability to accept signals of a specified amplitude without overloading or distortion.

Diode Detectors

The simplest detector for a-m is the diode. A germanium or silicon crystal is an imperfect form of diode (a small current can usually pass in the reverse direction), but the principle of detection in a semiconductor diode is similar to that in a vacuum-tube diode.

Circuits for both half-wave and full-wave diodes are given in Fig. 8-4. The simplified half-wave circuit at Fig. 8-4A includes the rf tuned circuit, L2C1, a coupling coil, L1, from which the rf energy is fed to L2C1, and the diode, CR1, with its load resistance, R1, and bypass capacitor, C2.

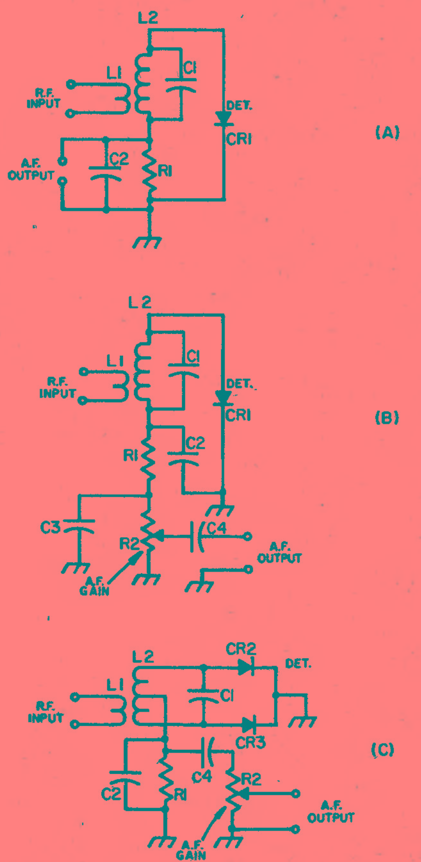


Fig. 8-4 – Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with rf filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L2C1, is tuned to the signal frequency; typical values for C2 and R1 in A and C are 250 pF and 250,000 ohms, respectively; in B, C2 and C3 are 100 pF each; R1, 50,000 ohms; and R2, 250,000 ohms. C4 is 0.1 μF and R3 may be 0.5 to 1 megohm.

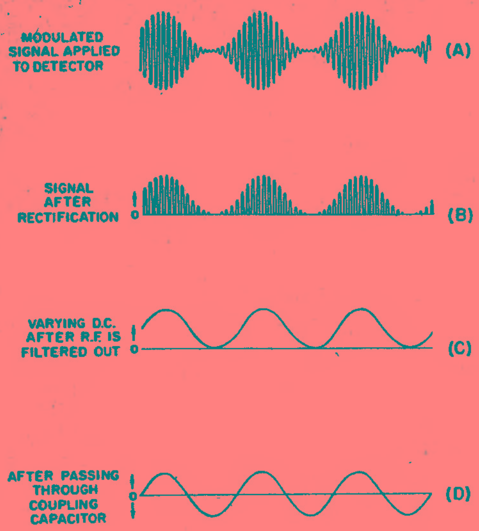


Fig. 8-5 – Diagrams showing the detection process.

The progress of the signal through the detector or rectifier is shown in Fig. 8-5. A typical modulated signal as it exists in the tuned circuit is shown at A. When this signal is applied to the rectifier, current will flow only during the part of the rf cycle when the anode is positive with respect to cathode, so that the output of the rectifier consists of half-cycles of rf. These current pulses flow in the load circuit comprised of R1 and C2, the resistance of R1 and the capacitance of C2 being so proportioned that C2 charges to the peak value of the rectified voltage on each pulse and retains enough charge between pulses so that the voltage across R1 is smoothed out, as shown in C. C2 thus acts as a filter for the radio-frequency component of the output of the rectifier, leaving a dc component that varies in the same way as the modulation on the original signal. When this varying dc voltage is applied to a following amplifier through a coupling capacitor (C4 in Fig. 8-4), only the *variations* in voltage are transferred, so that the final output signal is ac, as shown in D.

In the circuit at 8-4B, R1 and C2 have been divided for the purpose of providing a more effective filter for rf. It is important to prevent the appearance of any rf voltage in the output of the detector, because it may cause overloading of a succeeding amplifier stage. The audio-frequency variations can be transferred to another circuit through a coupling capacitor, C4. R2 is usually a "potentiometer" so that the audio volume can be adjusted to a desired level.

Coupling from the potentiometer (volume control) through a capacitor also avoids any flow of dc through the moving contact of control. The flow of dc through a high-resistance volume control often tends to make the control noisy (scratchy) after a short while.

The full-wave diode circuit at 8-4C differs in operation from the half-wave circuit only in that both halves of the rf cycle are utilized. The full-wave circuit has the advantage that rf filtering is easier than in the half-wave circuit. As a result, less attenuation of the higher audio frequencies will be obtained for any given degree of rf filtering.

The reactance of C2 must be small compared to the resistance of R1 at the radio frequency being rectified, but at audio frequencies must be relatively large compared to R1. If the capacitance of C2 is too large, response at the higher audio frequencies will be lowered.

Compared with most other detectors, the gain of the diode is low, normally running around 0.8 in audio work. Since the diode consumes power, the *Q* of the tuned circuit is reduced, bringing about a reduction in selectivity. The loading effect of the diode is close to one half the load resistance. The detector linearity is good, and the signal-handling capability is high.

Plate Detectors

The plate detector is arranged so that rectification of the rf signal takes place in the plate circuit of the tube or the collector of an FET. Sufficient negative bias is applied to the grid to bring the plate current nearly to the cutoff point, so that application of a signal to the grid circuit causes an increase in average plate current. The average plate current follows the changes in the signal in a fashion similar to the rectified current in a diode detector.

In general, transformer coupling from the plate circuit of a plate detector is not satisfactory, because the plate impedance of any tube is very high when the bias is near the plate-current cutoff point. The same is true of a JFET or MOSFET. Impedance coupling may be used in place of the

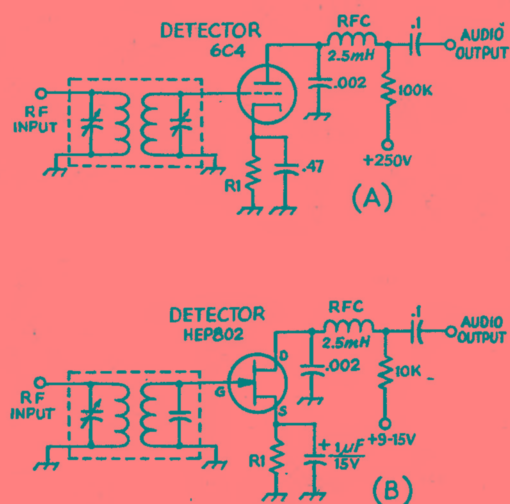


Fig. 8-6 – Circuits for plate detection. A, triode; B, FET. The input circuit, L2C1, is tuned to the signal frequency. Typical values for R1 are 22,000 to 150,000 ohms for the circuit at A, and 4700 to 22,000 ohms for B.

resistance coupling shown in Fig. 8-6. Usually 100 henrys or more of inductance are required.

The plate detector is more sensitive than the diode because there is some amplifying action in the tube or transistor. It will handle large signals, but is not 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.

Infinite-Impedance Detector

The circuit of Fig. 8-7 combines the high signal-handling capabilities of the diode detector with low distortion and, like the plate detector, does not load the tuned circuit it connects to. The circuit resembles that of the plate detector, except that the load resistance, 27kΩ, is connected between source and ground and thus is common to both gate and drain circuits, giving negative feedback for the audio frequencies. The source resistor is bypassed for rf but not for audio, while the drain circuit is bypassed to ground for both audio and radio frequencies. An rf filter can be connected between the cathode and the output coupling capacitor to eliminate any rf that might otherwise appear in the output.

The drain current is very low at no signal, increasing with signal as in the case of the plate detector. The voltage drop across the source resistor consequently increases with signal. Because of this and the large initial drop across this resistor, the gate usually cannot be driven positive by the signal.

HETERODYNE AND PRODUCT DETECTORS

Any of the foregoing a-m detectors becomes a heterodyne detector when a local-oscillator (BFO) is added to it. The BFO signal amplitude should be 5 to 20 times greater than that of the strongest incoming cw or ssb signal if distortion is to be minimized. These heterodyne detectors are frequently used in receivers that are intended for a-m as well as cw and ssb reception. A single detector can thus be used for all three modes, and elaborate switching techniques are not required. To receive a-m it is merely necessary to disable the BFO circuit.

The name product detector has been given to heterodyne detectors in which special attention has been paid to minimizing distortion and intermodulation (IM) products. Product detectors have been thought of by some as a type of detector whose output signal vanishes when the BFO signal is removed. Although some product detectors function that way, such operation is not a criterion. A *product* is something that results from the combination of two or more things, hence any heterodyne detector can rightfully be regarded as a product detector. The two input signals (i-f and BFO) are fed into what is essentially a mixer stage. The difference in frequency (after filtering out and removing the i-f and BFO signals from the mixer output) is fed to the audio amplifier stages and increased to speaker or headphone level. Although product detectors are intended primarily for use with cw and ssb signals, a-m signals can be copied satisfactorily on receivers which have good i-f selectivity. The a-m signal is tuned in as though it were an ssb signal. When properly tuned, the heterodyne from the a-m carrier is not audible.

A triode product-detector circuit is given in Fig. 8-8A. The i-f signal is fed to the grid of the tube, while the BFO energy is supplied to the cathode. The two signals mix to produce audio-frequency output from the plate circuit of the tube. The BFO voltage should be about 2 V rms and the signal should not exceed 0.3 V rms for linear detection. The degree of plate filtering required will depend on the frequency of operation. The values shown in Fig. 8-8A are sufficient for 450-kHz operation. At low frequencies more elaborate filtering is needed. A similar circuit using a JFET is shown at B.

In the circuit of Fig. 8-8C, two germanium diodes are used, though a 6AL5 tube could be substituted. The high back resistance of the diodes is used as a dc return; if a 6AL5 is used the diodes must be shunted by 1-megohm resistors. The BFO signal should be at least 10 or 20 times the amplitude of the incoming signal.

At Fig. 8-8D a two-diode circuit, plus one transistor, provides both a-m and product detection. This circuit is used in the Drake SPR-4 receiver. Balanced output is required from the BFO. The a-m detector is forward biased to prevent the self-squelching effect common to single-diode detectors (caused by signals of low level not exceeding the forward voltage drop of the diode). The IC detector given in Fig. 8-8E has several advantages. First, the BFO injection only needs to be equal to the input signal, because of the additional amplification of the BFO energy which takes place within the IC. Also, output filtering is quite simple, as the double-balanced design reduces the level of i-f signal and BFO voltage appearing in the output circuit. Motorola's MC1496G has a dynamic range of 90 dB and a conversion gain of about 12 dB, making it a good choice for use in a direct-conversion receiver.

A multipurpose IC i-f amplifier/detector/agc system, the National Semiconductor LM373, is shown in Fig. 8-8F. A choice of a-m, ssb, cw, and fm detection is available, as well as a 60-dB-range agc system and i-f amplification of 70 dB. Recovered audio is typically 120 mV. L1C1 tune to the i-f frequency.

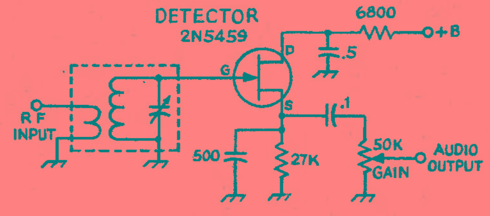


Fig. 8-7 — The infinite-impedance detector. The input circuit, L2C1, is tuned to the signal frequency.

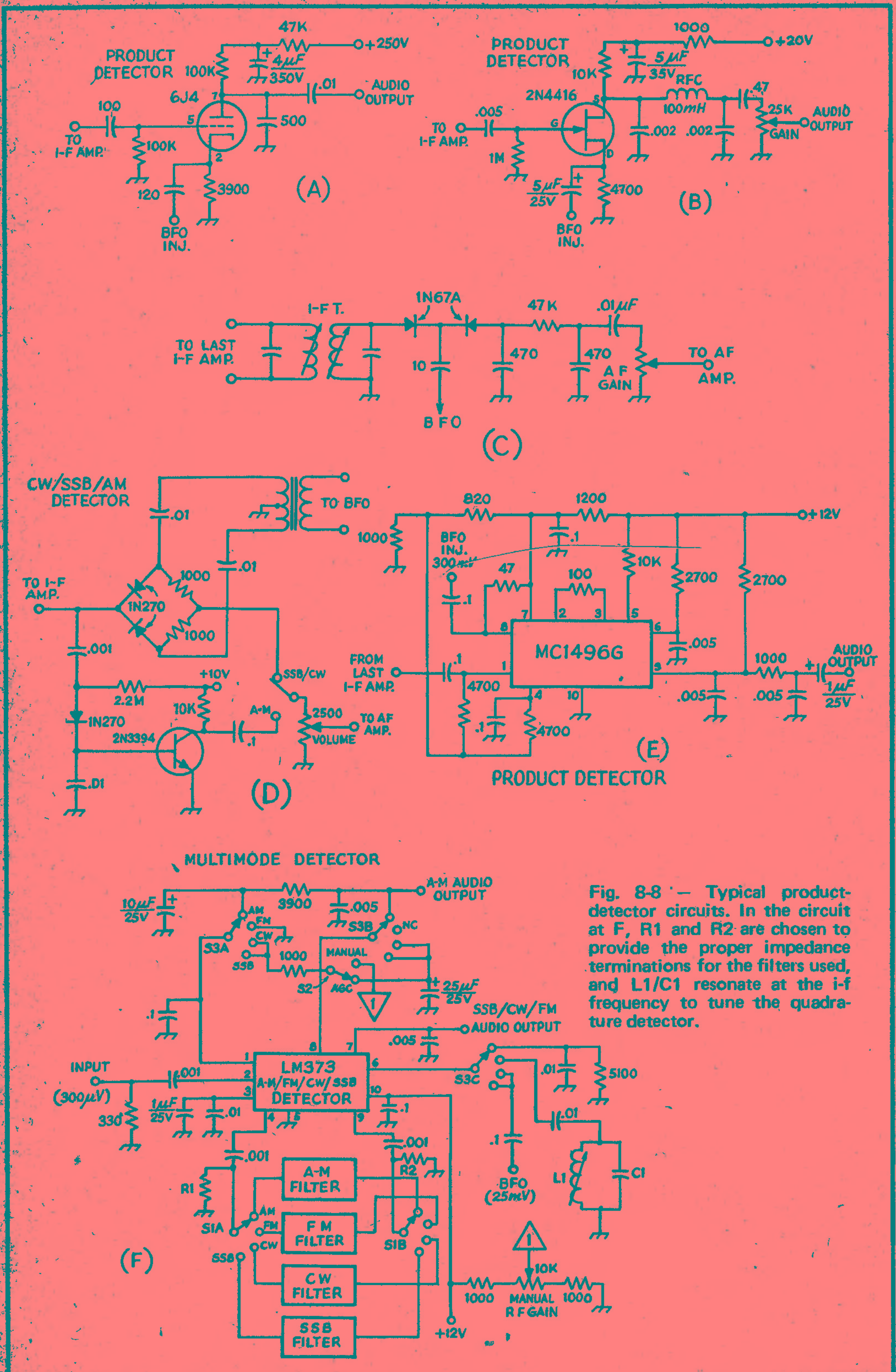


Fig. 8-8 — Typical product-detector circuits. In the circuit at F, R1 and R2 are chosen to provide the proper impedance terminations for the filters used, and L1/C1 resonate at the i-f frequency to tune the quadrature detector.

