I. F. AMPLIFIERS AND SECOND-DETECTOR CIRCUITS FOR COMMUNICATION RECEIVERS

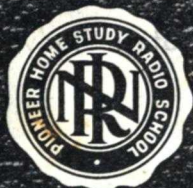*Finished Jan. 14, 1960*

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

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I. F. AMPLIFIERS AND SECOND-DETECTOR CIRCUITS FOR COMMUNICATION RECEIVERS

Choice of the I. F. Value

THE whole principle of the super-heterodyne is to convert all signals to a single frequency (the i.f.), which is usually lower than the signal frequency, so that high and relatively constant gain, high adjacent-channel selectivity, and relatively constant selectivity can be obtained. Once the i.f. is produced in the mixer, no further frequency changes are made until the i.f. carrier is demodulated after being delivered to the second detector. It is much easier to get the desired selectivity characteristics from an amplifier operating on a fixed frequency than from one which must tune over a range, because the problems of constant coupling and optimum L-to-C ratio are greatly simplified.

In the discussion of the preselector, it was pointed out that the lower the value of the i.f., the better the adjacent-channel selectivity of the receiver will be, as selectivity works on a percentage basis. Adjacent channel selectivity is the ability to separate signals which are very close in frequency on a percentage basis *at their original frequencies*. This percentage difference *increases* as the i.f. is lowered, and separation becomes much easier. Good adjacent channel selectivity is one of the big advantages of the superhet.

On the other hand, we found that as the i.f. is lowered, image rejection becomes poorer. This effect gets progressively worse as the signal frequency is raised. Thus, image rejection and the tuning range of the receiver must be balanced against adjacent-channel selectivity when the i.f. value is chosen.

Band Width. The band width required for the type of service is a factor to be considered when deciding on the proper i.f. value. For example, telegraph code reception at hand-keyed speeds requires less than a 100 c.p.s. band width. An i.f. as low as 10 kc. could be used for this service. This is done in the 15 to 200 kc. signal range. If it is used for higher ranges, poor image rejection will result. Voice signals require a band width of about 5000 c.p.s. for good intelligibility. This requires an i.f. of at least 175 kc. with modern components and use of two stages. High fidelity broadcast reception requires a band width of at least 20,000 c.p.s. This requires an i.f. of about 465 kc. Frequency modulation may require band widths of about 75 to 150 kc., which in turn require i.f values of about 4 or 5 mc. Television and pulse communication systems may require band widths of 6 mc. and i.f. values as high as 30 mc.

The band width of the i.f. amplifier is usually designated as the width of a rectangle equivalent to the area under a certain portion of the selectivity curve. This is approximately the width of the selectivity curve two times down from its optimum response (6 db down). In practice, the band width should be more than is theoretically required to allow for the detuning ef-

fect of oscillator drift, and the changes of the i.f. band width itself due to temperature or alignment changes with time.

For general purpose communication receivers covering the 0.54- to 30-mc. range, the i.f. values used range from 175 kc. to 1500 kc., with the most used i.f. probably in the 455-465 kc. region.

It is desirable that the i.f. be chosen so that there will not be any strong signal on its frequency to produce interference in case of insufficient i.f. rejection.

Older broadcast receivers produced squeals when i.f. harmonics beat with incoming signals. An i.f. ending in 5, such as 175 kc., produces harmonics of 350, 525, 700, 875, 1050, 1225, and 1400 kc., of which 700, 1050, and 1400 kc. are frequencies of stations in the broadcast band, where squeals may be produced. A 180 kc. i.f. produces harmonics of 360, 540, 720, 900, 1080, 1260, and 1440, where all but the first value are broadcast station frequencies. The i.f. ending in 5 gives only half as many squeal-producing points on the dial as one ending in 0. This difference in performance is practically eliminated with modern designs and tubes, and is relatively unimportant in communication receivers operating outside the broadcast band, where any frequency combination may come up. Thus, practically any value of i.f. may be used in the communication receiver, depending on the type of service for which it is used.

I. F. Response Curves

SIDE-BAND CUTTING

It might seem that the adjacent channel selectivity could be increased to any desired value by using plenty of i.f. circuits, or by using a very low i.f. value. There is a limitation on this other than image rejection or oscillator drift. It is known as side-band cutting. All radio signals used for communication must be modulated or varied in accordance with the intelligence to be transmitted. The process of modulation produces frequencies other than the original signal frequency. These other frequencies consist of the signal frequency plus the modulation frequencies, and the signal frequency minus the modulation frequencies. Fig. 1 shows a 1000 kc. broadcast station carrier sending a chime consisting of simultaneous 400, 1000, and 2000 c.p.s. tones. Frequencies of 998.0, 999.0, 999.6, 1000.0, 1000.4, 1001.0, and 1002.0 kc. are transmitted. The receiver selectivity, which in the case of a super is practically that of the i.f. amplifier, must thus pass a band of frequencies equal in width to 1002 minus 998, or
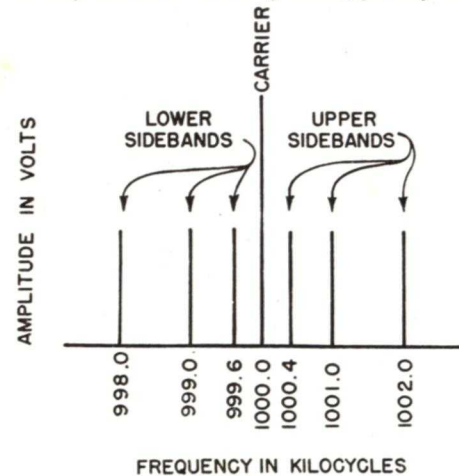

FIG. 1. Graphical representation of the components of a radio-telephone signal modulated with a chime containing 400, 1000, and 2000 c.p.s. tones. The band width required to transmit this signal without distortion is 1002—998, or 4 kc.

4 kc., to reproduce this signal without distortion. You know that changing the signal to the i.f. in the first detector does not change its modulation characteristics, so the band width required to pass it is the same as at the original signal or carrier frequency. All the tones shown in Fig. 1 are of equal intensity. If the i.f. amplifier will pass a band only 3 kc. wide, the highest and lowest side-band frequencies (corresponding to the 2000 c.p.s. tone) will be left out or will be greatly reduced, and this tone will be reproduced at
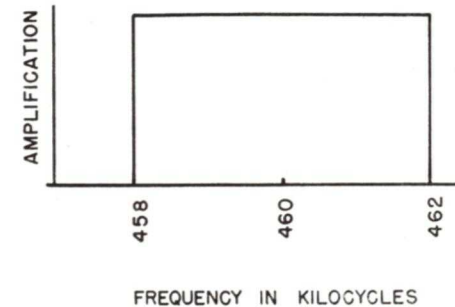

FIG. 2. Ideal response curve of a 460-kc. i.f. amplifier for the signal of Fig. 1, after it has been converted to 460 kc. in the first detector.

greatly reduced amplitude. This type of distortion is known as side-band cutting, and it limits the useful selectivity, especially on radiotelephone or television signals. The i.f. band width must be at least twice the highest frequency to be transmitted.

IDEAL SELECTIVITY CURVE

The ideal selectivity curve of a receiver is perfectly flat throughout the required band width, and has vertical sides at each end. Such a curve is shown in Fig. 2 for the signal of Fig. 1 after it has been converted to a 460 kc. i.f. This ideal characteristic cannot be obtained in practice. Typical curves obtained with one, two, and three tuned circuits arranged so that the signal must pass through them in cascade, are
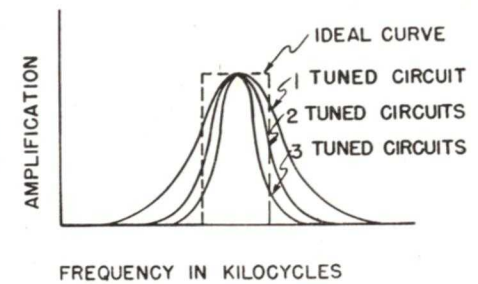
FIG. 3. The effect of tuned circuits on side band cutting.

shown in Fig. 3 along with the ideal response. It is clear that increasing the number of tuned circuits increases the selectivity and also the side-band cutting.

One way of overcoming side-band cutting without sacrificing selectivity is by tone correction, the principles of which are shown in Fig. 4. The higher audio frequencies available at the second detector have decreased in intensity because of side-band cutting as shown in the lower curve. They can be boosted by using a subsequent audio amplifier having a rising characteristic, as shown in the upper curve. The resultant uniform over-all response is shown in the middle curve.

An audio amplifier circuit which will produce increasing output at increasing frequencies is shown in Fig. 5. At very low frequencies, the imped-

FIG. 4. How a tone-correcting audio amplifier can overcome the effect of side-band cutting.

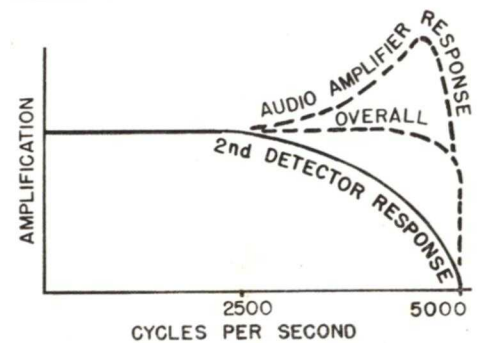ance of the L-C circuit is very low because the low reactance of L and the total tube load is the resistance R. As the frequency rises toward the resonant frequency of L-C (usually in the 5000 to 10,000 c.p.s. region), the im-
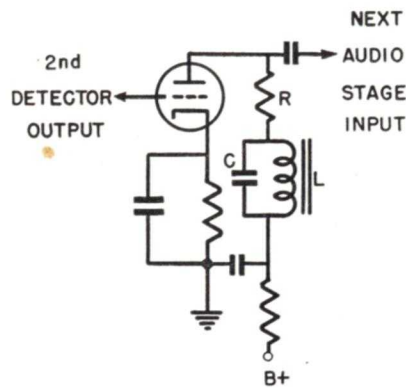

FIG. 5. Tone correction circuit.

pedance of the tuned circuit rises and adds to that of R. The output increases across this higher impedance. At still higher frequencies, the reactance of C becomes very low, and the circuit impedance is R shunted by the tube and wiring capacities, so the output drops off.

Compensation for side-band cutting by tone correction is not very satisfactory, because if the side-band is cut to the point where it is below the first circuit or tube noise, subsequent amplification of the cut frequency merely increases the noise without restoring the frequency that was lost. This system operates to restore only a moderate degree of side-band cutting.

A more satisfactory way to reduce side-band cutting is to vary the selectivity to fit the band width of the signal being received, or to change the shape of the selectivity curve. Since most of the selectivity is obtained in the i.f. amplifier, and it operates at a fixed frequency, its selectivity is logically that which must be varied.

