THE SUPERHETERODYNE RECEIVER

CONTENTS

INTRODUCTION

This pamphlet deals with the reception of amplitude-modulated waves only.

A radio receiver is designed to produce an audio-frequency output corresponding to the modulation of the wanted signal received at the aerial.

At any particular instant several signals of different frequencies may simultaneously exist in the aerial. The first function therefore, of a receiver is to select the wanted signal. This selection is most easily carried out by utilizing the resonant effects of tuned circuits.

In addition to selection, the required signal must also be amplified in order that the input to the detector is sufficient for distortionless detection.

A third requirement is that the audio frequency output of the detector be amplified in order to obtain sufficient power to operate the loudspeaker.

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The simplest way of satisfying these requirements is to use a Tuned Radio Frequency (T.R.F.) receiver, a block diagram of which is shown in Fig. 1.

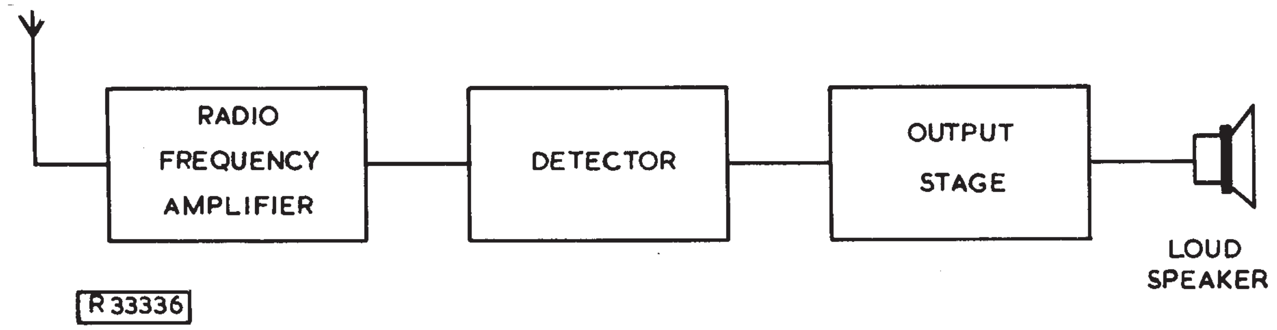


Fig. 1

It can be seen that a T.R.F. receiver consists of a radio-frequency amplifier, a detector, an output stage and a loudspeaker. The T.R.F. receiver, however, has a number of serious disadvantages which make it unsuitable for use with the crowded frequency bands of the present day.

The ability of the T.R.F. receiver to select the desired signal from others on adjacent frequencies is limited by the number of tuned circuits which can be incorporated without complicating the tuning arrangements of the receiver.

All the tuned circuits in this type of receiver have to be variable. Therefore if a number of controls is to be avoided, some means must be adopted for tuning all the circuits from one control. Unfortunately the greatest number of circuits which can be so controlled without excessive practical difficulties is four, and this number is seldom really adequate for the precise tuning required nowadays.

In addition, it must be remembered that an amplitude-modulated carrier wave consists of the carrier and sidebands and if reasonable quality reception is to be obtained these frequencies must be received with the relative magnitudes of the components substantially unchanged. A combination of single-tuned circuits tends, however, to select one frequency only and to reject the sidebands. Consequently, when single-tuned circuits are used, good selectivity is unobtainable without impairment of quality. It might be thought that band-pass circuits could be used but it is found that the bandwidth of such circuits varies with the frequency of operation.

A further disadvantage of the T.R.F. receiver is that the amount of r.f. amplification that can be employed is limited by the difficulties of avoiding oscillations in variable-tuned r.f. stages.

Most of the difficulties mentioned could be overcome if the selecting circuits worked at a fixed radio frequency and any desired signal frequency could be converted to this fixed frequency. This is the principle of the Superheterodyne Receiver, which combines the advantages of a fixed-frequency amplifier with variable tuning.

THE SUPERHETERODYNE PRINCIPLE

Most of the difficulties experienced with a T.R.F. receiver result from having to tune over a wide range of frequencies. The design of an amplifier for use at one fixed frequency is comparatively easy, variations of gain and selectivity with frequency no longer presenting any problem. If, therefore, it were possible to convert any desired signal frequency to a fixed frequency, a much more efficient receiver could be designed.

In a superheterodyne receiver the desired signal frequency is converted to a fixed radio frequency, known as the intermediate frequency (i.f.), by a process known as 'frequency changing'. In a frequency changer the desired signal frequency is combined with the output of a variable-frequency oscillator to give an output which contains frequencies equal to the sum and difference of the signal and oscillator frequencies. For reasons explained later in the pamphlet the difference is generally chosen as the intermediate frequency. Thus, by varying the frequency of the local oscillator any desired signal frequency may be converted to the intermediate frequency.

The choice of intermediate frequency is determined by a compromise between two conflicting factors. A low intermediate frequency is desirable in order to obtain high amplification with good selectivity without instability difficulties. A high intermediate frequency is desirable in order to obtain a good 'signal of image ratio' as will be seen later. In this country an intermediate frequency of between 460 kHz and 470 kHz is generally chosen, since these frequencies represent a reasonable compromise between the aforementioned factors. For simplicity, an intermediate frequency of 465 kHz will be considered throughout this pamphlet.

It has been stated that the intermediate frequency is equal to the difference between the signal frequency and the frequency of the local oscillator. Thus with an intermediate frequency of 465 kHz the local oscillator must be tuned to a frequency either 465 kHz greater than, or 465 kHz less than, the desired signal frequency. If the oscillator frequency, f_o, is higher than the signal frequency, f_s, the intermediate frequency will be $f_o - f_s$. If the oscillator frequency is lower than the signal frequency, the intermediate frequency will be $f_s - f_o$.

It would appear that the frequency of the local oscillator could be adjusted to either 465 kHz above the signal frequency or 465 kHz below the signal frequency. In practice however, the local oscillator frequency is made higher than the signal frequency since a considerable advantage is gained by doing so.

Consider that a particular receiver operates over a frequency range of 550 kHz to 1500 kHz. If the frequency of the local oscillator is set higher than the signal frequency the oscillator must vary from (550 + 465) kHz to (1500 + 465) kHz or 1015 kHz to 1965 kHz, a frequency ratio of $\frac{1965}{1015}$ or 1.94. If the frequency of the local oscillator is set lower than the signal frequency the oscillator must vary from (550 - 465) kHz to (1500 - 465) kHz or 85 kHz to 1035 kHz, a frequency ratio of $\frac{1035}{85}$ or 12.2.

There is a practical limitation to the range of frequencies which can be tuned by a single inductor and a variable capacitor, owing to the set minimum and maximum values of capacitance obtainable. A frequency ratio of 1.94, required when the frequency of the local oscillator is set higher than the signal frequency, is easily obtained but a frequency ratio of 12.2 would prove very difficult.

The difference frequency is generally chosen as the intermediate frequency because it is easier to obtain the desired amplification and selectivity at a fairly low radio frequency, and as explained in the previous paragraph, it is desirable for the signal frequency to be lower than the local oscillator frequency.

Second Channel Interference

A superheterodyne receiver has one disadvantage which necessitates the use of at least one signal-frequency tuned circuit before the frequency changer.

Consider that a signal frequency, f_s, is to be received. Then the frequency of the local oscillator must be f_o, where $f_o - f_s$ equals the intermediate frequency. There will, however, be another signal frequency, known as the image frequency, f_i, which will also combine with the local oscillator frequency to give an output at the intermediate frequency, i.f.

This frequency, f_i, will be such that $f_i - f_o = \text{i.f.}$

$$f_i = f_o + \text{i.f.}$$

$$f_o = f_s + \text{i.f.}$$

$$f_i = f_s + 2 \times \text{i.f.}$$

The image signal is separated from the wanted signal by twice the intermediate frequency. Thus, if the image signal is not to be received, variable tuned circuits must be provided, before the frequency changer, which will reject the image signal when the receiver is tuned to the signal frequency. These tuned circuits must be variably tuned because the desired signal frequency, and thus the image frequency will vary.

It is not difficult to provide the necessary rejection of the image frequency if a high intermediate frequency is employed, when the frequency spacing between the signal and image frequencies may be an appreciable fraction of the signal frequency. The usual intermediate frequency employed in broadcast receivers is 465 kHz, giving 930 kHz separation between the signal and image frequencies.

The suppression of second-channel interference is often inadequate with short-wave receivers. When a normal intermediate frequency of, say, 465 kHz, is employed it is difficult to obtain sufficient selectivity to give adequate rejection of image signals at very high frequencies since the 930 kHz separation will be only a few per cent off-tune.

To overcome this disadvantage receivers built especially for short-wave reception employ two different intermediate frequencies and two stages of frequency-changing. By using a higher frequency for the first i.f. stage there is a greater frequency difference between the wanted and image signals, and thus the response of the receiver to image signals is reduced.

Signal to image ratio

If the desired signal and the image signal are applied at equal strength, in turn, to the aerial of a receiver, and the receiver tuned to the frequency of the desired signal, the ratio of receiver output voltage obtained from the desired signal to that obtained from the image signal is known as the signal to image ratio. This ratio should be of the order of 60 dB for good image rejection.

The signal to image ratio is dependent upon the selectivity of the r.f. tuned circuits preceding the frequency changer and upon the value of the intermediate frequency, because an increase in the value of the intermediate frequency increases the frequency spacing between the desired signal and the image signal.

Second channel whistle and crosstalk

Consider the case of a receiver in which the r.f. tuned circuits do not completely prevent the passage of the image signal but merely attenuate it.

Assume the receiver is tuned to receive a signal of 1500 kHz, when the frequency of the local oscillator will be (1500 + 465) kHz or 1965 kHz. The frequency of the image signal will be (1500 + 930) kHz or 2430 kHz, and an unwanted signal at the intermediate frequency will be present in the output of the frequency changer. Thus the output of the receiver will contain two signals, one due to the desired signal and the other due to the image signal. However, the a.f. output due to the image signal is at a much lower level than the wanted output and thus appears in the form of overhearing or crosstalk.

Consider now that a signal at 2432 kHz appears in the aerial, i.e. 2 kHz from the image signal. This signal will be attenuated by the r.f. stage and will produce an intermediate frequency of (2432 - 1965) kHz or 467 kHz. This frequency is too near the intermediate frequency of 465 kHz to be rejected by the band-pass circuits of the i.f. amplifier and will appear at the input to the detector together with the wanted signal. The presence of these two frequencies at this point results in a 2 kHz whistle in the output of the receiver. Such a whistle is known as 'second channel whistle'.

In a similar manner, slight detuning of the local oscillator will also produce whistles in the output of the receiver. Assume that, as before, a signal of 1500 kHz is to be received and that the frequency of the local oscillator is 1966 kHz instead of 1965 kHz. Frequencies of (2430 - 1966) kHz and (1966 - 1500) kHz or 464 kHz and 466 kHz will be produced, resulting in a 2 kHz whistle in the output of the receiver.

The whistle will vary in frequency as the receiver is tuned over a small range of frequencies either side of the wanted signal. This is illustrated by Table 1.

TABLE 1

RECEIVER TUNED TO (kHz)	OSCILLATOR TUNED TO (kHz)	I.F. DUE TO 1500 kHz WANTED SIGNAL (kHz)	I.F. DUE TO 2430 kHz IMAGE SIGNAL (kHz)	FREQUENCY OF WHISTLE IN THE RECEIVER OUTPUT (kHz)
1504	1504 + 465 = 1969	469	461	8
1503	1968	468	462	6
1502	1967	467	463	4
1501	1966	466	464	2
1500	1965	465	465	0
1499	1964	464	466	2
1498	1963	463	467	4
1497	1962	462	468	6
1496	1961	461	469	8

In a well designed receiver, the r.f. tuned circuits will attenuate the image signal to an extent sufficient to prevent second channel whistle becoming noticeable.

Interference Due to Signals at, or Near, the Intermediate Frequency

Another form of interference to which a superheterodyne receiver is prone is that due to signals at, or near, the intermediate frequency of the receiver. If such a signal is able to penetrate as far as the i.f. amplifier, it will be amplified and cause a whistle to be heard superimposed on every station received.

To prevent such signals reaching the i.f. amplifier, a simple rejector circuit tuned to the intermediate frequency, may be placed in series with the aerial.

The r.f. tuned circuits cannot be used to block such signals since the low frequency end of the medium waveband is near to the 465 kHz intermediate frequency.

Intermodulation

Whistles may also be caused by signals, differing in frequency by the intermediate frequency, which, if allowed to reach the frequency changer, will produce a signal at the intermediate frequency. This signal will produce whistles as the receiver is tuned by beating with the intermediate frequency produced by the wanted signal.

Intermodulation is prevented by making the r.f. tuned circuits sufficiently selective to prevent the interfering signals reaching the frequency changer.

Sideband (Adjacent-Channel) Interference

In Europe the spacing of medium wave broadcast stations is only 9 kHz and a whistle at this frequency will be heard if the level of an adjacent signal is less than approximately 60 dB down on the level of the wanted signal at the input to the detector.

This form of interference can only be reduced by the band-pass circuits of the i.f. amplifier. This requires that the band-pass circuits employed must have a gain/frequency response that is uniform over a bandwidth of 465 kHz $\pm$ 4.5 kHz, i.e. a bandwidth of 9 kHz, and falls rapidly at frequencies outside this band.

Local Oscillator Radiation

Another common type of interference is caused by radiation of the local-oscillator signals from receivers. In many of the cheaper makes of superheterodyne receiver, not provided with an r.f. stage, a considerable amount of r.f. current at the oscillator frequency may find its way into the aerial circuit and be radiated from the aerial. The way in which this radiation causes interference with other receivers can be seen from the following example.

From certain areas of the country reports have been made of interfering whistles heard by listeners tuned to 1151 kHz. It has been found that the interference is caused by certain nearby receivers tuned to 692 kHz. In a receiver using an intermediate frequency of 460 kHz and tuned to 692 kHz, the frequency of the local oscillator is 1152 kHz. If this oscillator frequency is strongly radiated from the receiver aerial, it will be received by neighbouring receivers tuned to 1151 kHz.

The r.f. tuned circuits of the receivers will not be able to discriminate between signals of 1151 kHz and 1152 kHz and therefore both signals will reach the frequency changer and combine with the local oscillator frequency of 1611 kHz to produce frequencies of 460 kHz and 459 kHz. These two frequencies will beat together and produce an audio frequency of 1 kHz.

Radiation from the local oscillator of some receivers can cause interference up to 250 yards and in closely populated areas many receivers may be affected. Efficient screening of the oscillator circuit and the provision of an r.f. amplifier stage are essential for complete prevention of the interference radiation. Also in cases of persistent local interference of this nature the trouble may be overcome by altering the intermediate frequency of interfering receivers.

Harmonics of the Intermediate Frequency

Harmonics of the intermediate frequency are generated by the detector stage and, if the screening and layout of the r.f. circuits are poor, may be fed back into the r.f. circuits.

Consider a receiver having an intermediate frequency of 465 kHz and tuned to a frequency of 928 kHz. The second harmonic of the intermediate frequency is 930 kHz and if this appears in the r.f. circuits it will beat with the wanted signal and produce a 2 kHz whistle.

Harmonics of the Local Oscillator Frequency

Harmonics of the local oscillator frequency may beat with unwanted stations to produce signals at the intermediate frequency.

Consider a receiver tuned to receive a 600 kHz signal. The frequency of the local oscillator will be (600 + 465) kHz or 1065 kHz. The second harmonic of this is 2130 kHz. If signals at (2130 ± 465) kHz, i.e. 1665 kHz or 2595 kHz, are present at the frequency changer they will produce intermediate frequencies that will beat with the wanted intermediate frequency and produce a whistle that varies in pitch as the tuning of the receiver is altered.

This form of interference can be limited only by improving the selectivity of the r.f. tuned circuits and thus reducing the level of the interfering signals

Functions of the Various Stages of a Superheterodyne Receiver

A block diagram of a superheterodyne receiver is shown in Fig. 2. It can be seen that a superheterodyne receiver consists of a radio-frequency circuit, a frequency changer plus a local oscillator, an intermediate-frequency amplifier, a detector, an audio-frequency amplifier and a loudspeaker.

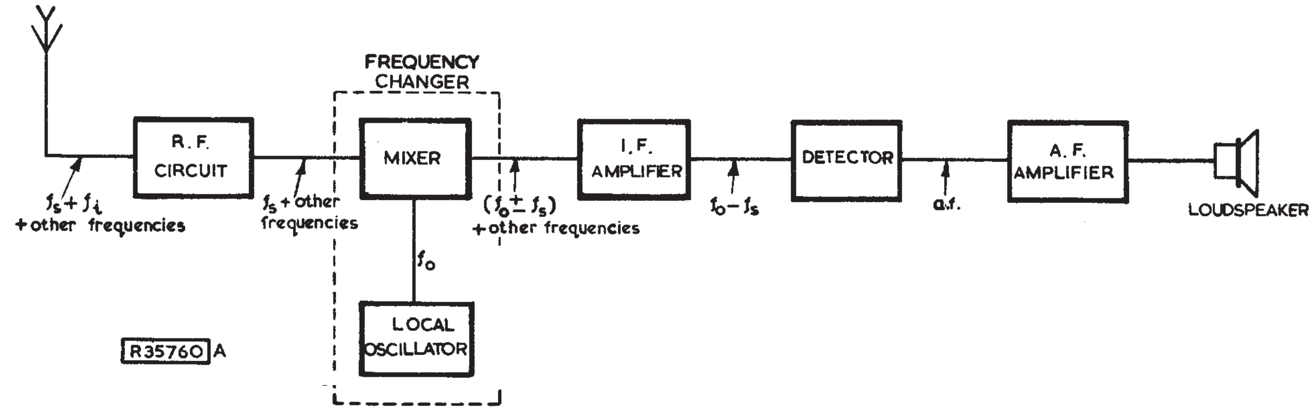


Fig. 2

The input to the r.f. circuit consists of all the frequencies simultaneously present in the aerial and one of the functions of the r.f. circuit is to suppress the image signal together with those signals which would cause intermodulation. In some cases the r.f. circuit also includes an r.f. amplifier thus improving the signal-to-noise ratio of the receiver.

The r.f. circuit also provides a means of efficiently coupling the aerial to the receiver proper in order to utilize as efficiently as possible the r.f. energy in the aerial.

The output of the r.f. circuit thus consists of the signal frequency, f_s, plus certain other frequencies not removed by the r.f. tuned circuits. These frequencies are fed into the frequency changer together with the output of the local oscillator.

The function of the frequency changer is to convert the signal frequency, f_s to the intermediate frequency of the receiver as efficiently as possible. It should be noted that, in practice, the frequency changer and local oscillator valves are often combined within a single envelope.

The output of the frequency changer contains the intermediate frequency, $f_o - f_s$, plus certain other frequencies, and is fed into the i.f. amplifier.

The i.f. amplifier provides most of the gain and selectivity of a superheterodyne receiver and has a gain and bandwidth that remain constant for all values of signal frequency. The amplifier employs double-tuned coupled circuits that have a bandwidth of approximately 9 kHz centred on 465 kHz and therefore the output of this stage consists of the intermediate frequency, $f_o - f_s$, only.

The function of the detector stage is to extract the audio frequency intelligence contained in the output of the i.f. amplifier. In modern receivers the detector is almost always a diode detector, since this type of detector introduces very little distortion and also provides a d.c. output which is proportional to the level of the carrier received and may be used for Automatic Gain Control.

The a.f. amplifier is a power amplifier and provides sufficient a.f. power to operate the loudspeaker.

Selectivity

The selectivity of a radio receiver is its ability to discriminate between the desired signal and signals at other frequencies.

The selectivity is fully expressed by a response curve, consisting of the output of the receiver plotted against the frequency of the input signal and referred to the point of maximum output. A typical selectivity curve is shown in Fig. 3.