REGENERATIVE DETECTORS

By providing controllable rf feedback (regeneration) in a triode, pentode, or transistorized-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 thus the selectivity. The grid-leak type of detector is most suitable for the purpose.

The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuit of Fig. 8-9A, the grid corresponds to the diode plate and the rectifying action is exactly the same as in a diode. The dc voltage from rectified-current flow through the grid leak, R1, biases the grid negatively, and the audio-frequency variations in voltage across R1 are amplified through the tube as in a normal af amplifier. In the plate circuit, R2 is the plate-load resistance and C3 and RFC a filter to eliminate rf in the output circuit.

A grid-leak detector has considerably greater sensitivity than a diode. The sensitivity is further increased by using a screen-grid tube instead of a triode. The operation is equivalent to that of the triode circuit. The screen bypass capacitor should have low reactance for both radio and audio frequencies.

The circuit in Fig. 8-9B is regenerative, the feedback being obtained by feeding some signal from the drain circuit back to the gate by inductive coupling. 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. The ciritical point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned. An oscillating detector can be detuned slightly from an incoming cw signal to give *autodyne* reception. The circuit of Fig. 8-9B uses a control which varies the supply voltage to control regeneration. If L2 and L3 are wound end to end in the same direction, the drain connection is to the outside of the "tickler" coil, L3, when the gate connection is to the outside end of L2.

Although the regenerative detector is more sensitive than any other type, its many disadvantages commend it for use only in the simplest receivers. The linearity is rather poor, and the signal-handling capability is limited. The signal-handling capability can be improved by reducing R1 to 0.1 megohm, but the sensitivity will be decreased. The degree of antenna coupling is often critical.

A bipolar transistor is used in a regenerative detector hookup at C. The emitter is returned to dc ground through a 1000-ohm resistor and a 50,000-ohm regeneration control. The 1000-ohm resistor keeps the emitter above ground at rf to permit feedback between the emitter and collector. A 5-pF capacitor (more capacitance might be required) provides the feedback path. C1 and L2 comprise the tuned circuit, and the detected signal is taken from the collector return through T1. A transistor with medinm or high beta works best in circuits of this type and should have a frequency rating which is well above the desired operating frequency. The same is true of the frequency rating of any FET used in the circuit at B.

Superregenerative detectors are somewhat more sensitive than straight regenerative detectors and can employ either tubes or transistors. An in-depth discussion of superregenerative detectors is given in Chapter 9.

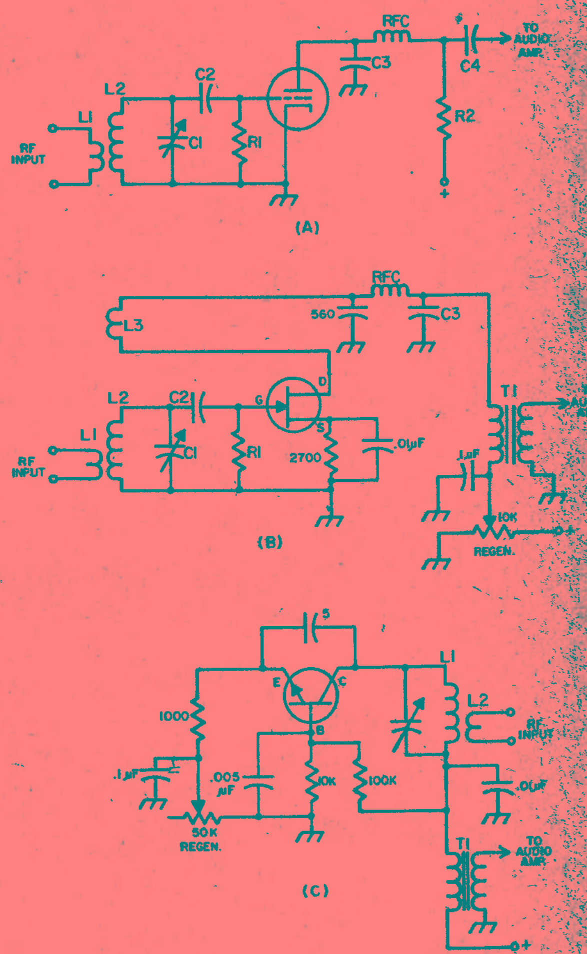


Fig. 8-9 — (A) Triode grid-leak detector combines diode detection with triode emplification. Although shown here with resistive plate load, R2, an audio choke coil or transformer could be used.

(B) Feeding some signal from the drain circuit back to the gate makes the circuit regenerative. When feedback is sufficient, the circuit will oscillate. The regeneration is edjusted by a 10,000-ohm control which varies the drain voltage.

(C) An npn bipolar transistor can be used as a regenerative detector too. Feedback occurs between collector and emitter through the 5-pF capacitor. A 50,000-ohm control in the emitter raturn sets the regeneration. Pnp transistors can also be used in this circuit, but the battery polarity must be reversed.

Tuning

For cw reception, the regeneration control is advanced until the detector breaks into a "hiss" which indicates that the detector is oscillating. Further advancing of the regeneration control will result in a slight decrease in the hiss.

Code signals can be tuned in and will give a tone with each signal depending on the setting of the tuning control. A low-pitched beat note cannot be obtained from a strong signal because the detector "pulls in" or "blocks."

The point just after the detector starts oscillating is the most sensitive condition for code reception. Further advancing the regeneration control makes the receiver less prone to blocking, but also less sensitive to weak signals.

If the detector is in the oscillating condition and an a-m 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.

TUNING METHODS

Tuning

The resonant frequency of a circuit can be shifted by changing either the inductance or the capacitance in the circuit. Panel control of inductance (permeability-tuned oscillator, or PTO) is used to tune a few commercial receivers, but most receivers depend upon panel-mounted variable capacitors for tuning.

Tuning Rate

For ease in tuning a signal, it is desirable that the receiver have a tuning rate in keeping with the type of signal being received and also with the selectivity of the receiver. A tuning rate of 500 kHz per knob revolution is normally satisfactory for a broadcast receiver, but 100 kHz per revolution is almost too fast for easy ssb reception – around 25 to 50 kHz being more desirable.

Band Changing

The same coil and tuning capacitor cannot be used for, say, 3.5 to 14 MHz because of the impracticable maximum-to-minimum capcitance ratio required. 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 capacitor usually is retained, but new coils are inserted in the circuit for each band.

One method of changing inductances is to use a switch having an appropriate number of contacts, which connects the desired coil and disconnects the others. The unused coils are sometimes short-circuited by the switch, to avoid undesirable self-resonances.

Another method is to use coils wound on forms that can be plugged into suitable sockets. These plug-in coils are advantageous when space is at a premium, and they are also very useful when considerable experimental work is involved.

Bandspreading

The tuning range of a given coil and variable capacitor will depend upon the inductance of the coil and the change in tuning capacitance. To cover a wide frequency range and still retain a suitable tuning rate over a relatively narrow frequency range requires the use of **bandspreading. Mechanical bandspreading** utilizes some mechanical means to reduce the tuning rate; a typical example is the two-speed planetary drive to be found in some receivers. **Electrical bandspreading** is obtained by using a suitable circuit configuration. Several of these methods are shown in Fig. 8-10.

In A, a small bandspread capacitor, C1 (15- to 25-pF maximum), is used in parallel with capacitor C2, which is usually large enough (100 to 140 pF) to cover a 2-to-1 frequency range. The setting of C2 will determine the minimum capacitance of the circuit, and the maximum capacitance for bandspread tuning will be the maximum capacitance of C1 plus the setting of C2. The inductance of the coil can be adjusted so that the maximum-minimum ratio will give adequate bandspread. It is almost impossible, because of the nonharmonic relation of the various band limits, to get full bandspread on all bands with the same pair of capacitors. C2 is variously called the **bandsetting** or **main-tuning** capacitor. It must be reset each time the band is changed.

If the capacitance change of a tuning capacitor is known, the total fixed shunt capacitance (Fig. 8-10A) for covering a band of frequencies can be found from Fig. 8-11.

Example: What fixed shunt capacitance will allow a capacitor with a range of 5 to 30 pF to tune 3.45 to 4.05 MHz?

$$(4.05 - 3.45) \div 4.05 = 0.148$$

From Fig. 8-11, the capacitance ratio is 0.38, and hence the minimum capacitance is $(30 - 5) \div 0.38 = 66$ pF. The 5-pF minimum of the tuning capacitor, the tube capacitance and any stray capacitance must be included in the 66 pF.

Fig. 8-10 — Essentials of the three basic bandspread tuning systems.

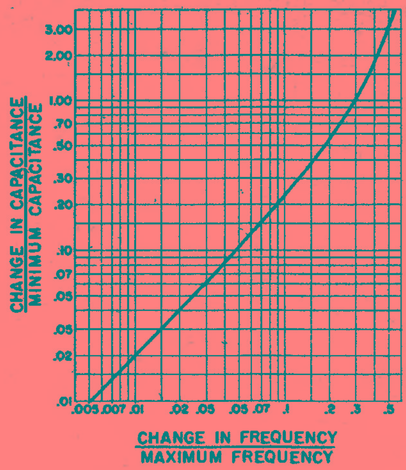


Fig. 8-11 — Minimum circuit capacitance required in the circuit of Fig. 8-10A as a function of the capacitance change and the frequency change. Note that *maximum* frequency and *minimum* capacitance are used.

The method shown at Fig. 8-10B makes use of capacitors in series. The tuning capacitor, C1, may have a maximum capacitance of 100 pF or more. The minimum capacitance is determined principally by the setting of C3, which usually has low capacitance, and the maximum capacitance by the setting of C2, which is in the order of 25 to 50 pF. This method is capable of close adjustment to practically any desired degree of bandspread. Either C2 or C3 must be adjusted for each band or separate preadjusted capacitors must be switched in.

The circuit at Fig. 8-10C also gives complete spread on each band. C1, the bandspread capacitor, may have any convenient value; 50 pF is satisfactory. C2 may be used for continuous frequency coverage ("general coverage") and as a bandsetting capacitor. The effective maximum-minimum capacitance ratio depends on C2 and the point at which C1 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 C2 is set at higher capacitance. C2 may be connected permanently across the individual inductor and preset, if desired. This requires a separate capacitor for each band, but eliminates the necessity for resetting C2 each time.

Ganged Tuning

The tuning capacitors of the several rf 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 for a given setting of the tuning control.

True tracking can be obtained only when the inductance, tuning capacitors, and circuit inductances and minimum and maximum capacitances are identical in all "ganged" stages. A small trimmer or padding capacitor may be connected across the coil, so that various minimum capacitances can be compensated. The use of the trimmer necessarily increases the minimum circuit capacitance but is a necessity for satisfactory tracking. Midget capacitors having maximum capacitances of 15 to 30 pF are commonly used.

The same methods are applied to bandspread circuits that must be tracked. The inductance can be trimmed by using a coil form with an adjustable brass (or copper) core. This core material will reduce the inductance of the coil, raising the resonant frequency of the circuit. Powdered-iron or ferrite core material can also be used, but will lower the resonant frequency of the tuned circuit because it increases the inductance of the coil. Ferrite and powdered-iron cores will raise the *Q* of the coil provided the core material is suitable for the frequency being used. Core material is now available for frequencies well into the vhf region.

The Superheterodyne

In a superheterodyne receiver, the frequency of the incoming signal is heterodyned to a new radio frequency, the intermediate frequency (abbreviated "i-f"), then amplified, and finally detected. The frequency is changed by modulating the output of a tunable oscillator (the high-frequency, or local oscillator) by the incoming signal in a mixer or converter stage to produce a side frequency equal to the intermediate frequency. The other side frequency is rejected by selective circuits. The audio-frequency signal is obtained at the detector. Code signals are made audible by heterodyne reception at the detector stage; this oscillator is called the "beat-frequency oscillator" or BFO. Block diagrams of typical single- and double-conversion receivers are shown in Fig. 8-12.

As a numerical example, assume that an intermediate frequency of 455 kHz is chosen and that the incoming signal is at 7000 kHz. Then the high-frequency oscillator frequency may be set to 7455 kHz in order that one side frequency (7455 minus 7000) will be at 455 kHz. The high-frequency oscillator could also be set to 6545 kHz and give the same difference frequency. To produce an audible code signal at the detector of, say, 1000 Hz, the heterodyning oscillator would be set to either 454 or 456 kHz.

The frequency-conversion process permits rf amplification at a relatively low frequency, the i-f. High selectivity and gain can be obtained at this frequency, and this selectivity and gain are constant. The separate oscillators can be designed for good stability and, since they are working at frequencies considerably removed from the signal frequencies, they are not normally "pulled" by the incoming signal.

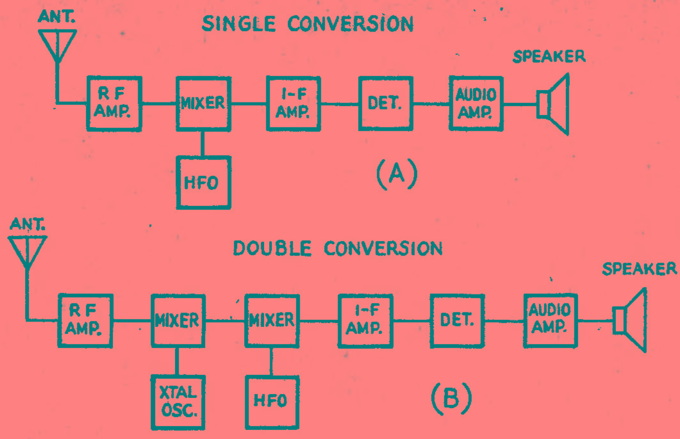


Fig. 8-12 — Block diagrams of a (A) single- and (B) double-conversion superheterodyne receiver.

Images

Each hf 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 kHz to tune to a 7000-kHz signal, for example, the receiver can respond also to a signal on 7910 kHz, which likewise gives a 455-kHz beat. The undesired signal is called the image. It can cause unnecessary interference if it isn't eliminated.

The radio-frequency circuits of the receiver (those used before the signal is heterodyned to the i-f) normally are tuned to the desired signal, so that the selectivity of the circuits reduces or eliminates the response to the image signal. 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 rf tuned circuits preceding the mixer tube. Also, the higher the intermediate frequency, the higher the image ratio, since raising the i-f increases the frequency separation between the signal and the image and places the latter further away from the resonance peak of the signal-frequency input circuits.

The Double-Conversion Superheterodyne

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 kHz. To reduce image response the signal frequently is converted first to a rather high (1500, 5000, or even 10,000 kHz) intermediate frequency, and then – sometimes after further amplification – converted to a lower i-f where higher adjacent-channel selectivity can be obtained. Such a receiver is called a **double-conversion superheterodyne** (Fig. 8-12B).

Other Spurious Responses

In addition to images, other signals to which the receiver is not 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 pickup by any means other than the antenna. When a strong signal is received, the harmonics generated by rectification in the detector may, by stray coupling, be introduced into the rf 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 by careful mechanical design.