STAGGER TUNING

A selectivity curve more nearly approaching the ideal rectangular shape than that obtained with cascaded tuned circuits can be obtained by slightly overcoupling the i.f. amplifier tuned circuits in pairs, or by "staggering" their tuning, that is, tuning successive circuits to slightly different frequencies, so that the peaks do not coincide as they do in Fig. 3. Curves produced by over-coupling or stagger-tuning are shown in Fig. 6. This procedure is sometimes used in "high fidelity" receivers where flat band widths of 20 kc. are desired, but steep sides are needed to reduce adjacent channel interference and noise. Over-coupling or stagger-tuning is seldom used in communication receivers in the 0.5- to 30-mc. region, because of the increased difficulty of alignment, and the comparatively small band widths required for communication purposes. The alignment difficulty can be seen from Fig. 6, where there are two peaks, but neither of them is at the resonant frequency. The simplest alignment procedure involves the use of a sweeping signal and a cathode ray oscilloscope
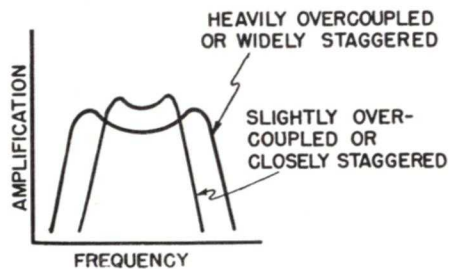

FIG. 6. Selectivity curve obtained by over-coupling or stagger-tuning two tuned circuits

to view the whole curve at one time to see the effect of each adjustment.

Skirt-to-Nose Ratio. A "figure of merit" expressing the quality of the i.f. amplifier selectivity curve would be one way of expressing the degree to

which the curve approaches the ideal rectangle shape. This is called the skirt-to-nose ratio, and is usually determined by dividing the band width in kilocycles of the curve at 1000 times down from resonance (60 db down) by the band width in kilocycles of the curve at 2 times down (6 db). This 60/6 db band width ratio is usually between 2 and 5 for a good communication receiver. Increasing the number of tuned circuits helps this ratio, because the sides are steepened faster than the nose. The ideal rectangle curve would, of course, have a ratio of 1, so the smaller this ratio, the better the curve, and the better the adjacent channel selectivity without sacrifice of desired side bands.

Since the major portion of the receiver gain is usually in the i.f. amplifier, there may be some difficulty preventing feedback from its output to its input. This feedback may cause regeneration or even oscillation in severe cases, giving a squeal on all signals tuned in, as well as severe distortion. This is usually prevented by isolating each stage from the common power supply impedance with R-C filters in all plate and screen leads, and by careful layout and shielding where necessary. Regeneration is present if selectivity curves measured at different gain settings give different shapes. The nose gets sharper and the skirts get broader, giving a much poorer skirt-to-nose ratio at higher gain settings.

Variable-Selectivity I.F. Amplifiers

We have seen that for optimum sensitivity and maximum freedom from interference, noise, and oscillator drift effects, for code, voice, facsimile, etc. at various carrier frequencies, different receiver band widths are required. It is thus desirable that the band width or selectivity be adjustable to fit receiving conditions and the degree of

side-band cutting which is permissible. We have found that the best portion of the receiver in which to incorporate variable selectivity is the i.f. amplifier.

TYPICAL I.F. AMPLIFIER CIRCUITS

Two typical i.f. amplifier circuits are shown in Fig. 7. The circuit of Fig. 7A uses only one tuned circuit (the
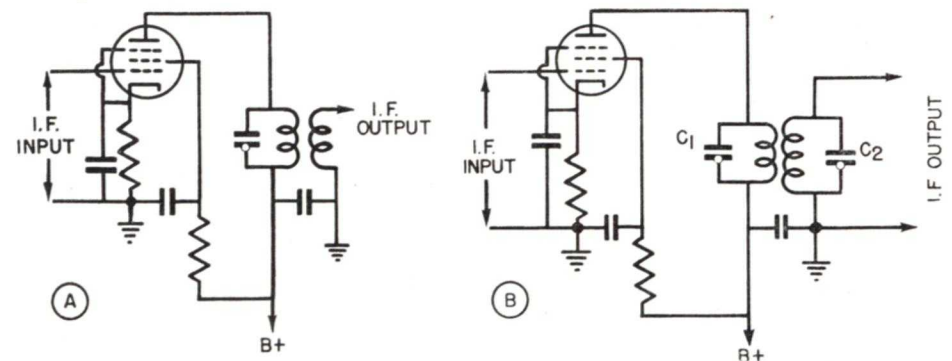
FIG. 7. Typical i.f. amplifier circuits. A, single-tuned i.f. amplifier using close-coupled secondary; B, double-tuned i.f. amplifier. When coupled loosely, it gives a single peak with steeper sides than two separate equivalent circuits. When over-coupled, it gives a "band-pass" double-humped response with very steep sides. This circuit may also be stagger-tuned to give a band-pass response by tuning C_1 below resonance and C_2 above resonance by the same amount.

primary). The secondary is tightly coupled, and is thus effectively tuned to some extent by the same condenser. The losses of both circuits are combined, so the sharpness of tuning is reduced from that obtained with double tuning, and the skirt-to-nose ratio is poorer. This circuit is more often used in low-cost broadcast receivers than in those designed for communication purposes.

pass" response. One circuit may be tuned somewhat below the i.f. value, and the other may be tuned somewhat above it. This "stagger-tuning" will also produce a band-pass response, but at a considerable loss of gain. If the amplifier to be tuned to the i.f. value is out of alignment, excessive interference or low gain may be produced.

Variable Selectivity Methods. The method used to accomplish variable

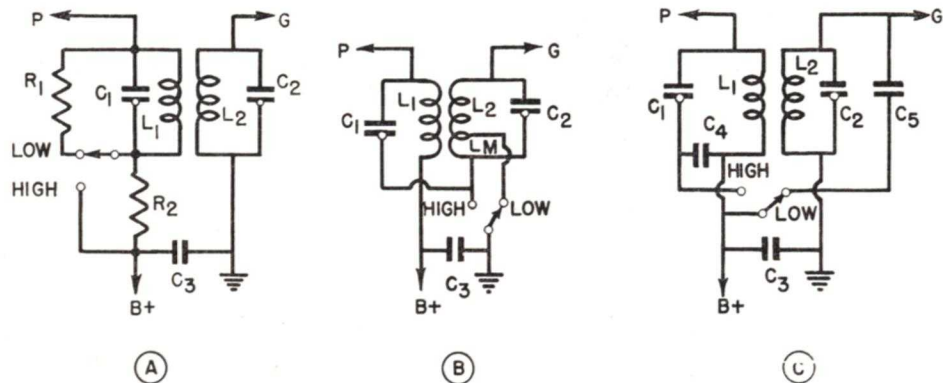lower impedance, hence the stage gain is reduced. By placing R_2 in series with the tuned circuit when R_1 shunts it, and by properly proportioning the two resistance values, the impedance of the circuit can be made the same in either the broad or the sharp position of the switch. This keeps the gain constant, and there is no shift of center frequency. The same procedure can be used on the L_2-C_2 circuit and all the other i.f. circuits if desired, and all the switches can be ganged. Other

humped tuning, and an increase in band width. The extent of the widening is determined by the amount of the tapped portion L_M. Several steps of selectivity may be provided with this arrangement by having several taps on L_2. All the i.f. stages may be similarly switched simultaneously by ganging the switches.

Another system is shown in Fig. 8C. The coupling is such that with the switch position set for high selectivity, both C_1 and C_2 are tuned to resonance,
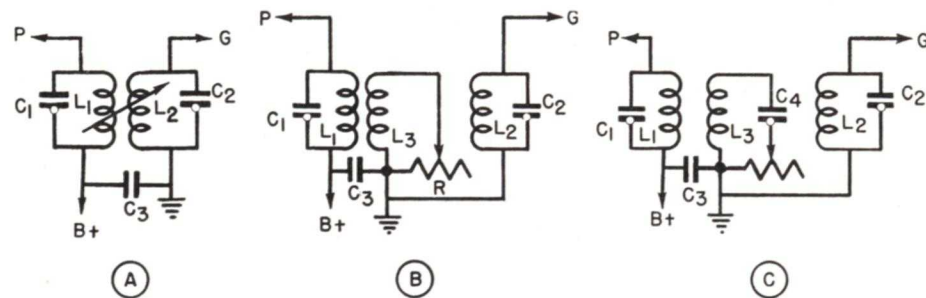

FIG. 8. Methods of varying selectivity in steps.

FIG. 9. Methods of continuously varying selectivity.

Critical Coupling. Fig. 7B shows the most widely used i.f. amplifier circuit. It uses two tuned coupled circuits. When both these circuits are tuned to the same frequency, and the coupling is increased, the point at which maximum response is obtained with a single response peak is called "critical" coupling. Usually the coupling is adjusted slightly below this value, to be sure it does not exceed it during extremes of temperature, humidity, vibration, or when tubes are changed. If the coupling is less than critical, the selectivity is only slightly improved, and the gain falls off rapidly. If the coupling is greater than critical (overcoupling), a "double humped" curve such as that of Fig. 6 is obtained. The resonant frequency gain is then reduced. This is one method of obtaining a "band-

i.f. selectivity should maintain the same general shape of curve. Side-band cutting may not be reduced when the curve is broadened, if the broadening takes place mainly in the skirts. The variable selectivity method should not shift the center of the curve, or the upper side bands will be amplified a different amount from the corresponding lower ones, and excessive distortion, loss of gain, or interference from signals to which the curve shifts, will result.

Three methods of obtaining variable selectivity in steps, are shown in Fig. 8. The method of Fig. 8A broadens the selectivity by shunting the tuned-circuit inductance L_1 with a resistance R_1. This resistance loading reduces the Q of the circuit and broadens its response. The lower Q circuit also has

resistance combinations can be added to give several degrees of band width. This system is not in common use, because it broadens the skirts too rapidly with respect to the nose, giving a poor shape (or poor skirt-to-nose ratio) on the broad positions.

The circuit of Fig. 8B uses a tapped coil in one of the tuned coupled circuits. With the switch in the high selectivity position, C_1 and C_2 are set for the optimum single peak response, as the coupling is adjusted at, or slightly below the critical point. With the switch in the low-selectivity position, the L_M portion of L_2 is common to both tuned circuits. This increased common or mutual inductance of the coils appears to add more inductance to L_1-C_1, tuning it to a lower frequency, and to take inductance away from L_2-C_2, tuning it to a higher frequency. There is thus a tendency toward double-

giving a single peak. With the switch on the low-selectivity position, condenser C_4 is in series with C_1, giving a smaller resultant capacity than before. This tunes L_1 to a higher frequency. Condenser C_5 is added in parallel to C_2, giving a larger resultant capacity than before. This tunes L_2 to a lower frequency. By properly choosing the capacities of C_4 and C_5, the band width can be increased symmetrically by any desired amount. Several degrees of selectivity can be obtained by employing several capacity combinations. This system can be used on all i.f. stages by ganging the selectivity switches.