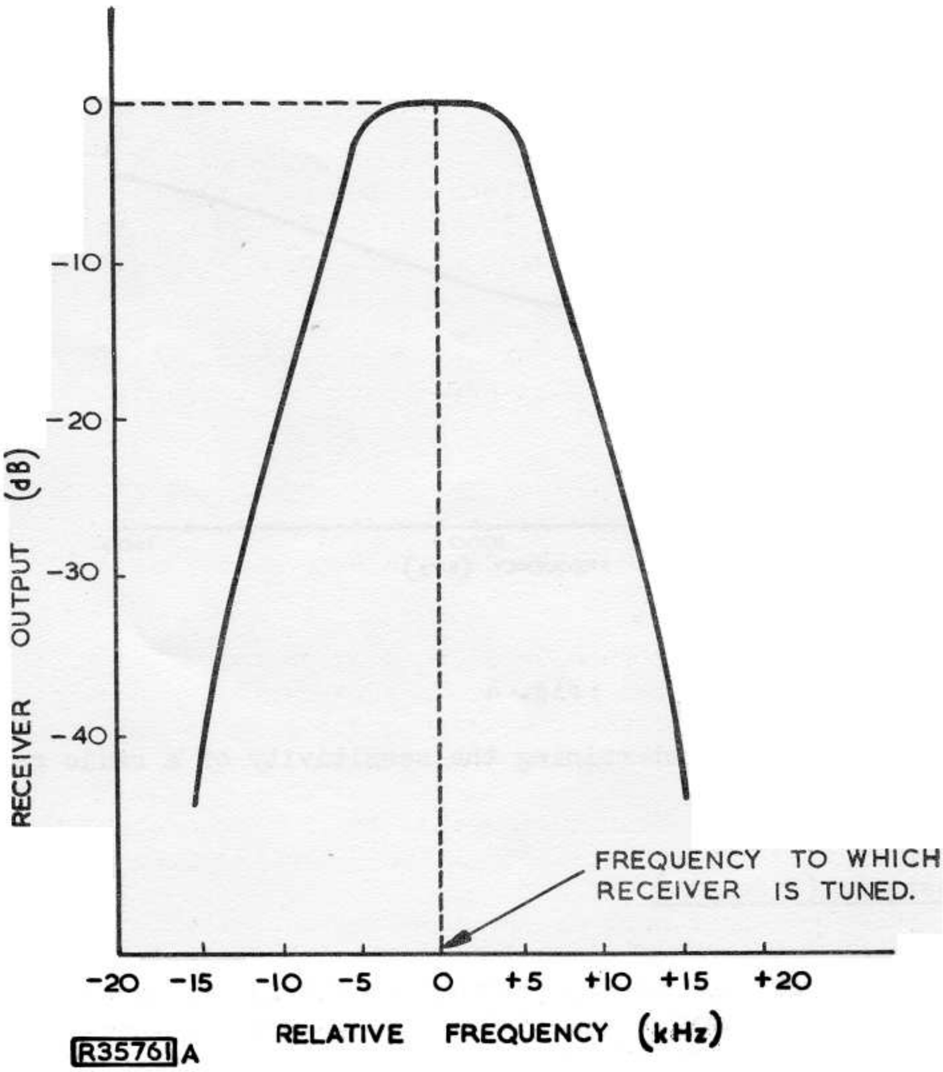


Fig. 3

It can be seen that the curve exhibits a reasonably flat top and fairly steep sides. The flat top means that the response of the receiver will be the same for the extreme side frequency components of the modulated input signal as for the carrier component. The steepness of the sides of the curve is advantageous since adjacent channel interference will be minimized.

The selectivity of a receiver is determined mainly by the characteristics of the band-pass circuits incorporated in the i.f. amplifier.

Sensitivity

The sensitivity of a radio receiver is a measure of the least r.f. input which will produce a specified a.f. output under given conditions. It is usually specified that the a.f. output should be 50 mW in a load of equivalent impedance to the loudspeaker, with a signal-to-noise ratio of 15 dB, and that the input signal should be 30% modulated at 400 Hz. The sensitivity is expressed as the input in microvolts applied under the prescribed conditions.

A typical sensitivity curve for a superheterodyne receiver is shown in Fig. 4.

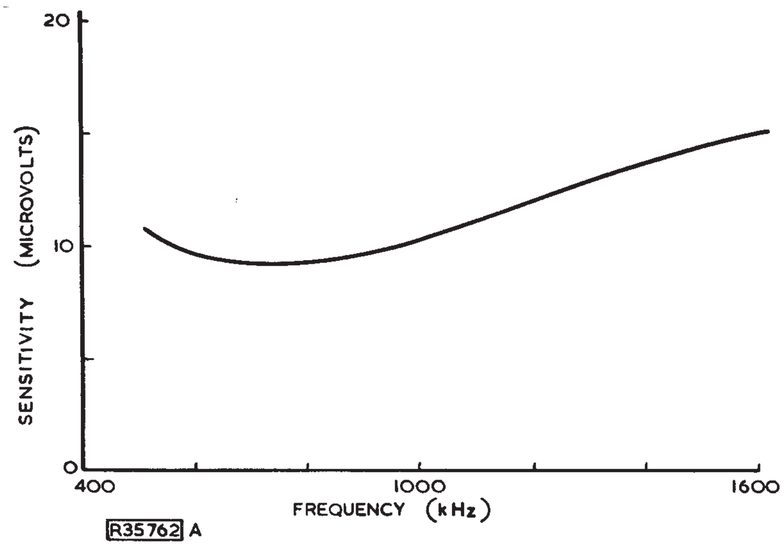


Fig. 4

The most important factor determining the sensitivity of a radio receiver is the gain of the i.f. amplifier.

Gain-Frequency Response (Fidelity)

The gain-frequency response of a receiver is the characteristic obtained by plotting the a.f. output of the receiver against modulating frequency at a fixed modulation depth, with the receiver terminated in a pure resistance. Or in other words, the fidelity of a receiver is the degree to which the a.f. output of the receiver resembles the modulation of the modulated signal received at the aerial. Fig. 5 shows the gain-frequency response of a typical receiver. It can be seen that the high frequency end of the characteristic falls rapidly at approximately 4.5 kHz, thus further reducing any vestige of the adjacent channel signal that has passed through the i.f. stages.

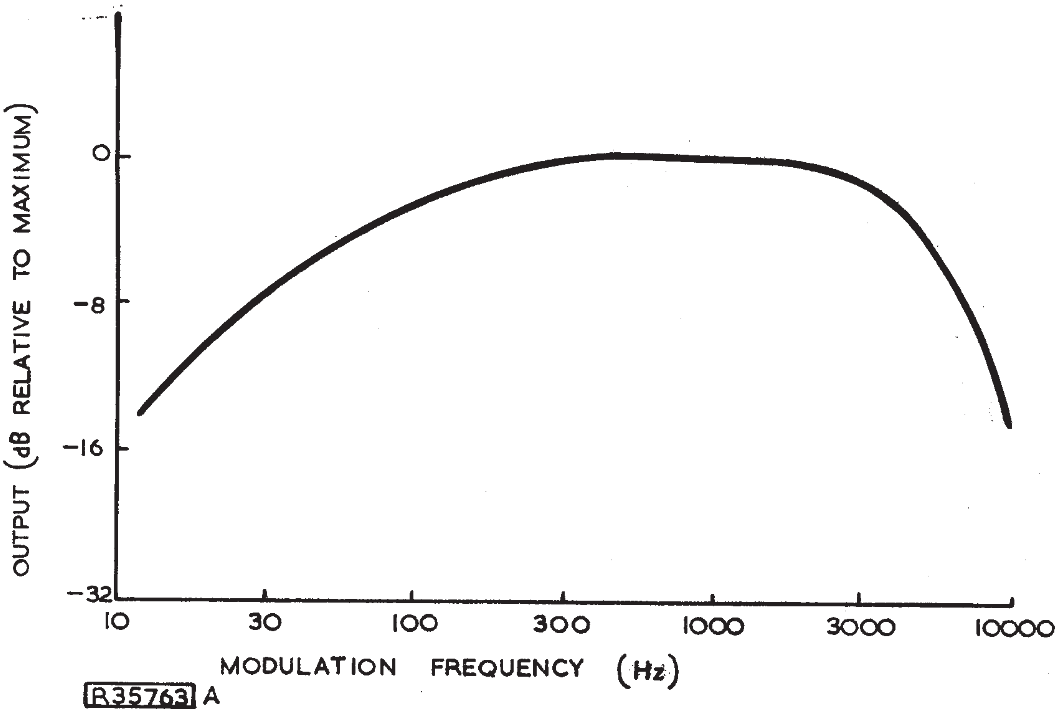


Fig. 5

The gain-frequency response of the receiver at the lower modulation frequencies is mainly determined by the l.f. characteristics of the a.f. amplifier. The response at the higher modulation frequencies is determined partly by the h.f. characteristics of the a.f. amplifier and partly by the attenuation suffered by the upper side frequency components in the i.f. amplifier.

Ganging

When two or more variable capacitors are arranged to be simultaneously rotated by a single control they are said to be 'ganged'.

Ganging of the variable tuned circuits of a superheterodyne receiver is employed in order to simplify tuning of the receiver. With one or more r.f. tuned circuits to be adjusted to a single frequency, and since the wanted signal frequency and the frequency of the local oscillator must differ by the intermediate frequency, the sensitivity and selectivity of a receiver depend upon the preciseness of the tuning.

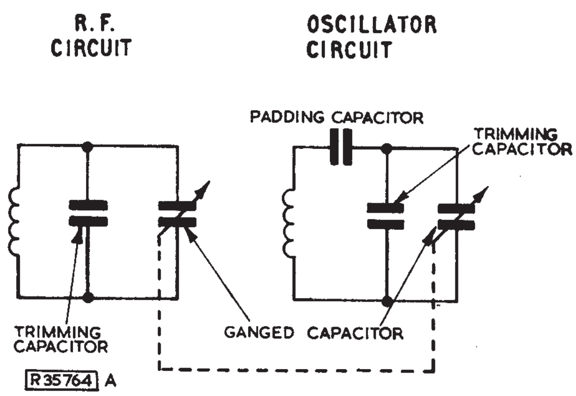


Fig. 6

Tuning of a receiver is easiest when the r.f. tuned circuits are identical and the tuning capacitors of the r.f. circuit and the local oscillator are ganged. It is desirable that the sections of the ganged capacitor are identical and any difference in the value of capacitance required to be obtained by placing small, pre-set, capacitors either in series with, or in parallel with, the ganged capacitors as shown in Fig. 6.

A capacitor placed in series with the ganged capacitor is known as a 'padding' capacitor and a capacitor placed in parallel as a 'trimming' capacitor.

Ganging will be dealt with in greater detail in a later section of this pamphlet.

THE RADIO-FREQUENCY CIRCUIT

r.f. circuit of a superheterodyne receiver provides

a means of coupling the aerial to the receiver, and

(b) some measure of selectivity.

The primary function of the r.f. circuit is to accept the wanted signal and reject the image frequency. This signal is separated from the wanted signal by twice the intermediate frequency, and since the usual intermediate frequency is 465 kHz, this separation is 930 kHz. This is an appreciable fraction of the frequency of the wanted signal when this signal is in either the long or medium wavebands. This makes adequate image rejection with a single r.f. tuned circuit readily obtainable. At the higher frequencies of short waves however, twice the intermediate frequency is only a few per cent off tune. Thus the short-wave performance of such a receiver is worse than the performance on the long and medium wavebands owing to the increased image interference. Another disadvantage of short-wave reception with such a receiver is that many of the short-wave signals are received at so low a level that they are 'swamped' by the inherent noise of the frequency changer.

To overcome these disadvantages some receivers employ at least one stage of r.f. amplification before the frequency changer.

Coupling the Aerial

The wanted signal must be transferred from the aerial to the signal grid of the first valve, and at the same time the image signal must be rejected.

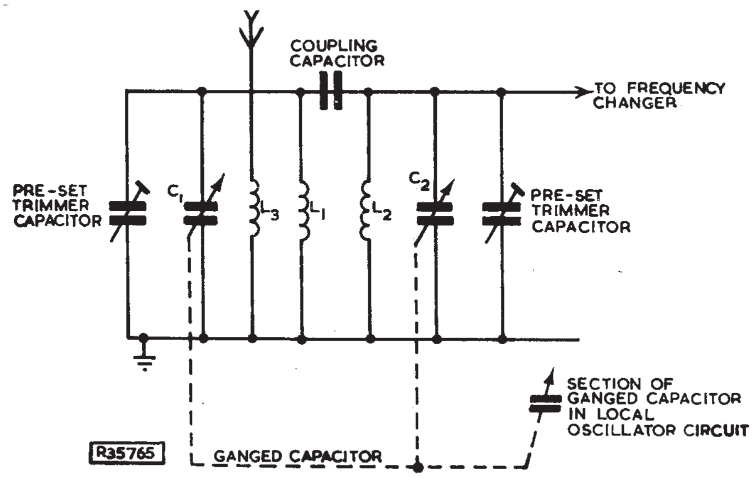


Fig. 7

The usual method of achieving these requirements is to employ some form of aerial coupling transformer which will apply the wanted signal to the first valve with a peak value greater than that of any other frequency which may be present in the aerial.

One form of aerial coupling circuit is shown in Fig. 7 which illustrates a typical r.f. circuit for use on the long and medium wavebands. The aerial is coupled via inductor L_3 to the primary coil, L_1, of a double-tuned r.f. transformer. The coefficient of coupling is arranged to be a little less than the optimum value, hence the transformer has a single peaked response which is fairly broad at the peak but which falls off sharply at frequencies several kHz off resonance. Both primary and secondary windings are tuned by similar sections of a ganged capacitor, a typical value of which is 400 pF. A third section of this capacitor tunes the local oscillator. Capacitors C_1 and C_2 are pre-set trimmer capacitors which enable minor adjustments to be made to the circuit.

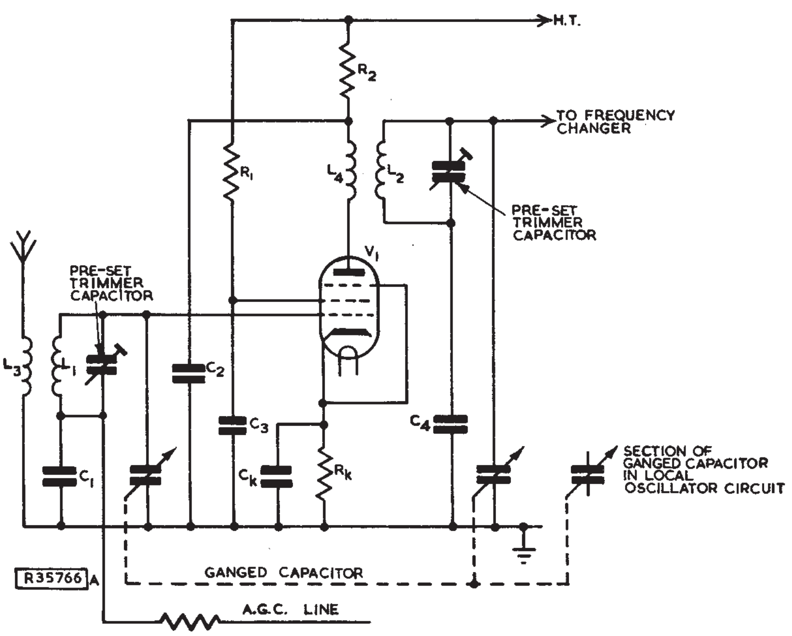


Fig. 8

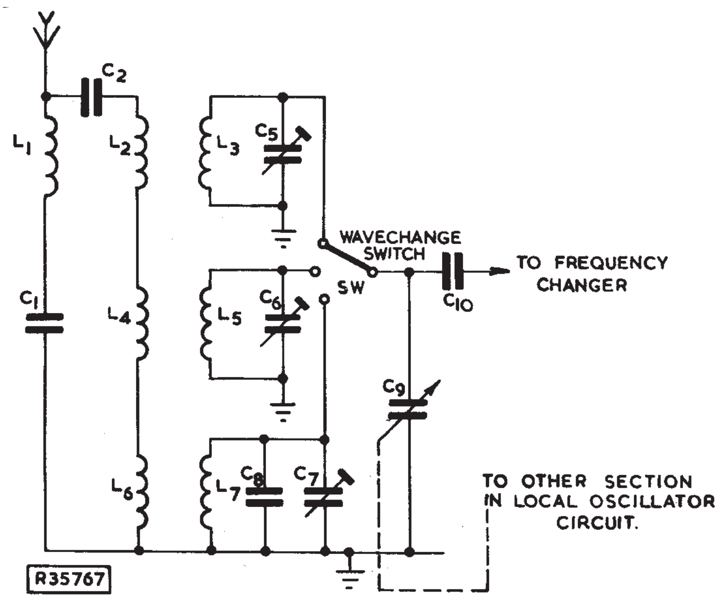


Fig. 9

The r.f. circuit of a superheterodyne receiver employing an r.f. amplifier is shown in Fig. 8. Such a circuit is often used in receivers designed for operation on all three wavebands.

Coils L_1 and L_2 are identical and are tuned by similar sections of the ganged capacitor. The trimmer capacitors provide a means of making small adjustments to the tuned circuits and ensure that the two circuits always tune to the same frequency. The third section of the ganged capacitor tunes the local oscillator, the coil inductance of which cannot be the same as L_1 and L_2 since the oscillator must tune to a higher frequency.

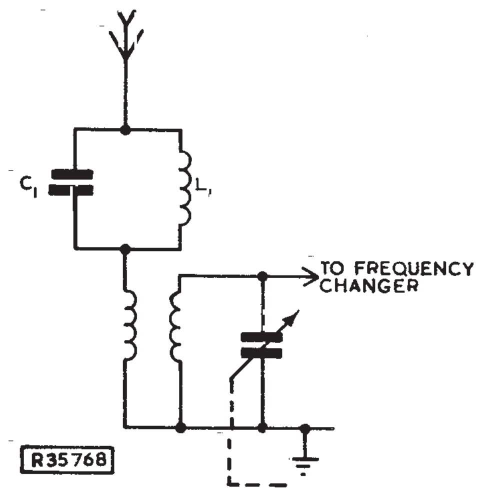
The amplifier is a tuned-secondary r.f. amplifier whose main function is to increase the signal-to-noise ratio of the receiver.

Fig. 9 shows an r.f. circuit employing an i.f. trap (often known as a wave trap) and separate tuned circuits for the short, medium and long wavebands.

Signals from the aerial are fed via capacitor C_2 to the short, medium and long wave series input transformers, each secondary of which is shunted by a pre-set trimmer capacitor. The required circuit is selected by switch SW and connected across the aerial tuning capacitor C_9. Inductor L_1 and capacitor C_1 form an effective i.f. trap against unwanted signals close to the intermediate frequency of the receiver. The wanted signals are then fed via the coupling capacitor C_{10} to the frequency changer.

Intermediate-Frequency Traps

As mentioned earlier in this pamphlet superheterodyne receivers are prone to interference from stations working at or near the intermediate frequency of the receiver. In order to prevent such signals reaching the frequency-changer, i.f. traps are often fitted.



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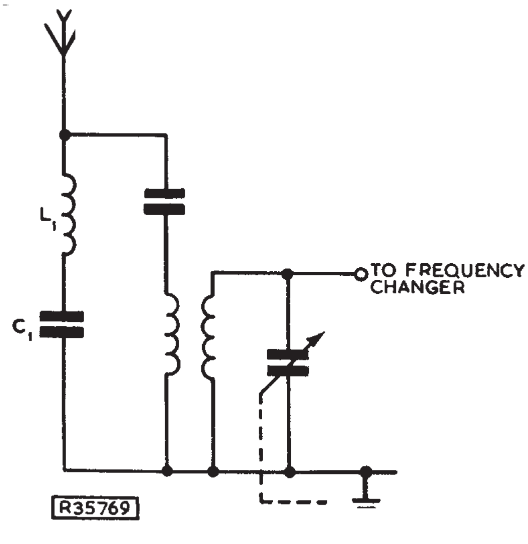
One form of wave trap in which a parallel-tuned circuit, L_1, C_1, is connected in the aerial lead is shown in Fig. 10. The tuned circuit is tuned to the intermediate frequency and forms a rejector circuit. If the losses of the coil and capacitor used are low and the ratio $\frac{L_1}{C_1}$ is large, the tuned circuit will offer a very high impedance to all frequencies at, or near, its resonant frequency, and thus reject any signals at the intermediate frequency.

An alternative method of preventing i.f. interference is shown in Fig. 11, which shows a series-tuned acceptor circuit, L_1 C_1. The acceptor circuit is designed to have a low impedance at the intermediate frequency and a high impedance at all other frequencies. Thus signals at the intermediate frequency pass straight to earth whilst all other signals are unaffected.



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GANGING AND TRACKING

Superheterodyne receivers would be extremely difficult to tune to a particular station if all the variable capacitors were rotated by separate tuning controls. To simplify the tuning of such a receiver it is usual to arrange that all the variable capacitors in the r.f. and local oscillator circuits are rotated simultaneously by a single control. This can be achieved by using a 'ganged' capacitor in which all the sections are arranged to have very nearly equal capacitance at all settings. Ganging simply means connecting two or more variable capacitors to a single shaft. Thus ganged circuits are a number of tuned circuits linked mechanically so that their resonant frequencies can be adjusted by a single control.

Since, however, the frequency difference between the wanted signal and the local oscillator must be kept constant, the oscillator must be tuned in a different manner from the r.f. tuned circuits. It is a relatively simple matter to make the frequency difference equal to the intermediate frequency at two points in the tuning range.