MIXER PRODUCTS

Additional spurious products are generated during the mixing process, and these products are the most troublesome of all, as it is difficult indeed to eliminate them unless the frequencies chosen for the mixing scheme are changed. The tables and chart given in Fig. 8-13 will aid in the choice of spurious-free frequency combinations, and they can be used to determine how receiver "birdies" are being generated. Only mixer products that fall close to the desired frequency are considered, as they are the ones that normally cause trouble. The horizontal axis of the chart is marked off in steps from 3 to 20, and the vertical axes from 0 to 14. These numbers can be taken to mean either kilohertz or megahertz, depending on the frequency range used. Both axes must use the same reference; one cannot be in kHz and the other in MHz.

Spurious Response Chart

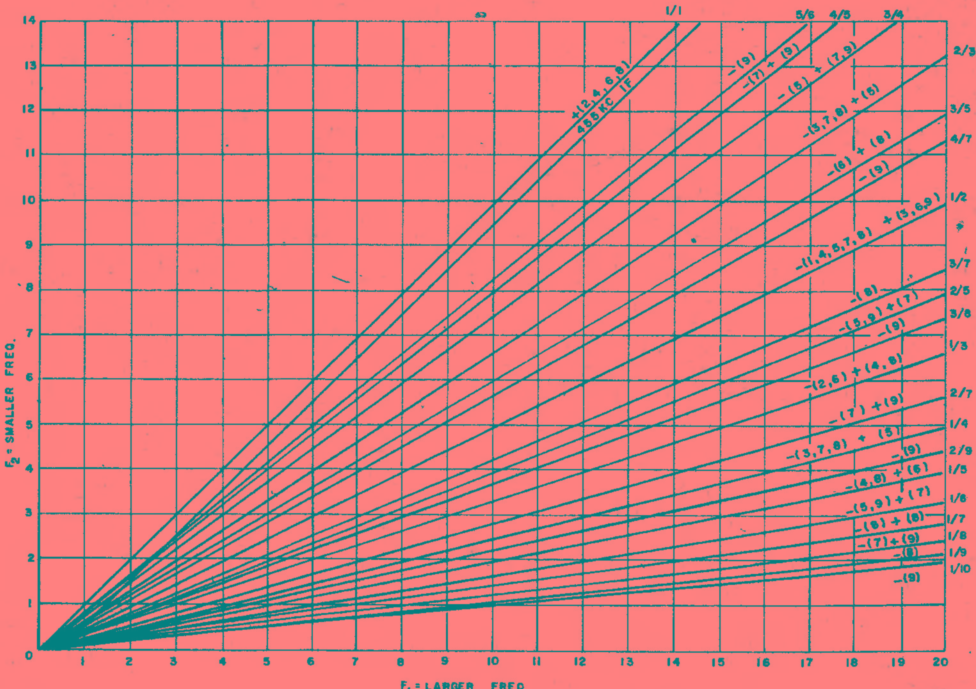


TABLE 1

$F_2 \sim F_1$

ORDER	1	2	3	4	5	6	7	8	9
1/1		•20 •02		•13 •31		•24 •42		•35 •53	
1/2	10		•12 •30	31	32	•33 •51	52	53	•54 •72
1/3		20		•22 •40		42 51		•53 •71	
1/4			30		•32 •50		52	71	
1/5				40		•42 •60		62	
1/6					50		•52 •70		72
1/7						60		•62 •80	
1/8							70		•72 •90
1/9								80	
1/10									90
2/3			21		•23 •41		43	53	

TABLE 2

$F_2 \sim F_1$

ORDER	1	2	3	4	5	6	7	8	9
2/5				41			•43 •61		63
2/7							61		•63 •81
2/9									81
3/4					32		•34 •52		54
3/5						42		•44 •62	
3/7								62	
3/8									72
4/5							43		•45 •63
4/7									63
5/6									54

• INDICATES SUM MIXING
OTHER — DIFF MIXING

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Fig. 8-13 — Chart to aid in the calculation of spurious frequencies generated during the mixing process.

To demonstrate the use of the chart, suppose an amateur wanted to mix a 6- to 6.5-MHz VFO output with a 10-MHz ssb signal to obtain output in the 80-meter band (the same problem as with a receiver that tunes 3.5 to 4 MHz, using a 6- to 6.5-MHz VFO to heterodyne to a 10-MHz i-f). Thus, F1 is 10 MHz and F2 is 6 to 6.5 MHz. Examination of the chart shows the intersection of these frequencies to be near the lines marked 2/3 and 3/5. In the case of the transmitter, difference (subtractive) mixing is to be used. The order of the products that will be close to the desired mixer output frequency is given on each line in parentheses. A plus sign in front of the parentheses indicates the product order in a sum (additive) mix, and a minus sign the order of a difference mix. For this example, the chart indicates the 3rd-, 7th-, and 8th-order products in a 2/3 relationship are going to be near the 80-meter band, plus the 6th-order product of the 3/5 relationship.

The exact frequencies of these products can be found with the help of the two small tables in Fig. 8-13. The product orders from 1 to 9 are given for all the product lines on the chart. The first digit of each group in a box is the harmonic of the lower frequency, F2, and the second digit is the harmonic of the larger frequency, F1. The dot indicates sum mixing and no dot indicates products in a difference mix. In the example, the chart shows that the 2/3 relationship will yield a 3rd-order product 2F2-F1, a 7th-order product 4F2-3F1, and an 8th-order product 5F2-3F1.

(Continued on next page)

(2 X 6) — 10 = 2
(2 X 6.5) — 10 = 3 (3rd order)

(4 X 6) — (3 X 10) = -6
(4 X 6.5) — (3 X 10) = -4 (7th order)

(5 X 6) — (3 X 10) = 0
(5 X 6.5) — (3 X 10) = 2.5 (8th order)

The 3/4 relationship produces a 6th-order product 4F2-2F1.

(4 X 6) — (2 X 10) = 4
(4 X 6.5) — (2 X 10) = 6

Thus, the ranges of spurious signals near the desired output band are 2 to 3 MHz, 6 to 4 MHz, 0 to 2.5 MHz, and 4 to 6 MHz. The negative sign indicates that the 7th-order product moves in the opposite direction to the normal output frequency, as the VFO is tuned. In this example proper mixer operation and sufficient selectivity following the mixer should keep the unwanted products sufficiently down in level without the use of filters or traps. Even-order products can be reduced by employing a balanced or doubly balanced mixer circuit, such as shown in Fig. 8-16.

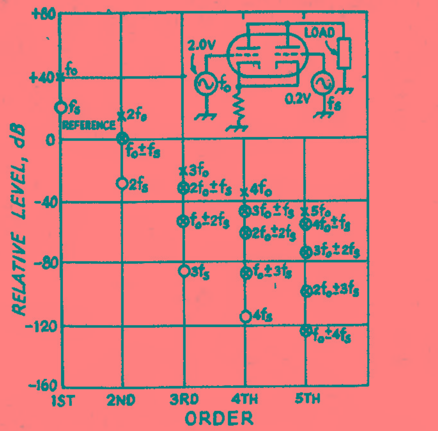


Fig. 8-14 — Chart showing the relative levels of spurious signals generated by e 12AU7A mixer.

The level of spurious products to be found in the output of a 12AU7 have been calculated by V. W. Bolie, using the assumption that the oscillator injection voltage will be 10 times (20 dB) greater than the input signal. This information is given in Fig. 8-14 for 1st- to 5th-order products. It is evident from the chart that multiples of the oscillator voltage produce the strongest of the undesired products. Thus, it follows that using a balanced-mixer design which reduces the level of oscillator signal in the output circuit will decrease the strength of the unwanted products.

MIXERS

A circuit tuned to the output frequency is placed in the plate circuit of the mixer, to offer a high impedance load for the output current that is developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are rejected by the selectivity of this circuit. The output tuned circuit should have low impedance for these frequencies, a condition easily met if neither is close to the output frequency.

The conversion efficiency of the mixer is the ratio of output voltage from the plate circuit to rf signal voltage applied to the grid. High conversion efficiency is desirable. The device used as a mixer also should be low noise if a good signal-to-noise ratio is wanted, particularly if the mixer is the first active device in the receiver.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called pulling. Pulling should be minimized, because the stability of the whole receiver or transmitter depends critically upon the stability of the hf oscillator. Pulling decreases with separation of the signal and hf-oscillator frequencies, being less with higher output frequencies. Another type of pulling is caused by lack of regulation in the power supply. Strong signals cause the voltage to change, which in turn shifts the oscillator frequency.

Circuits

If the mixer and high-frequency oscillator are separate tubes or transistors, the converter portion is called a "mixer." If the two are combined in one tube envelope (as is often done for reasons of economy or efficiency), the stage is called a "converter." In either case the function is the same.

Typical mixer circuits are shown in Figs. 8-15 and 8-16. The variations are chiefly in the way in which the oscillator voltage is introduced. In 8-15A, a pentode functions as a plate detector at the output frequency; the oscillator voltage is capacitance-coupled to the grid of the tube through C2. 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. The circuit is a sensitive one and makes a good mixer, particularly with high-transconductance tubes like the 6CY5, 6EJ7 or 6U8A (pentode section). Triode tubes can be used as mixers in grid-injection circuits, but they are commonly used at 50 MHz and higher, where mixer noise may become a significant factor. The triode mixer has the lowest inherent noise, the pentode is next, and the multigrid converter tubes are the noisiest.

In the circuit of Fig. 8-15A the oscillator voltage could be introduced at the cathode rather than at the control grid. If this were done, C3 would have to be removed, and output from the oscillator would be coupled to the cathode of the mixer through a .001-μF capacitor. C2 would also be discarded. Generally, the same rules apply as when the tube uses grid injection.

It is difficult to avoid "pulling" in a triode or pentode mixer, and a pentagrid mixer tube provides much better isolation. A typical circuit is shown in Fig. 8-15B, and tubes like the 6BA7 or 6BE6 are commonly used. The oscillator voltage is introduced through an "injection" grid. Measurement of the rectified current flowing in R2 is used as a check for proper oscillator-voltage amplitude. Tuning of the signal-grid circuit can have little effect on the oscillator frequency because the injection grid is isolated from the signal grid by a screen grid that is at rf ground potential. The pentagrid mixer is much noisier than a triode or pentode mixer, but its isolating characteristics make it a very useful device.

Penagrid tubes like the 6BE6 or 6BA7 are somtimes used as "converters" performing the dual function of mixer and oscillator. The usual circuit resembles Fig. 8-15B except that the No. 1 grid connects to the top of a grounded parallel-tuned circuit by means of a larger grid-blocking capacitor, and the cathode (without R1 and C3) connects to a tap near the grounded end of the coil. This forms a Hartley oscillator circuit. Correct location of the cathode tap is indicated by the grid current; raising the tap increases the grid current because the strength of oscillation is increased.

The effectiveness of converter tubes of the type just described becomes less as the signal frequency is increased. Some oscillator voltage will be coupled to the signal grid through "space-charge" coupling, an effect that increases with frequency. If there is relatively little frequency difference between oscillator and signal, as for example a 14- or 28-MHz signal and an i-f of 455 kHz, this voltage can become considerable because the selectivity of the signal circuit will be unable to reject it. If the signal grid is not returned directly to ground, but instead is returned through a resistor or part of an agc system, considerable bias can be developed which will cut down the gain. For this reason, and to reduce image response, the i-f following the first converter of a receiver should be not less than 5 or 10 percent of the signal frequency.

Diodes, FETs, ICs, and bipolar transistors can be used as mixers. Examples are given in Figs. 8-15 and 8-16. A single-diode mixer is not shown here since its application is usually limited to circuits operating in the uhf region and higher. A discussion of diode mixers, plus a typical circuit, is given in Chapter 9.

Oscillator injection can be fed to the base or emitter elements of bipolar-transistor mixers, Fig. 8-15C. If emitter injection is used, the usual emitter bypass capacitor must be removed. Because the dynamic characteristics of bipolar transistors prevent them from handling high signal levels, FETs are usually preferred in mixer circuits, although they do not provide the high conversion gain available with bipolar mixers. FETs (Fig. 8-15D and E) have greater immunity to cross-modulation and overload than bipolar transistors, and offer nearly square-law performance. The circuit at D uses a junction FET, N-channel type, with oscillator injection being supplied to the source. The value of the source resistor should be adjusted to provide a bias of approximately 0.8 volts. This value offers a good compromise between conversion gain and good intermodulation-distortion characteristics. At this bias level a local-oscillator injection of approximately 1.5 volts is desirable for good conversion gain. The lower the oscillator-injection level, the lower the gain. High injection levels improve the mixers immunity to cross-modulation.

A dual-gate MOSFET is used as a mixer at E. Gate 2 is used for injecting the local-oscillator signal while gate 1 is supplied with signal voltage.

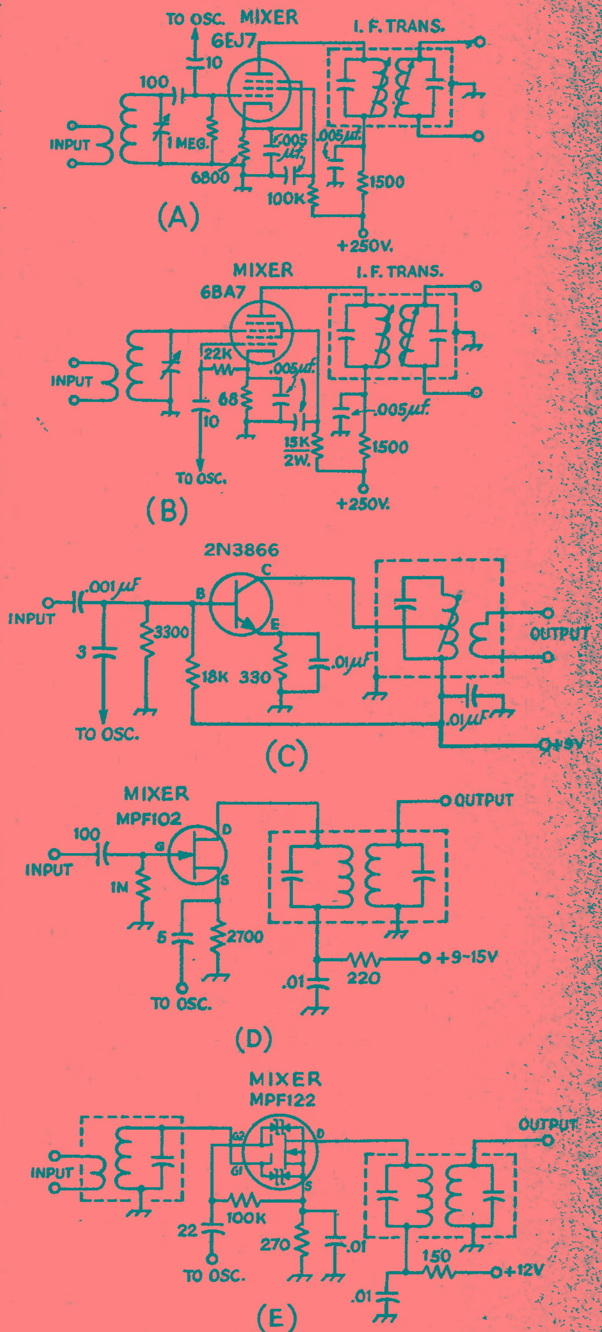


Fig. 8-15 — Typical single-ended mixer circuits.