In some cases, it may be desirable to be able to vary the selectivity continuously to fit receiving conditions. Three methods of doing this are shown in Fig. 9.

In the circuit of Fig. 9A, the cou-

pling between the coils, and hence their mutual inductance is varied by mechanically moving one of them with respect to the other. In some cases only a portion of the coil may be movable to get a finer degree of control. The trimmers C_1 and C_2 are adjusted for a maximum single peak response when the coupling is minimum (coils farthest apart), since this gives critical or below critical coupling. When the coils are more closely coupled, the two tuned circuits interact, and tend to push each other off resonance an equal and opposite amount, and the response curve widens and becomes double humped if the coupling is made tight enough at the broadest setting.

When L_2-C_2 is tuned to resonance, its impedance is high, and is a pure resistance. When L_1-C_1 is also tuned to resonance, the current flowing in L_2-C_2 caused by the induced voltage will, in turn, induce a voltage in L_1-C_1, which is opposite in phase to the effective series L_1-C_1 voltage. This opposing L_1-C_1 voltage reduces the L_1-C_1 current at resonance. Since the actual applied L_1-C_1 voltage has not changed, the reduction of current can be explained by an apparent increase of the series resistance of L_1. The parallel resistance decreases at resonance because of the decreased Q caused by the increased series resistance.

When L_2-C_2 is not at resonance, as when a slightly off-resonance frequency is applied, the voltage induced in L_1-C_1 by the L_2-C_2 current will not be exactly out of phase with the voltage acting in L_1-C_1, and reactance appears to be coupled into L_1-C_1, retuning it to a new frequency. At frequencies far off resonance, little current will flow in L_2-C_2 and this effect will be slight. There will thus be two frequencies off resonance where maximum current flows in L_2-C_2, and there will

be a drop in current at exact resonance and frequencies very close by.

The mechanical coupling of several such i.f. stages may be ganged to a single selectivity control. This system produces the most symmetrical response at different band widths, of any in use. The mechanical design problems involved are somewhat difficult, and it is not often used on low-cost receivers.

The arrangement of Fig. 9B uses a third winding L_3 coupled equally to L_1 and L_2. Winding L_3 is shunted with resistance R. When R is an open circuit or a very high resistance, L_3 has no appreciable effect on L_1 and L_2, and these circuits are tuned to resonance with about critical coupling between them. When R is reduced, power is dissipated in it because of the voltage impressed on it from L_3. This increased loading effectively loads both tuned circuits with resistance, due to the common coupling, and the selectivity is broadened. The gain decreases as the band width increases.

Another method is shown in Fig. 9C. It appears similar to that of Fig. 9B, except that L_3-C_4 are tuned to resonance, as are the other two tuned circuits when R is set to zero resistance (not maximum as in Fig. 9B). Then circuit L_3-C_4 absorbs energy from L_1-C_1 and passes it to circuit L_2-C_2. When R is increased, energy is dissipated in it, and because all three circuits are about critically coupled, they are all loaded, reducing the selectivity and gain. With either this arrangement or that of Fig. 9B, the variable resistors of several similar stages could be ganged.

It may have occurred to you that an even closer approximation to the ideal rectangle selectivity curve than is obtained by the double-humped curve of Fig. 6 could be obtained by

superimposing a single tuned circuit response at the middle of the "sag" to give a triple-humped curve with small variations and steep sides. Such a curve can be obtained with the circuit of Fig. 9C when the resistance is zero, and the coupling of the third winding is properly adjusted. This is then called a triple-tuned i.f. transformer. It is most readily tuned with the aid of a sweep generator and oscilloscope, and gives more losses than the double tuned type, but these can be overcome with tube gain.

The Use of Crystal Filters

Audio Image. In continuous wave (c.w.) code reception, it is usual to beat another local oscillator with the i.f. in the second detector to give an audible difference frequency. For example, if a 465 kc. i.f. is used, the beat oscillator frequency might be 464 kc., giving a 1000-cycle beat note. If an interfering signal appears at 463 kc., it will also be heterodyned by the beat oscillator, and will produce a 1000-cycle beat note. The 463 kc. signal is known as an *audio image*. It is produced in exactly the same manner as the high-frequency image previously discussed. It can be reduced in exactly the same way as high-frequency images are reduced, by increasing the selectivity, in this case, of the i.f. amplifier.

The manner of producing an audio image is illustrated in Fig. 10 for the example given above. The rejection of the audio image for a typical i.f. band width of 6 kc. is seen from the dotted curve to be 1.43 to 1 or 3 db. This is negligible, hence two signals can appear at all points on the dial, or the same signal is heard at either side of zero beat, and we have "double signal" reception. When the i.f. selectivity is made to follow the very sharp solid curve of Fig. 10, the audio image is suppressed over 1000 to 1 or over 60 db, which makes it completely inaudible.

Tuning the receiver over the band with an i.f. response like the solid curve

of Fig. 10 produces a strong response on one side of zero beat only, instead of one on each side as in a less selective receiver. No audio image interference is ever present. This is known as "single-signal" reception.

The sharp i.f. selectivity required for single-signal reception can be obtained by using a very low i.f. (and losing image rejection) or by using a prohibitive number of circuits (raising the cost and complexity). It is obtained in practice either by the use of regeneration or by the use of a crystal filter.

Regeneration. With the use of regeneration, a single 465 kc. i.f. stage response can be made as sharp as 1 kc.
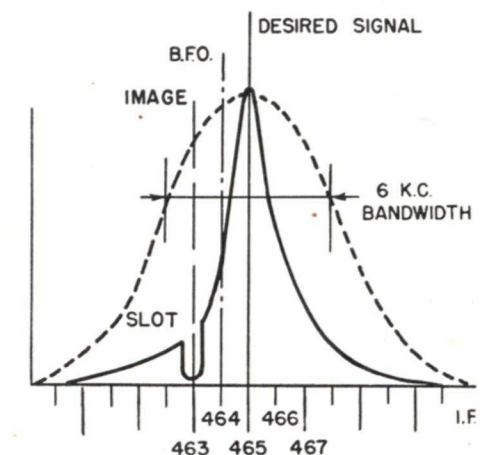

FIG. 10. The sharp dip or slot in the crystal curve is obtained by adjusting the phasing condenser as described in the text.

wide at 10 times down, and 5 kc. wide at 100 times down when adjusted just short of oscillation. This reduces the audio image by about 80 times with a 1000-cycle beat note. The i.f. can be made regenerative by providing a very
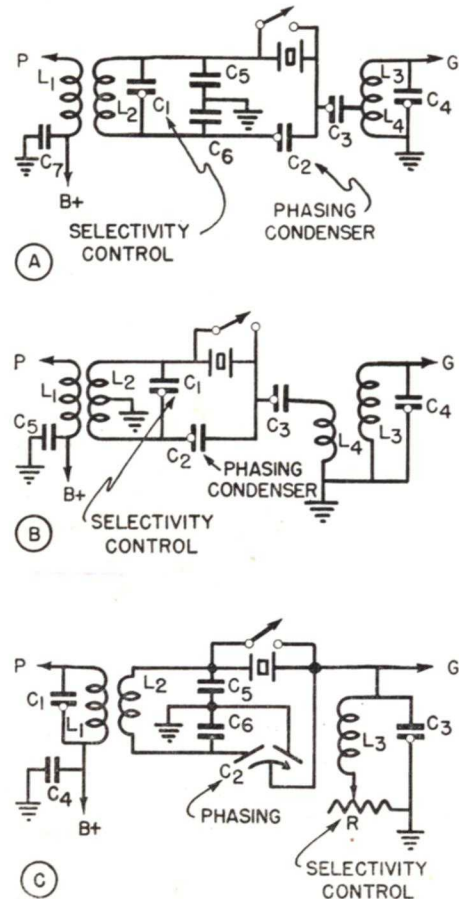

FIG. 11. Crystal filter circuits having variable band widths.

small feedback capacity between the grid and the plate of one of the i.f. amplifier tubes. A lead from the grid to the vicinity of the plate is usually enough. Feedback is usually controlled by varying the grid bias, which is the usual method of gain control. With a regenerative i.f., the stage is usually operated with high bias and thus re-

duced gain, and regeneration is relied on to bring up the signal strength. This reduces overload and increases selectivity.

The selectivity obtained by i.f. regeneration reduces noise outside the band width just as would similar selectivity obtained by other means. The main disadvantages of regeneration are the difficulty of adjusting and holding a given setting, and the fact that selectivity varies with signal strength, being less on strong signals as a result of change in regenerative gain.

Crystal Filters. The most satisfactory method of obtaining a high selectivity required for single-signal reception and rejection of very close interference frequencies, is by use of a piezo-electric quartz crystal as a selective filter in the i.f. amplifier. A quartz crystal when properly grounded and mounted, has electrical properties similar to a tuned circuit, but it has a Q which may be in the thousands instead of 100 or so like that of a typical i.f. tuned circuit. The difference in selectivity is shown in Fig. 10. The crystal filter is made to resonate at the intermediate frequency, and is connected so signals must pass through it before reaching the i.f. amplifier output. The very narrow band width obtained reduces the receiver response to external noise, and the random first circuit and tube noise generated in the receiver itself.

Crystal Filter Circuits. Three typical crystal filter circuits are shown in Fig. 11. Those shown at A and B perform about the same and differ only in that the first i.f. transformer secondary center is effectively grounded for i.f. by use of two equal condensers in series in A. The output tuned circuit is tapped to match the crystal impedance in A and a separate primary accomplishes the same thing in B. The

crystal is connected in a bridge circuit, the four arms of which are the two halves of the i.f. transformer secondary, the crystal, and variable condenser C_2, called the *phasing condenser*, whose maximum capacity is slightly higher than that of the crystal in its holder. When C_2 is set equal to the crystal holder capacity, the resonance curve is quite symmetrical, and the crystal acts as a series resonant circuit of high Q, allowing only signals of the desired frequency to be fed through C_3 to the output transformer. If C_2 were not used, undesired frequencies would flow through the crystal holder capacity, and the crystal would be by-passed and ineffective at some frequencies.

The phasing control accomplishes another purpose besides that of neutralizing the crystal holder capacity. The crystal may be made parallel resonant at a frequency very close to its normal series resonant frequency, and part of the parallel resonant capacity is that of its holder. Signals at the parallel resonant frequency cannot get through the crystal to the output circuit. As the phasing control varies the effect of the holder capacity, it permits the parallel resonant frequency to be shifted over a considerable range without affecting the series resonant frequency, or without greatly upsetting the phasing adjustment. The parallel resonant frequency is sharply rejected, and this characteristic is sometimes called the "rejection slot," as it looks like a slot in the crystal filter i.f. selectivity curve shown in Fig. 10. Any nearby signal producing undue interference may be rejected by adjusting the phasing condenser, or it may be set to the audio image frequency as shown in Fig. 10, where the image is rejected far more than it would be for a symmetrical resonance curve.