Consider a receiver that is required to tune over the frequency range 500 kHz to 1500 kHz and assume that the identical sections of the ganged capacitor have a capacitance range of 400 pF. The required frequency ratio is $\frac{1500}{500}$ or 3:1, hence a capacitance change of 9:1 is needed which can be achieved with a total minimum capacitance of 50 pF. The capacitance of the sections of the ganged capacitor will therefore vary from 50 pF to 450 pF. The inductance required in each r.f. tuned circuit in order to tune the receiver over the frequency range required is 225 μH.

If the receiver has an intermediate frequency of 465 kHz the local oscillator must tune from (500 + 465) kHz to (1500 + 465) kHz or 965 kHz to 1965 kHz. Thus the ratio of highest frequency to lowest frequency is $\frac{1965}{965}$ or 2.036:1, which requires a capacitance ratio of 4.14:1. Such a range can still be obtained if the minimum capacitance is increased to approximately 128 pF, giving a capacitance range of 128 pF to 528 pF.

Unfortunately however, merely increasing the minimum tuning capacitance and using the appropriate inductance will not maintain the frequency of the local oscillator 465 kHz above that to which the r.f. circuits are tuned. This is shown by Table 2.

TABLE 2

R.F. TUNED CIRCUITS (C_{min} = 50 pF : L = 225 μH)		LOCAL OSCILLATOR (C_{min} = 128 pF : L = 51.23 μH)		OSCILLATOR FREQUENCY MINUS INTERMEDIATE FREQUENCY (kHz)	ERROR IN TUNING OF R.F. CIRCUIT (kHz)
TUNING CAPACITANCE (pF)	FREQUENCY TUNED TO (kHz)	TUNING CAPACITANCE (pF)	FREQUENCY TUNED TO (kHz)		
50	1500	128	1965	1500	0
100	1060	178	1666	1201	-141
200	750	278	1334	869	-119
300	612	378	1144	679	-67
400	530	478	1017	552	-22
450	500	528	968	503	-3

Table 2 is based on the assumption that the total tuning capacitance in the local oscillator circuit is always 78 pF greater than the tuning capacitance of the r.f. circuits. If this difference were slightly adjusted the error at the low frequency end could also be made zero.

Thus the effect of increasing the minimum value of the oscillator tuning capacitance is to give correct tracking at both ends of the tuning range only. Tracking is an arrangement by which the resonant frequency of one of a number of ganged circuits is maintained at a constant difference from that of the other circuits.

An alternative method of reducing the ratio of maximum to minimum capacitance is to add capacitance in series with the tuning capacitor.

Consider that the r.f. tuned circuit capacitance varies from 50 pF to 450 pF as before. If a padding capacitor, C_p, is placed in series with the tuning capacitor of the local oscillator the total capacitance will vary from:-

$$\frac{50\ C_p}{50 + C_p} \text{ to } \frac{450\ C_p}{450 + C_p}$$

The capacitance range required is 4.14:1, thus

$$\frac{450\ C_p}{450 + C_p} = 4.14 \left(\frac{50\ C_p}{50 + C_p}\right)$$

$$\frac{450\ C_p}{450 + C_p} = \frac{207\ C_p}{50 + C_p}$$

$$243\ C_p^2 = 70650\ C_p$$

$$C_p = 291 \text{ pF}$$

The effective oscillator tuning capacitance then varies from:-

$$\frac{291 \times 50}{291 + 50} \text{ pF to } \frac{291 \times 450}{291 + 450} \text{ pF}$$

or 42.7 pF to 176.5 pF

The value of inductance required to tune the local oscillator may be found by use of the expression $f = \frac{1}{2\pi\sqrt{LC}}$ Hz to give $L = 153.6\ \mu H$.

Table 3 shows the tuning error between the oscillator and r.f. tuning circuits for several settings of the ganged capacitor and illustrates that correct tracking is again obtained only at the two ends of the tuning range.

TABLE 3

R.F. TUNED CIRCUITS (C_{min} = 50 pF : L = 225 μH)		LOCAL OSCILLATOR CIRCUIT (C_{min} = 42.7 pF : L = 153.6 μH)		OSCILLATOR FREQUENCY MINUS INTERMEDIATE FREQUENCY (kHz)	ERROR IN TUNING R.F. CIRCUITS (kHz)
TUNING CAPACITANCE (pF)	FREQUENCY TUNED TO (kHz)	TUNING CAPACITANCE (pF)	FREQUENCY TUNED TO (kHz)		
50	1500	42.7	1965	1965 - 465 = 1500	0
100	1060	74.4	1489	1489 - 465 = 1024	+36
200	750	118.5	1180	1180 - 465 = 715	+35
300	612	148	1055	1055 - 465 = 590	+22
400	530	168.5	989	989 - 465 = 524	+6
450	500	176.5	965	965 - 465 = 500	0

In both Table 2 and Table 3, the tracking error has been expressed as an error in the tuning of the r.f. tuned circuits.

Fig. 12 shows the frequencies at which the r.f. tuned circuits and the local oscillator are in tune, plotted against the values of capacitance of the ganged capacitor.

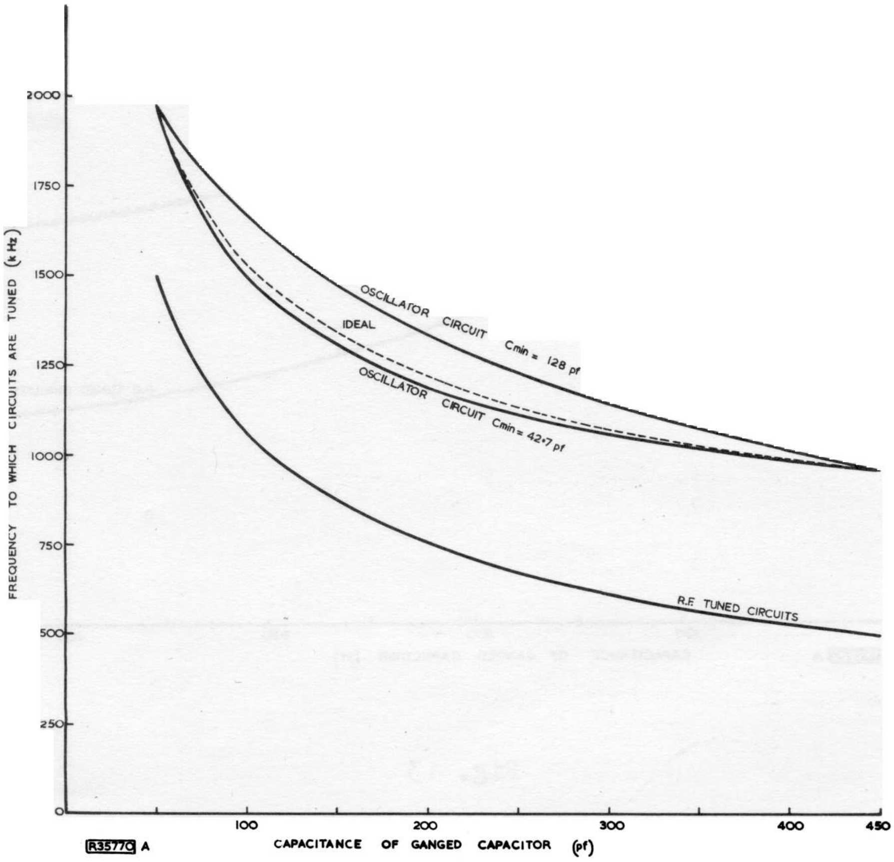


Fig. 12

It can be seen that at each end of the tuning range the local oscillator is tuned to almost the same frequency for either value of C_{min} and the difference between the frequency of the local oscillator and the frequency to which the r.f. tuned circuits are tuned is nearly 465 kHz (the intermediate frequency).

At any other setting of the ganged capacitor the frequency difference is not 465 kHz. The difference is greater than 465 kHz when C_{min} is 128 pF and less than 465 kHz when C_{min} is 42.7 pF.

It is extremely difficult with equal section ganged capacitors to arrange that the r.f. tuned circuits and the local oscillator are always tuned to frequencies differing by an amount equal to the intermediate frequency.

However, by a combination of the two methods considered, i.e. the simultaneous use of padding and trimming capacitors, 'three point tracking' may be obtained and the tracking errors made very small. It is possible to select the component values such that the intermediate frequency is correct at three points in the tuning range instead of two. These points are marked A, B and C in Fig. 13. It can be seen that the frequency error over the remainder of the tuning range is appreciably reduced.

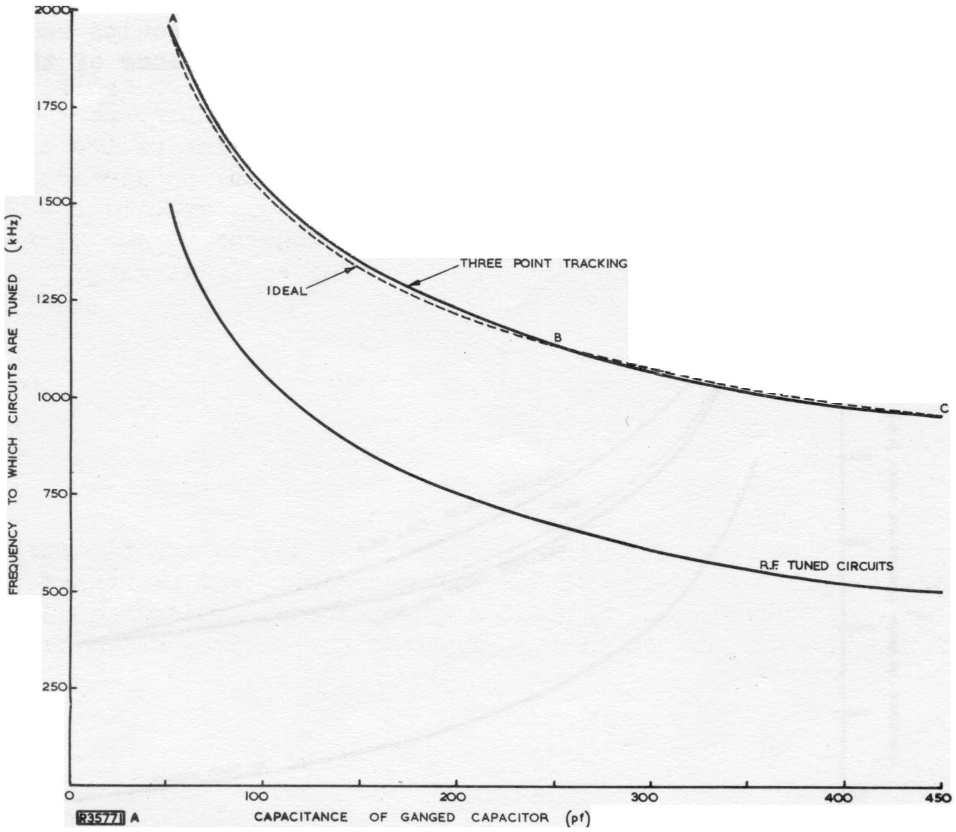


Fig. 13

The fact that small tracking errors result even from three-point tracking does not mean that an incorrect intermediate frequency will be produced. The user, in tuning the receiver by ear, will set the tuning control so that the wanted signal is correctly centred in the pass-band of the i.f. amplifier and the error will be in the tuning of the r.f. circuit. An error of a few kHz in tuning the r.f. circuit will not cause any noticeable change of amplification or selectivity. Or in other words, the response of the i.f. amplifier is so much sharper than that of the r.f. circuit, that the effect of the r.f. circuit is negligible for frequencies near the wanted signal.

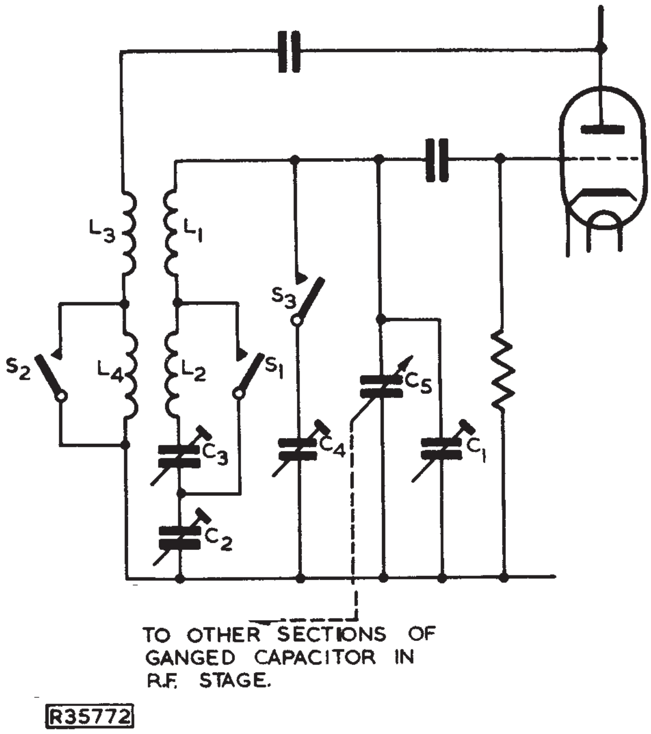


Fig. 14

Fig. 14 shows a part of the circuit of an oscillator for use on the medium and long wavebands. The circuit shows a typical arrangement of padding and trimming capacitors to obtain three point tracking. With correct values for these capacitors, practically perfect tracking can be achieved.

Capacitor C_5 is a section of the ganged capacitor. For reception of medium-wave signals switches S_1 and S_2 are closed and switch S_3 open. Capacitor C_1 then increases the minimum capacitance of the circuit and capacitor C_2 decreases the maximum capacitance. The actual values of C_1 and C_2 are critical for accurate ganging.

For long-wave reception, switches S_1 and S_2 are opened and S_3 closed. Opening switch S_1 increases the inductance of the tuned circuit, and decreases the capacitance owing to capacitors C_2 and C_3 now being in series. Switch S_2 increases the inductance of the circuit and switch S_3 increases the minimum capacitance of the circuit.

FREQUENCY CHANGING

Frequency changing is a process by which two voltages at different frequencies may be combined to give voltages at the sum and difference of the two frequencies.

There are two main methods of combining the two voltages, additive mixing, in which the two voltages are added together and the resultant voltage rectified: and multiplicative mixing, in which the amplitude of one signal is varied at the frequency of the other.

The majority of valve h.f. superheterodyne radio receivers employ multiplicative mixing since additive mixing suffers from the disadvantage that interaction between local oscillator and input signal circuits is difficult to avoid. Multiplicative mixing is the only form of mixing considered in this pamphlet.

In a multiplicative mixer the two voltages to be combined are applied to different grids of a multi-electrode valve. The variation of the potential on one grid causes the mutual conductance of the other grid to anode to vary, the resulting anode current containing components at the sum and difference frequencies.

The anode circuit of the valve is tuned to select the desired frequency, which is usually the difference frequency.

Frequency Changing by the Use of Two Control Grids

Consider a pentode valve in which the anode, screen grid and signal grid voltages with respect to the cathode are all constant, and in which the voltage applied between the suppressor grid and the cathode increases from zero to a high negative value.

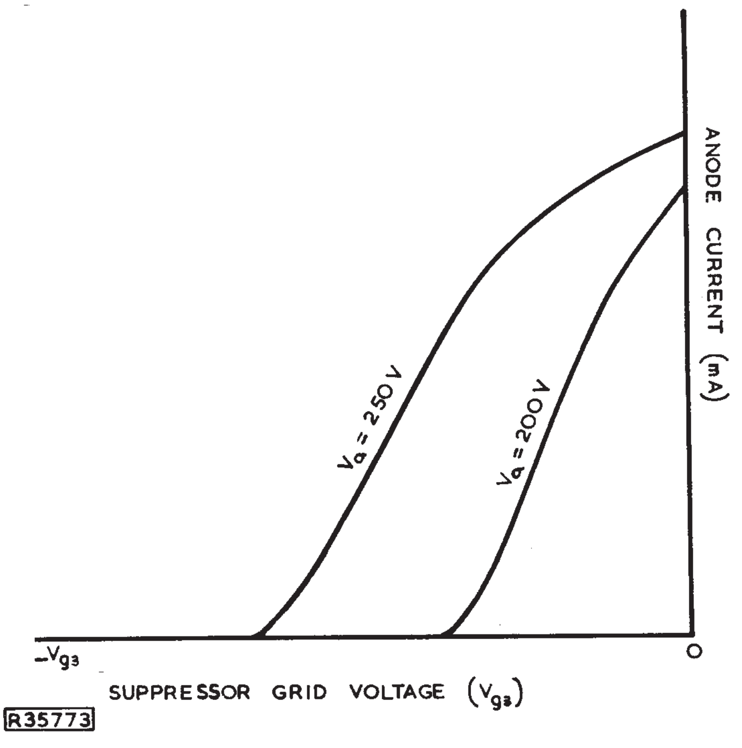


Fig. 15

As the suppressor grid voltage, V_{g3}, becomes more negative the anode current of the valve falls until at a particular value of V_{g3} it is negligible. Fig. 15 shows curves of anode current plotted against suppressor grid voltage for a typical pentode valve. The cathode current of the valve remains unaffected by the variation of V_{g3}, any decrease in anode current being made up by a corresponding increase in screen grid current.

Any decrease in anode current will obviously reduce the slope of the I_a/V_{g1} characteristics of the valve. Thus the mutual conductance, gm_1, of the valve between signal grid and anode is dependent upon the value of V_{g3} ($gm = \frac{\delta I_a}{\delta V_g}$, V_a constant).

Fig. 16 shows values of the mutual conductance, gm_1, plotted against values of suppressor grid voltage, V_{g3}.

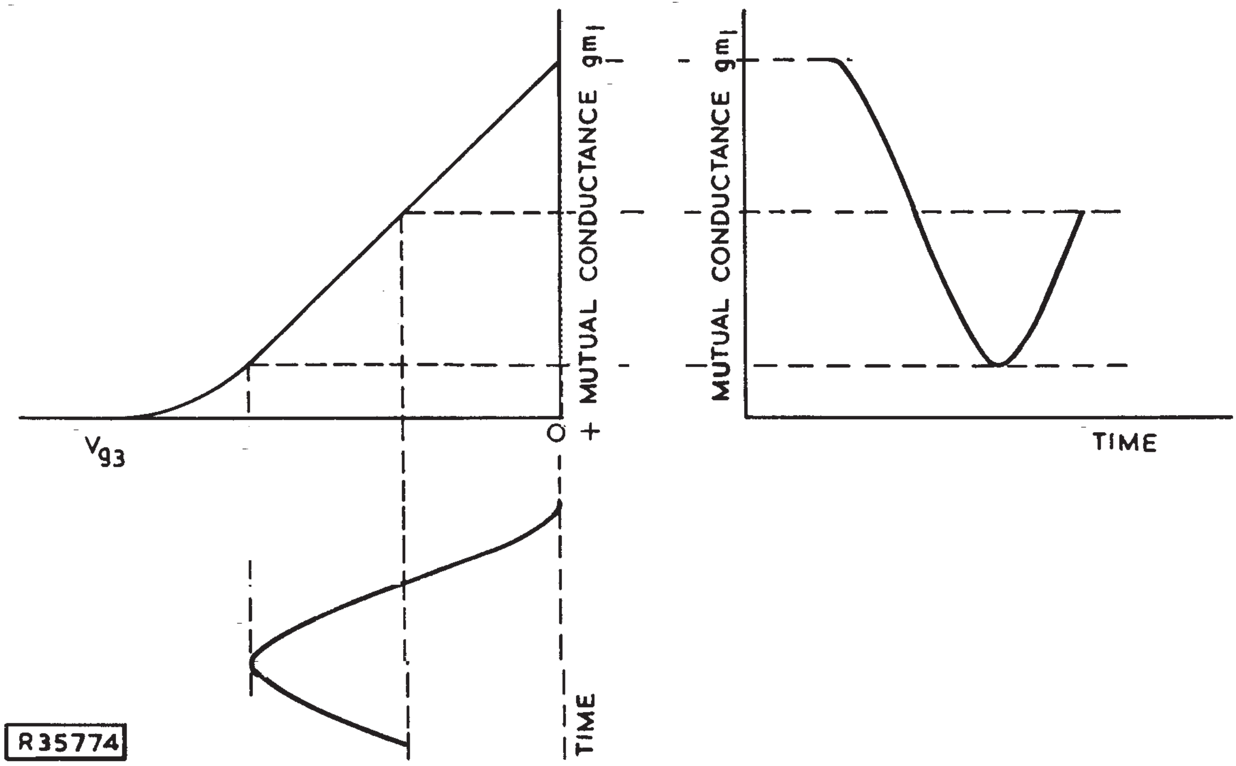


Fig. 16

If the slope of the gm_1/V_{g3} characteristic is linear then a sinusoidal variation in the peak value of V_{g3} will result in the mutual conductance of the valve varying sinusoidally.