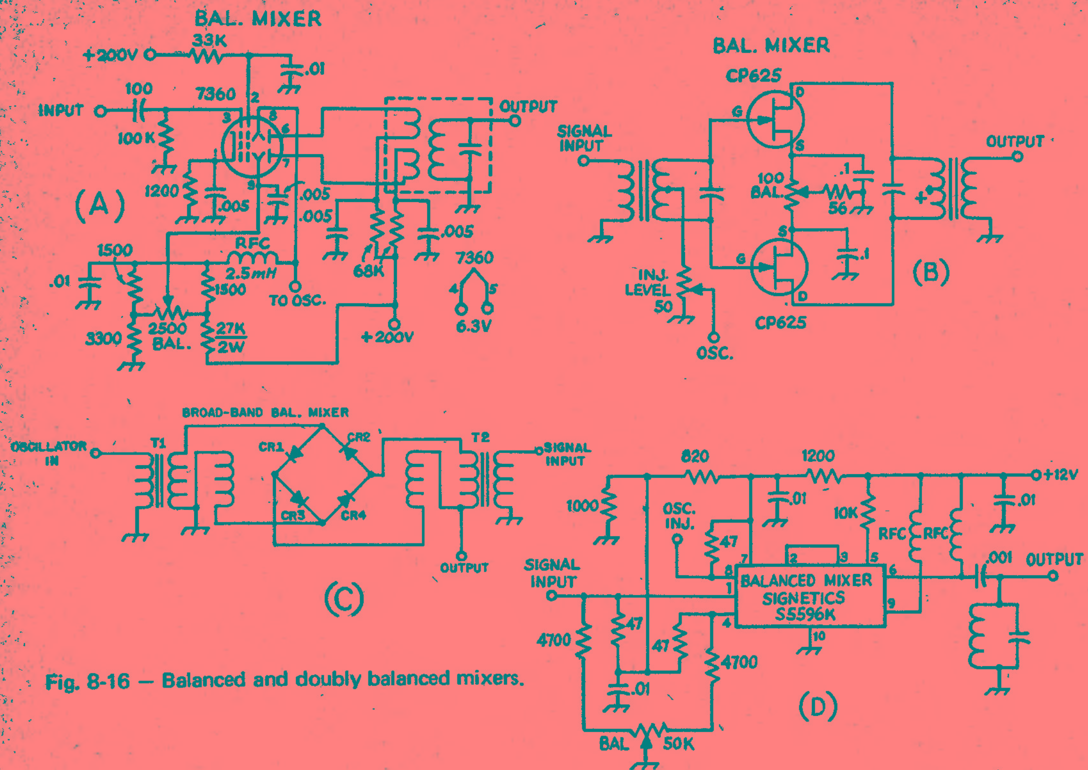


Fig. 8-16 — Balanced and doubly balanced mixers.

This type of mixer has excellent immunity to cross-modulation and overload. It offers better isolation between the oscillator and input stages than is possible with a JFET mixer. The mixers at D and E have high-Z input terminals, while the circuit at C has a relatively low-Z input impedance. The latter requires tapping the base down on the input tuned circuit for a suitable impedance match.

BALANCED MIXERS

The level of input and spurious signals contained in the output of a mixer may be decreased by using a balanced or doubly balanced circuit. The balanced mixer reduces leakthrough and even-order harmonics of one input (usually the local oscillator) while the doubly balanced designs lower the level of spurious signals caused by both the signal and oscillator inputs. One type of balanced mixer uses a 7360 beam-deflection tube, connected as shown in Fig. 8-16A. The signal is introduced at the No. 1 grid, to modulate the electron stream running from cathode to plates. The beam is deflected from one plate to the other and back again by the BFO voltage applied to one of the deflection plates. (If oscillator radiation is a problem, push-pull deflection by both deflection plates should be used.) At B, two CP625 FETs are used; these devices have a large dynamic range, about 130 dB, making them an excellent choice for either a transmitting or receiving mixer. Dc balance is set with a control in the source leads. The oscillator energy is introduced at the center tap of the input transformer.

In the circuit of Fig. 8-16C, hot-carrier diodes are employed as a broad-band balanced mixer. With careful winding of the toroid-core input and output transformers, the inherent balance of the mixer will provide 40- to 50-dB attenuation of the oscillator signal. The transformers, T1 and T2, having trifilar windings – using No. 32 enamel wire, 12 turns on a 1/2-inch core will provide operation on any frequency between 500 kHz and 100 MHz. Using Q3 cores the upper-frequency range can be extended to 300 MHz. CR1 to CR4, inc, comprise a matched quad of Hewlett-Packard HPA 5082-2805 diodes. Conversion loss in the mixer will be 6 to 8 dB.

Special doubly balanced mixer ICs are now available which can simplify circuit construction, as special balanced transformers are not required. Also, the ICs produce high conversion gain. A typical circuit using the Signetics S5596K is shown in Fig. 8-16D. The upper frequency limit of this device is approximately 130 MHz.

THE HIGH-FREQUENCY OSCILLATOR

Stability of the receiver is dependent chiefly upon the stability of the tunable hf oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to mechanical shock and changes in voltage and loading. Thermal effects (slow change in frequency because of tube, transistor, or circuit heating) should be minimized. See Chapter 6 for sample circuits and construction details.

THE INTERMEDIATE-FREQUENCY AMPLIFIER

One major advantage of the superhet is that high gain and selectivity can be obtained by using a good i-f amplifier. This can be a one-stage affair in simple receivers, or two or three stages in the more elaborate sets.

Choice of Frequency

The selection of an intermediate frequency is a compromise between 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. A low i-f also increases pulling of the oscillator frequency. On the other hand, a high i-f is beneficial to both image ratio and pulling, but the gain is lowered and selectivity is harder to obtain by simple means.

An i-f of the order of 455 kHz gives good selectivity and is satisfactory from the standpoint of image ratio and oscillator pulling at frequencies up to 7 MHz. The image ratio is poor at 14 MHz when the mixer is connected to the antenna, but adequate when there is a tuned rf amplifier between antenna and mixer. At 28 MHz and on the very high frequencies, the image ratio is very poor unless several rf stages are used. Above 14 MHz, pulling is likely to be bad without very loose coupling between mixer and oscillator. Tuned-circuit shielding also helps.

With an i-f of about 1600 kHz, satisfactory image ratios can be secured on 14, 21 and 28 MHz with one rf stage of good design. For frequencies of 28 MHz and higher, a common solution is to use double conversion, choosing one high i-f for image reduction (5 and 10 MHz 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 by the i-f wiring. Shifting the i-f or better shielding are the solutions to this interference problem.

Fidelity; Sideband Cutting

Amplitude modulation of a carrier generates sideband frequencies numerically equal to the carrier frequency plus and minus the modulation frequencies present. If the receiver is to give a faithful reproduction of modulation that contains, for instance, audio frequencies up to 5000 Hz, it must at least be capable of amplifying equally all frequencies contained in a band extending from 5000 Hz above or below the carrier frequency. In a superheterodyne, where all carrier frequencies are changed to the fixed intermediate frequency, the i-f amplification must be uniform over a band 5-kHz wide, when the carrier is set at one edge. If the carrier is set in the center, at 10-kHz band is required. The signal-frequency circuits usually do not have enough overall selectivity to affect materially the "adjacent-channel" selectivity, so that only the i-f-amplifier selectivity need be considered.

If the selectivity is too great to permit uniform amplification over the band of frequencies occupied by the modulated signal, some of the sidebands are "cut." While sideband cutting reduces fidelity, it is frequently preferable to sacrifice naturalness of reproduction in favor of communications effectiveness.

The selectivity of an i-f-amplifier, and hence the tendency to cut sidebands increases with the number of tuned circuits and also is greater the lower the intermediate frequency. From the standpoint of communication, sideband cutting is never serious with two-stage amplifiers at frequencies as low as 455 kHz. A two-stage i-f-amplifier at 85 or 100 kHz will be sharp enough to cut some of the higher frequency sidebands, if good transformers are used. However, the cutting is not at all serious, and the gain in selectivity is worthwhile in crowded amateur bands as an aid to QRM reduction.

Circuits

I-f amplifiers usually consist of one or more stages. The more stages employed, the greater the selectivity and overall gain of the system. In double-conversion receivers there is usually one stage at the first i-f, and sometimes as many as three or four stages at the second, or last, i-f. Most single-conversion receivers use no more than three stages of i-f amplification.

A typical vacuum-tube i-f stage is shown in Fig. 8-17 at A. The second or third stages would simply be duplicates of the stage shown. Remote cutoff pentodes are almost always used for i-f amplifiers, and such tubes are operated as Class-A amplifiers. For maximum selectivity, double-tuned transformers are used for interstage coupling, though single-tuned inductors and capacitive coupling can be used, but at a marked reduction in selectivity.

Agc voltage can be used to reduce the gain of the stage, or stages, by applying it to the terminal marked AGC. The agc voltage should be negative. Manual control of the gain can be effected by lifting the 100-ohm cathode resistor from ground and inserting a potentiometer between it and ground. A 10,000-ohm control can be used for this purpose. A small amount of B-plus voltage can be fed through a dropping resistor (about 56,000 ohms from a 250-volt bus) to the junction of the gain control and the 100-ohm cathode resistor to provide an increase in tube bias in turn reducing the mutual conduction of the tube for gain reduction.

An integrated-circuit i-f amplifier is shown at B. A positive-polarity agc voltage is required for this circuit to control the stage gain. If manual gain control provisions are desired, a potentiometer can be used to vary the plus voltage to the agc terminal of the IC. The control would be connected between the 9-volt bus and ground, its movable contact wired to the agc terminal of the IC.

A dual-gate MOSFET i-f amplifier is shown at B. Application of negative voltage to gate 2 of the

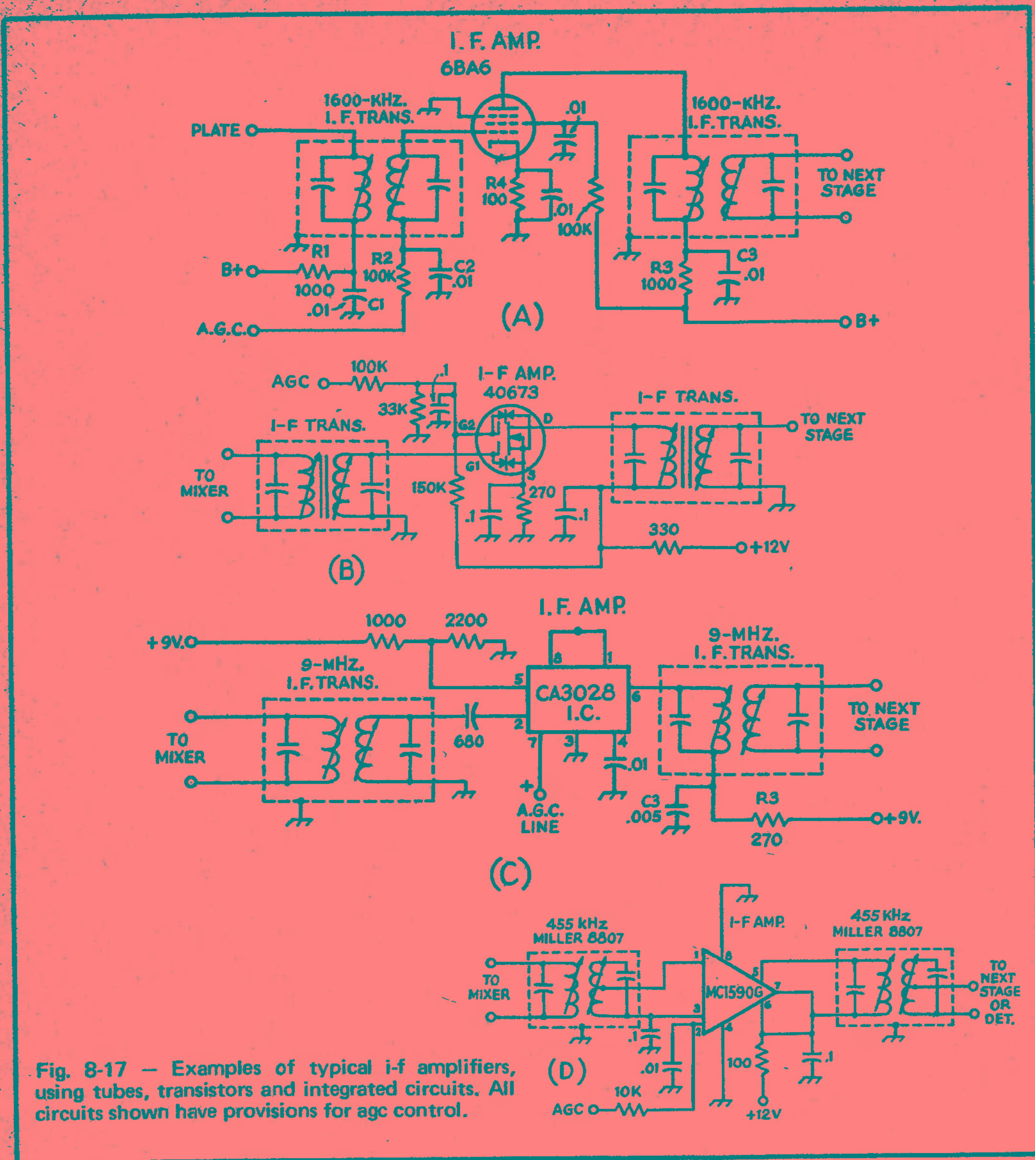


Fig. 8-17 — Examples of typical i-f amplifiers, using tubes, transistors and integrated circuits. All circuits shown have provisions for agc control.

device reduces the gain of the stage. To realize maximum gain when no agc voltage is present, it is necessary to apply approximately 3 volts of positive dc to gate 2. Neutralization is usually not required with a MOSFET in i-f amplifiers operating up to 20 MHz. Should instability occur, however, gate 1 and the drain may be tapped down on the i-f transformer windings.

High-gain linear ICs have been developed specifically for use as receiver i-f amplifiers. A typical circuit which uses the Motorola MC1590G is shown at D; 70 dB of gain may be achieved using this device. Agc characteristics of the IC are excellent. A 4-volt change at the agc terminal produces 60-dB change in the gain of the stage. Agc action starts at 5 volts, so a positive agc system with a fixed dc level must be employed.

Tubes for I-f Amplifiers

Variable-μ (remote cutoff) pentodes are almost invariably used in i-f amplifier stages, since grid-bias gain control 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 normally has no effect on the signal-to-noise ratio, since this is determined by the preceding mixer and rf amplifier.

The 6BA6, 6BJ6 and 6BZ6 are recommended for i-f work because they have desirable remote cutoff characteristics.

When two or more stages are used the high gain may tend to cause troublesome instability and oscillation, so that good shielding, bypassing, and careful circuit arrangement to prevent stray coupling between input and output circuits are necessary.

When vacuum tubes are used, the plate and grid leads should be well separated. When transistors are used, the base and collector circuits should be well isolated. With tubes it is advisable to mount the screen-bypass capacitor directly on the bottom of the socket, crosswise between the plate and grid pins, to provide additional shielding. As a further precaution against capacitive coupling, the grid and plate leads should be "dressed" close to the chassis.