Variable Selectivity. The sharpest crystal filter selectivity is so sharp that it is useful only for c.w. reception, and even then the receiver tunes so sharply that signals may be passed by without being noticed if extreme care in tuning is not exercised. The side-band cutting is so severe that voice reception is impossible. Some means of broadening the selectivity is desirable.

In the circuits of Fig. 11A and Fig. 11B, variable selectivity is obtained by varying the input impedance effectively in series with the crystal. This is done by varying C_1, used as a selectivity control, which tunes the balanced secondary L_2. At resonance L_2-C_1 is purely resistive and thus is at its maximum impedance. As L_2 is effectively center tapped, one-fourth of the circuit impedance is in series with the crystal through C_3 and L_4 to ground. This lowers the crystal Q and gives minimum selectivity, while applying maximum voltage to the crystal circuit. It may seem a bit confusing that tuning L_2-C_1 to resonance *reduces* the selectivity. Detuning it reduces the impedance and also reduces its resistive component. The selectivity becomes higher as a result of reduced crystal loading, and the applied voltage decreases. At first, the resistance decreases faster than the applied voltage, so the output from the filter *increases* as the selectivity increases. Still further detuning of C_1 causes a still greater increase of selectivity, and the output falls off gradually. Even the broadest selectivity obtainable with the circuits of Fig. 11A and Fig. 11B is much too sharp for voice reception, so the selectivity range is not great enough for many purposes. The crystal may be short circuited, which gives the normal tuned circuit response, but values in between may be desired.

The circuit of Fig. 11C gives a wider

range of selectivity. The principle of operation is similar to that of the circuits of Fig. 11A or 11B. A higher resistance can be introduced in the crystal circuit to reduce the selectivity. The secondary L_2 has low impedance and is not tuned. The phasing condenser is a split-stator type, so its total (rotor to both stators) capacity does not change when it is adjusted. Thus it does not affect the resonant frequency. Since output circuit L_3-C_3 is of high Q when R is set to zero, the selectivity is increased by increasing R and decreasing the Q and the impedance of L_3-C_3. As this entire impedance is in series with the crystal and the source of input, a voltage divider action takes place, tending to keep the output voltage to the next grid constant. Since the impedance of L_3-C_3 drops as a result of its Q falling from an increase of R, the grid voltage tends to decrease. The decrease of crystal loading under these conditions increases the crystal Q, and the voltage at the grid tends to increase.

Tuning Arrangements. Because i.f. transformers operate at a fixed frequency, each circuit is individually tuned for optimum performance, and no attempt is made to gang the adjustments. The coils are usually wound on ceramic or phenolic forms and are spaced to give the desired degree of coupling. The coils should be wax impregnated and wax dipped to protect them from the effects of humidity and condensed moisture. Two basic methods of tuning are used. These are variable capacity and variable inductance systems.

When a fixed inductance on an insulating form material is used (often called an air-core coil, because such a form behaves electrically about the same as air), it is tuned either by a condenser having fixed metal plates,

thin mica dielectric insulation, and springy variable plates which are compressed toward the mica by a threaded screw, or by a small condenser having rotor and stator plates with air between, and some form of insulation at the ends such as phenolic or ceramic. The mica compression trimmer is compact and economical, but its capacity is affected considerably by temperature and humidity, so it is not used in communication receivers of high quality, as too-frequent realignment would be required, or receiver performance would suffer. The air dielectric trimmer is quite stable in capacity when a good ceramic insulation is used as a mechanical support. Sometimes such a trimmer is made in two parts; one fixed, and one variable, so some "bandspread" is obtained, making adjustment less critical.

An i.f. transformer in which the inductance is varied usually has a fixed-sealed condenser with a silvered-mica dielectric, which is stable under temperature and humidity variations. There is a slug made of a very finely powdered iron and insulating material mixture placed inside the coil form, arranged so it can be moved in or out of the coil by turning a threaded screw, to vary the inductance. As you probably know, changing from an air core to an iron core greatly increases the inductance of a coil. If the core were solid iron, the magnetic losses due to the reversal of the magnetic polarity of the iron on each cycle (hysteresis loss), and the eddy current losses would be so high at i.f. values that the coil Q would be uselessly low. Breaking up the iron into sufficiently fine particles insulated from each other, reduces the circulating current and hysteresis losses, but the inductance decrease is much less. The inductance increases faster than the losses, with proper core

design, so that a higher Q coil can be made with a powdered-iron core than with an air core in the same size. The core could be fixed and a capacity trimmer could be used, but the powdered-iron core is less affected by changes in temperature and humidity, so it is usual to have the core variable as a tuning element when iron-core coils are used. They are used in most modern communication receivers.

It will be noted in most of the i.f. amplifier circuits shown, that both
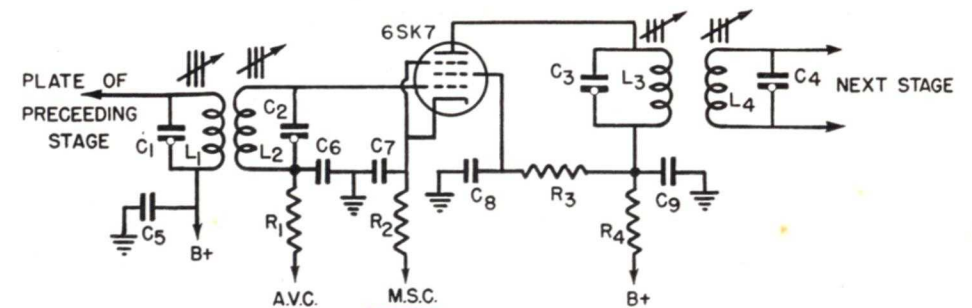

FIG. 12. A typical permeability-tuned i.f. amplifier stage.

sides of the primary tuning condenser are above ground potential by the plate voltage of the tube. The operator or technician aligning the receiver could receive an unpleasant shock if he used a metal aligning tool. With iron-core tuning, the drive screw can be placed at ground potential, elminating this hazard.

I. F. transformers are practically always placed in a metal shield can of copper or aluminum to confine the field around the coils. This makes for higher stability and also protects the coils from dust, dirt, and mechanical damage.

For i.f. values up to 1500 kc., solenoid windings, (in which the turns are wound adjacent to each other on a plane perpendicular to the coil axis) produce relatively high voltages between turns, and particularly between layers, giving rise to capacity currents

between turns through the wire insulation. The distributed capacity of such a coil is high, and since the dielectric is not of particularly good quality, the Q of such a coil is rather low. In this frequency range a special "universal winding" is used. The coil is wound in layers, but each turn weaves up and down on the form as it goes around, so that all wires cross each other at an angle, and the potential difference and distributed capacity effects are reduced.

A complete i.f. amplifier stage with powdered iron core (sometimes called permeability) tuning, is shown in Fig. 12. A typical receiver would probably incorporate a crystal filter. The cathode of this stage is fed through minimum bias resistor R_2 to the cathodes of the other i.f. and r.f. stages similarly arranged, and to the manual sensitivity control (m.s.c.). Resistor R_1, combined with C_6 acts as the a.v.c. filter. A separate screen-dropping resistor R_3 and screen by-pass condenser C_8 should be used on each i.f. stage to minimize interaction of feedback possibilities due to common interstage coupling. Decoupling resistor R_4 and by-pass condenser C_9 isolate the stage from the plate power supply.

Tubes Used. Automatic volume control in the i.f. amplifier requires remote cut-off tubes to prevent severe distortion and possible cross modula-

tion on strong signals. The type of tube used has practically no effect on the signal-to-noise ratio, which has already been established in the preselector and mixer circuits. High mutual conductance tubes gives the greatest gain, but usually reduce the selectivity somewhat by loading. This is not usually important with modern components, and the tube used is a compromise between high mutual conductance and good remote cut-off characteristics.

The 6SK7 is most popular, with the 6AB7 also in wide use. When a single-ended tube is used, a special grounded metal shield across the tube socket between plate and grid pins may be required to prevent regeneration. To accomplish this, the screen by-pass condenser can be mounted across the socket with its outside foil connected to ground, but this makes the socket less accessible. In some sockets by-pass condensers are built in the socket.

The Second Detector

The second detector takes the output of the i.f. amplifier and delivers the intelligence superimposed on the i.f. signal. This is the same function performed by the detector in the simple receivers that were used before the superheterodyne. The second detector of a super operates at a high level of signal input, because the r.f. converter and i.f. gain precedes it. It must be able to handle this level effectively. Any of several types of detector can be used as the second detector of a super, and the reasons for the choice will be more apparent after analyzing the operation of the most common ones.

In early receivers, the *sensitivity* of the detector was very important. Sensitivity is the ratio of audio-frequency output to radio-frequency input. With the high level operation of the super, this is not now so important. Detector *linearity* is the degree to which the audio output is a true reproduction of the modulation on the applied r.f. signal. This is important in voice reception. Detector *overload* is a lack of linearity on strong signals while there is good linearity on weak ones. The second detector of a super must be able to handle very strong signals with-

out overloading. The detector *impedance* is the load the detector places on its associated r.f. or i.f. circuit. A lower impedance loads the circuit more heavily, and reduces its selectivity. This is usually not much of a limitation in the super, for there are so many tuned circuits ahead of the detector that loading the last one has little effect on the over-all selectivity.

TYPES OF DETECTORS

Diode Detector. This two-element tube (cathode and plate) is the simplest of all detectors and operates because current will flow in only one direction through it. Thus, when an a.c. voltage such as a modulated i.f. signal is applied to the diode, only the positive peaks pass through. In order to be useful, this rectified current must pass through some kind of impedance across which it develops a voltage proportional to the current. This impedance is called the *diode load*.

A simple diode detector circuit is shown in Fig. 13. The circuit L_1-C_1 is tuned to the incoming carrier frequency (the i.f. in the case of a super second detector). The modulated r.f.

signal such as that shown in curve A in Fig. 13 is applied between the plate and the cathode of the diode through the diode load resistor R. If we assume that this resistor is not shunted with capacity, the resultant rectified current wave or the voltage wave across the diode load looks like that of curve B in Fig. 13, which is the upper halves of the input cycles. Condenser C_2 performs two useful functions. First, it by-passes the diode load at the carrier frequency so that the full modulated r.f. voltage is applied between the plate and the cathode of the diode, thus increasing the rectified current and the efficiency of detection. Condenser C_2 also filters the r.f. "bumps" or variations due to the individual r.f. cycles, as shown in curve C of Fig. 13. This condenser is charged to the peak value on each cycle. It discharges between peaks at a rate determined by the RC_2 time constant until the next peak appears. This set of curves is somewhat crude, and exaggerates the variations. Usually there are more r.f. cycles per audio cycle, and there is more r.f. filtering to smooth out the variations between this point and the first audio amplifier grid.