If an alternating voltage is applied to the suppressor grid which is biased negatively with respect to the cathode, the magnitude of the output waveform of the valve will vary at the frequency of this alternating voltage, owing to the changes in mutual conductance. In the next section of the pamphlet this waveform is shown to contain the difference frequency, i.e. a frequency corresponding to the difference between the frequencies applied to the control and suppressor grids.

Mathematical treatment of frequency-changing

Let the anode current/suppressor grid characteristic of a pentode be represented by the expression:-

$$\frac{I_a}{I_K} = a + b\,V_{g3} + c\,V_{g3}^2 + \ldots\ldots\ldots\ldots \quad (1$$

where a, b, c, etc. are constants, since only a certain proportion of the cathode current, I_K, reaches the anode, this proportion depending upon the value of V_{g3}.

Let the cathode current/signal grid characteristic of a pentode be represented by the expression:-

$$I_K = a' + b' V_{g1} + c' V_{g1}^2 + \ldots\ldots \qquad (2)$$

where a', b', c', etc. are constants.

Substitution of the expression for I_K in equation (1) gives

$$I = (a + b V_{g3} + c V_{g3}^2 + \ldots\ldots)(a' + b' V_{g1} + c' V_{g1}^2 + \ldots\ldots)$$

If it may be assumed that the valve is operated over the linear part of its characteristic then very little error is introduced by writing

$$I_a = (a + b V_{g3})(a' + b' V_{g1})$$

Let the voltage applied between suppressor grid and cathode, V_{g3}, be

$$V_{gb3} - A_o \sin \omega_o t$$

where V_{gb3} = bias voltage

and
$A_o \sin \omega_o t$ = output voltage of local oscillator.

Let the voltage applied between signal grid and cathode, V_{g1}, be

$$V_{gb1} - A_s \sin \omega_s t$$

where V_{gb1} = bias voltage

and
$A_s \sin \omega_s t$ = signal voltage.

Then $I_a = \left[a + b(V_{gb3} - A_o \sin \omega_o t)\right]\left[a' + b'(V_{gb1} - A_s \sin \omega_s t)\right]$
and contains the product,

$$bb'(V_{gb3} - A_o \sin \omega_o t)(V_{gb1} - A_s \sin \omega_s t)$$

Expanding gives:-

$$bb'(V_{gb3} V_{gb1} - V_{gb3} A_s \sin \omega_s t - V_{gb1} A_o \sin \omega_o t + A_o A_s \sin \omega_o t \sin \omega_s t).$$

Now a trigonometric identity is 2 sin A sin B = cos (A - B) - cos (A + B) and thus the last term $bb' A_o A_s \sin \omega_o t \sin \omega_s t$ may be re-written

$$\tfrac{1}{2} bb' A_o A_s \left[\cos(\omega_o - \omega_s)t - \cos(\omega_o + \omega_s)t\right]$$

showing that the output of the frequency changer contains the sum and difference frequencies, i.e. $f_o + f_s$ and $f_o - f_s$.

Conversion Conductance

In order to measure the efficiency of a frequency-changer a term known as the 'conversion conductance' is used. The conversion conductance of a frequency-changer is the ratio

$$\frac{\text{peak value of intermediate-frequency component of anode current}}{\text{peak value of input signal voltage}}$$

expressed in milliamps per volt.

Consider the expression for the intermediate frequency component of the anode current $\frac{1}{2}\, bb' \, A_oA_s \cos 2\pi\, (f_o - f_s)t$. The peak value of this component is $\frac{1}{2}\, bb' \, A_oA_s$ and thus the conversion conductance is

$$\frac{bb' \, A_oA_s}{2\, A_s} = \frac{1}{2}\, bb' \, A_o$$

This expression shows that the conversion conductance of a frequency changer is proportional to the voltage of the local oscillations applied to the signal grid. It might seem desirable therefore, to make the oscillator voltage as large as possible but unfortunately there is a practical limit to the oscillatory voltage that can be applied to the suppressor grid. If the peak value of the oscillator voltage exceeds half the cut-off voltage of the suppressor grid, the valve will merely cut-off for part of each cycle of the oscillator output.

Any further increase in A_o is then ineffective in increasing the value of the conversion conductance. Fig. 17 shows how the conversion conductance of a typical frequency-changer varies with the local oscillator voltage, A_o. It can be seen that while the conversion conductance increases almost linearly with A_o for small values of A_o, it falls gradually as A_o exceeds a certain value.

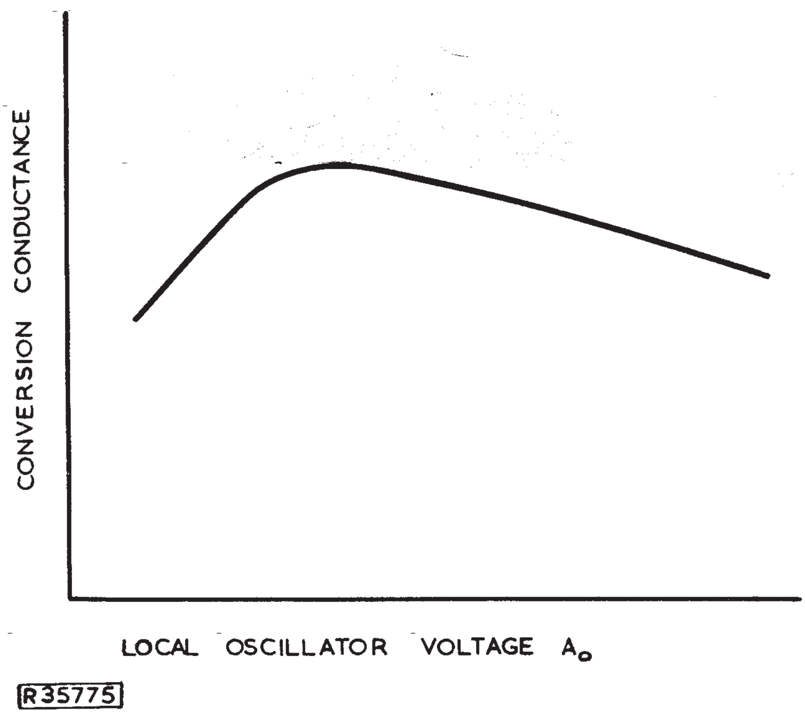


Fig. 17

Conversion impedance

The conversion impedance of a frequency-changing valve is analogous to the anode a.c. resistance of a triode or pentode valve and is given by the ratio,

$$\frac{\text{change of anode voltage at the intermediate frequency}}{\text{change of anode current at the signal frequency}}$$

Frequency-Changing Valves

As has been previously stated in this pamphlet, the frequency-changing and local oscillator valves are often combined within a single envelope. This is advantageous in that space is saved and less wiring is required.

It will have been appreciated from the foregoing description of the frequency-changing process that a valve having two control grids is required.

The pentode valve, used to illustrate the frequency-changing process, is unsuitable owing to the interaction that occurs between the suppressor grid and the anode.

The hexode valve

A hexode valve is one which has six electrodes arranged as shown in Fig. 18. The valve has two control grids, G_1 and G_3, G_3 being screened from both G_1 and the anode by screen grids G_2 and G_4. The screen grids are connected together and to positive potential with respect to the cathode.

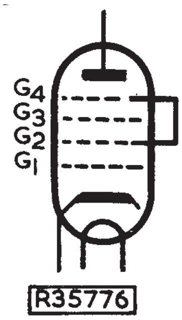


Fig. 18

The triode-hexode valve

The hexode valve is rarely encountered nowadays by itself but is usually found combined within a single envelope with a triode valve. Such a valve is known as a 'triode-hexode' and is the most frequently used of all frequency-changing valves.

The arrangement of the electrodes may take one of two forms as shown in Fig. 19 and 20, the more common arrangement being that of Fig. 20. The triode section of the valve is a part of the local oscillator circuit, the oscillatory voltage appearing at its grid being applied directly to either G_3 (Fig. 19) or G_1 (Fig. 20). The signal voltage is applied to G_1 (Fig. 19) or G_3 (Fig. 20) and electrodes G_2 and G_4 act as screen grids.

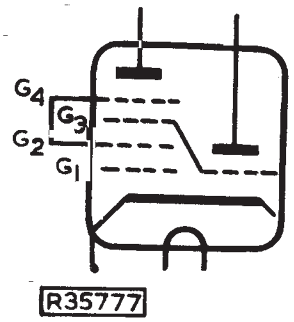


Fig. 19

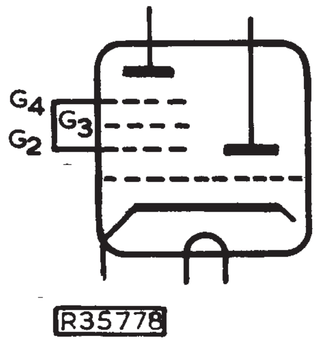


Fig. 20

The electrode arrangement of Fig. 20 has the advantage of simpler construction, but is not as efficient as that shown in Fig. 19 since the grid nearest the cathode, G_1, exerts the greatest control on the anode current and should therefore be controlled by the smaller of the two input voltages.

Fig. 21 shows the construction of the simpler type of triode-hexode valve (Fig. 20). The valve has an oval cathode which is surrounded by the triode grid and the first grid of the hexode section. It can be seen that these are physically one. On one side of the cathode is placed the triode anode and on the other side are the signal and screening grids and the anode of the hexode section. The two shields are made of metal and are included to improve the screening of the valve.

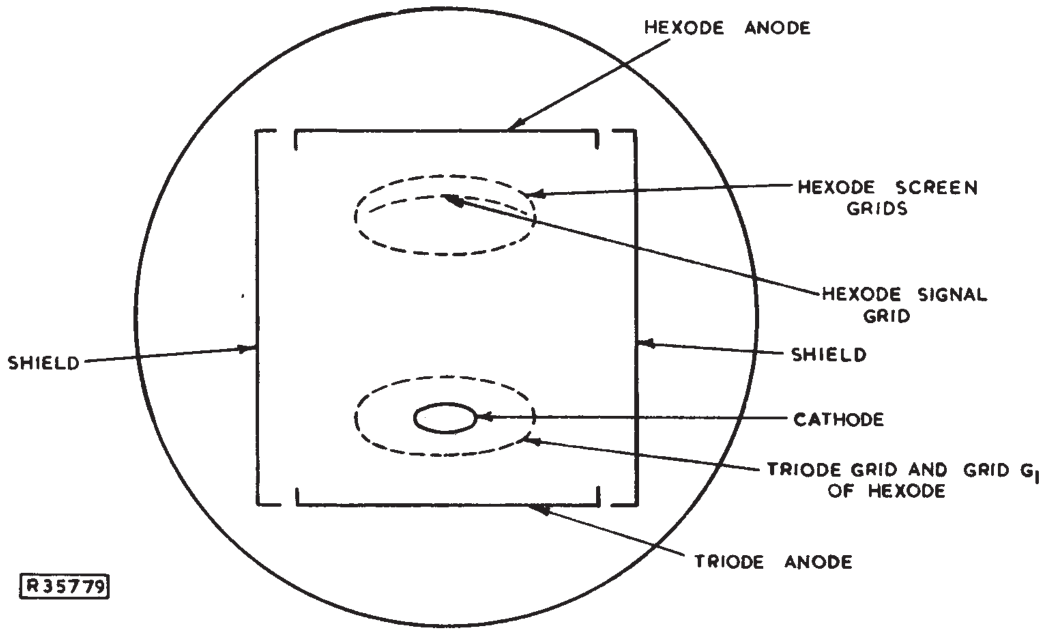


Fig. 21

The main advantages of the triode-hexode valve are:-

(a) Very effective gain control may be obtained since the hexode portion has a variable μ characteristic.

The valve can be satisfactorily operated at very high frequencies.

Interaction between signal and local oscillator circuits is negligible.

Radiation of the local oscillator frequency from the aerial is negligible.

The heptode valve

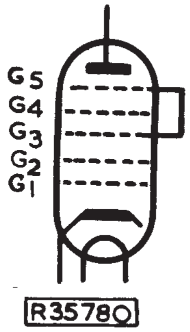


Fig. 22

A heptode valve is one with seven electrodes, the arrangement of which is shown in Fig. 22. Such a valve can be used simultaneously as both the local oscillator valve and the mixer valve. The cathode and grids G_1 and G_2 form a triode valve for use in the local oscillator circuit, grid G_2 acting as the anode. Thus the potential of G_2 varies at the frequency of the local oscillator. Grids G_3 and G_5 are connected together and screen signal G_4 from both anode and grid G_2.

The heptode valve has the following advantages

(a) Radiation of the local oscillator frequency from the aerial is negligible.

(b) Interaction between signal frequency and local oscillator circuits is negligible.

(c) Very high conversion conductance.

Frequency-Changing Circuits

Triode-hexode frequency-changers

Fig. 23 shows the circuit of a typical triode-hexode frequency-changing stage.

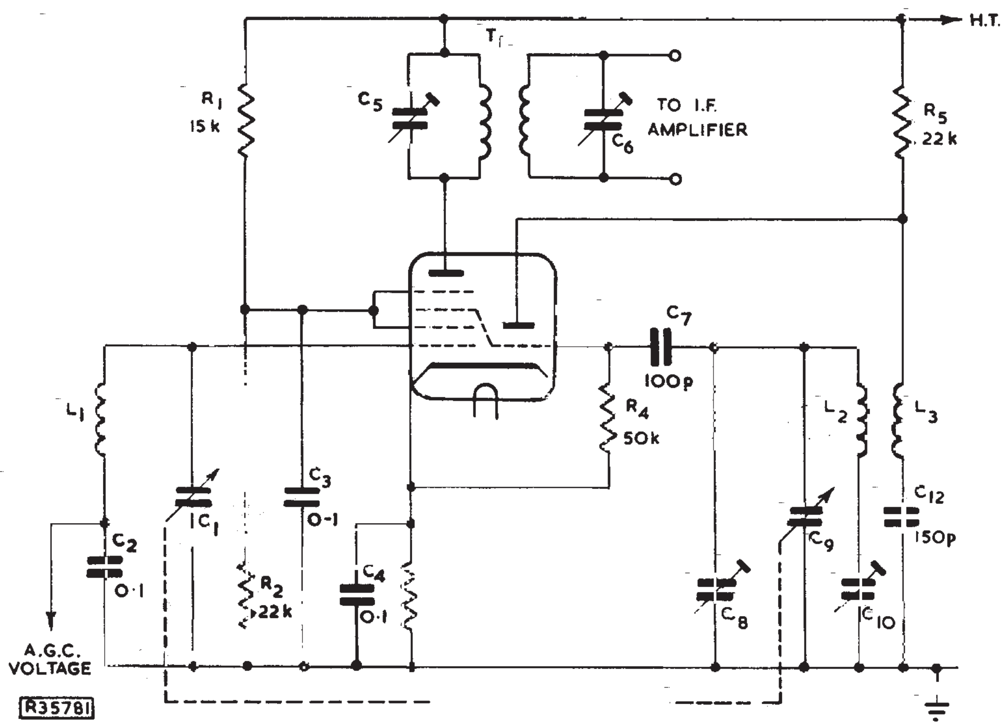


Fig. 23

The input circuit, L_1C_1, applies the incoming signals to the control grid of the hexode section of the valve. The triode section of the valve is part of a tuned grid oscillator, coupling between anode and grid circuits being obtained by means of coils L_2 and L_3. The frequency of oscillation is determined by the values of coil L_2 and ganged capacitor C_9. Grid bias for the triode section of the valve is obtained from the grid leak circuit formed by resistor R_4 and capacitor C_7, and grid bias for the hexode section is obtained from resistor R_3.

The intermediate frequency component of the anode current is selected by the i.f. transformer, T_1, the windings of which are tuned by capacitors C_5 and C_6. The screen grid voltage for the hexode section is obtained via a potential divider formed by resistors R_1 and R_2, enabling a constant screen grid voltage to be applied.

The circuit of a triode-hexode frequency-changer in which a tuned-anode oscillator is employed is shown in Fig. 24.

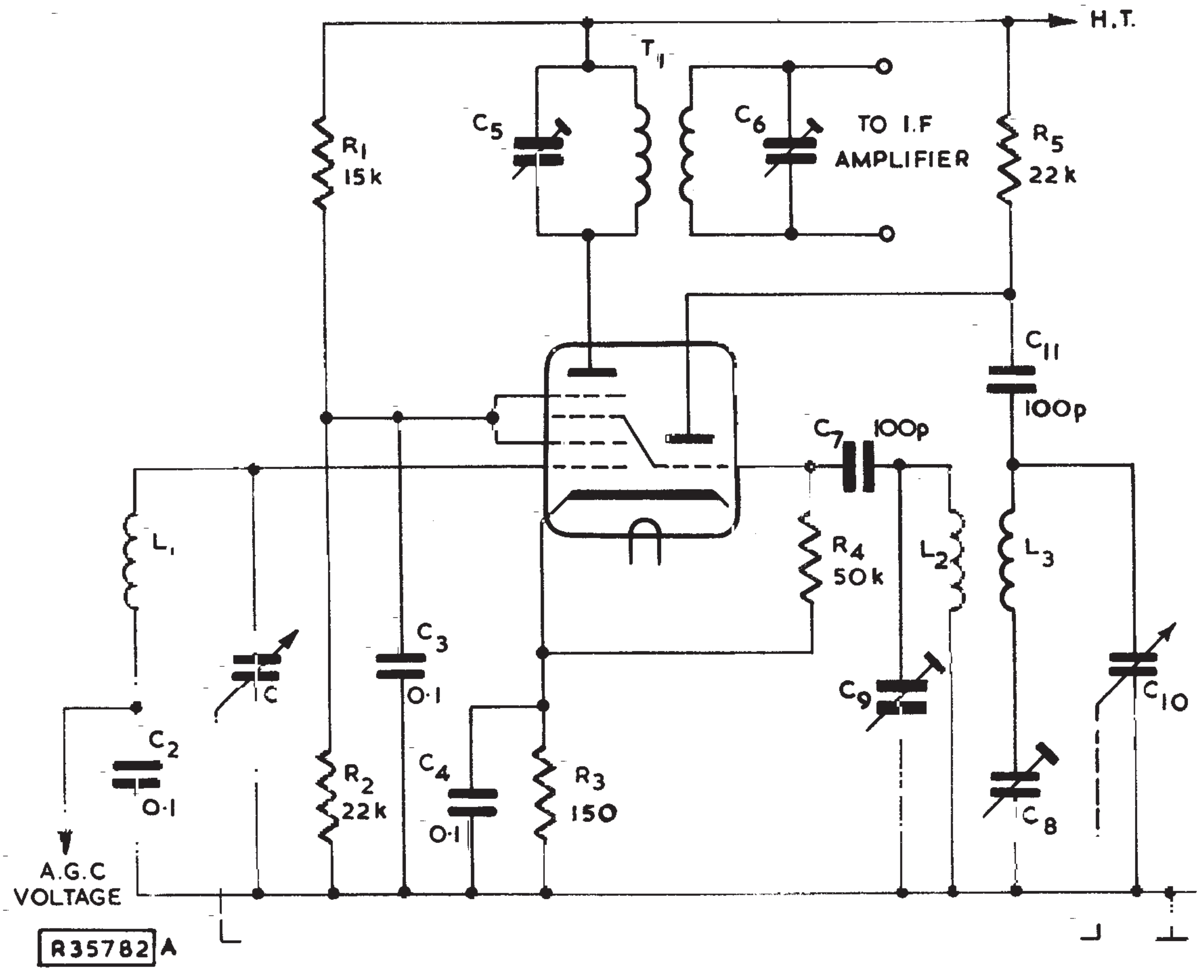


Fig. 24

The use of a tuned-anode oscillator has the following advantages compared with the tuned-grid circuit:-

(a) since the triode anode is electrically further removed from the hexode signal grid than the triode grid, any tendency for interaction to occur is reduced.

(b) the frequency stability of the oscillator is better.

Heptode frequency-changers

The circuit of a frequency-changer employing a heptode valve is shown in Fig. 25.