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 capacitors are mounted. Both air-core and powered-iron-core universal-wound coils are used, the latter having somewhat higher *Q*s and hence greater selectivity and gain. In universal windings the coil is wound in layers with each turn traversing the length of the coil, back and forth, rather than being wound perpendicular to the axis as in ordinary single-layer coils. In a straight multilayer winding, a fairly large capacitance can exist between layers. Universal winding, with its "criss-crossed" turns, tends to reduce distributed-capacitance effects.

For tuning, air-dielectric tuning capacitors are preferable to mica compression types because their capacitance 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-capacitor tuning can be obtained by use of high-stability fixed mica or ceramic capacitors. 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.

The normal interstage i-f transformer is loosely coupled, to give good selectivity consistent with adequate gain. A so-called **diode transformer** is similar, but the coupling is tighter, to give sufficient transfer when working into the finite load presented by a diode detector. Using a diode transformer in place of an interstage transformer would result in loss of selectivity; using an interstage transformer to couple to the diode would result in loss of gain.

Besides the conventional i-f transformers just mentioned, special units to give desired selectivity characteristics have been used. For higher-than-ordinary adjacent-channel selectivity, triple-tuned transformers, with a third tuned circuit inserted between the input and output windings, have been made. 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.

Selectivity

The overall selectivity of the i-f amplifier will depend on the frequency and the number of stages. The following figures are indicative of the bandwidths to be expected with good-quality circuits in amplifiers so constructed as to keep regeneration at a minimum:

Tuned Ckts.	*Freq.*	*Circuit Q*	*Bandwidth, kHz −6 dB*	*−20 dB*	*−60 dB*
4	50 kHz	60	0.5	0.95	2.16
4	455 kHz	75	3.6	6.9	16
6	1600 kHz	90	8.2	15	34

THE BEAT OSCILLATOR AND DETECTOR

The detector in a superheterodyne receiver functions the same way as do the simple detectors described earlier in this chapter (Fig. 8-4), but usually operates at a higher input level because of the amplification ahead of it. The detectors of Fig. 8-4 are satisfactory for the reception of a-m signals. When copying cw and ssb signals it becomes necessary to supply a beat-oscillator (BFO) signal to the detector stage as described in the earlier section on product detectors. Suitable circuits for variable-frequency and crystal-controlled BFOs are given in Chapter 6.

AUTOMATIC GAIN CONTROL

Automatic regulation of the gain of the receiver in inverse proportion to the signal strength is an operating convenience in phone reception, since it tends to keep the output level of the receiver constant regardless of input-signal strength. The average rectified dc voltage, developed by the received signal across a resistance in a detector circuit, is used to vary the bias on the rf and i-f amplifier stages. 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 and the output more constant as the number of stages to which the agc bias is applied is increased. Control of at least two stages is advisable.

Carrier-Derived Circuits

A basic diode-detector/agc-rectifier circuit is given at Fig. 8-18A. Here a single germanium diode serves both as a detector and an agc rectifier, producing a negative-polarity agc voltage. Audio is taken from the return end of the i-f transformer secondary and is filtered by means of a 47,000-ohm resistor and two 470-pF capacitors.

At B, CR1 (also a germanium diode) functions as a detector while CR2 (germanium) operates as an agc rectifier. CR2 furnishes a negative agc voltage to the controlled stages of the receiver. Though solid-stage rectifiers are shown at A and B, vacuum-tube diodes can be used in these circuits. A 6AL5 tube is commonly used in circuits calling for two diodes (B), but a 1-megohm resistor should be shunted across the right-hand diode if a tube is used.

The circuit at C shows a typical hookup for agc feed to the controlled stages. S1 can be used to disable the agc when this is desired. For tube and FET circuits the value of R1 and R2 can be 100,000 ohms, and R3 can be 470,000 ohms. If bipolar transistors are used for the rf and i-f stages being controlled, R1 and R2 will usually be

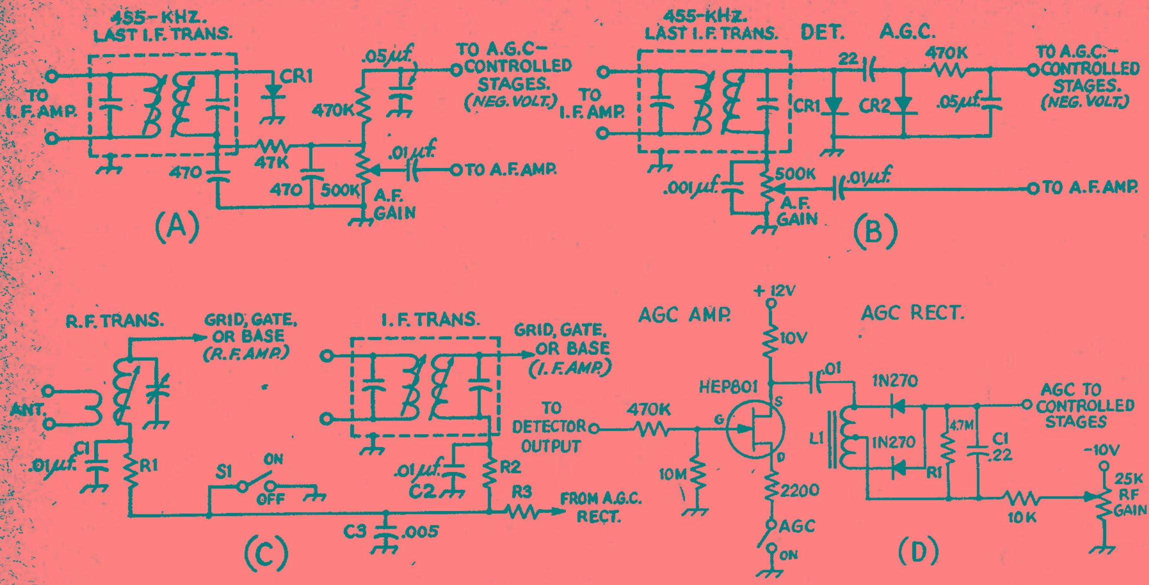


Fig. 8-18 — Methods for obtaining rectified voltage. At A the detector furnishes agc voltage. 8 shows separate diodes being used for the detector and agc circuits. C illustrates how negative agc voltage is fed to the rf and i-f stages of a typical receiver. D shows an audio-derived agc scheme. S1 is used to disable the agc when desired. R1, R2 and R3 in combination with C1, C2, and C3, are used for rf decoupling. Their values are dependent upon the device being used — tube or transistor. CR1 and CR2 at A and 8 are germenium diodes.

between 1000 and 10,000 ohms, depending upon the bias network required for the transistors used. R3 will also be determined by the bias value required in the circuit.

Agc Time Constant

The time constant of the resistor-capacitor combinations in the agc circuit is an important part of the system. It must be long enough so that the modulation on the signal is completely filtered from the dc output, leaving only an average dc component which follows the relatively slow carrier variations with fading. Audio-frequency variations in the agc voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal. But the time constant must not be too long or the agc will be unable to follow rapid fading. The capacitance and resistance values indicated in 8-18A will give a time constant that is satisfactory for average reception.

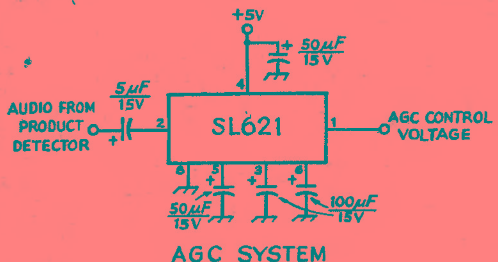


Fig. 8-19 — An IC agc system.

Cw and Ssb

Agc can be used for cw and ssb reception but the circuit is usually more complicated. The agc voltage must be derived from a rectifier that is isolated from the beat-frequency oscillator (otherwise the rectified BFO voltage will reduce the receiver gain even with no signal coming through). This is done by using a separate agc channel connected to an i-f amplifier stage ahead of the second detector (and BFO) or by rectifying the audio output of the detector. If the selectivity ahead of the agc rectifier isn't good, strong adjacent-channel signals may develop agc voltages that will reduce the receiver gain. When clear channels are available, however, cw and ssb agc will hold the receiver output constant over a wide range of signal inputs. Agc systems designed to work on these signals should have fast-attack and slow-decay characteristics to work satisfactorily, and often a selection of time constants is made available.

Audio-Derived Agc

Agc potential for use in a cw/ssb receiver may also be obtained by sampling the audio output of the detector and rectifying this signal. A typical circuit is shown in Fig. 8-18D. The JFET stage amplifies the audio signal; the output of the HEP801 is coupled to the secondary of an audio transformer, L1. The time constant of the agc line is established by R1C1. Manual gain control can be accomplished by adding a variable negative voltage to the common lead of the audio rectifier.

An improved audio-derived agc circuit is shown in Fig. 8-19, using the Plessey Microelectronics SL-621 integrated circuit. This design provides the fast-attack, slow-decay time constant required for ssb reception. High-level pulse signals that might "hang up" the agc system are sampled by the IC input circuit, activating a trigger which provides a fast-discharge path for the time-constant capacitor. Thus, noise bursts will not produce a change in the level of agc output voltage.

NOISE REDUCTION

Types of Noise

In addition to tube and circuit noise, much of the noise interference experienced in reception of high-frequency signals is caused by domestic or industrial 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 dc and series-wound ac motors, while the "shot" type results from separated spark discharges (ac power leaks, switch and key clicks, ignition sparks, and the like).

The only known approach to reducing tube and circuit noise is through the choice of low-noise front-end active components and through more overall selectivity.

Impulse Noise

Impulse noise, because of the short duration of the pulses compared with the time between them, must have high 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 to allow the desired signal to pass through the receiver unaffected, but to make the receiver inoperative for amplitudes greater than that of the signal. The greater the amplitude of the pulse compared with its time of duration, the more successful the noise reduction.

Another approach is to "silence" (render inoperative) the receiver during the short duration time of any individual pulse. The listener will not hear the "hole" because of its short duration, and very effective noise reduction is obtained. Such devices are called "blankers" rather than "limiters."

In passing through selective receiver circuits, the time duration of the impulses is increased, because of the *Q* of the circuits. Thus, the more selectivity ahead of the noise-reducing device, the more difficult it becomes to secure good pulse-type noise suppression. See Fig. 8-22.

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 during fading. These output-limiter systems are simple, and they are readily adaptable to most receivers without any modification of the receiver itself. However, they cannot prevent noise peaks from overloading previous stages.

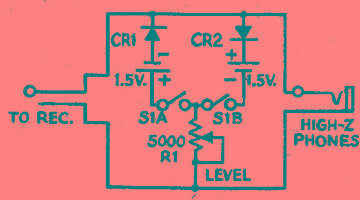


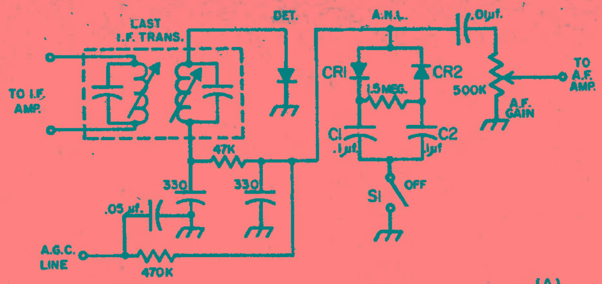
Fig. 8-20 — Circuit of a simple audio limiter/clipper. It can be plugged into the headphone jack of the receiver. R1 sets the bias on the diodes, CR1 and CR2, for the desired limiting level. S1 opens the battery leads when the circuit is not being used. The diodes can ba 1N34As or similar.

NOISE-LIMITER CIRCUITS

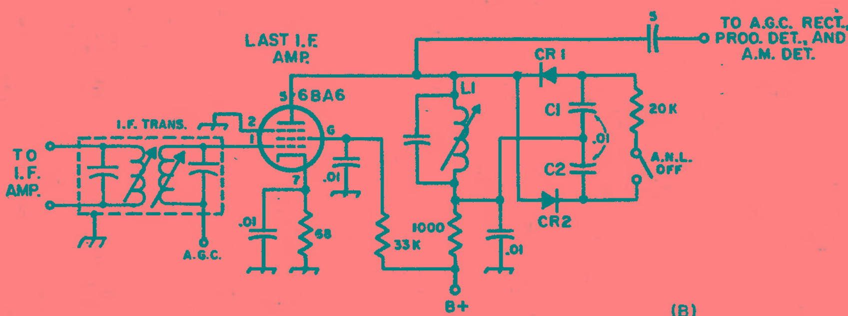
Pulse-type noise can be eliminated to an extent which makes the reception of even the weakest of signals possible. The noise pulses can be clipped, or limited in amplitude, at either an rf or af point in the receiver circuit. Both methods are used by receiver manufacturers; both are effective.

A simple audio noise limiter is shown at Fig. 8-20. It can be plugged into the headphone jack of the receiver and a pair of headphones connected to the output of the limiter. CR1 and CR2 are wired to clip both the positive and negative peaks of the audio signal, thus removing the high spikes of pulse noise. The diodes are back-biased by 1.5-volt batteries to permit R1 to serve as a clipping-level control. This circuit also limits the amount of audio reaching the headphones. When tuning across the band, strong signals will not be ear-shattering and will appear to be the same strength as the weaker ones. S1 is open when the circuit is not in use to prevent battery drain. CR1 and CR2 can be germanium or silicon diodes, but 1N34As are generally used. This circuit is usable only with high-impedance headphones.

The usual practice in communications receivers is to use low-level limiting, Fig. 8-21. The limiting can be carried out at rf or af points in the receiver, as shown. Limiting at rf does not cause poor audio quality as is sometimes experienced when using series or shunt af limiters. The latter limits the normal af signal peaks as well as the noise pulses,



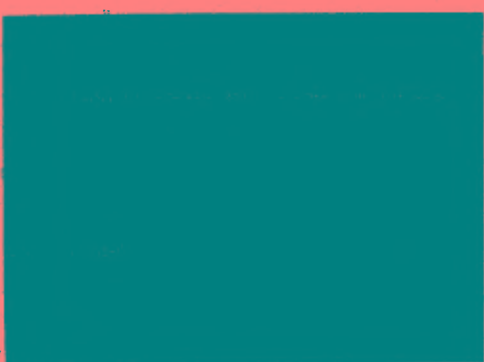
A.F. SHUNT (A)



(B)

R.F. SHUNT

Fig. 8-21 — Typical rf and af anl circuits. A shows the circuit of a self-adjusting af noisa limiter. CR1 and CR2 are self-biased silicon diodes which limit both the positive and negative audio and noise-pulse peaks. S1 turns tha limiter on or off; 8 shows an rf limiter of the same type as A, but this circuit clips the positive and negative rf peaks and is connected to the last i-f stage. This clrcuit does not degrade the audio quality of the signal as doas the circuit of A.

(A) 

 (B)

(C) 

Fig. 8-22 — The delay and lengthening of a noise pulse when passed through a 2-kHz wide amplifier with good skirt selectivity (4 kHz at −60 d8). (B) a 3.75-MHz carrier modulated 30 percent, interfered with by noise pulses. The noise pulses were originally 1000 times the amplitude of the signal; they have been reduced (and lengthened) by overload in the i-f. The i-f bandwidth is 5 kHz. Sweep speed = 1 millisecond/cm. (C) Same as B but with a noise blanker on.

giving an unpleasant audio quality to strong signals.