The d.c. voltage across the diode load has the polarity shown, and it varies with the modulation. The value of R is large compared with the diode resistance, so variations in the diode characteristics do not affect the diode load voltage, and a very small rectified current gives a rather high output voltage. This detector is quite linear for reasonably large applied voltages (above about 2 volts). The varying d.c. voltage of curve C is usually applied to an audio amplifier through a coupling condenser, so that only the audio component is transmitted.

The diode detector has quite low sensitivity, as it possesses no amplifi-

cation. It takes power from the tuned circuit, and this power is dissipated in the diode load. This loading of the tuned circuit reduces its Q and its selectivity. When a diode is used, the net load on the tuned circuit is about one-half the resistance of the diode load. The diode will not overload until its
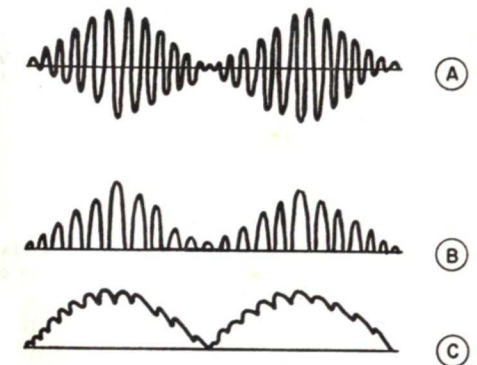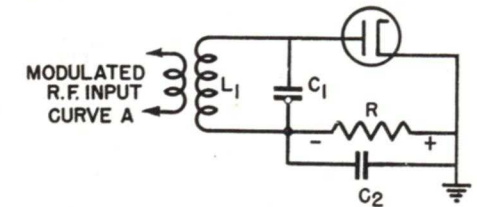
VOLTAGE WAVEFORMS

A. AT INPUT
B. ACROSS R WITH $C_2 = 0$
C. ACROSS R WITH $C_2 =$ OPTIMUM

FIG. 13. Simple diode detector. The output voltage across diode load R looks like curve B without C_2, and like curve C with C_2 added. The variations in curve C are not as large in practice because there are many more r.f. cycles per audio cycle.

emission is insufficient or its cathode-to-plate insulation breaks down, neither of which will happen in receiver applications. It may be considered as being completely free of overload troubles.

A practical diode detector circuit is shown in Fig. 14. The diode load is divided into two parts, R_1 and R_2, and

an extra by-pass condenser has been added, so that C_2, R_1, and C_3 form a filter network which filters out more of the r.f. "bumps" than are taken out in the circuit of Fig. 13 as shown in curve C. The i.f. carrier must be filtered out, or it may overload the first audio amplifier, and cause distortion. Condenser C_4 couples the audio voltage to the audio amplifier and prevents the d.c. voltage from affecting its bias. Potentiometer R_3 is used to vary the applied audio input. This is the audio volume control.
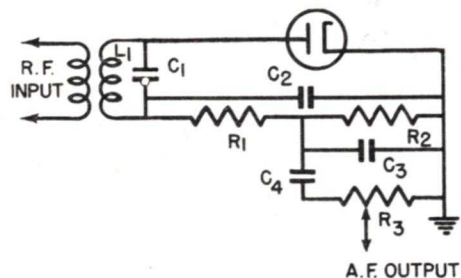

FIG. 14. Practical diode detector with r.f. filtering and audio output coupling.

A full-wave diode detector circuit is shown in Fig. 15. It makes use of both halves of the r.f. cycle as does the full-wave rectifier commonly used in the receiver power supply. The main advantage of this circuit is that very little r.f. voltage appears across R_1, the diode load, which is connected between the zero-potential tuned circuit center tap and ground. Very little r.f. filtering is thus required.

The major source of distortion in the diode detector is the time constant of the diode load and its associated by-pass condenser. The reactance of the condenser must be small compared to the diode load resistance at the carrier frequency, but it must be large compared to this resistance at audio frequencies. If this condenser is too large, the time constant will be such that it will not be able to charge as

fast as the voltage rises on a signal such as that of curve C in Fig. 13. Then it can be seen that the positive peak will be clipped, and distortion will result. It is usually possible to proportion this time constant to keep the distortion down to less than 2% at the higher audio frequencies, which is satisfactorily low.

Grid Leak Detector. The grid leak detector is a triode or multi-grid vacuum tube connected as shown in Fig. 16. The modulated r.f. signal from tuned circuit L_1-C_1 is applied between the tube grid and the cathode through resistor R_1 (called the grid leak) in parallel with grid condenser C_2. This arrangement is exactly the same as the diode detector, and operates the same. The grid acts as a diode plate, and the grid leak is the diode load. Condenser C_2 must be a good r.f. by-pass, but should not by-pass audio frequencies. The flow of rectified current through the grid leak to ground produces a d.c. voltage on the grid, making it negative with respect to the grounded cathode. The audio voltage developed across the grid leak is amplified as in any audio amplifier, and audio output is taken from the plate load resistor R_2. The tube also amplifies the r.f. signal, but it is by-passed by C_3 between plate and ground to prevent possible overload of subsequent amplifiers. Condenser C_3 must be a good r.f.

by-pass, but must have a high reactance at a.f. compared to the resistance of the plate load resistor R_2 to prevent excessive audio by-passing. Condenser C_4 couples the audio variations to the next stage, and removes the d.c. plate voltage component from the output.

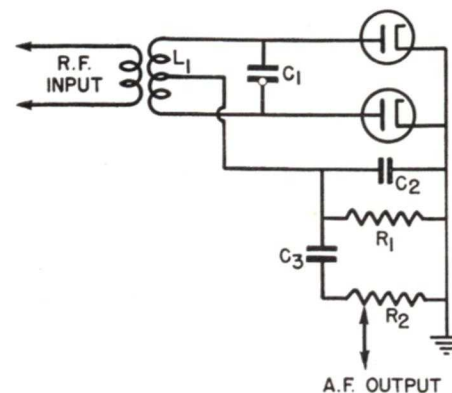The circuit of Fig. 16A shows a tri-

FIG. 15. For full-wave or push-pull diode detector much less r.f. filtering is needed than in the half-wave circuit of Fig. 14.

ode grid leak detector. R_2 and C_4 may be replaced by an audio output coupling transformer in some cases to reduce the d.c. plate voltage drop in the plate load resistor. The sensitivity of the detector is naturally increased if the gain of the audio amplifier portion is increased. This is accomplished by using a screen grid tube or a pentode tube as shown in Fig. 16B. Operation is the same as for the triode circuit. The screen by-pass condenser C_5 must have low reactance for both r.f. and audio frequencies. Resistors R_3 and R_4 are used as a voltage divider across the plate supply to give the proper screen voltage. This is usually considerably lower than it would be for conventional amplifier operation, because with no signal input, there is no grid bias, and the tube might otherwise draw excessive current.

The screen grid or pentode tube has

such a high plate impedance that the impedance of any practical audio transformer would be far too low to match the tube, and very little output could be taken from it. For this reason only resistance or impedance (an audio choke of about 500 henries inductance) coupling is used.

As the grid bias (the d.c. voltage across C_2, R_1) for audio amplifier operation increases with signal strength, strong signals may cut the plate current off completely, and the output becomes zero. The detector is then said to be "blocked." When close to this condition, the negative audio signal peaks may be cut, producing severe distortion. A tube used as a grid leak detector can deliver only about one-third the undistorted output voltage
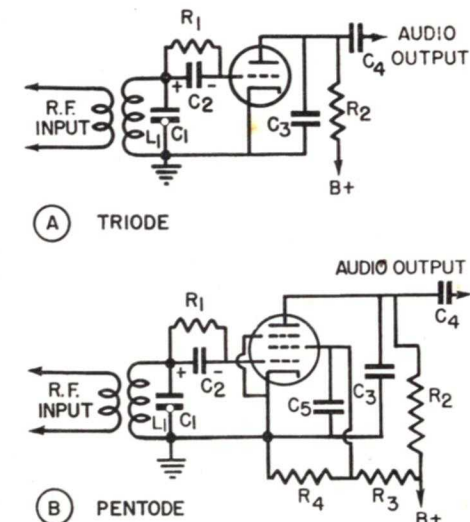

(A) TRIODE

(B) PENTODE

FIG. 16. Grid-leak detector circuits.

that it can deliver as a class A audio amplifier.

The grid leak detector has the highest sensitivity of all non-regenerative types used. It loads the associated tuned circuit as the diode does. It has poor linearity, and overloads easily. The sensitivity of this detector made

it very popular before the superheterodyne appeared. Sensitivity can now be obtained in other portions of the receiver, and the limited signal-handling ability of this detector makes it unsuitable for a super second detector where the signal level is high.

The Plate Detector. The plate detector uses a vacuum-tube amplifier whose grid bias is adjusted so that the plate current is practically cut off. A modulated r.f. signal applied to the grid results in no plate current change on negative swings, but the plate current can and does increase on the positive swings in proportion to the input voltage. There is a pulse of plate current on the positive half cycles of grid voltage, and no current change during the negative half cycles. This action is illustrated in Fig. 17. The average value
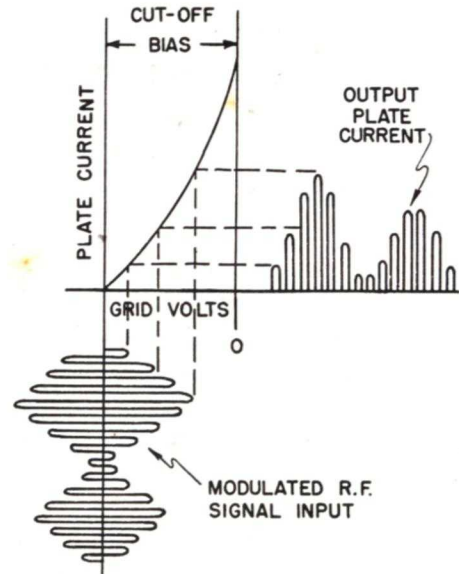

FIG. 17. Method of operation of a plate detector.

of plate current varies in accordance with the positive half of the modulation envelope, and reproduces the intelligence on the incoming signal. As in the diode, it is necessary to place a load in the plate circuit, so that this varying current will produce a corresponding varying voltage which constitutes the useful detector output.