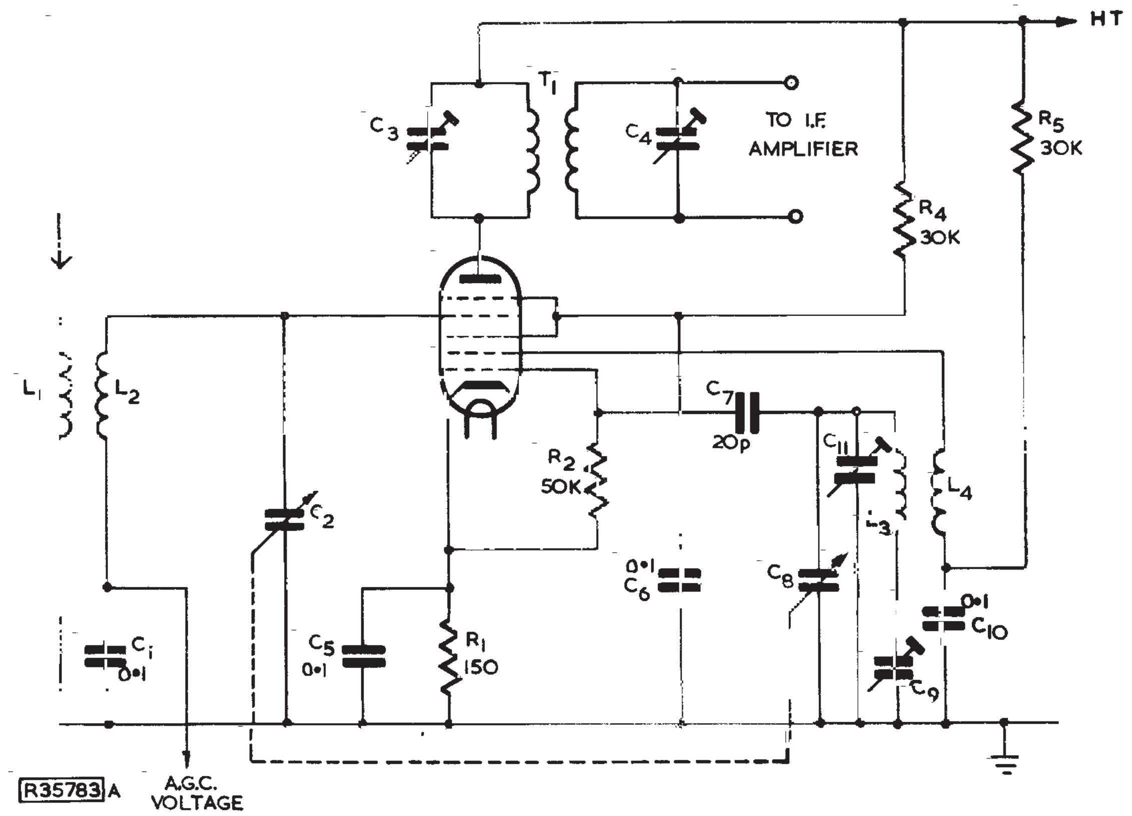


Fig. 25

The cathode and first two grids of the valve form a part of a triode tuned-grid oscillator, and the second grid acting as the anode of the triode.

This means that the total emission of the valve is controlled by the oscillatory voltage present on the first grid. The input signal voltage is tuned by L_2, C_2, and applied to the fourth grid causing further variation of the electron stream and thus producing the sum and difference frequencies.

The third and fifth grids function as screen grids and screen the fourth grid from the third grid and the anode. A further grid is sometimes inserted between the fifth grid and the anode to act as a suppressor grid. A valve with such an extra grid is known as an Octode and has a higher anode slope resistance than a heptode and for this reason is sometimes preferred.

Relative advantages of triode-hexode and heptode frequency-changers

Both triode-hexode and heptode frequency-changers are to be met with in modern superheterodyne receivers. The triode-hexode is the more common and has the following advantages compared to the heptode:-

(a) greater conversion conductance and thus its conversion efficiency is greater.

(b) it can be shown that the noise generated within a frequency-changing valve is inversely proportional to the square of the conversion conductance. Thus, from (a), the triode-hexode introduces less noise than does the heptode. This means that fewer r.f. circuits before the frequency-changer are required for a given signal-to-noise ratio for weak signals.

(c) triode-hexodes can be employed at much higher frequencies than can the heptode.

The heptode however, has the advantage of possessing a higher conversion impedance than the triode-hexode. This reduces damping of the i.f. transformer making the gain/frequency response of the transformer sharper.

INTERMEDIATE-FREQUENCY AMPLIFIERS

An intermediate-frequency (i.f.) amplifier is a special case of an r.f. amplifier operating at a fixed frequency and provides most of the gain and selectivity of a superheterodyne receiver. The values of L and C forming the tuned circuits are not limited by the need for a variable range of capacitance values, but may be chosen to give the best performance.

High selectivity requires high Q values for the coils whilst the gain is largely determined by the dynamic resistance of the i.f. tuned circuits, $\frac{L}{CR}$ ohms.

There is a practical limit to the minimum value of C since the valve and stray capacitances must not form too large a proportion of the total tuning capacitance, as too small a value for C means greater liability to variation in the value of the total tuning capacitance during the life of the receiver.

Either inductive or capacitive tuning may be employed, but inductive tuning is preferred, since a variable inductance is less affected by changes in temperature than a variable capacitance. Inductance tuning is generally obtained by means of a movable screwed iron-dust core at the centre of the coil.

I.F. amplifiers generally employ double-tuned 'band-pass filter' coupled circuits since such circuits have a gain/frequency characteristic that approximates to the ideal case of a flat pass-band with infinite attenuation outside the band.

Very often the band-pass filters employed have identical L and C values in both primary and secondary since this eases manufacturing problems.

Reasons for Choice of Intermediate Frequency

There are several factors which must be considered when selecting a particular intermediate frequency for a superheterodyne receiver.

These factors are:-

(a) if a low intermediate frequency is used the band-pass circuits employed in the i.f. amplifier will be too selective since bandwidth is proportional to the mid-band frequency employed.

(b) the lower the intermediate frequency the more difficult it becomes to eliminate the image frequency.

(c) the intermediate frequency should not fall within the tuning range of the receiver since this would lead to instability and increased interference.

(d) too high a value of intermediate frequency will lead to a reduction in the gain and selectivity of the amplifier, and

(e) too high an intermediate frequency increases the difficulty of obtaining reasonable tracking.

As a compromise between these conflicting requirements the majority of modern receivers employ an intermediate frequency falling within the range 460 kHz to 470 kHz.

Band-Pass Filters

If two resonant circuits tuned to the same frequency are suitably coupled, the secondary current will be fairly constant over a band of frequencies in the neighbourhood of resonance, while currents of all other frequencies will be sharply discriminated against. A pair of coupled circuits adjusted to give this result is called a 'band-pass filter' because currents of frequencies within a certain band or range are transmitted almost equally well while currents of all other frequencies are suppressed. This is in contrast with the resonance curves of series and parallel circuits, which have rounded tops that do not give uniform response over a range of frequencies. Band-pass characteristics can also be exhibited by the primary current but because the effect is not so pronounced as on the secondary this is practically never used.

The important characteristics of a band-pass filter are the width of the pass-band, the uniformity of response to the different frequencies within the band, and the response to the frequencies lying within the band compared with the response to those that are outside. When a band-pass filter is formed by coupling two tuned circuits, the width of the pass-band depends primarily upon the coefficient of coupling and very little upon other factors, while the uniformity of response within the pass-band is fixed by the relation which the Q of the resonant circuits bears to the coefficient of coupling. Thus in a given circuit, the coupling being fixed, high Q circuits produce very pronounced double humps which result in serious attenuation of the mid-band frequencies while low Q circuits cause the band-pass characteristics to be replaced by a curve with a rounded top similar to the resonance curve of a simple or parallel circuit. In between these two extremes a circuit Q giving the desired band-pass characteristic with substantially equal response to a band of frequencies is possible.

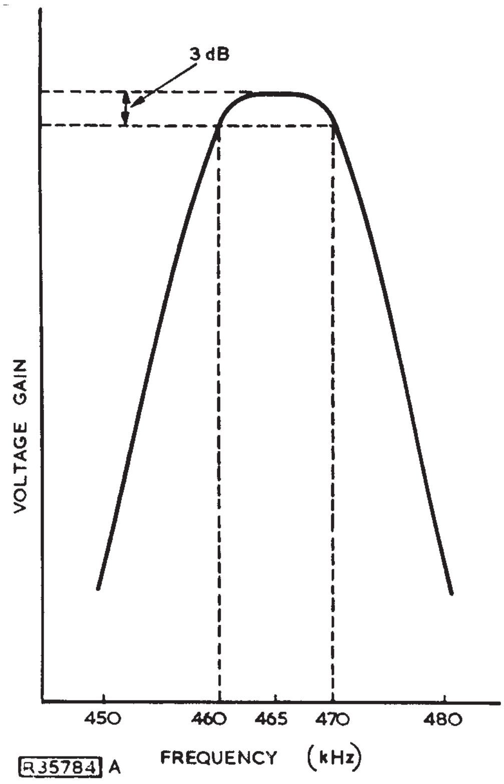


Fig. 26

Fig. 26 shows a typical gain/frequency characteristic for such a circuit. It can be seen that the curve exhibits a response that is more or less flat over a bandwidth of 10 kHz falling off rapidly outside this band.

Fig. 27 shows how a band-pass filter may be used to couple the output of the frequency-changing stage to an i.f. amplifier. The mid-band frequency of the band-pass filter is made equal to the intermediate frequency of the receiver and thus selects the difference frequency component of the output of the frequency-changer.

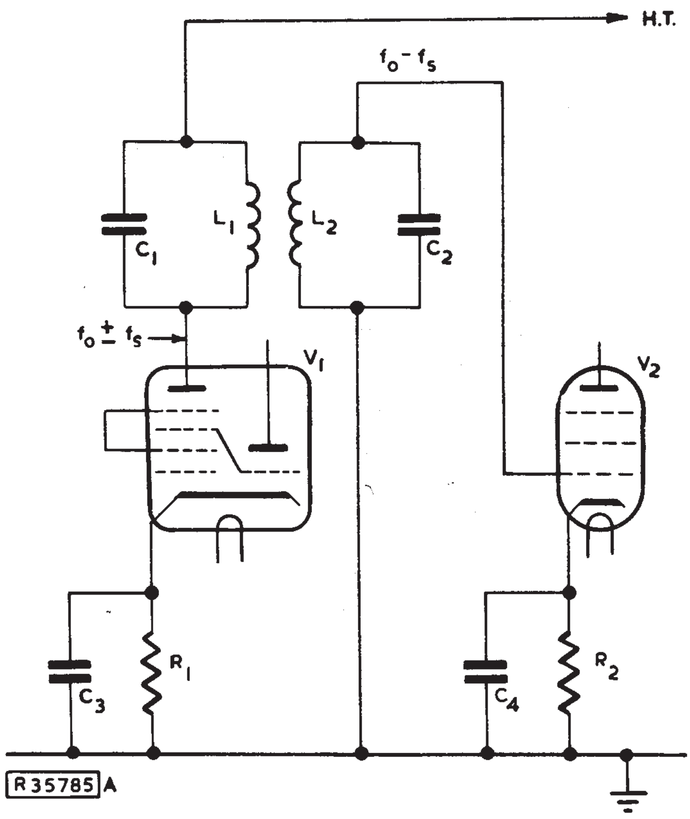


Fig. 27

Communication receivers often incorporate a crystal filter in the i.f. amplifier circuit so that very narrow bandwidths can be obtained when receiving continuous wave telegraph signals.

A Practical I.F. Amplifier Circuit

The circuit of an i.f. amplifier is shown in Fig. 28 together with typical component values.

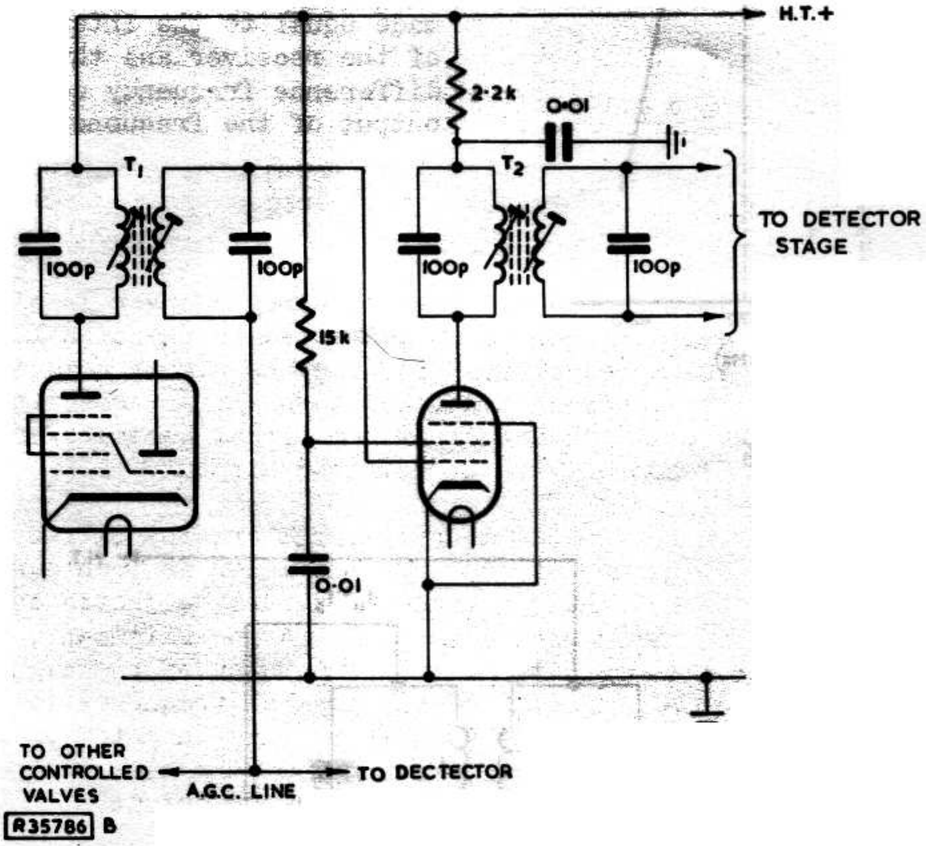


Fig. 28

The values of L and C for the band-pass filters, T_1 and T_2, are obtained from the expression $f_r = \frac{1}{2\pi\sqrt{LC}}$ Hz. In practice, C must be fairly large compared with the valve inter-electrode capacitances, in order to avoid detuning as a result of variations in these capacitances, and is arbitrarily chosen. The value of L is then chosen to give resonance at the desired frequency. It can be seen from Fig. 28 that a value of 100 pF has been chosen for both T_1 and T_2. Final adjustment to the circuit is made by pre-set adjustment of the inductance of the coils.

Alignment of I.F. Amplifiers

To align the i.f. amplifier of a superheterodyne receiver a signal generator capable of producing the intermediate frequency of the receiver is required together with a high grade a.c. voltmeter.

The alignment procedure is then as follows:- The signal generator is adjusted to the intermediate frequency of the receiver and the earthed terminal connected to the chassis of the receiver. The other lead is connected to the signal grid of the last valve in the i.f. amplifier and the a.c. voltmeter connected across the output transformer.

The secondary of the last i.f. transformer is then adjusted for maximum voltage indication on the meter. Once maximum output has been obtained, the primary is adjusted, again for maximum output.

The non-earthed lead of the signal generator is then removed from the grid of the last i.f. valve and connected to the grid of the last but one valve, and the second from last i.f. transformer brought into alignment by adjustment of its primary and secondary circuits.

This process is continued until all i.f. transformers have been aligned.

It will be necessary to reduce the level of the signal generator output as more of the i.f. amplifier is brought into use otherwise the increased gain may cause overloading of the amplifier and consequent inaccurate results.

The i.f. transformer in the anode of the frequency-changer is aligned with the non-earthed lead of the signal generator connected to the signal grid of the frequency-changer. Since the tuned circuit feeding the signal grid of the frequency-changer is tuned to a higher frequency than the intermediate frequency, it may effectively short-circuit the output of the signal generator. It may therefore be necessary to disconnect the signal grid circuit.

If the i.f. amplifier incorporates a crystal filter, the filter should be switched out of circuit. Alignment of the i.f. amplifier is then carried out as previously described, with the signal generator frequency adjusted as closely as possible to the resonant frequency of the crystal. After alignment, the crystal should be re-inserted into the circuit, and the frequency of the signal generator varied over a small range on either side of the crystal frequency to find its exact frequency, which will be indicated by a sharp rise in output. The signal generator frequency should be left at the frequency of the peak in the crystal output and the i.f. transformers re-adjusted for maximum output.

More accurate and efficient alignment of an i.f. amplifier may be performed by using a signal generator, in which the oscillator frequency is varied over a suitable range at a low repetition rate; and a cathode ray oscilloscope, the horizontal sweep of which is synchronized with the repetition rate of the signal generator so that the horizontal deflexion is a function of frequency. The audio frequency output of the diode detector is connected to the vertical deflexion plates of the oscilloscope. The spot on the screen therefore traces a curve proportional to the receiver response in terms of the instantaneous value of the signal generator frequency. This response curve is continuously visible and thus the effect of any adjustments of the amplifier may be readily observed.

To align an i.f. amplifier by means of an oscilloscope the same procedure as before is followed, the oscilloscope being connected across the output transformer, each transformer being adjusted to give maximum output consistent with the required gain/frequency response characteristic.

AUTOMATIC GAIN CONTROL

The sensitivity of the modern receiver is such that the reception of signals transmitted over great distances is common. The intelligibility and entertainment value of these signals is usually limited by "fading". Fading consists of a series of unpredictable fluctuations of the strength of a signal from a distant radio transmitter. The variations may occur as the result of the transmitted signal arriving at the receiver via several different and variable paths. The voltage applied to the receiver will vary according to the relative phases of the incoming waves; the voltage will be a maximum when the waves are in phase and a minimum when out of phase. Fading can be minimized by using more than one aerial, but this method is impracticable except for commercial point-to-point radio circuits. In most receivers, however, automatic control, by the incoming signal, of the gain of the radio-frequency and intermediate-frequency stages is employed; this method is known as "automatic gain control" or a.g.c. In a receiver fitted with a.g.c. the intensity of the signals from all radio transmitters tends to be reduced to a common level, so that the level of a local transmitter is made to approach that of a more distant transmitter. Besides minimizing fading, a.g.c. reduces the necessity for continual readjustment of the gain when the receiver is being tuned from one station to another.

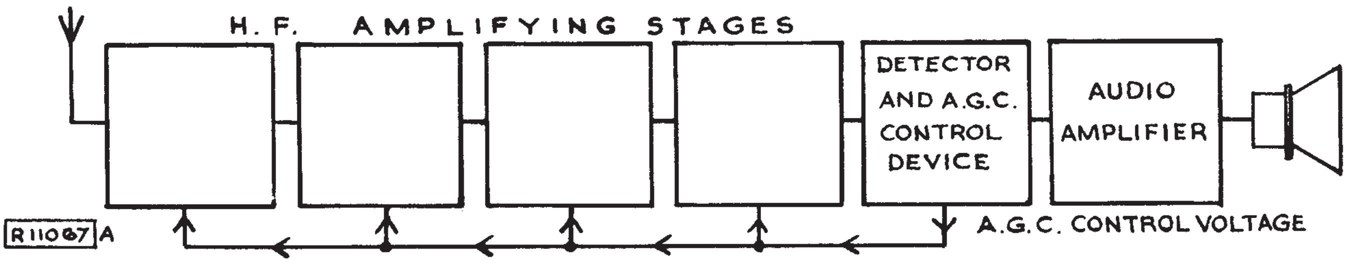


Fig. 29

The operation of the a.g.c. system is controlled by the level of the carrier component of the signal which is being reproduced; the modulation depth is not affected, so that the relative intensities of the audio components remain the same as at the transmitter. The general scheme of control is shown in Fig. 29. The input passes through the amplifying stages to the detector and a.g.c. device. Here a control voltage is developed which is proportional to the carrier strength. This control voltage is fed back to the amplifying stages and reduces the gain of each stage, again proportionately to the carrier strength.

A disadvantage of a.g.c. is that selective fading (uneven fading of the side-band and carrier components of a signal) is accentuated by its use. This is because the a.g.c. voltage is controlled by the strength of the carrier component, which may increase, causing a reduction in gain, at the instant when the strength of sidebands is reduced, or vice versa. Another disadvantage is that the sensitivity of the receiver is greatest when it is not tuned to a signal, and the background noise rises to a high level during alterations to the tuning.

The valves used in the various amplifying stages (radio frequency, frequency changing, and intermediate frequency) have variable-mu characteristics, i.e. the mutual or conversion conductance of the valve is reduced as the negative bias on the control grid is increased.