In a series-limiting circuit, a normally conducting element (or elements) is connected in the circuit in series and operated in such a manner that it becomes nonconductive above a given signal level. In a shunt limiting circuit, a nonconducting element is connected in shunt across the circuit and operated so that it becomes conductive above a given signal level, thus short-circuiting the signal and preventing its being transmitted to the remainder of the amplifier. The usual conducting element will be a forward-biased diode, and the usual nonconducting element will be a back-biased diode. In many applications the value of bias is set manually by the operator; usually the clipping level will be set at about 1 to 10 volts.

The af shunt limiter at A, and the rf shunt limiter at B operate in the same manner. A pair of self-biased diodes are connected across the af line at A, and across an rf inductor at B. When a steady cw signal is present the diodes barely conduct, but when a noise pulse rides in\on the incoming signal, it is heavily clipped because capacitors C1 and C2 tend to hold the diode bias constant for the duration of the noise pulse. For this reason the diodes conduct heavily in the presence of noise and maintain a fairly constant signal output level. Considerable clipping of cw signal peaks occurs with this type of limiter, but no apparent deterioration of the signal quality results. L1 at C is tuned to the i-f of the receiver. An i-f transformer with a conventional secondary winding could be used in place of L1, the clipper circuit being connected to the secondary winding; the plate of the 6BA6 would connect to the primary winding in the usual fashion.

I-F NOISE SILENCER

The i-f noise silencer circuit shown in Fig. 8-23 is designed to be used ahead of the high-selectivity section of the receiver. Noise pulses are amplified and rectified, and the resulting negative-going dc pulses are used to cut off an amplifier stage during the pulse. A manual "threshold" control is set by the operator to a level that only permits rectification of the noise pulses that rise above the peak amplitude of the desired signal. The clamp transistor, Q3, short circuits the positive-going pulse "overshoots." Running the 40673 controlled i-f amplifier at zero gate 2 voltage allows the direct application of agc voltage. See July 1971 *QST* for additional details.

SIGNAL-STRENGTH AND TUNING INDICATORS

It is convenient to have some means by which to obtain *relative* readings of signal strength on a communications receiver. The actual meter readings in terms of S units, or decibels above S9, are of little consequence as far as a meaningful report to a distant station is concerned. Few signal-strength meters are accurate in terms of decibels, especially across their entire indicating range. Some manufacturers once established a standard in which a certain number of microvolts were equal to S9 on the meter face. Such calibration is difficult to maintain when a number of different receiver circuits are to be used. At best, a meter can be calibrated for one receiver – the one in which it will be used. Therefore, most S meters are good only as relative indicating instruments for comparing the strength of signals at a given time, on a given amateur band. They are also useful for "on-the-nose-tuning" adjustments with selective receivers. If available, a signal generator with an accurate output attenuator can be used to calibrate an S meter in terms of microvolts, but a different calibration chart will probably be required for each band because of probable differences in receiver sensitivity from band to band. It is helpful to establish a 50-μV reading at midscale on the meter so that the very strong signals will crowd the high end of the meter scale. The weaker signals will then be spread over the lower half of the scale and will not be compressed at the low end. Midscale on the meter can be called S9. If S units are desired across the scale, below S9, a marker can be established at every 6 dB point.

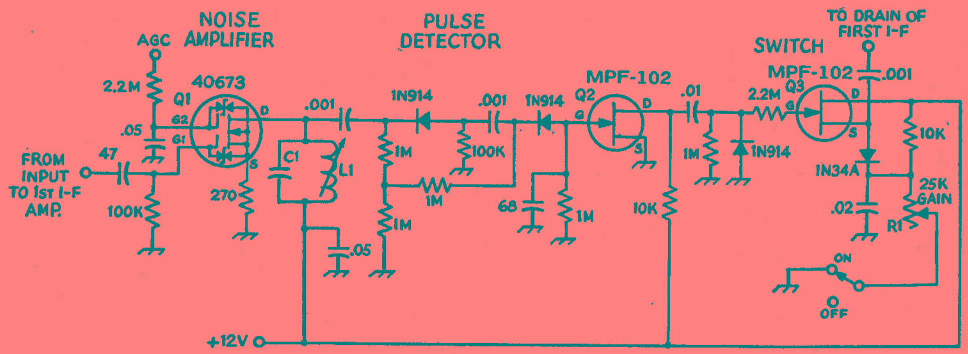
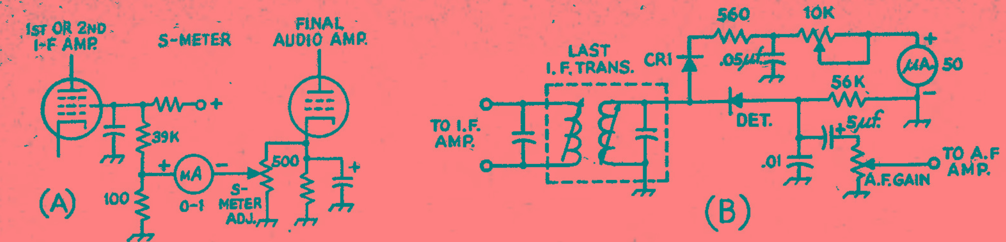


Fig. 8-23 – Diagram of the noise blanker. L1 and C1 ara chosen to resonata at the desired i-f.



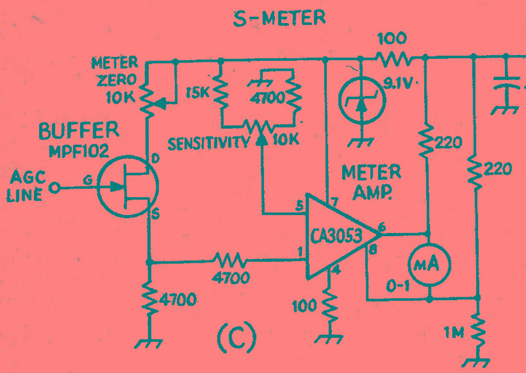


Fig. 8-24 – Practical examples of S-meter circuits. At A, a meter measures the change in screen voltage of en i-f amplifier stage, caused by changes in the agc voltage epplied to the grid of the amplifier. At B, the i-f signel is rectified by CR1 and is fed to a 50-μA meter. A 10,000-ohm control sets the sensitivity and also functions as e "linearizing" resistor to make the meter less subject to the square-law response of CR1. At C, an FET semples the voltage on the agc line end drives an IC amplifier which provides the required current swing to operate a 1-mA meter.

S-METER CIRCUITS

A very simple meter indicator is shown at Fig. 8-24B. Rectified i-f is obtained by connecting CR1 to the take-off point for the detector. The dc is filtered by means of a 560-ohm resistor and a .05-μF capacitor. A 10,000-ohm control sets the meter at zero reading in the absence of a signal and also serves as a "linearizing" resistor to help compensate for the nonlinear output from CR1. The meter is a 50-μA unit, therefore consuming but a small amount of current from the output of the i-f.

Another simple approach is to meter the change in screen voltage of an i-f amplifier stage. The swing in screen potential is caused by changes in the agc voltage applied to the stage. A reference voltage is obtained from the cathode of the audio-output stage. A 1-mA meter is suitable for the circuit shown in Fig. 8-24A. At C, a more complex design is employed which can operate directly from the agc line of a transistorized receiver. The sensitivity of the metering circuit is adjusted by changing the gain of the IC meter amplifier. An FET buffer is employed to insure that loading of the agc line will be negligible.

IMPROVING RECEIVER SELECTIVITY

INTERMEDIATE-FREQUENCY AMPLIFIERS

One of the big advantages of the superheterodyne receiver is the improved selectivity that is possible. This selectivity is obtained in the i-f amplifier, where the lower frequency allows more selectivity per stage than at the higher signal frequency. For normal a-m (double-sideband) reception, the limit to useful selectivity in the i-f amplifier is the point where too many of the high-frequency sidebands are lost. The limit to selectivity for a single-sideband signal, or a double-sideband a-m signal treated as an ssb signal, is about 2000 Hz, but reception is much more normal if the bandwidth is opened up to 2300 or 2500 Hz. The correct bandwidth for fm or pm reception is determined by the deviation of the received signal; sideband cutting of these signals results in distortion. The limit to useful selectivity in code work is around 150 or 200 Hz for hand-key speeds, but this much selectivity requires excellent stability in both transmitter and receiver, and a slow receiver tuning rate for ease of operation.

Single-Signal Effect

In heterodyne cw (or ssb) 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 454 kHz (the i-f being 455 kHz) to give a 1000-Hz beat note. Now, if an interfering signal appears at 453 kHz or if the receiver is tuned to heterodyne the incoming signal to 453 kHz, it will also be heterodyned by the beat oscillator to produce a 1000-Hz beat. Hence every signal can be tuned in at two places that will give a 1000-Hz beat (or any other low audio frequency). The audio-frequency image effect can be reduced if the i-f selectivity is such that the incoming signal, when heterodyned to 453 kHz, is attenuated to a very low level.

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 not obtained with nonregenerative amplifiers using ordinary tuned circuits unless a low i-f, or a large number of circuits, is used.

Regeneration

Regeneration can be used to give a single-signal effect, particularly when the i-f is 455 kHz or lower. The resonance curve of an i-f stage at critical regeneration (just below the oscillating point) is extremely sharp, a bandwidth of 1 kHz at 10 times down and 5 kHz 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-Hz beat note (image 2000 Hz 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 feedback may be controlled by a cathode-resistor gain control. When the i-f is regenerative, it is preferable to operate the tube at reduced gain (high bias) and depend on regeneration to bring up the signal strength. This prevents overloading and increases 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. However, the regenerative gain varies with signal strength, being less on strong signals.

Crystal Filters; Phasing

A simple means for obtaining high selectivity is by the use of a piezoelectric quartz crystal as a selective filter in the i-f amplifier. Compared to a good tuned circuit, the *Q* of such a crystal is extremely high. The crystal is ground resonant at the i-f and used as a selective coupler between i-f stages. For single-signal reception, the audio-frequency image can be reduced by 50 dB or more. Besides practically eliminating the af image, the high selectivity of the crystal filter provides good discrimination against adjacent signals and also reduces the broadband noise.

BAND-PASS FILTERS

A single high-*Q* circuit (e.g., a quartz crystal or regenerative stage) will give adequate single-signal cw reception under most circumstances. For phone reception, however, either single-sideband or a-m, a band-pass characteristic is more desirable. A band-pass filter is one that passes without unusual attenuation a desired *band* of frequencies and rejects signals outside this band. A good band-pass filter for single-sideband reception might have a bandwidth of 2500 Hz at -6 dB and 4 kHz at -60 dB; a filter for a-m would require twice these bandwidths if both sidebands were to be accommodated, thus assuring suitable fidelity.

The simplest band-pass crystal filter is one using two crystals, as in Fig. 8-25A. The two crystals are separated slightly in frequency. If the frequencies are only a few hundred Hz apart the characteristic is a good one for cw reception. With crystals about 2 kHz apart, a reasonable phone characteristic is obtained. Fig. 8-2 shows a selectivity characteristic of an amplifier with a bandpass (at -6 dB) of 2.4 kHz, which is typical of what can be expected from a two-crystal bandpass filter.

More elaborate crystal filters, using four and six crystals, will give reduced bandwidth at -60 dB without decreasing the bandwidth at -6 dB. The resulting increased "skirt selectivity" gives better rejection of adjacent-channel signals. "Crystal-lattice" filters of this type are available commercially for frequencies up to 40 MHz or so, and they have also been built by amateurs from inexpensive transmitting-type crystals. (See Vester, "Surplus-Crystal High-Frequency Filters," *QST*, January, 1959; Healey, "High-Frequency Crystal Filters for SSB," *QST*, October, 1960.)

Two half-lattice filters of the type shown at Fig. 8-25A can be connected back to back as shown at B. The channel spacing of Y1 and Y2 will depend upon the receiving requirements as discussed in the foregoing text. Ordinarily, for ssb reception (and nonstringent cw reception) a frequency separation of approximately 1.5 kHz is suitable. The overall i-f strip of the receiver is tuned to a frequency which is midway between Y1 and Y2. C1 is tuned to help give the desired shape to the passband. L1 is a bifilar-wound toroidal inductor which tunes to the i-f frequency by means of C1. The values of R1 and R2 are identical and are determined by the filter response desired. Ordinarily the ohmic value is on the order of 600 ohms, but values as high as 5000 ohms are sometimes used. The lower the value of resistance, the broader and flatter will be the response of the filter. Though the circuit at B is shown in a transistorized circuit, it can be used with vacuum tubes or integrated circuits as well. The circuit shows an i-f frequency of 9 MHz, but the filter can be used at any desired frequency below 9 MHz by altering the crystal frequencies and the tuned circuits. Commercial versions of the 9-MHz lattice filter are available at moderate cost.[1] War-surplus FT-241 crystals in the 455-kHz range are inexpensive and lend themselves nicely to this type of circuit.

Mechanical filters can be built at frequencies below 1 MHz. They are made up of three sections; an input transducer, a mechanically resonant filter

[1] Spectrum International, P. O. Box 87, Topsfield, MA 01983. Also, McCoy Electronics Co., Mount Holly Springs, PA.

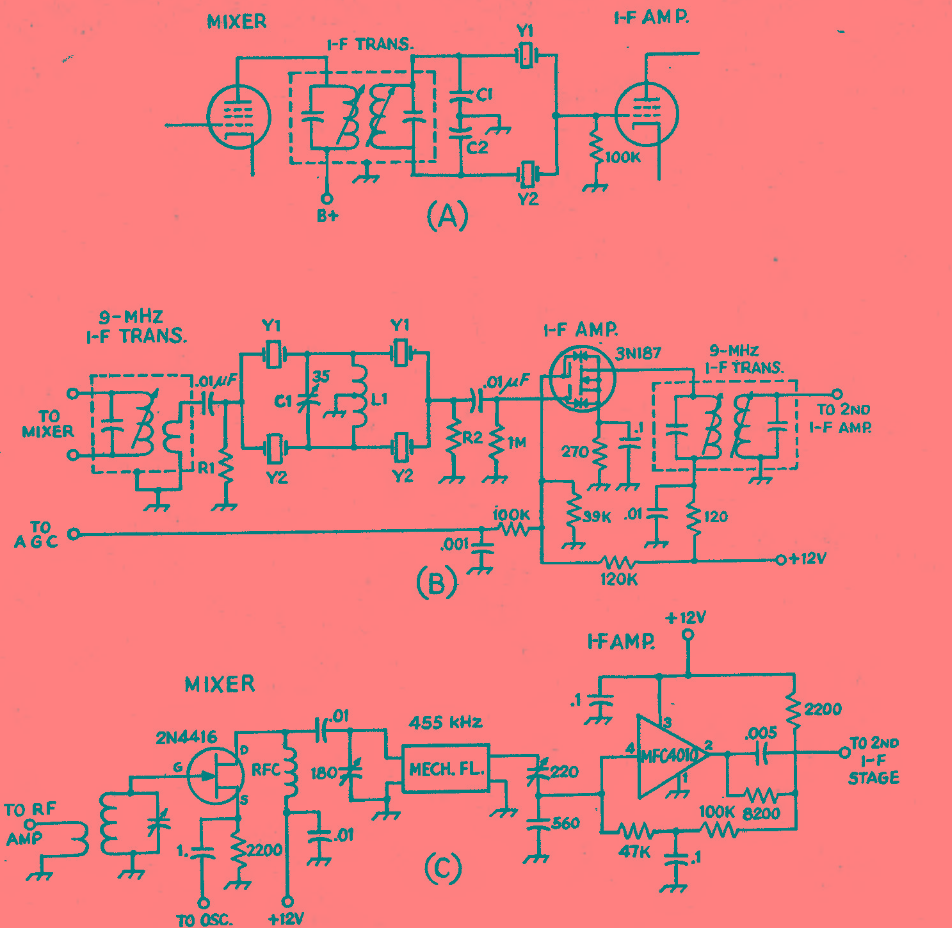


Fig. 8-25 — A half-lattice bandpass filter at A; B shows two half-lattice filters in cascade; C shows a mechanical filter.

section, and an output transducer. The transducers use the principle of magneto-striction to convert the electrical signal to mechanical energy, then back again. The mechanically resonant section consists of carefully machined metal disks supported and coupled by thin rods. Each disk has a resonant frequency dependent upon the material and its dimensions, and the effective *Q* of a single disk may be in excess of 2000. Consequently, a mechanical filter can be built for either narrow or broad passband with a nearly rectangular curve. Mechanical filters are available commercially and are used in both receivers and single-sideband transmitters. They are moderately priced.