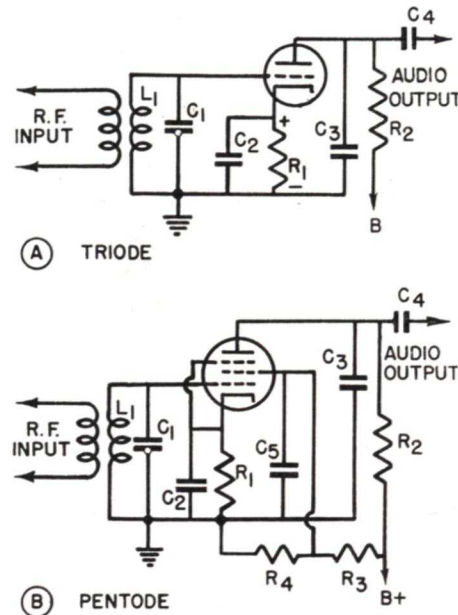A typical triode plate detector circuit is shown in Fig. 18A. The modu-

FIG. 18. Plate detector circuits.

lated r.f. signal in the tuned circuit L_1-C_1 is applied between grid and cathode through r.f. by-pass C_2. The cathode is grounded through R_1, and as the plate current must flow through it, the voltage drop across it is of proper polarity to place a negative bias on the tube. By making R_1 very much larger than is usual for amplifier operation, a bias near enough to cut-off is obtained to give operation similar to that shown in Fig. 17. Condenser C_2 must by-pass R_1 for both audio and radio frequencies to prevent excessive drop of either across R_1. Plate by-pass condenser C_3 removes most of the r.f. from the output, but it must be small enough so that it won't by-pass the higher audio frequencies excessively. The plate load resistor R_2 has the use-

ful audio output voltage developed across it from the plate current variations. Condenser C_4 couples the audio variations to the audio amplifier. The plate impedance is so high when biased nearly to cut-off, that impedance or resistance coupling is usual for either triode or pentode plate detectors. The extra components in the pentode circuit of Fig. 18B serves the same purposes as the corresponding ones in the grid leak pentode detector of Fig. 16B. Operation is similar to that of the triode plate detector, but with increased sensitivity.

The sensitivity of the plate detector is between that of the diode and the grid leak types. The linearity is good with the cathode bias system shown in Fig. 18, because as the grid is driven more positive, more cathode current flows, producing more negative bias. This detector does not overload as readily as the grid leak detector, but the grid can be driven positive on strong signals, whereupon it takes power from the tuned circuit, reducing the selectivity and producing distortion.

Degenerative or "Infinite Impedance" Detector. This circuit, shown in Fig. 19, resembles that of the plate detector, except that the audio output is taken from the cathode instead of the plate. Cathode resistor R_1 is by-passed by C_2 for r.f., but not for audio frequencies. As R_1 is common to both grid and plate circuits, there is negative audio feedback (degeneration). The output of this detector is thus less than its audio input, or in other words, it has a slight loss, similar to that of the diode. Condenser C_3 by-passes the plate for both audio and radio frequencies. The R_2-C_3 combination isolates the plate from the power supply.
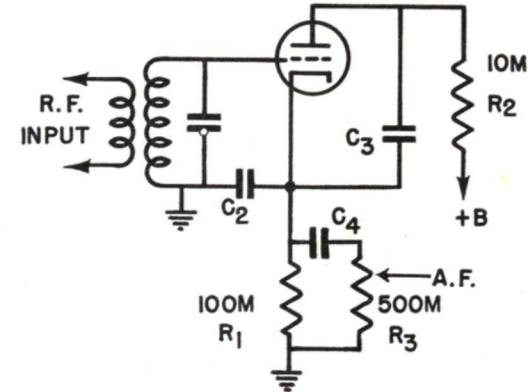

FIG. 19. Degenerative or "infinite impedance" detector.

With no signal, the plate current is nearly cut off, and it increases as the signal becomes larger, as in the plate detector. Because of the large initial bias which increases as the signal increases, it would take a very large signal to drive the grid positive, and this does not usually occur. As no grid current is drawn, no load is placed on the tuned circuit, and the input impedance is infinite.

This detector is insensitive like the diode, but is very linear, and will handle quite high inputs with low distortion. It is somewhat more complicated than the diode, and requires a d.c. plate supply. It is not much used as a super second detector because it does not lend itself as well to automatic volume control as the diode does.

Regenerative Detector. The analysis of the various preceding types of detectors indicates that signal frequency currents as well as the desired modulation frequencies flow in the plate circuit. A filter or by-pass condenser is usually used to keep these signal frequency currents out of the

(A) VARIABLE INDUCTIVE COUPLING

(B) VARIABLE CAPACITIVE COUPLING

(C) VARIABLE MUTUAL CONDUCTANCE ON PENTODE TUBE

FIG. 20. Regenerative detector circuits.

output. The regenerative detector with a triode or pentode tube, makes use of these currents, and feeds them back in proper phase to reinforce those in the input circuit. The increased sensitivity achieved in this way may be hundreds of times that of the same detector without regeneration.

Regeneration effectively reduces the resistance of the input circuit. The Q is thus raised, giving a larger voltage step-up equivalent to higher gain, and the selectivity is greatly sharpened.

Although the regenerative detector appears to give "something for nothing," it has several disadvantages. As the gain increases, so does the selectivity, and severe side-band cutting and distortion of voice signals result. To get maximum benefit from regeneration, it must be carried to a high degree, which is a delicate adjustment process requiring a fair degree of skill on the part of the operator. The setting is not stable, and performance may suffer with line voltage, temperature, or humidity changes, or when any mechanical jarring takes place. Excess regeneration produces oscillation, resulting in annoying audio beats or squeals. This occasional accidental maladjustment is nearly impossible to prevent. Optimum regeneration settings vary if the signal frequency is changed, so the regeneration control must be constantly readjusted if the detector tuning is changed. These disadvantages have resulted in the abandonment of this type of detector in modern communication receivers.

Typical regenerative detector circuits are shown in Fig. 20. The grid leak detector is most suitable for this purpose, and is used in all these circuits. That shown at A employs a triode that has in the plate circuit, a special winding which is coupled to the tuned circuit inductance of the grid

circuit. This special winding L_2 is known as the tickler. By varying the coupling between the coils, the feedback (regeneration) can be controlled. A control which rotates one coil with respect to the other, or which varies their relative spacing can thus be used as a regeneration control. The grid leak R_1 and grid condenser C_2 serve the same purpose as in the diode or grid leak detector. Condenser C_3 by-passes the low r.f. potential end of the tickler to ground for r.f., but must be small enough not to by-pass the higher audio frequencies excessively. The audio output voltage is developed across R_2, which is coupled to the audio amplifier through C_4.

The circuit shown in Fig. 20B also uses a grid leak detector, but the grid leak R_1 is connected from grid to ground instead of across the grid condenser C_2. This arrangement is optional on the grid leak detector and gives similar performance. In the plate circuit is a tickler coil L_2 with fixed coupling to L_1. The low potential side of the tickler is by-passed to ground through variable condenser C_5 which is used as the regeneration control. As the capacity increases, more r.f. flows through L_2, and the coupling and regeneration increase. The detected a.f. signal is fed to the plate load resistor R_2 through an r.f. choke. This is a coil having a high r.f. impedance and a low audio impedance. Condenser C_3 by-passes the other end of the r.f. choke to ground at r.f., but not at audio frequencies. The audio voltage is developed across R_2 and is coupled to the audio amplifier through C_4.

A pentode regenerative detector circuit is shown in Fig. 20C. The tube is used as a grid leak detector with feedback provided by the portion of L_1 between the cathode and ground, which corresponds to the tickler of the other

circuits shown. The screen voltage is used to set the amount of cathode current, and the screen voltage is varied by adjusting the tap on R_3, which is part of a voltage divider consisting of R_3 and R_4 across the plate supply. Varying R_3 varies the mutual conductance of the tube, and hence the amount of feedback current. This method of regeneration control can be made very smooth. Condenser C_5 by-passes the screen for audio and radio frequencies, and suppresses noise caused when the regeneration control R_3 is varied. The plate is isolated from the rest of the circuit by the r.f. grounded screen of the tube and is coupled only by the electron stream. Useful audio output voltage is developed across R_2 which is coupled to the output through C_4. Condenser C_3 by-passes r.f. from the plate circuit, but is too small to by-pass audio.

The regenerative detector is extremely sensitive, but is quite non-linear, unstable, and subject to blocking on strong signals.

Autodyne Detector. The regeneration or feedback of any of the regenerative detector circuits of Fig. 20 can be adjusted so that the detector oscillates at the resonant frequency of L_1-C_1. If this frequency is slightly different from that of an incoming signal, the audio beat frequency difference between the two will appear in the detector output, because of rectification. The grid leak detector is always used because it is sensitive, and the proper bias is automatically set by the grid leak - condenser combination. This method of operation is used for code reception so an audible tone will be produced. This detector is extremely sensitive when used in this manner if the regeneration is just sufficient to maintain oscillation. Because the tube supplies its own heterodyne voltage the
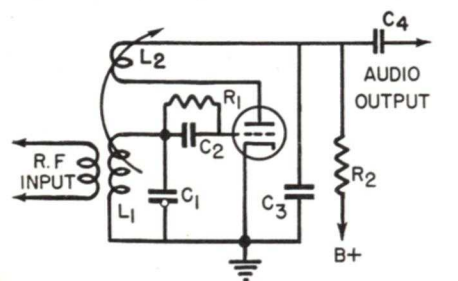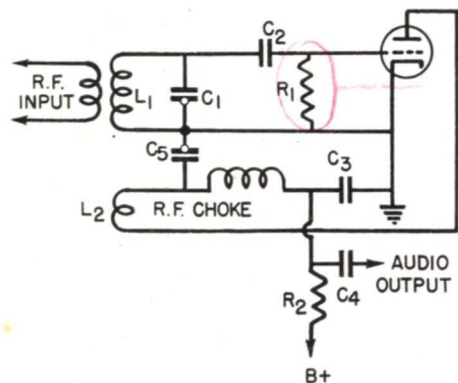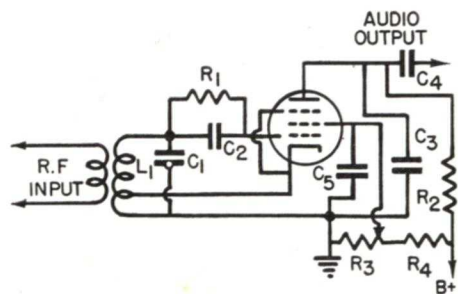
name "autodyne" or "self-heterodyne" is applied to this type of detector.

A low-frequency audio beat note cannot be obtained on strong signals, because the detector "locks in" with the incoming signal, or synchronizes its frequency with it, even though L_1-C_2 is tuned to a slightly different frequency. This effect can be reduced by increasing the regeneration so the detector oscillates more vigorously and has more of a "mind of its own." Or the incoming signal can be reduced by decreasing antenna coupling or r.f. gain if any r.f. stages are used. Either of these methods of reducing "locking in" or blocking, reduces the sensitivity to weak signals.

As the autodyne detector must be slightly detuned from the signal, it is not satisfactory at very low frequencies, because here the percentage of detuning must be so much greater than at high frequencies that an excessive loss of signal strength occurs. At these frequencies, a separate heterodyne oscillator should be used.

The control of feedback near the point of oscillation should be smooth. A hissing sound is heard just when oscillation starts with no signal present. Increasing regeneration past the point at which oscillation starts, decreases the intensity of the hiss, indicating a decrease of sensitivity. If the coupling is too loose and the voltage too high, oscillation starts violently with a "plop," and the detector is relatively insensitive. If oscillation starts with too low a voltage, conversion efficiency is low as a result of low plate current.