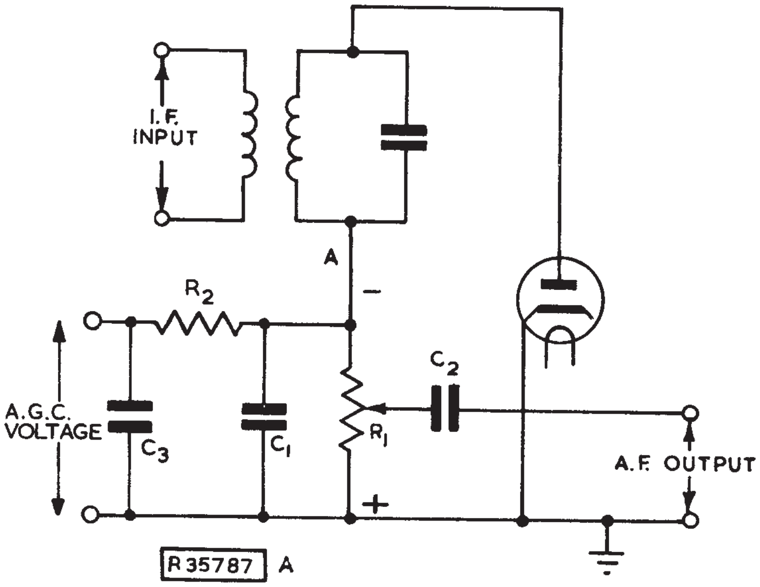


Fig. 30

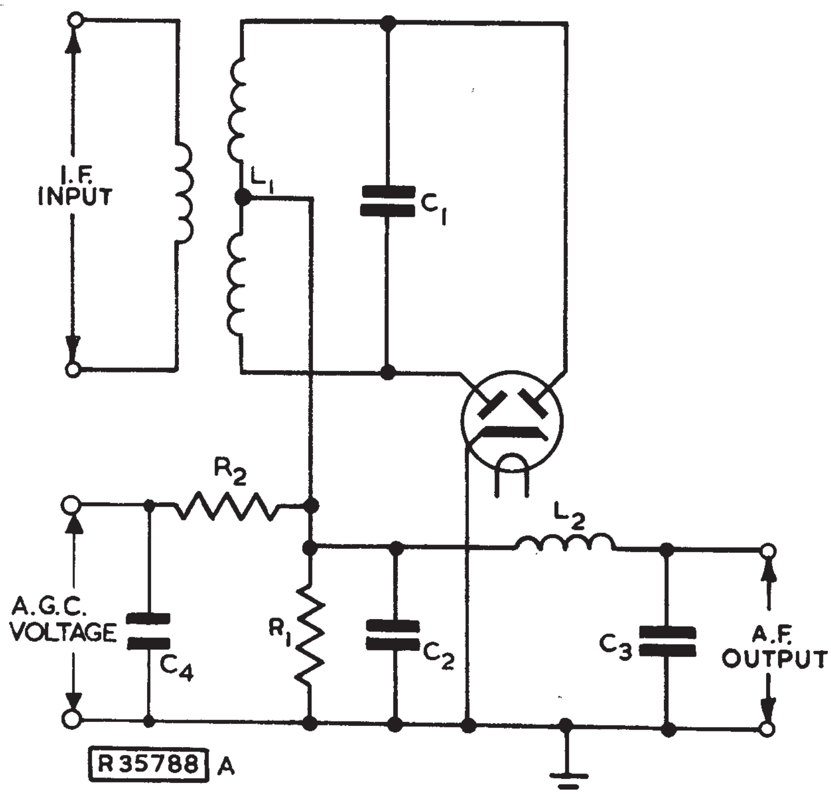


Fig. 31

Fig. 30 shows a simplified a.g.c. circuit employing a single diode valve. A direct voltage, proportional to the mean value of the carrier (465 kHz), is developed across R_1, and audio-frequency variations are superimposed on this voltage. The audio-frequency variations are passed to the next stage via capacitor C_2. The d.c. voltage is blocked from the next stage by C_2 and is used to control the gain of the previous stages.

If the carrier voltage at the detector increases, the direct voltage developed across R_1 increases proportionally providing an increasingly negative voltage at point A. Since this point is connected to the grids of the valves to be controlled, the grid bias of these valves increases, reducing the overall gain of the receiver and thus checking the increase in the carrier level at the detector input.

A complete levelling-out of the carrier characteristic cannot be obtained because the detector voltage must increase slightly in order to produce the necessary control voltage. The larger the number of controlled stages in front of the detector, however, the more sensitive is the control and with commercial receiving circuits this is one of the reasons why a large number of stages is used.

Another a.g.c. circuit which also combines the function of a detector is shown in Fig. 31. The circuit employs a double-diode as a full-wave rectifier, greater efficiency obtaining as a result. The a.g.c. voltage developed across R_1 is equal to the peak carrier voltage across one-half of the input transformer.

Biased (or delayed) a.g.c.

The circuits shown in Fig. 30 and Fig. 31 develop an a.g.c. voltage from any signal which will result in a rectified current. Now it is desirable that with very weak signals the maximum gain of the valves should be available, and therefore these signals should not develop any a.g.c. voltage. A practical method of obtaining this condition is to arrange that no bias voltage is generated by signals with a strength less than that necessary to produce the normal output. This requires the use of a separate valve to produce the a.g.c. voltage; this valve is given a bias to prevent current flowing until the input reaches a predetermined value.

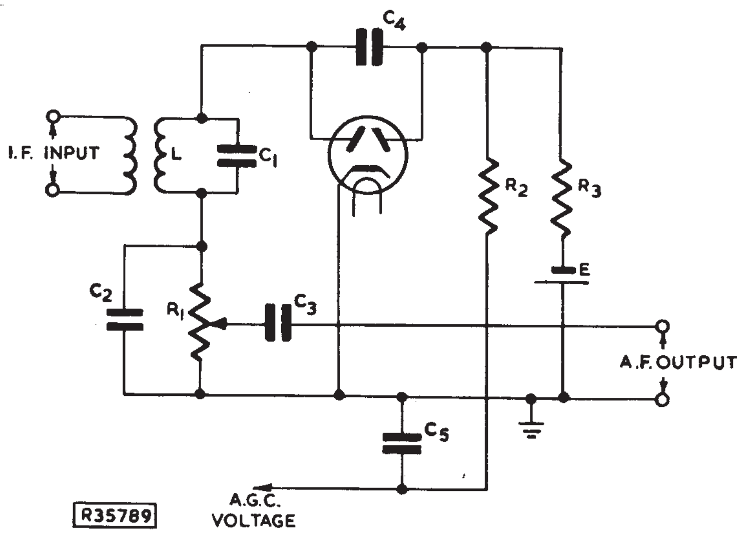


Fig. 32

A circuit providing this facility is shown in Fig. 32. The tuned circuit, L C_1, is directly connected to the signal diode and the audio frequency output voltage developed across R_1 is fed to the next stage via C_3. The signal is fed to the a.g.c. diode via coupling capacitor C_4. The a.g.c. diode is connected as a shunt detector and the a.g.c. voltage is developed across load resistor R_3. In this circuit the delay facility is obtained by biasing the a.g.c. diode negative with respect to earth by means of battery E. The a.g.c. diode will not conduct until the peak signal voltage exceeds the negative voltage on the a.g.c. diode anode. The negative delay voltage is also passed to the controlled valves via resistor R_2 and sometimes eliminates the need for normal bias on the controlled valves. Capacitor C_5 in conjunction with R_2 filters out any a.c. components in the a.g.c. line.

Fig. 33 shows a delayed a.g.c. circuit which has been used in some commercial radio receivers. The operation of the double-diode valve is similar to that just described. The audio-frequency output voltage is developed across R_1 and fed to the grid of a.f. amplifier, V_2, via capacitor C_3.

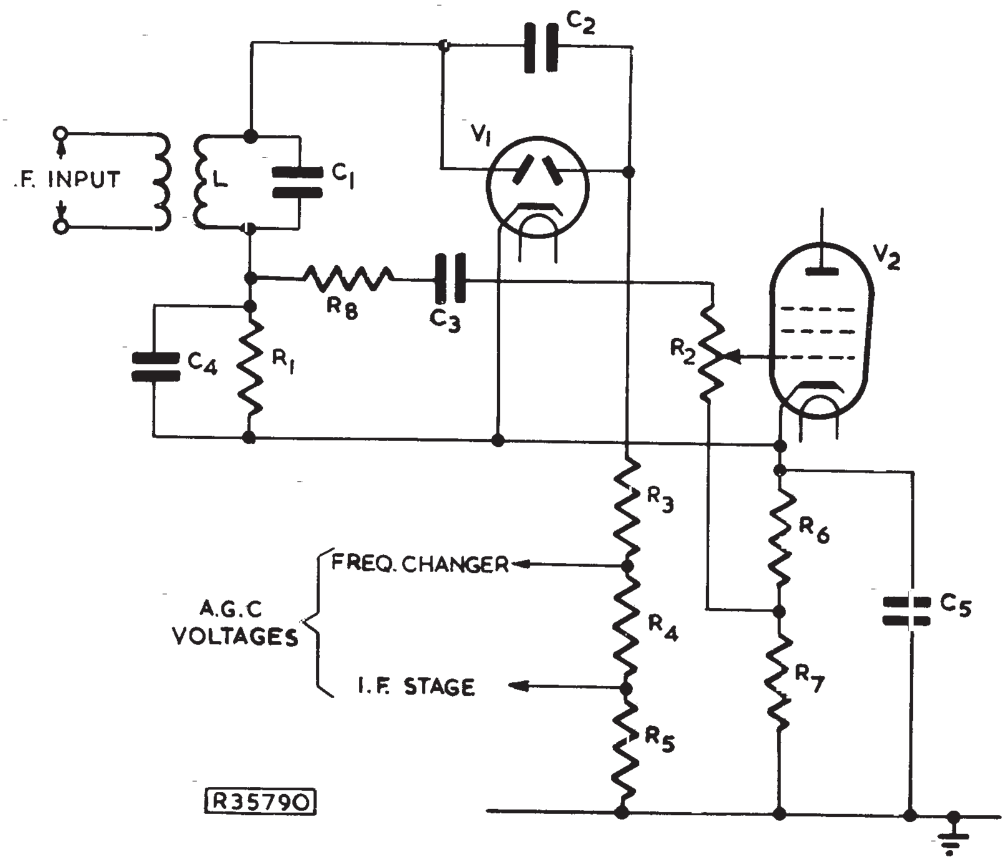
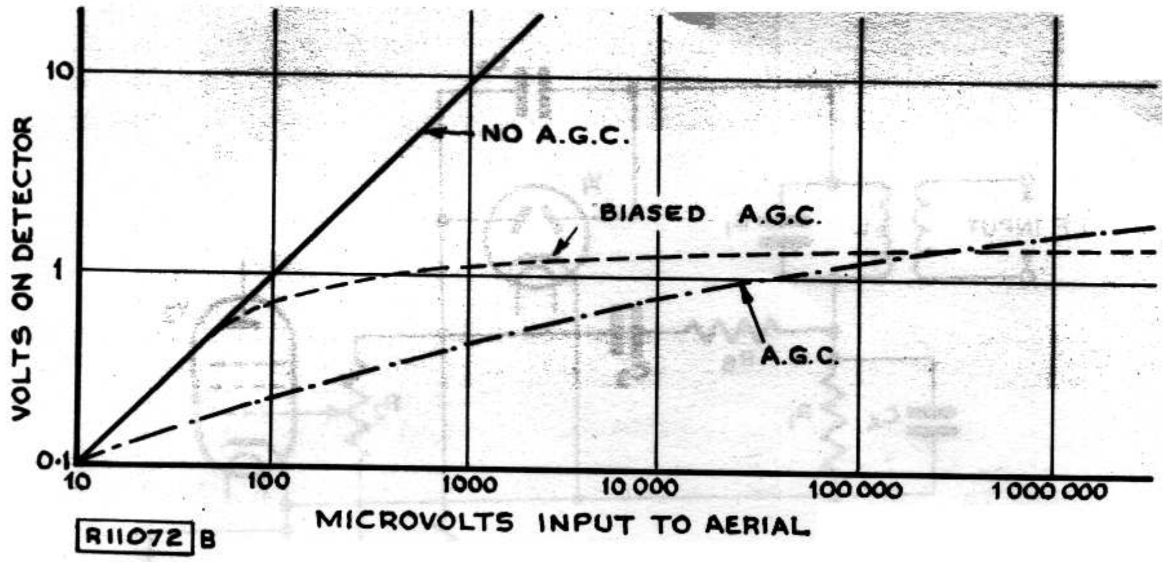


Fig. 33

Three resistors R_3, R_4, and R_5 comprise the load for the a.g.c. diode. The values of these resistors are such that the voltage developed across R_5 is suitable for controlling the i.f. amplifier, while the voltage across R_4 and R_5 together is suitable for the frequency-changer.

The cathode current of V_2 flows through R_6 and R_7 but only the voltage developed across R_6 is used as grid bias voltage for this valve. The full voltage developed across R_6 and R_7 is the delay voltage, the cathode of the a.g.c. diode being biased positively by this voltage.

Fig. 34 shows the gain characteristics of typical radio receivers incorporating (a) no a.g.c., (b) a.g.c., and (c) biased a.g.c.



34

The rate of response of the a.g.c. circuit is determined by its time constant e.g. R_2 C_5 in Fig. 32. This time constant must be sufficiently small to prevent the a.g.c. voltage varying with the modulation envelope, but must not be too great or the a.g.c. voltage will not follow the more rapid variations in the received signal due to fading.

In designing an a.g.c. system the aim is to obtain a good control ratio, i.e. a large control voltage for a small change in signal strength. To some extent biased a.g.c. automatically provides an improvement, as the curves in Fig. 34 shows.

The full curve represents the output voltage in terms of input, without a.g.c. If simple a.g.c. is used, this becomes operative at once and cannot therefore be allowed to be too fierce in its operation or the output would never reach its normal value. The chain-dotted curve shows a typical performance without delay. It will be noted that the gain on weak signals is appreciably reduced.

The third curve shows biased a.g.c. which does not operate until the output is nearly up to the full amount and then comes into play with full action, giving an almost level output and a much better control ratio.

To Calculate the Degree of Control

Let the gain of the controlled stages be inversely proportional to the grid bias applied to the valves and let the maximum gain from aerial to detector be A.

If A_1 is the gain when a negative bias of v volts is applied to the valves, then $A_1 = A \times \frac{k}{v}$

where k is a constant dependent upon the characteristics of the valves

e.g. if the a.g.c. reduces the gain of a receiver from A to $\frac{A}{1000}$ for a change in bias voltage of 10 volts then:

$$A_1 = \frac{A}{1000} = A \times \frac{k}{10}$$

$$k = \frac{1}{100}$$

With delayed a.g.c. no controlling bias is developed until a certain signal level at the detector is exceeded. Let this delay voltage be 14 volts. The signal voltage at the detector input, that will generate a steady voltage of 14 volts, is 10 volts r.m.s. since the detector delivers an output voltage approximately equal to the peak value of the input signal voltage. This means that no control bias will be generated until the signal level at the detector input reaches 10 volts r.m.s.

Let the input signal level at the input terminals of the receiver, which corresponds to the level necessary to develop a signal at the detector to produce a d.c. potential equal to the delay potential, be V_1 volts. The signal level at the detector will then be $V_1^1 = AV_1$.

Let the input signal required at the input terminals of the receiver, before the a.g.c. will reduce the gain from A to $\frac{A}{1000}$, be V_2. The signal at the detector is then $V_2^1 = V_2 \frac{A}{1000}$.

$$V_1 = \frac{V_1^1}{A}$$

$$V_2 = \frac{V_2^1}{\frac{A}{1000}} \text{ or } \frac{1000\ V_2^1}{A}$$

$$\frac{V_2}{V_1} = 1000 \frac{V_2^1}{V_1^1}$$

If the input to the detector rises to 17 volts r.m.s. corresponding to a peak voltage of $17\sqrt{2}$ or 24 volts, producing a bias of (24 - 14) or 10 volts; then $V_2^1 = 17$ volts and $V_1^1 = 10$ volts.

$$\frac{V_2}{V_1} = 1000 \frac{17}{10}$$

1700 or 64.6 dB

Thus 64.6 dB variation in the level of the input signal carrier component gives rise to only $\frac{17}{10}$ or 4.6 dB variation in audio output.

Example

A radio receiver has two i.f. amplifier valves, and one r.f. amplifier valve, each having an a.g.c. control of 3 dB/volt, and a frequency-changer valve with a control of 2 dB/volt. The receiver is to be such that an input range of 5µ volts to 100 millivolts will give rise to only 6 dB increase in output. Find the necessary a.g.c. delay voltage assuming 100% detection efficiency.

The total gain control per volt is the sum of the individual gain controls, i.e. 11 dB/volt.

The range of input levels is from 5×0^{-6} volts to 0.1 volt, or expressed in dB, 86 dB variation.

The output of the receiver must not, therefore, increase by more than 6 dB when the input to the receiver is increased by 86 dB, so that the required change in gain, given by the difference between the input and output level ranges, is 80 dB.

The change in bias required is therefore $\frac{80}{11}$ or <u>7.273 volts</u>.

If the signal level at the input to the a.g.c. diode when the input is 5µ V is V_o r.m.s. then the signal level for an input of 100 mV is 2 V r.m.s., since 6 dB corresponds to a voltage ratio of 2:1.

Since 100% detection efficiency may be assumed, the d.c. component of the detector output is given by:

$$\text{direct voltage} = \sqrt{2} \times \text{r.m.s. voltage of input}$$

Therefore the required change in bias for 2:1 change in output level

$$2\sqrt{2}\,V_o - \sqrt{2}\,V_o = 7.273$$

$$\sqrt{2}\,V_o = 7.273 \text{ volts}$$

If it is assumed that the 5 µV input is the level below which the output of the receiver is proportional to the input, then 7.273 volts must be the delay voltage, i.e. the voltage that must be exceeded by the peak output of the detector before any a.g.c. bias voltage is produced.

Amplified a.g.c.

There are two main ways in which the performance of an a.g.c. system may be improved. The first is to increase the factor k: k may be increased by increasing the rate of change of gain of the valves with change of bias. There is a limit to this however, since the valves will tend to introduce excessive distortion because of the curvature of their characteristics. The factor k is also increased if the number of controlled stages is also increased.

The second method of improving the performance of an a.g.c. system is to amplify the bias potential. The bias potential may be amplified by amplification of the direct voltage by means of a d.c. amplifier or by providing a separate i.f. stage to feed the a.g.c. detector as shown in block diagram form in Fig. 35 or by feeding the a.g.c. voltage directly from the anode of the last i.f. amplifier valve.

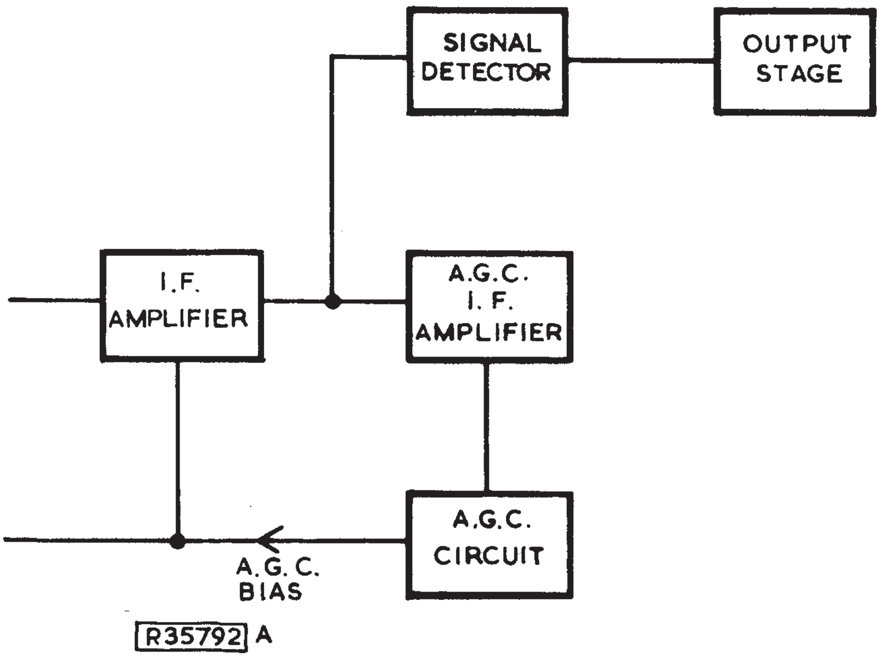


Fig. 35

Various forms of amplified a.g.c. have been devised, in which the a.g.c. voltage is applied to a d.c. amplifier which develops an adequate output. Most of these circuits are fairly simple if an extra valve is used, but in broadcast receivers cost precludes the use of another valve and such single-valve systems that have been used have proved rather unstable.

Practical a.g.c. Circuits

In broadcast receivers a double-diode triode valve is often employed, enabling the functions of detection, a.g.c. and audio-frequency amplification to be carried out by a single valve. Fig. 36 shows a typical circuit employing a double-diode triode valve.