The signal-handling capability of a mechanical filter is limited by the magnetic circuits to from 2 to 15 volts rms, a limitation that is of no practical importance provided it is recognized and provided for. Crystal filters are limited in their signal-handling ability only by the voltage breakdown limits, which normally would not be reached before the preceding amplifier tube was overloaded. A more serious practical consideration in the use of any high-selectivity component is the prevention of coupling "around" the filter, externally, which can only degrade the action of the filter.

The circuit at Fig. 8-25C shows a typical hookup for a mechanical filter. FL1 is a Collins 455-FB-21, which has an ssb band-pass characteristic of 2.1 kHz. It is shown in a typical solid-state receiver circuit, but can be used equally as well in a tube-type application.

Placement of the BFO signal with respect to the passbands of the three circuits at A, B, and C, is the same. Either a crystal-controlled or self-excited oscillator can be used to generate the BFO signal and the usual practice is to place the BFO signal at a frequency that falls at the two points which are approximately 20 dB down on the filter curve, dependent upon which sideband is desired. Typically, with the filter specified at C, the center frequency of FL1 is 455 kHz. To place the BFO at the 20-dB points (down from the center-frequency peak) a signal at 453 and 456 kHz is required.

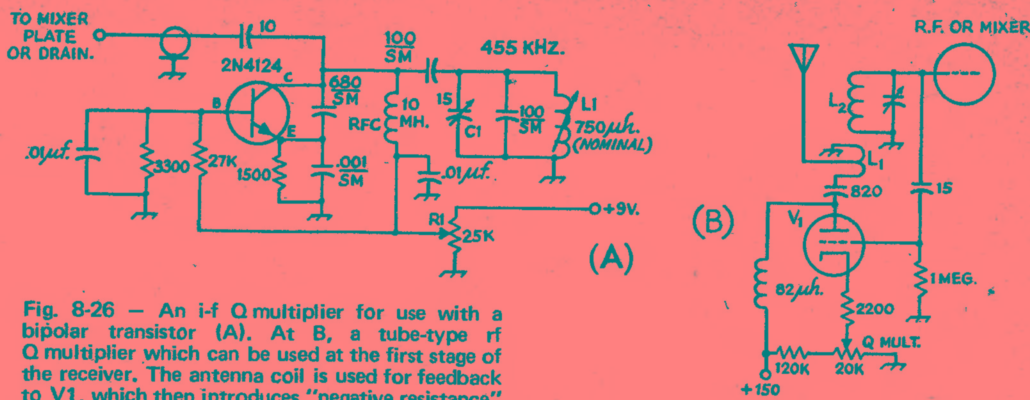


Fig. 8-26 — An i-f Q multiplier for use with a bipolar transistor (A). At B, a tube-type rf Q multiplier which can be used at the first stage of the receiver. The antenna coil is used for feedback to V1, which then introduces "negative resistance" to L2.

Q Multiplier

The "*Q* Multiplier" is a stable regenerative stage that is connected in parallel with one of the i-f stages of a receiver. In one condition it narrows the bandwidth and in the other condition it produces a sharp "null" or rejection notch. A "tuning" adjustment controls the frequency of the peak or null, moving it across the normal passband of the receiver i-f amplifier. The *shape* of the peak or null is always that of a single tuned circuit (Fig. 2-42) but the effective *Q* is adjustable over a wide range. A *Q* Multiplier is most effective at an i-f of 500 kHz or less; at higher frequencies the rejection notch becomes wide enough (measured in Hz) to reject a major portion of a phone signal. Within its useful range, however, the *Q* Multiplier will reject an interfering carrier without degrading the quality of the desired signal.

In the "peak" condition the *Q* Multiplier can be made to oscillate by advancing the "peak" (regeneration) control far enough and in this condition it can be made to serve as a beat-frequency oscillator. However, it cannot be made to serve as a selective element and as a BFO *at the same time.* Some inexpensive receivers may combine either a *Q* Multiplier or some other form of regeneration with the BFO function, and the reader is advised to check carefully any inexpensive receiver he intends to buy that offers a regenerative type of selectivity, in order to make sure that the selectivity is available when the BFO is turned on.

A representative circuit for a transistorized *Q*-multiplier is given in Fig. 8-26A. The constants given are typical for i-f operation at 455 kHz. L1 can be a J. W. Miller 9002 or 9102 slug-tuned inductor. A 25,000-ohm control, R1, permits adjustment of the regeneration. C1 is used to tune the *Q*-multiplier frequency back and forth across the i-f passband for peaking or notching adjustments. With circuits of this type there is usually a need to adjust both R1 and C1 alternately for a peaking or notching effect, because the controls tend to interlock as far as the frequency of oscillation is concerned. A *Q*-multiplier should be solidly built in a shielded enclosure to assure maximum stability.

Q multipliers can be used at the front end of a receiver also, as shown at B in Fig. 8-26. The enhancement of the *Q* at that point in a receiver greatly reduces image problems because the selectivity of the input tuned circuit is increased markedly. The antenna coil, L1, is used as a feedback winding to make V1 regenerative. This in effect adds "negative resistance" to L2, increasing its *Q*. A 20,000-ohm control sets the regeneration of V1, and should be adjusted to a point just under regeneration for best results. Rf *Q* multiplication is not a cure for a poor-quality inductor at L2, however.

T-Notch Filter

At low intermediate frequencies (50 – 100 kHz) the T-notch filter of Fig. 8-27 will provide a sharp tunable null.

The inductor L resonates with C at the rejection frequency, and when $R = QX_L/4$ the rejection is maximum. (X_L is the coil-reactance and Q is the coil Q.) In a typical 50-kHz circuit, C might be 3900 pF making L approximately 2.6 mH. When R is greater than the maximum-attenuation value, the circuit still provides some rejection, and in use the inductor is detuned or shorted out when the rejection is not desired.

At higher frequencies, the T-notch filter is not sharp enough with available components to reject only a narrow band of frequencies.

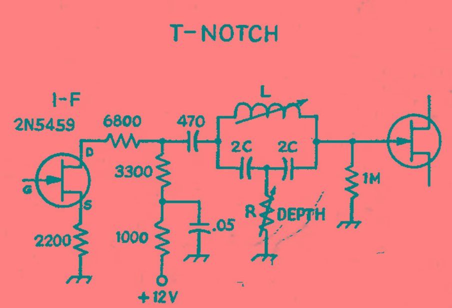


Fig. 8-27 — Typical T-notch (bridged-T) filter, to provide a sharp notch at a low i-f. Adjustment of L changes the frequency of the notch; adjustment of R controls the notch depth.

RADIO-FREQUENCY AMPLIFIERS

While selectivity to reduce audio-frequency images cau be built into the i-f amplifier, discrimination against radio-frequency images can only be obtained in tuned circuits or other selective elements ahead of the first mixer or converter stage. These tuned circuits are usually used as the coupling networks for one or more vacuum tubes or transistors, aud the combinations of circuits aud amplifying devices are called radio-frequency amplifiers. The tuned circuits contribute to the rf-image rejection and the amplifying device(s) determines the noise figure of the receiver.

Knowing the Q of the coil in each tuned circuit between the antenna and the first mixer or converter stage, the image rejection capability can be computed by using the chart in Fig. 2-50. The Q of the input tuned circuit (coupled to the antenna) should be taken as about one-half the unloaded Q of that circuit, aud the Q of auy other tuned circuit can be assumed to be the unloaded Q to a first approximation (the vacuum tubes will reduce the circuit Q to some extent, especially at 14 MHz and higher).

In general, receivers with an i-f of 455 kHz can be expected to have some noticeable image response at 14 MHz and higher if there are only two tuned circuits (or one rf stage) ahead of the mixer or converter. Regeneration in the rf amplifier will reduce image response, but regeneration usually requires frequent readjustment when tuning across a baud. Regeneration is, however, a useful device for improving the selectivity of au rf amplifier without requiring a multiplicity of tuned circuits.

With three tuned circuits between the antenna and the first mixer, and an i-f of 455 kHz, no images should be encountered up to perhaps 25 MHz. Four tuned circuits or more will eliminate any images at 28 MHz when an i-f of 455 kHz is used.

Obviously, a better solution to the rf selectivity problem (elimination of image response) is to use an i-f higher than 455 kHz, and most modern receivers use an i-f of 1600 kHz or higher. The owner of a receiver with a 455-kHz i-f amplifier can enjoy image-free reception on the higher frequencies by using a crystal-controlled converter ahead of the receiver and utilizing the receiver as a "tunable i-f amplifier" at 3.5 or 7.0 MHz.

For best selectivity rf amplifiers should use high-Q circuits and tubes with high input and output resistance. Variable-μ pentodes and field-effect transistors (JFET and MOSFET) are practically always used, although triodes (neutralized or otherwise connected so that they won't oscillate) are often used on the higher frequencies because they introduce less noise. However, their lower plate resistance will load the tuned circuits. Pentodes and FETs are better where maximum image rejection is desired, because they have less loading effect on the tuned circuits.

Representative Circuits

An example of a typical vacuum-tube rf amplifier using a remote-cutoff pentode aud agc is given in Fig. 8-28 at A. The manual rf gain control, R1, varies the bias on the stage, thereby changing the gain of the tube.

In the circuit at B, two junction field-effect transistors are used as a cascade rf amplifier. If sufficient isolation is provided between the input and output tuned circuits, neutralization is seldom required below 30 MHz. Agc potential is applied to the gate of the second JFET. For efficient operation as an rf amplifier, the transistors chosen should have an f_T rating somewhat above the desired operating frequency.

A dual-gate MOSFET with built-in transient protection is used in the circuit at C. Negative agc voltage is applied to gate 2. Zener diodes contained within the 40673, bridged between the gates and the source/substrate connection, provide protection from transient voltages (up to 10 V pk-pk) that might otherwise damage the device. The gate-1 aud drain connections are tapped down on their associated tuned circuits, a technique which produces stability without neutralization with only a slight reduction in gain.

The rf amplifier shown in Fig. 8-28D is a broad-band type which produces approximately 12 dB of gain from 0.5 to 50 MHz. Because of the high drain-to-gate capacitance of high-current FETs such as the CP651, the transistor is operated in a grounded-gate circuit to avoid the need for neutralization. A passive input circuit and 4-to-1 balun output transformer are employed. Details of the construction of balun transformers are given in Chapter 6.

FEEDBACK

Feedback giving rise to regeneration and oscillation can occur in a single stage or it may appear as an overall feedback through several stages that are on the same frequency. To avoid feedback in a single stage, the output must be isolated from the input in every way possible, with the vacuum tube or transistor furnishing the only coupling between the two circuits. An oscillation can be obtained in an rf or i-f stage if there is any undue capacitive or inductive coupling between output and input circuits, if there is too high an impedance between cathode and ground or screen and ground, or if there is any appreciable impedance through which the grid and plate currents can flow in common.

To avoid overall feedback in a multistage amplifier, attention must be paid to avoid running any part of the output circuit back near the input circuit without first filtering it carefully. Since the signal-carrying parts of the circuit can't be filtered, the best design for any multistage amplifier is a straight line, to keep the output as far away from

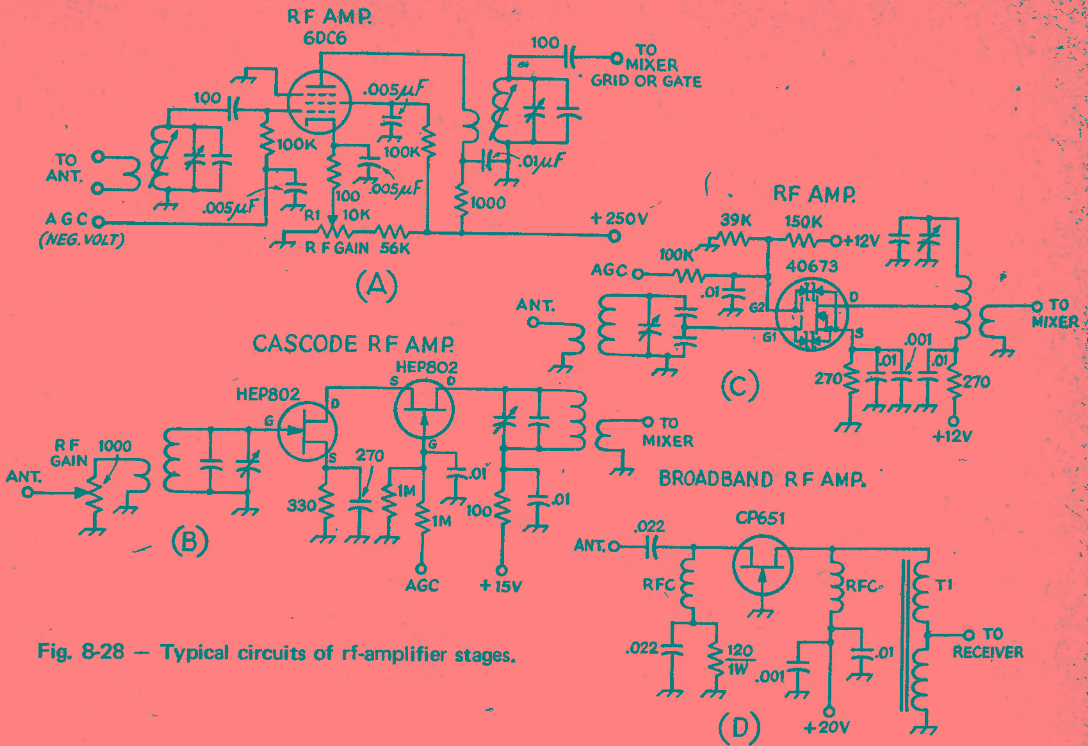


Fig. 8-28 — Typical circuits of rf-amplifier stages.

the input as possible. For example, an rf amplifier might run along a chassis in a straight line, run into a mixer where the frequency is changed, and then the i-f amplifier could be run back parallel to the rf amplifier, provided there was a very large frequency difference between the rf and the i-f amplifiers. However, to avoid any possible coupling, it would be better to run the i-f amplifier off at right angles to the rf amplifier line, just to be on the safe side. Good shielding is important in preventing overall oscillation in high-gain-per-stage amplifiers, but it becomes less important when the stage gain drops to a low value. In a high-gain amplifier, the power leads (including the heater circuit) are common to all stages, and they can provide the overall coupling if they aren't properly filtered. Good bypassing and the use of series isolating resistors will generally eliminate any possibility of coupling through the power leads. Rf chokes, instead of resistors, are used in the heater leads where necessary.