Heterodyne Detector. Any type of non-oscillating detector may be used as a heterodyne detector by feeding in voltage from a local oscillator along with that of the desired signal. The considerations which apply to this

method of detection have already been covered in an earlier Lesson. The heterodyne method of reception is superior to the autodyne method because there is less chance of strong signals detuning the local oscillator, and better over-all sensitivity to both strong and weak signals is achieved. Also less skill is required on the part of the operator.

Super-regenerative Detector. The limit of useful regenerative amplification of a regenerative detector is reached when the tube breaks into oscillation. When the voltage on the tube is swung intermittently above and below that required to sustain oscillation, the regeneration can be pushed still higher, and the tube starts and stops oscillating at the rate of swing, which is usually a low radio frequency, between 15 and 300 kc. Oscillation starts to build up during the time the plate voltage is increasing, but dies out during the time the plate voltage is decreasing. This dying out of oscillations is called "quenching." The purpose of the quench circuit is to prevent continuous oscillation and thus permit the regenerative detector to operate at peak efficiency. Oscillations should completely stop before starting again, for proper operation. This method of operation is called super-regeneration, and by its use, signals are amplified enormously before detection.

Typical super-regenerative detector circuits are shown in Fig. 21. That shown at A is essentially a triode oscillator employing what is called an "ultra-audion" circuit, where both plate and grid are above ground r.f. potential, and the coil center tap is nearly at ground potential. The r.f. choke keeps any residual r.f. out of the rest of the circuit. The useful audio voltage is developed across plate load resistor R_2 as in other detector circuits. The quench frequency oscillator has a con-

ventional circuit, and the voltage on its plate is that fed to the detector plate, thus plate-modulating the detector. Condenser C_4 by-passes r.f. voltage, but should not by-pass audio and quench voltages. Condenser C_3 couples the audio to the audio amplifier. Resistor R_4 varies both the quench and detector voltages together and is used as the regeneration control.

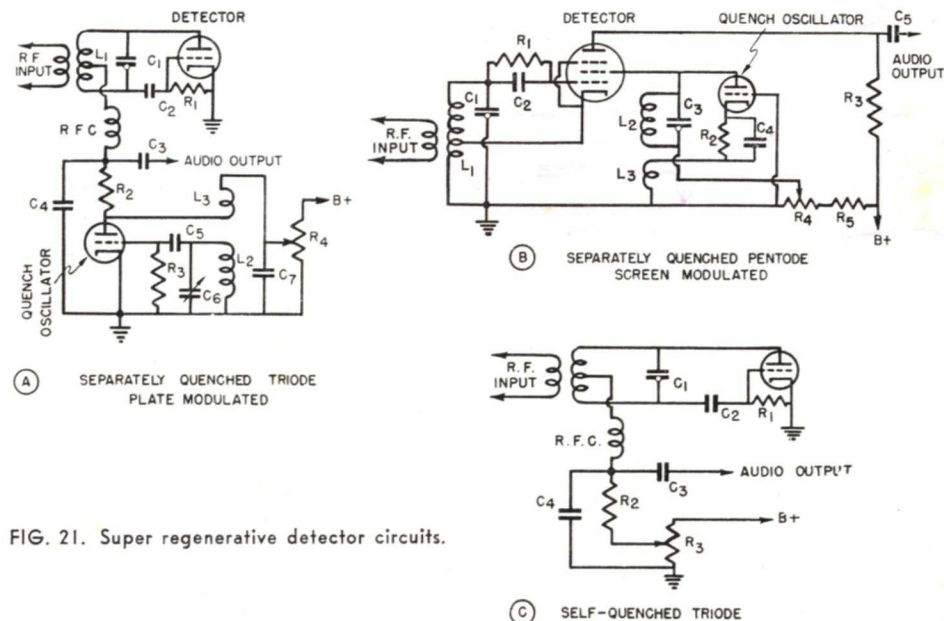The circuit of Fig. 21B has a pentode

FIG. 21. Super regenerative detector circuits.

tube connected as a regenerative grid leak detector with an electron-coupled output plate circuit, with audio developed across R_3 and coupled to the next amplifier through C_5. The quench oscillator has a grounded grid circuit, and its plate is tied to the detector screen. The screen voltage is thus varied at the quench frequency, which modulates the detector plate current. Resistor R_4 varies the quench and detector screen voltage together and is used as the regeneration control.

The circuit of Fig. 21C has no quench oscillator. In it, the values of

the grid leak R_1 and grid condenser C_2 are higher than usual so that after several cycles of oscillation, the negative charge on C_2 builds up faster than it can leak off through R_1, and the tube is blocked, stopping oscillation. When the charge finally leaks off, oscillation starts, and the process repeats. By proper choice of R_1 and C_2, this blocking and unblocking is made to occur at the proper speed for satisfactory quenching. This is thus known as a self-quenching circuit. Resistor R_3 varies the plate voltage, and is used for regeneration control. The output is by-passed and coupled to the next stage as in Fig. 21A.

The start of oscillations of the super-regenerative detector is triggered by random first circuit noise in the absence of a signal. This random rate produces a loud hiss in the output, which is much stronger than that of a similar regenerative circuit. A signal triggers the oscillations at a definite r.f. rate, and the stronger the signal

is, the more the hiss disappears. There is thus a marked automatic volume control effect, the output tending to be constant regardless of signal strength. This effect practically eliminates peaks of noise such as that from auto ignition systems, and is one of the big advantages of this type of detector. The many side bands produced by the quenching process require a large band width, so this detector is effective only at rather high frequencies (above about 20 mc.). The strong hiss noise in the absence of a signal is often a disadvantage. The self-quenched circuit gives such great sensitivity and output for such a small number of components that it is often used in ultra-high frequency "walkie-talkies," and other portable equipment.

SECOND DETECTOR FOR COMMUNICATIONS RECEIVERS

We have found that the requirements of the second detector of a super are those of high signal handling capability with freedom from overload, and good linearity. Sensitivity and light loading of the tuned circuit are desirable, but not at all necessary. From the preceding analysis of detector types, it is seen that either the diode or some type of plate detector appears to fit the requirements best. As we shall see later, the requirements of automatic volume control (a.v.c.) are much more easily met by the use of a diode than a plate detector. This and the greater freedom from overload have resulted in the nearly universal use of the diode as the communication receiver second detector.

Three typical complete second detector circuits are shown in Fig. 22. That shown at A employs a double diode, only one section of which is shown. The other section is used for the noise limiter which will be dis-

cussed in another Lesson. The i.f. signal is inductively coupled to the tuned circuit L_1-C_1, which is the secondary of the last i.f. transformer. It is applied to the diode detector through the diode load consisting of R_1, R_2, R_3, and the tuning microammeter M in series, all of which is by-passed at intermediate frequencies, but not at audio frequencies by C_2. Condensers C_3, C_4, and C_5 are similar by-passes for portions of the diode load, which produce better filtering action. Resistor R_4 feeds the noise limiter circuit, to be described later elsewhere. The audio output at the junction of R_2 and R_3 is coupled to audio gain control R_6 through C_7. Voltage from this same junction is filtered by R_5 and C_6, and is used for automatic volume control. The other noise limiter connection is also taken from this junction. The full audio voltage developed across the diode load is not used in this circuit, in order to get better filtering of the i.f. The current of meter M increases with carrier increase, but cannot follow the relatively fast audio variations. It is used to tune in signals by observing maximum deflection, or it can be used to compare signal strengths when calibrated, even though the a.v.c. may make different strength signals sound nearly equal in strength.

The circuit of Fig. 22B is similar to that of 22A. The diode load consists of only two resistors R_1 and R_2 in series, by-passed by C_2. The audio voltage at the tap between R_1 and R_2 is coupled through C_4 to the audio volume control R_5, from which any value less than the maximum may be applied to the audio amplifier. The audio voltage variations at the tap between R_1 and R_2 are filtered out by resistor R_4 and condenser C_5 to give a negative d.c. voltage proportional to the carrier level. This d.c. voltage is used for a.v.c.

purposes, being applied to the control grids of the i.f. mixer and r.f. tubes. Thus the receiver sensitivity will be reduced on strong signals and will be high on weak signals, as they produce little a.v.c. voltage. When switch SW_2 is closed, the a.v.c. voltage is shorted to ground, but the diode load is not shorted, because of the decoupling provided by the relatively high resistance of R_4. This switch is closed in c.w. code reception to prevent raising the receiver gain, filling the intervals between keying with noise. This is the a.v.c. ON-OFF switch. The R_3-C_3 combination is used for a peak noise limiter circuit, the rest of which is not shown. When a beat note is desired for c.w. reception, heterodyne voltage from the beat-frequency oscillator is fed into the diode through condenser C_6. Usually the beat oscillator plate voltage is turned on with a switch ganged to SW_2, which simultaneously disables the a.v.c.

Fig. 22C illustrates the use of a double diode and a pentode in one tube envelope, where one diode is used as the second detector, the other is used for a.v.c., and the pentode is used for the last i.f. amplifier. The pentode section acts as a conventional i.f. amplifier getting its bias from the a.v.c. diode load. The i.f. plate load L_2-C_2 is inductively coupled to untuned secondary L_3. The low potential end of L_3 is by-passed to ground by C_8. The high potential end is connected to the a.v.c. diode plate and is capacitively coupled to the a.f. diode plate through C_3. In many circuits where two diodes are used for these two functions, the a.f. diode is directly connected, and the a.v.c. is capacitively coupled. It makes little difference. The common diode and pentode cathode is by-passed to ground through C_7. The a.v.c. diode load consists of R_3, R_4, and R_5 in series. The
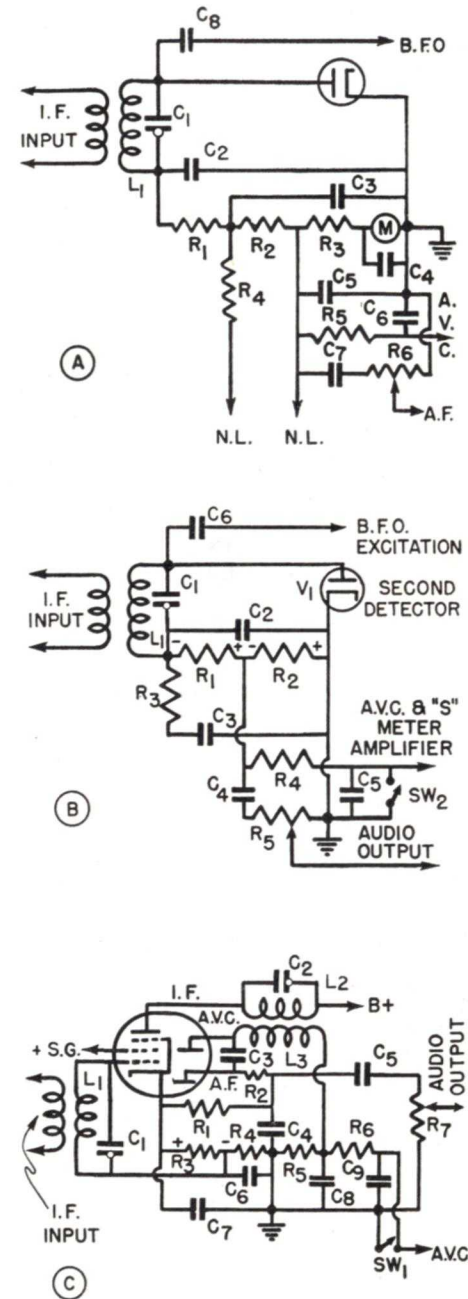

FIG. 22. Typical communication receiver second detector circuits. A, Hammerlund HQ-120 second detector circuit. B, Hallicrafters SX-28 second detector circuit. C, RCA AR-60-S second detector circuit.

cathode is positive with respect to ground, as a result of the flow of plate and screen current in the pentode section of the tube. The drop across R_3 is used as bias on the pentode. The voltage across R_3 and R_4 acts as delay bias on the a.v.c. diode. When the a.v.c. diode voltage exceeds the delay bias, it adds to the bias on the pentode section, reducing its gain. Filter R_6-C_9 removes audio variations from the a.v.c. voltage used on the remainder of the receiver. The a.v.c. is disabled for c.w. reception by closing switch SW_1. The audio diode load consists of R_1 and R_2 in series. Their junction is by-passed for i.f. but not for audio by C_4. This point is coupled to volume control R_7 by condenser C_5.