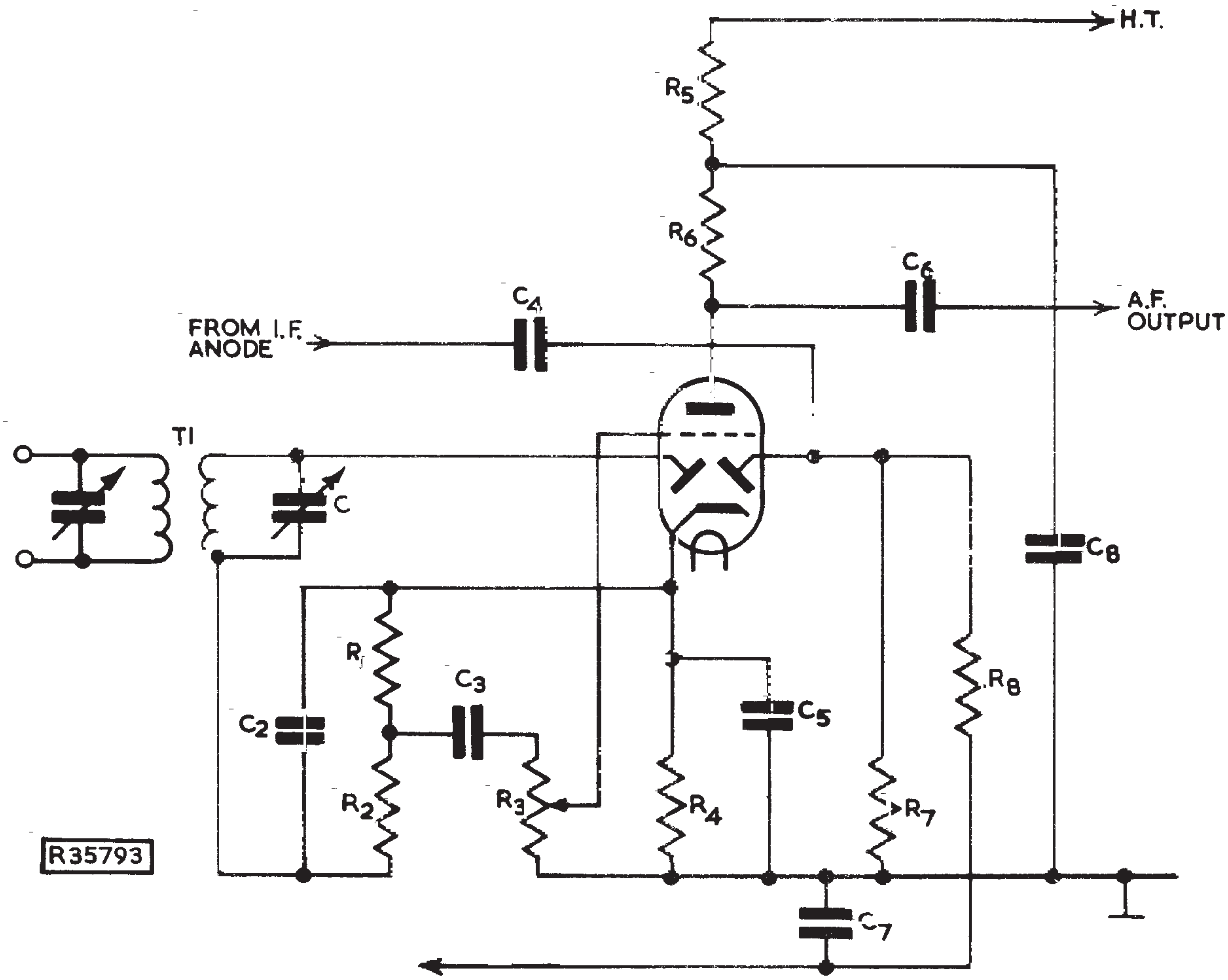


Fig. 36

Detection of the signal takes place in the left-hand diode and the a.f. output voltage developed across $R_1 + R_2$ is fed via capacitor C_3 and appears across potentiometer R_3. A portion of this voltage is then tapped off and fed to the grid of the triode section of the valve.

A voltage at the intermediate frequency is fed to the right-hand diode from the anode of the last valve in the i.f. amplifier. The voltage available at this point is greater than at the secondary of T_1. The delay voltage is developed across resistor R_4 and as soon as the i.f. voltage at the right-hand diode exceeds this delay voltage, the diode conducts, and the a.g.c. voltage is developed across resistor R_7.

Another typical circuit employing a double-diode triode valve is shown in Fig. 37.

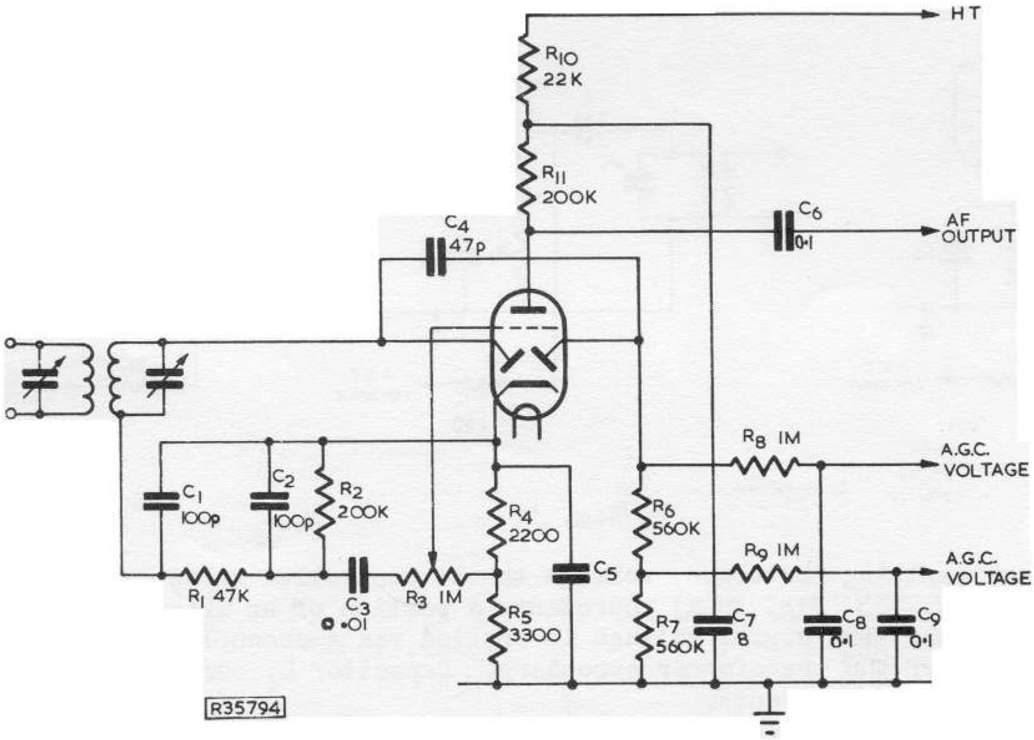


Fig. 37

The functions of the components are as follows:-

R_1, R_2 and C_2 : signal diode load

R_1, C_1 : intermediate-frequency filter

C_3 : blocks d.c. from triode grid

R_3 : signal gain control

C_4 : a.g.c. diode coupling capacitor

R_4 : triode cathode bias resistor

R_4, R_5 : the a.g.c. delay bias voltage is developed across these resistors

R_6, R_7 : a.g.c. diode load

R_8, R_9, C_8, C_9 : a.g.c. line filter circuits

C_6 : blocks h.t. from a.f. output terminals

R_{11} : anode load for triode

R_{10}, C_7 : anode decoupling circuit.

Methods of Applying the a.g.c. Voltage

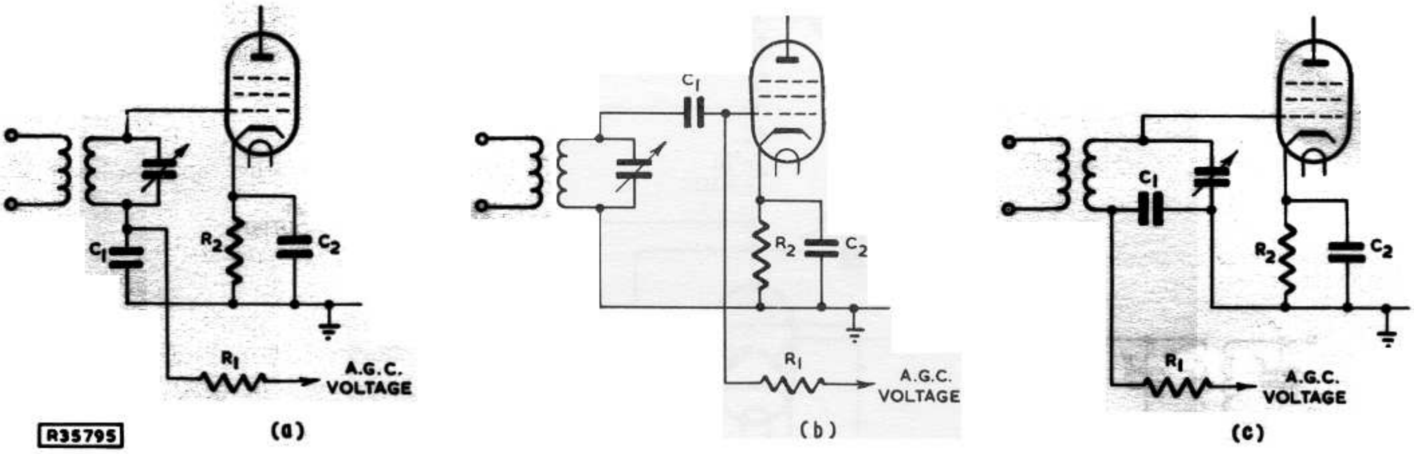


Fig. 38

Methods of applying the a.g.c. voltage to the controlled valves are illustrated in Fig. 38. Fig. 38(a) represents a portion of an intermediate-frequency amplifier. The a.g.c. voltage is applied via a decoupling resistor R_1 to the earthy end of the transformer secondary. Capacitor C_1 completes the grid circuit for alternating currents.

In tuned radio-frequency stages the variable capacitor has normally one set of plates at earth potential. Thus the circuit of Fig. 38(a) cannot be used. Either the circuit of Fig. 38(b) or 38(c) may be used. Series feed is shown in Fig. 38(c), the a.g.c. voltage is applied to the earthy end of the tuning inductor; the tuned circuit is completed via capacitor C_1. If C_1 has too small a value its presence in the tuned circuit will upset the ganging as it effectively reduces the value of the variable capacitor. A usual value for C_1 is 0.1 microfarad. Fig. 38(b) illustrates shunt feed; it is not very popular owing to the damping effect on the tuned circuit.

In all these circuits R_1 is the a.g.c. decoupling resistance and has a value between 10000 ohms and 1 megohm. It must never be greater than the value recommended by the valve manufacturer as the maximum grid to cathode resistance. Resistor R_2 provides the standing grid bias until an a.g.c. voltage is developed and is chosen to give maximum stage gain.

COMMUNICATION RECEIVERS

This pamphlet, so far, has discussed superheterodyne radio receivers with particular emphasis on the domestic broadcast receiver. Communication receivers are generally more elaborate than broadcast receivers and may often employ two intermediate frequencies in order to obtain adequate image rejection.

The main performance requirements of a communication receiver are as follows:-

(a) Selectivity: in order to achieve the best signal-to-noise ratio it is desirable to use a receiver having the narrowest bandwidth that will satisfactorily receive the wanted signal. In practice, this requirement is met by providing a range of bandwidths varying from, say, 100 Hz to 6 kHz, the bandwidths being chosen so as to suit the different types of signal the receiver may be called upon to receive.

(b) Sensitivity: the receiver must be able to work with very weak signals, preferably down to the level of the local radio noise.

(c) Tuning: when used to receive h.f. signals with a narrow bandwidth, it is important that the tuning of the receiver should not drift and that the tuning scale should be sufficiently large and accurate.

(d) Gain: high-frequency signals are subject to fading and thus an efficient a.g.c. system is required. The receiver should be suitable for working as one of a group of two or more receivers if a diversity system is required.

(e) Beat oscillator: if the receiver is to receive continuous wave telegraph signals a beat oscillator is necessary.

In general, communication receivers may be divided into two main types:-

(i) those employed as general communication receivers, i.e. police, army, etc Such receivers may be called upon to handle various signals, varying in frequency and bandwidth, and generally handle double-sideband signals.

(ii) those used for point-to-point communication circuits which are often designed to handle single-sideband transmissions.

A block diagram of a communication receiver, designed to operate over a frequency range of 60 kHz to 30 MHz, and to have five different bandwidths of 6 kHz, 3 kHz, 1.2 kHz, 300 Hz and 100 Hz respectively is shown in Fig. 39.

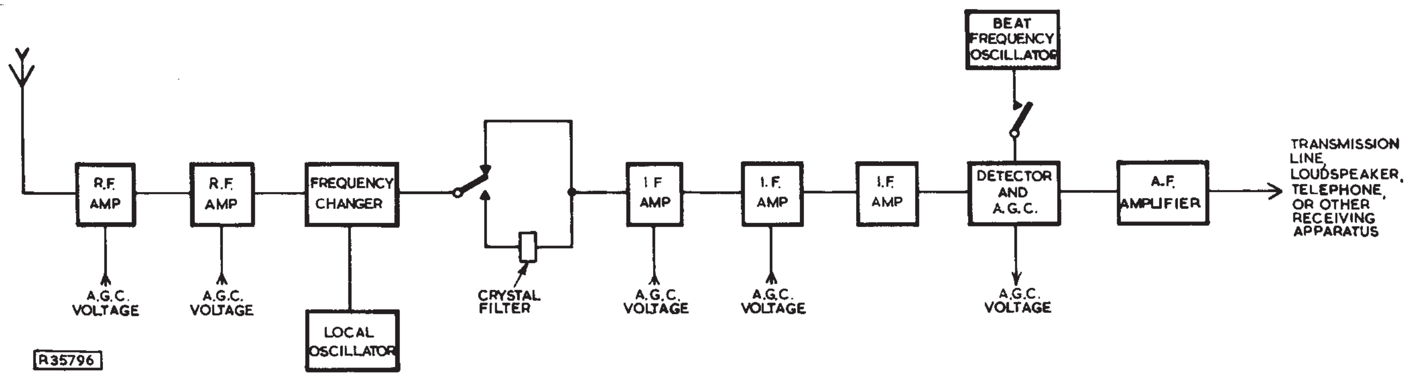


Fig. 39

The receiver includes two r.f. amplifier stages working into the frequency-changing stage. The output of the frequency-changer, 465 kHz, is passed through the three-stage i.f. amplifier. The output of the i.f. amplifier operates a double-diode triode incorporating the detector, the a.g.c. diode and a stage of a.f. gain. A beat frequency oscillator may be connected to the detector by means of a switch when it is required to receive continuous wave telegraphy signals.

Radio-frequency stages

The r.f. amplifier stages follow conventional design and are arranged to have the same stage gain at all frequencies within the tuning range of the receiver.

Intermediate-frequency amplifier

The i.f. amplifier incorporates a crystal filter circuit inserted between the frequency changer and the first i.f. amplifier valve. The crystal filter is inoperative on the 6 kHz and 3 kHz pass-bands, selection between the two pass-bands being obtained by altering the coupling of the i.f. amplifier band-pass filter circuits.

For the 1.2 kHz and 300 Hz pass-bands, the crystal filter is introduced; change in the bandwidth being obtained by altering the phase and impedance of the associated reactive components.

For 100 Hz bandwidth the i.f. circuits are left in the 300 Hz position and further selectivity obtained by inserting, in the audio circuits, a band-pass filter having a mid-band frequency of 1000 Hz. The additional selectivity introduced by this filter is restricted to the reception of unmodulated signals since the beat frequency oscillator must be used to produce the frequency change necessary to give the selected low frequency beat output.

A.G.C. circuits

The input of the a.g.c. diode is taken from the primary of the last i.f. transformer. The a.g.c. voltage produced is applied to both the r.f., and the first two i.f. valves. Time constants suitable for both telephony and telegraphy may be selected, the time constant for telegraphy being long enough to prevent surges of noise between morse elements when working at comparatively slow hand speeds.

Beat-frequency oscillator

The beat-frequency oscillator is electron-coupled to the signal diode and has a peak value such that it is capable of modulating the highest signal level at the diode, but is well below the a.g.c. delay voltage, and therefore, does not operate the a.g.c.

The gain/frequency characteristics of this receiver, when adjusted for bandwidths of 6 kHz, 3 kHz, 1.2 kHz and 300 Hz, are shown in Fig. 45. The 100 Hz bandwidth curve is not shown since it would need an expanded horizontal scale to be clearly seen.

Single-sideband receivers

Fig. 40 shows a block diagram of a typical single-sideband receiver capable of handling an audio frequency bandwidth of 100 to 3000 Hz.

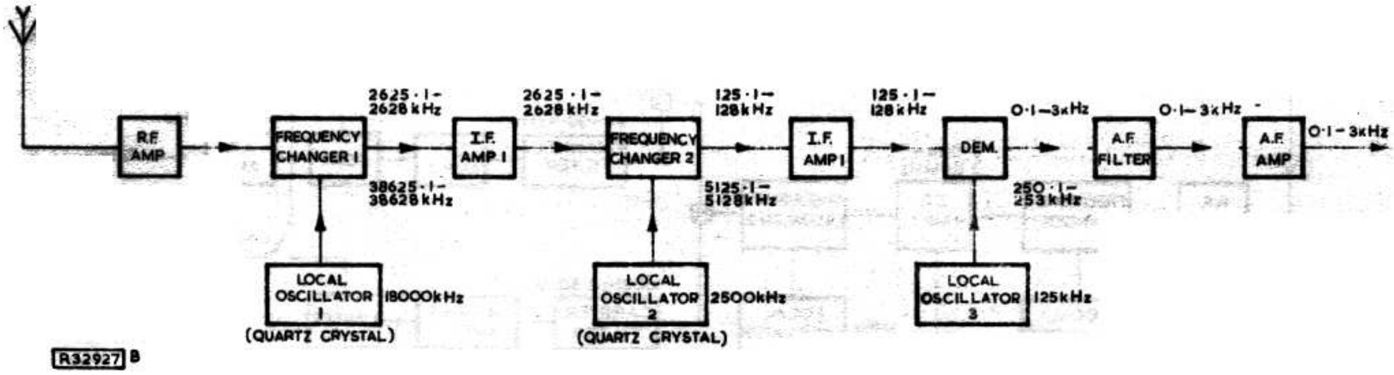


Fig. 40

It can be seen that the circuit is essentially the same as that of a receiver for a normal double-sideband, transmitted carrier. The main difference lies in the use of a carrier fed demodulator instead of a detector not requiring a local source of carrier, such as a diode detector.

No provision is made for a pilot carrier to synchronize the re-introduced carrier which must be within 15 Hz of the transmitter carrier for commercial quality speech, and no provision is made for a.g.c.

It would be possible to synchronize the carriers, by using quartz crystal oscillators at the transmitter and receiver, and by taking the usual precautions to eliminate frequency drift, e.g. either temperature controlled crystal ovens or use of crystals with a low temperature-frequency coefficient, but it is difficult to provide reasonable gain control without using a pilot carrier or pilot frequency. For this reason the circuit of Fig. 40 is not practical for long distance communication where fading occurs in the transmission path.

Independent-sideband receivers

In the form of single-sideband transmission employed by the British Post Office, two audio channels modulate carrier supplies having the same frequency and the upper sideband of one and the lower sideband of the other are transmitted, together with the carrier frequency at a reduced level. This type of transmission is known as 'Independent-Sideband' and combines the advantages of double-sideband and single-sideband working.

Fig. 41 shows a block schematic diagram of a typical independent sideband receiver. In common with practically all single-sideband receivers this circuit employs the double-superheterodyne principle, i.e. two stages of frequency-changing. It should be noted that the first stage of i.f. amplification has a bandwidth of 28 kHz and the two second stages each have a bandwidth of 6 kHz.