CROSS MODULATION

Since a one- or two-stage rf amplifier will have a bandwidth measured in hundreds of kHz at 14 MHz or higher, strong signals will be amplified through the rf amplifier even though it is not tuned exactly to them. If these signals are strong enough, their amplified magnitude may be measurable in *volts* after passing through several rf stages. If an undesired signal is strong enough after amplification in the rf stages to shift the operating point of a tube or transistor (by driving the grid into the positive region), the undesired signal will modulate the desired signal. This effect is called cross modulation, and is often encountered in receivers with several rf stages working at high gain. It shows up as a superimposed modulation on the signal being listened to, and often the effect is that a signal can be tuned in at several points. It can be reduced or eliminated by greater selectivity in the antenna and rf stages (difficult to obtain), the use of FETs or variable-μ tubes in the rf amplifier, reduced gain in the rf amplifier, or reduced antenna input to the receiver. The 6BJ6, 6BA6 and 6DC6 are recommended for rf amplifiers where cross modulation may be a problem.

A receiver designed for minimum cross modulation will use as little gain as possible ahead of the high-selectivity stages, to hold strong unwanted signals below the cross-modulation point. Cross modulation often takes place in double-conversion superheterodynes at the *second* converter stage because there is insufficient selectivity up to this point and at this point the signals have quite appreciable amplitudes. Whenever interference drops out quite suddenly with a reduction in the setting of the gain control, cross modulation should be suspected. Normally, of course, the interference would reduce in amplitude in proportion to the desired signal as the gain setting is reduced.

RF Gain Control

To avoid cross modulation and other overload effects in the mixer and rf stages, the gain of the rf stages is usually made adjustable. This is accomplished by using variable-μ tubes and varying the dc grid bias, either in the grid or cathode circuit. If the gain control is automatic, as in the case of agc, the bias is controlled in the grid circuit. Manual control of rf gain is generally done in the cathode circuit. A typical rf amplifier stage with the two types of gain control is shown in schematic form in Fig. 8-28A. The agc control voltage (negative) is derived from rectified carrier or signal at the detector before the audio amplifier, or in the case of a cw or ssb receiver it can be derived from rectified audio. The manual-gain control voltage (positive with respect to chassis) is usually derived from a potentiometer across the B+ supply, since the bias can be changed even though little plate current is being drawn.

Tracking

Tracking refers to the ability of a receiver to have all of its front-end stages – usually the rf amplifier, the mixer, and the oscillator – tune over a given range while each stage remains tuned to its proper frequency at any specified point in the tuning range. This arrangement provides a single tuning control for bandset and bandspread adjustments. To achieve proper tracking, it is usually necessary to have variable inductors and variable trimmer and padder capacitors for each of the tuned circuits. A two- or three-section variable capacitor is used for the tuning control.

Most modern receivers use a separate tuning control for the local oscillator and this is called the "main tuning." The rf and mixer stages are tracked and use a two-section variable for front-end peaking adjustments. This control is frequently called "preselector tuning." If the main tuning control is moved, the preselector is readjusted for a peak signal response at the new frequency.

REDUCING BROADCAST STATION INTERFERENCE

Some receivers, particularly those that are lacking in front-end selectivity, are subject to cross talk and overload from adjacent-frequency ham or commercial stations. This condition is particularly common with simple receivers that use bipolar transistors in the rf and mixer stages. With the latter, the range of linear operation is small compared to that of vacuum tubes. Large signals send the transistors into the nonlinear operating region, causing severe crosstalk.

The most common cross-talk problem in ham radio is that which is caused by the presence of nearby broadcast stations in the 550- to 1600-kHz range. In some regions, the ham bands – when tuned in on even the best receivers – are a mass of distorted "pop" music, garbled voices, and splatter. It should be pointed out at this juncture that the broadcast stations themselves seldom are at fault, (although in isolated instances they are capable of generating spurious output if operating in a faulty manner).

The most direct approach to the problem of broadcast-station interference is to install a rejection filter between the antenna feed line and

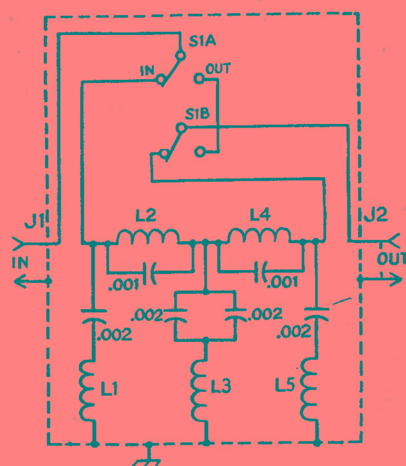


Fig. 8-29 – Inside view of the broadcast trap.

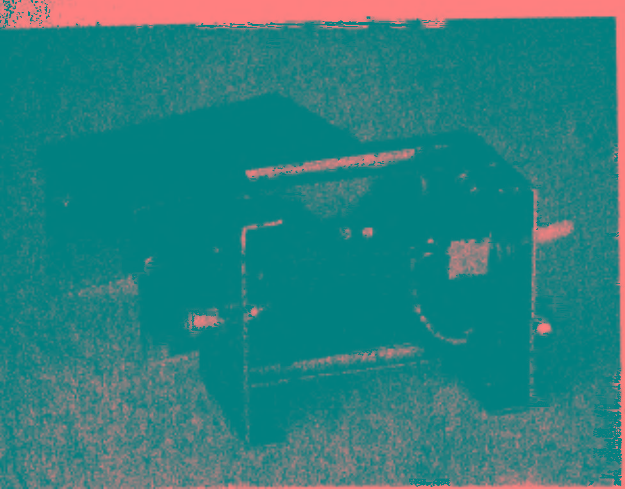


Fig. 8-30 – Capacitance is in pF. Capacitors are disk or tubular ceremic.

J1, J2 – Phono jack.
L1, L5 – 10-μH inductor, 43 turns, No. 26 enam. wire on Amidon T-50-2 toroid core (available from Amidon Associates, 12033 Ostego St., North Hollywood, CA 91607).
L2, L4 – 33-μH inductor, 75 turns, No. 30 enam. wire on Amidon T-68-2 toroid core.
L3 – 4.7-μH inductor, 30 turns, No. 26 enam. wire on Amidon T-50-2 toroid core.
S1 – Spst toggle.

the input terminals of the receiver. Such a filter, if capable of providing sufficient attenuation, prevents the broadcast-station signals from reaching the ham receiver's front end, thus solving the cross-talk problem.

An effective band-rejection filter, containing two constant-*k* sections in cascade, is shown in Fig. 8-30. It offers sharp rejection to signals in the 500- to 1600-kHz range but does not impair reception above or below the broadcast band. It is designed for use in low-impedance lines, particularly those that are 50 to 75 ohms.

The band-rejection filter is housed in a 3 1/2 X 2 1/8 X 1 5/8-inch Minibox. Phono connections are used for J1 and J2 – an aid to cost reduction. Different-style fittings can be used if the builder wishes. Standard-value components are used throughout the filter and the values specified must be used if good results are to be had.

In situations where a *single* broadcast station is involved in the cross-talk problem, a simple series- or parallel-tuned wave trap, tuned to the frequency of the interfering station, may prove adequate in solving the problem. (Such a trap can be installed as shown in Fig. 8-31.) The trap inductors can be made from ferrite-bar broadcast radio loop antennas and tuned to resonance by means of a 365-pF variable capacitor. Traps of this type should be enclosed in a metal box, as is true of the band-rejection filter.

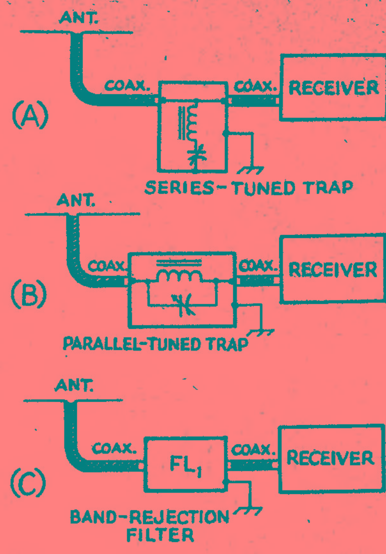


Fig. 8-31 – Examples of series- and parallel-tuned single-frequency traps (installed) are shown at A and B. At C, FL1 represents the band-rejection filter described in the text. If possible, the filter used should be bolted to the chassis or case of the receiver. The receiver should have e good earth ground connected to it.

FRONT-END OVERLOAD PROTECTION FOR THE RECEIVER

It is not uncommon to experience front-end overloading when the station receiver is subjected to an extremely strong signal. Frequentiy, it becomes necessary to install some type of external attenuator between the antenna and the input of the receiver to minimize the bad effects caused by the strong signal, or signals. Ideally, such an attenuator should be designed to match the impedance of the antenna feed line and the input impedance of the receiver. Also, the attenuator should be variable, enabling the user to have some control over the amount of attenuation used. Manufacturers of some modern receiving equipment build attenuators into the front end of their receivers, offering benefits that are not available from the normal rf gain-control circuit.

Examples of two such attenuators are given in Figs. 8-34 and 8-35. In Fig. 8-35 a ladder-type attenuator gives a 0- to 40-decibel range of control in five steps. A precision step attenuator is illustrated in Fig. 8-34. The latter offers an attenuation range of 3 to 61 decibels in 3-dB steps by closing one or more of five toggle switches. Both units are designed for use in low-impedance lines. The one in Fig. 8-35 is designed for a midrange impedance of 60 ohms, making it satisfactory for use with receivers having a 50- or 75-ohm input. Although designed for an impedance of 50 ohms, the attenuator of Fig. 8-34 will work satisfactorily with 75-ohm receiver inputs if accurate attenuation steps are not required.

Standard-value 1/2-watt resistors are used in the simple attenuator, which will give good results from the broadcast band to 30 MHz. Isolation between sections is not good enough to make this unit particularly effective above 30 MHz. The

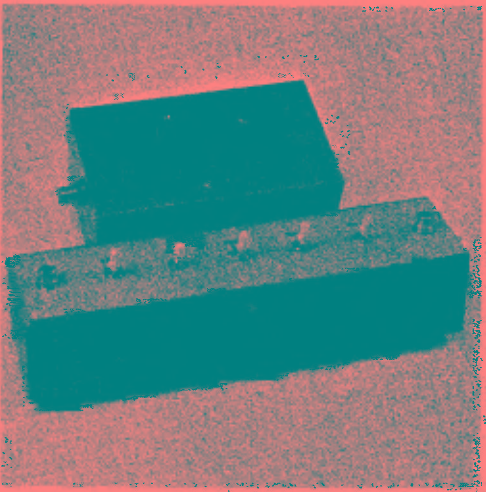


Fig. 8-32 – Two attenuators for receiver front-end protection.

Fig. 8-33 — Inside view of the attenuators. In the upper unit the resistors are mounted directly on the switch, using short pigtails wherever possible. Wide strips of copper are used for the input and output leads. The lower unit has each attenuator section individually shielded. The entire assembly is made up of double-sided circuit board material, cut to form the necessary sections and soldered on all abutting edges. All resistors should be connected with the shortest possible leads. A U-shaped piece of aluminum forms the base.

precision step attenuator, if carefully constructed to reduce leakage to a minimum, will be effective to 150 MHz or higher. The smaller 1/4-watt resistors are used as they have less inductance than the 1/2-watt types.

Either attenuator can be used ahead of the receiver, or can be built into the receiver as an integral part of the circuit. Such a device is particularly useful ahead of receivers that do not have an rf gain control, such as simple regenerative receiving sets.

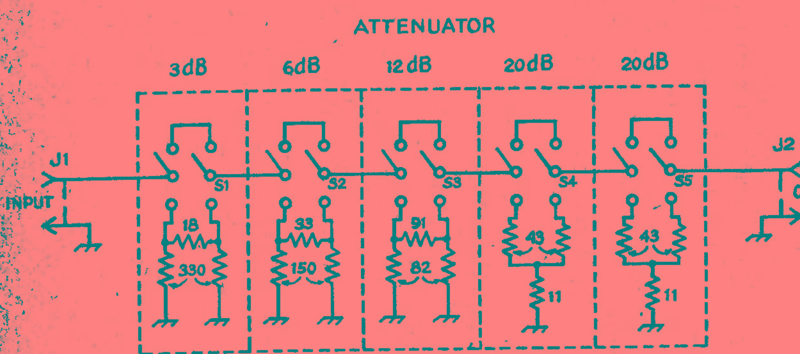


Fig. 8-34 — Circuit diagram of tha step attenuator. All resistors are 1/4-wett composition, 5-percent tolerance.
J1, J2 — Phone plugs, or similar.
S1-S5 — Miniature toggle switch.

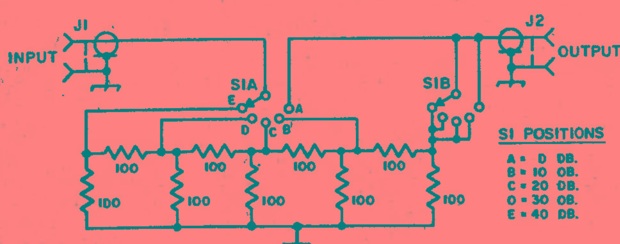


Fig. 8-35 — Schematic of the attenuator. Resistance is in ohms. Resistors are 1/2-watt composition, 10-percent tolerance. S1 is a phenolic rotary 1-section, 2-pole, 5-position switch. J1 and J2 are standard coax connectors. Approximate attenuation in decibels is given for each switch position.

IMPROVING RECEIVER SENSITIVITY

The sensitivity (signal-to-noise ratio) of a receiver on the higher frequencies (above 20 MHz) is dependent upon the bandwidth of the receiver and the noise contributed by the front end of the receiver. Neglecting the fact that image rejection may be poor, a receiver with no rf stage is generally satisfactory, from a sensitivity point, in the 3.5- and 7-MHz bands. However, as the frequency is increased and the atmospheric noise becomes less, the advantage of a good front end becomes apparent. Hence at 14 MHz and higher it is worth while to use at least one stage of rf amplification ahead of the first detector for best sensitivity as well as image rejection. The multigrid converter tubes have very poor noise figures, and even the best pentodes and triodes are three or four times noisier when used as mixers than they are when used as amplifiers.

If the purpose of an rf amplifier is to improve the receiver noise figure at 14 MHz and higher, a good FET, or a high-g_m pentode or triode should be used. Among the pentodes, the best tubes are the 6EH7, 6BZ6, and 6AK5. Of the triodes, the 6AN4, 6CW4, and 6DS4 are best. Among the better field-effect transistors are the MPF102, 2N4417, 3N128, CP625, 3N200, and 40673.