Second detector circuits such as that just described look rather complex compared to the elementary diode circuit, but a great deal of the extra components are associated with the a.v.c. circuits which cannot very well be divorced from the diode circuits. Most of the extra elements are merely for filtering and decoupling purposes and do not affect the basic operation of the circuit.

The Beat Frequency Oscillator

For radiotelegraph code reception or when searching for very weak signals, it is necessary to be able to detect the presence or absence of a carrier, even though it is not modulated. This is done by obtaining an audible heterodyne beat between the i.f. and a special local oscillator known as the "beat frequency oscillator" (b.f.o.).

This oscillator is employed to give heterodyne detection at the second detector in the same manner that the i.f. is obtained by heterodyne detection in the first detector. When the signal is not present, no beat is produced. Hence no audio output is obtained, even though the beat oscillator is operating. Thus, the code keying of a signal is reproduced by an audio tone whose pitch or frequency depends upon the difference frequency between the b.f.o. and the i.f. The audio beat may be obtained by having the b.f.o. either above or below the i.f., and as we have seen, audio images may be produced in a manner similar to that in which high-frequency images are produced.

Since the i.f. is essentially constant with different frequency signals, the b.f.o. can operate on a relatively fixed frequency, with just enough tuning adjustment to vary the beat note to suit the operator, to make it fit an audio filter, or to correct for signal or receiver oscillator frequency drift. The tuning range is usually set to give a 10 kc. total range with 5 kc. on each side of zero beat. Thus, the b.f.o. of a receiver having a 465 kc. i.f. would cover from 460 to 470 kc., and would normally be set near either 464 or 466 kc. to give about a 1-kc. beat note when the signal was tuned for optimum output.

The harmonic content of the b.f.o. must be low, and it is desirable to operate the detector so that harmonic generation in the detection process is minimized. If this is not done, harmonics of the b.f.o. may become coupled into the front end of the receiver, and will be tuned in just as any other incoming r.f. signal. These spurious steady signals will always be found at the same places on the dial, and can be easily checked, as they are still there when the antenna is removed. To minimize this effect, the b.f.o. should not deliver any more voltage than is necessary for proper beating action. It should have a comparatively large capacity and small inductance to minimize harmonic generation. It should be carefully shielded and decoupled from the power supply to prevent stray coupling to the front end of the receiver. When all these precautions are taken, no harmonic responses detectable above the receiver noise level are obtained.

The i.f. voltage at the second detector may be 50 volts or more for a strong signal. The b.f.o. voltage at this point must be equal to or greater than the i.f. voltage, otherwise the audio output will be limited to the b.f.o. voltage value. If the b.f.o. excitation is less than the i.f. signal voltage, the signal may have a coughing sound as if the beat note were being smothered, and copying the keying may be made more difficult. If the excitation is excessive, there may be more tendency toward spurious harmonic responses, as previously explained, and the noise level is raised due to beating of the b.f.o. with all the noise components coming through the receiver pass-band. There is usually a definite increase in "hiss" when the b.f.o. is turned on with no signal present. The b.f.o. is always run at a level to give proper beating with the strongest signal, in the same way that the high-frequency local oscillator is operated.

The b.f.o. voltage is usually coupled to the high i.f. potential point of the second detector through a very small (about 5 micromicrofarad) coupling condenser. Often the coupling is merely stray wiring capacity, but this usually shows poor design and shielding practice, leading to the "birdie" spurious responses from b.f.o. harmonics.

Occasionally the b.f.o. is inductively coupled to the second detector by means of a special coil near the last i.f. tuned circuit. Another system is to
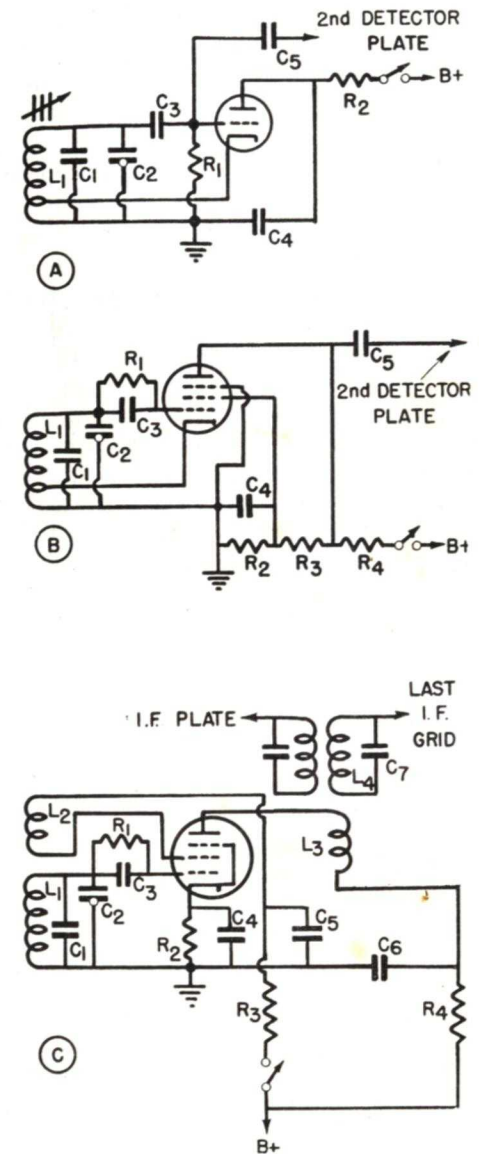

FIG. 23. Typical b.f.o. circuits. A, Triode grounded-plate Hartley with capacity coupling. B, electron-coupled pentode with capacity coupling. C, tickler-feedback, electron-coupled pentode with inductive coupling to last i.f. stage.

couple, either capacitively or inductively, the b.f.o. voltage to the grid of the last i.f. stage. The b.f.o. voltage is then amplified and fed to the second detector along with the desired i.f. signal.

Typical b.f.o. circuits are shown in Fig. 23. That at A has a triode tube in a grounded plate Hartley circuit. The cathode is tapped up from the grounded end of tuning inductance L_1. Fixed condenser C_1 is of fairly large capacity to give a high C/L ratio. The inductance is varied by moving an iron-core slug on a threaded screw, until the frequency is the desired value with the condenser C_2 at mid-scale. Variable condenser C_2 is rather small, giving "band-spread" action. It is often operated by a front panel knob marked "pitch control." It is usually capable of varying the beat note about plus or minus 5 kc. from zero beat. Grid leak R_1 and grid condenser C_3 serve the same purposes as for high-frequency oscillators. Condenser C_4 grounds the plate for r.f., and with R_2 acts as a filter to keep r.f. out of the power supply. Condenser C_5 is only a few micro-microfarads, and couples the high r.f. potential oscillator grid to the plate of the second detector diode.

The circuit of Fig. 23B is similar to that of 23A except that an electron-coupled pentode is used, and the necessary screen-dropping resistors and by-pass condensers are added.

The circuit of Fig. 23C has a tickler-feedback, electron-coupled pentode with some cathode bias furnished by the cathode current flowing through R_2. A special coupling coil L_3 in the plate circuit is coupled to the last i.f. amplifier grid circuit L_4-C_7, so that this stage amplifies both i.f. and b.f.o. voltages simultaneously. This circuit can operate with very low b.f.o. output, as it is subsequently amplified. Care must be taken not to overload the amplifier with excessive excitation.

Lesson Questions

Be sure to number your Answer Sheet 34RC.

Place your Student Number on *every* Answer Sheet.

Most students want to know their grade as soon as possible, so they mail their set of answers immediately. Others, knowing they will finish the next Lesson within a few days, send in two sets of answers at a time. Either practice is acceptable to us. However, don't hold your answers too long; you may lose them. Don't hold answers to send in more than two sets at a time or you may run out of Lessons before new ones can reach you.

1. A super is adjusted to 2738 kc. The i.f. is 475 kc.; what is the frequency to which the input circuit of the second detector must be tuned?

2. What is the chief advantage to be gained in the utilization of high intermediate frequencies in a superheterodyne receiver?

3. What determines the minimum i.f. band width in a receiver to be used for amplitude-modulated reception?

4. What are the advantages of using powdered-iron cores in i.f. transformers?

5. How is the position of the "slot" in the response curve of a crystal filter varied?

6. Why is it necessary to provide a radio frequency by-pass condenser in the plate circuit of a grid-leak condenser detector tube?

7. Draw a simple schematic diagram of a regenerative detector that uses variable inductive coupling.

8. What is the purpose of the quench circuit in a super-regenerative detector?

9. Why is it desirable to remove a.v.c. action for c.w. code reception?

10. What is the purpose of the b.f.o.?

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SHALL I CHANGE JOBS?

Before you can answer this question, you must set your goal in life. Do you want a professional career as a broadcast or shipboard operator, or as an aviation communications expert?

With a goal in mind, apply these three questions to your present job: 1. Does it give reasonable pay for the present? 2. Does it give knowledge, training or experience which will be worth money to you in the future? 3. Does it give you prestige, or bring you in touch with men who can help you to attain your goal?

Judge each new job opportunity also by these three questions. Hold each job only long enough to learn what is needed for the next job. Then, if a vacancy up ahead is unlikely, you are justified in changing to a corresponding job somewhere else.

The important thing is to keep going ahead. Learn something every day. When you reach the point where you are no longer learning, no longer progressing, then you are already moving backward, and it's time for you to change jobs.

First, though, be sure you really are prepared for a better job. Be sure your training is complete, and be sure you are able to apply it.

J. E. SMITH