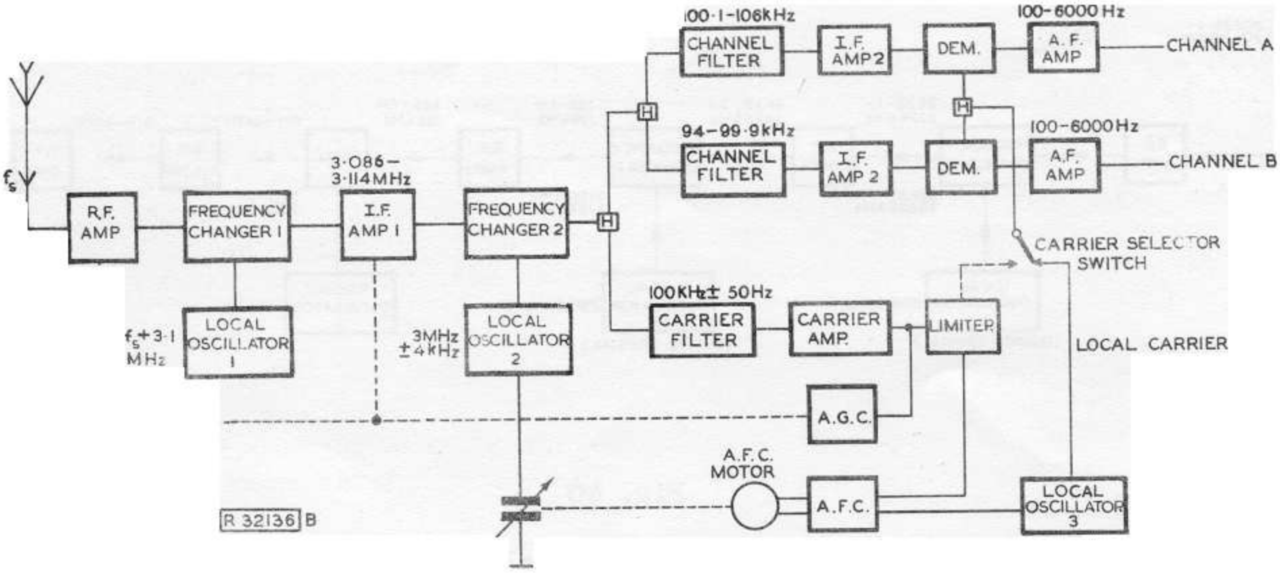


Fig. 41

Referring to Fig. 41; the aerial input is connected to a radio-frequency amplifier then to the first frequency-changer which produces an intermediate frequency of 3.1 kHz. The radio-frequency amplifier assists in increasing the signal-to-noise ratio at the output and the high intermediate frequency provides a high image response ratio.

Effective filtering of the carrier and sideband signals can best be carried out at a low intermediate frequency so that, after a stage of 3.1 kHz intermediate frequency amplification, a second frequency changer is used to translate the 3.1 MHz ± 14 kHz of the first intermediate frequency to 100 kHz ± 6 kHz. Channel filters which are replicas of those used in the transmitter are used to separate channels *A* and *B*, and a carrier filter with a pass-band of 100 Hz selects the 100 kHz carrier. The two channels are then amplified in their respective second intermediate frequency amplifiers and applied to the channel demodulators.

The signal applied to each demodulator consists of one sideband and the low-level pilot carrier which has been further attenuated by the channel filter. To obtain a relatively undistorted audio-frequency output from the demodulator it is necessary to have a strong carrier. In this case the carrier frequency is supplied either from a local carrier source, or from the amplified and reconditioned pilot carrier. The demodulator is usually of the linear balanced type using diode valves and the carrier is supplied at a level approximately ten times the peak sideband level. By using the balanced type of demodulator unwanted amplitude modulation on the carrier is prevented from producing an audio-frequency noise output.

The pilot carrier which is selected by the carrier filter is amplified in the carrier amplifier, and used to provide automatic gain control, automatic frequency control, and carrier reinsert to the channel demodulators or control of the frequency of the local carrier oscillator which under certain conditions provides the carrier reinsert to the demodulators.

Linear class A amplifiers are used in the first few stages of the carrier amplifier so that the output of the automatic gain control circuit may vary linearly with the input carrier level. A carrier of approximately constant magnitude is desirable for application to the automatic frequency control and the demodulators, and thus the last stage of the carrier amplifier functions as a limiter, the output of which does not vary appreciably until the input is reduced by about 20 dB below the level normally set by the automatic gain control.

Automatic gain-control

The function of the automatic gain control is to keep the receiver output level at approximately the same value for all levels of signal input above the threshold level at which the control begins to operate. Fig. 41 shows the control voltage applied to the r.f. and first i.f. stages of the receiver, and it is so arranged that a gain of 80 dB in the input will produce 10 volts bias which will reduce the receiver gain. There will be a delay voltage on the a.g.c. so that it will not come into operation on weak incoming signals; the a.g.c. will therefore only come into operation after the incoming signal has risen sufficiently to overcome the delay voltage. If the incoming signal rises 80 dB above this threshold level the 10 volts bias produced will decrease the receiver gain but obviously if the a.g.c. is to continue to function it cannot reduce the gain to give the same output as existed before the input rose. In practice there will be approximately

6 dB increase in the input of the i.f.$_2$ amplifiers; this may be compensated for by applying part of the a.g.c. control voltage to the grids of the i.f.$_2$ amplifiers; a.g.c. applied in this manner is known as forward-acting automatic-gain control.

A fairly long time constant is required on discharge to prevent changes in gain of the receiver on typical selective fades. A discharge time-constant of between 5 and 50 seconds, depending upon the type and period of the selective fading encountered, allows sustained changes in signal level to be corrected but prevents the receiver gain from changing appreciably on short term selective fades. The carrier level does not rise much above its average level during fading therefore a smaller charging time-constant may be used and a value of time-constant of about 2 seconds is commonly employed.

From the above it can be seen that while the a.g.c. system described is very effective in controlling relatively slow changes in the average level of the received signal it does not correct the more rapid changes of level due to propagation conditions. These rapid changes in level generally do not exceed 15 dB and are frequently corrected by using a constant-volume audio-frequency amplifier and by the use of spaced-aerial diversity-receiving equipment.

Automatic frequency-control (a.f.c.)

The purpose of the pilot carrier is to enable the received signal to be demodulated to produce the original audio-frequencies, and to provide a means by which satisfactory automatic gain-control may be applied to the receiver. A narrow-band carrier filter is used to separate the carrier from the sidebands and the primary function of the a.f.c. is to keep the second intermediate carrier-frequency output from the second frequency changer within the carrier-filter pass-band of 100 kHz ± 50 Hz. The frequency may be changed by frequency drift in the local oscillators. Alteration of frequency due to this cause is corrected by application of a.f.c. to the second frequency changer.

The reconditioned carrier, i.e. the carrier after being amplified and limited in magnitude, may be fed directly to the demodulators, or if the conditions of propagation are such that the carrier itself is not suitable for this purpose it may be used, via the a.f.c. circuit, to control the frequency of a local oscillator which supplies the carrier reinsert frequency. In Fig. 41 the carrier is shown as being supplied from local oscillator 3.

There are two forms of a.f.c. which have been employed for this purpose:-

An electronic system using a frequency-discriminator and a reactance valve.

An electro-mechanical system employing a motor drive tuning capacitor.

SPECIFYING AND MEASURING RECEIVER PERFORMANCE

Sensitivity

The preferred definition of sensitivity is:- the characteristic of a receiver which determines the minimum usable input, i.e. the least input which produces an output which satisfies certain specified requirements, including generally a specified output signal-to-noise ratio.

To enable reliable results to be obtained under exactly repeatable conditions certain standards are laid down for use when sensitivity tests are being made. If these conditions are adhered to then the results are of value in comparing the performances of different types of receiver.

A signal generator (with facilities for modulating the carrier 30% by 400 Hz) with its output variable in steps of one microvolt from one microvolt to one volt is used to provide the input to the receiver. Matching of the signal generator output to the receiver input is achieved by connecting a standard dummy aerial between them. The connexions for measurement of receiver sensitivity are shown in Fig. 42 while the circuit of the standard dummy aerial and a graph of its impedance-frequency characteristics are shown in Fig. 43.

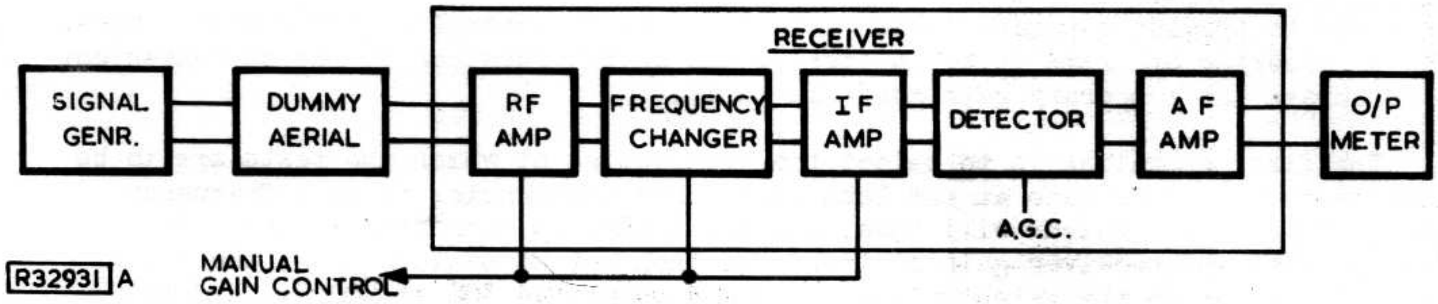


Fig. 42

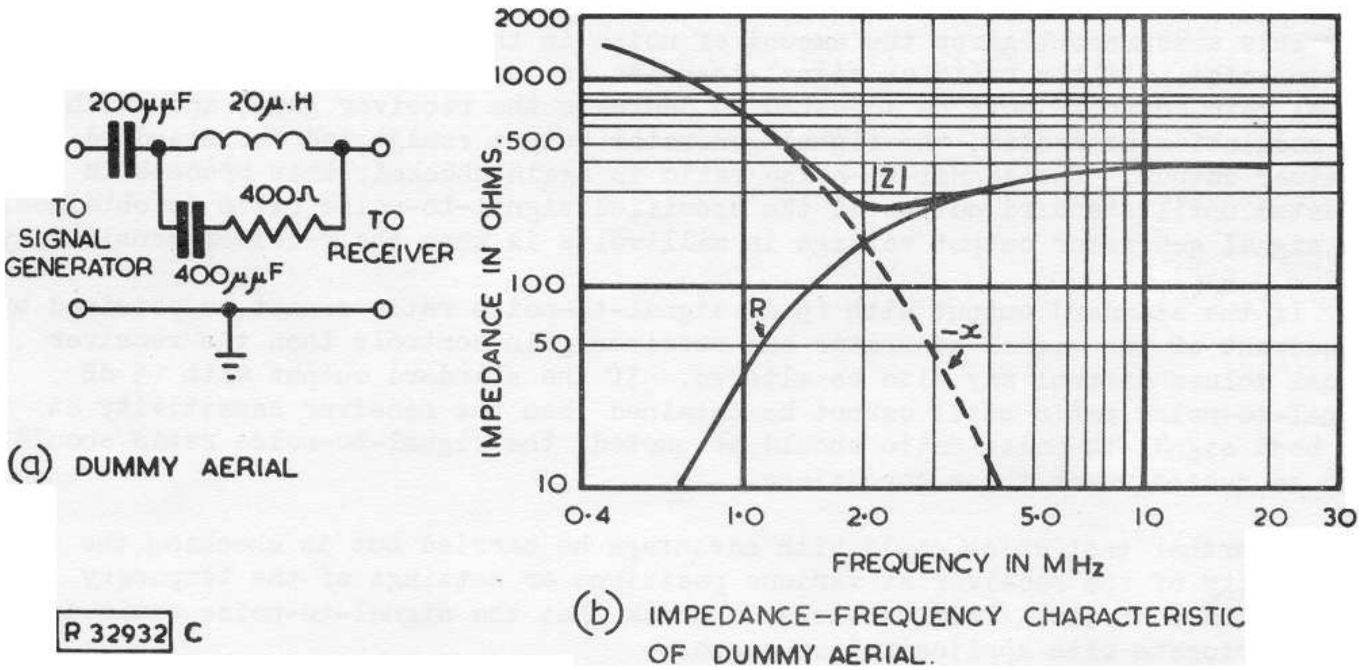


Fig. 43

For the types of receiver under consideration the specification of sensitivity is:- that input in millivolts, applied as described above, required to produce the standard output with a minimum signal-to-noise ratio of 15 dB in the output. Standard output for all types of receiver using loud speakers is taken as 50 mW for a 30% modulated signal, i.e. approximately 500 mW for 100% modulation; for receivers using headphones standard output is 0.1 mW for 30% modulated signal, i.e. 1 mW for 100% modulated signal.

Method of measurement using a modulated carrier

The sensitivity test may be made with the a.g.c. in action but it is more easily carried out if the a.g.c. is put out of action and substituted by a manual gain control as shown in Fig. 42. In some circumstances, e.g. if less a.g.c. is fed to the intermediate-frequency amplifier than to the other stages, it may be desirable to provide separate manual gain controls to control the different stages. Also if a series of tests is to be carried out on the receiver it may be convenient to calibrate the temporary gain controls in dB.

The first operation is to select the frequencies at which the tests are to be made; tests should be made at not less than three frequencies in each frequency band to which the receiver will tune, and the apparatus connected as shown in Fig. 42. All the receiver gain controls are then set at maximum gain, the signal generator set up on the selected frequency and modulated 30% at 400 Hz, and the receiver tuned for maximum output (measured by the output meter). Care must be taken here to keep the output of the signal generator low enough to avoid distortion in the receiver. The receiver manual volume control is then set to its normal operating position and the signal generator output adjusted for standard receiver output. Signal-to-noise ratio at this input is then measured by disconnecting the modulation in the signal generator and measuring the receiver output.

This measurement gives the amount of noise in the output when the carrier only is connected. If the ratio of signal-to-noise is less than 15 dB the temporary manual gain controls must be adjusted to decrease the receiver gain, and, with the 30% modulation connected, the signal generator output readjusted for standard receiver output. The signal-to-noise ratio is again checked; this process is repeated until standard output at the specified signal-to-noise ratio is obtained. The signal generator output voltage in millivolts is then the receiver sensitivity.

If the standard output with 15 dB signal-to-noise ratio cannot be obtained by adjustment of the signal generator and receiver gain controls then the receiver manual volume control may also be altered. If the standard output with 15 dB signal-to-noise ratio still cannot be obtained then the receiver sensitivity at the best signal-to-noise ratio should be quoted; the signal-to-noise ratio should also be quoted under these conditions.

A further test which could with advantage be carried out is checking the sensitivity of the receiver at various positions or settings of the temporary manual gain controls. This serves as a check that the signal-to-noise ratio does not deteriorate with application of a.g.c.

Method of measurement using an unmodulated carrier

Most communication receivers are fitted with beat oscillators to enable them to receive continuous waves. This facility enables the sensitivity of the receiver to be measured using an unmodulated signal generator. The beat oscillator in the receiver is adjusted to give a 400 Hz note with the frequency at the mid-point of the i.f. band. The same procedure may be followed as previously, except that the beat oscillator is switched off to measure the noise power, and that care must be taken to ensure that a linear relation exists between the r.f. input and the a.f. output; if this is not so a reduced value of a.f. gain should be used.

If a comparison of the results obtained is made with results obtained on the same receiver using a modulated input with the beat oscillator switched off, it will be found that the sensitivity as measured with the C.W. signal is greater than that measured with the modulated input. (N.B. a greater sensitivity implies a smaller input in millivolts thus the figure for the sensitivity will be lower).

Image channel selectivity

The image channel selectivity of a receiver is the measure of its ability to reject the image channel which is twice the intermediate frequency above or below the wanted channel. If the receiver has more than one stage of frequency changing there will be more than one image channel. Image channel selectivity is measured by measuring the sensitivity of the receiver in the normal manner then adjusting the signal generator frequency to the second channel of the test frequency, without altering the receiver tuning or gain controls and again measuring the sensitivity of the receiver. Then:-

$$\text{image response ratio} = \frac{\text{Sensitivity of receiver to image signal}}{\text{Sensitivity of receiver to wanted signal}}$$

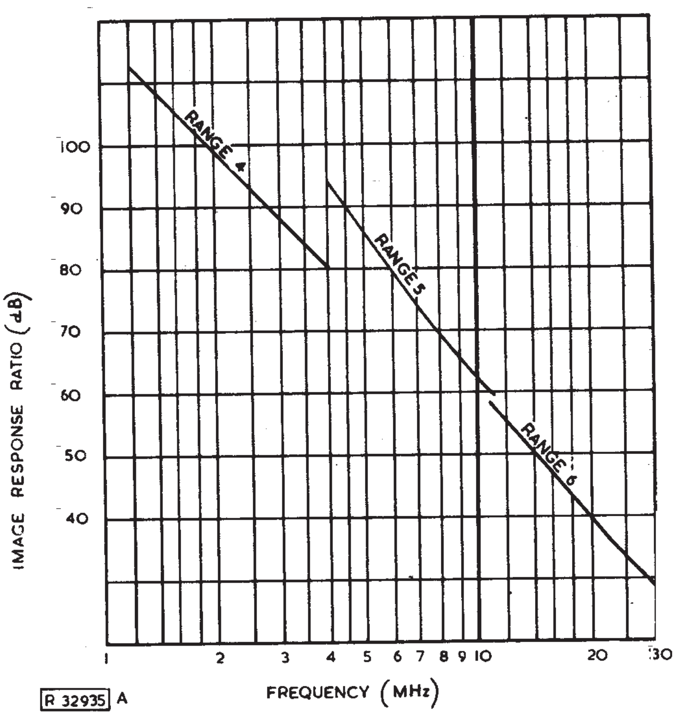


Fig. 44

The image response ratio will vary over the tunable range of the receiver decreasing as the test frequency is increased. It should be measured at not less than three frequencies in each band but its measurement is much more important at the higher end of the receiver frequency-scale. Fig. 44 shows typical image channel selectivity curve for a communication receiver which has two radio-frequency stages. The frequency range of the receiver is 60 kHz to 30 MHz covered in six ranges. Switch positions 1, 2, and 3 cover the range 60 kHz to 1.4 MHz. These ranges are not shown on the graph as the image response ratio is greater than 100 dB at carrier frequencies less than 1.4 MHz. Ranges, 4, 5 and 6 are covered by the three remaining positions of the range switch. Inspection of the graph will show how the image response ratio falls sharply as the carrier frequency is increased. The intermediate frequency of the receiver is 465 kHz.

Selectivity

The selectivity of a receiver is its ability to discriminate between the desired signal and signals at other frequencies. It may be measured by the same method as used for sensitivity and the results of the measurements are usually given in graphical form.

For the test the a.g.c. of the receiver is disconnected and for convenience each a.g.c. line, may be taken to a temporary manual gain control. Fig. 42 shows the connexions of the equipment for the test. A modulated signal generator may be used and the sensitivity at the test frequency measured. The sensitivity to frequencies on both sides of the test frequency is then measured without altering the receiver tuning or gain controls. The results when plotted in graphical form give a selectivity curve.

The above method is not entirely satisfactory for normal receivers and is unsatisfactory for highly selective receivers. It is preferable to use an unmodulated carrier input and adjust the output of the signal generator for constant current in the diode load.

Most of the selectivity of a receiver is obtained in the i.f. stages except at low carrier frequencies, and with the exception of some telegraph receivers which may have selective a.f. circuits. The selectivity response is therefore largely intermediate frequency response except at low carrier frequencies, and it may often be more convenient to measure the selectivity of the i.f. stages directly. Great care must be taken during the test to avoid overloading any stage by too large a test signal, or by allowing a large interfering signal to reach it.

Selectivity should be measured at a number of frequencies in each frequency band and it may also be desirable to measure it at different settings of the gain controls. This serves as a check on whether the selectivity of the receiver is altered by application of the a.g.c. causing effective inter-electrode capacitances and valve characteristics to alter.

Fig. 45 shows a set of selectivity curves for a typical communication receiver. The receiver has two radio-frequency stages, one frequency-changer stage, and three intermediate-frequency amplifiers. There are five ranges of selectivity which are selected as required by a switch. The switch positions are as follows:-

	Band-pass	00 Hz not shown on graph	
2.		300 kHz curve 1 on graph	
3.		.2 kHz	2
4.		3 kHz	3
5.		6 kHz	4

The 1.2 kHz and lower bands utilize a crystal filter with an additional low-frequency filter in the 100 Hz position.

PRACTICAL SUPERHETERODYNE RECEIVER CIRCUITS

The complete basic circuits of two superheterodyne radio receivers, omitting the power supplies, are shown in Figs. 46 and 47. The receivers are designed for use on the long, medium, and short wavebands, the necessary alterations in circuit values being carried out at the switching points marked X.

The circuit shown in Fig. 46 uses a single stage of r.f. amplification which feeds into a triode-hexode frequency-changing stage. It can be seen that the frequency-changer uses a tuned-anode oscillator, such an oscillator having a good frequency stability.

The signal diode is fed from the secondary of the last i.f. transformer whilst the a.g.c. diode is fed via a capacitor from the anode of V_3.

R_1, R_2 and C_1 form the load for the signal diode; R_2 and C_2 form an intermediate frequency filter and C_3 blocks the flow of d.c. The signal voltage is developed across R_3 and applied to the grid of the triode section of V_4. The delay voltage is developed across R_4 and the a.g.c. voltage is fed via R_5.

In the circuit of Fig. 47 a radio-frequency amplifier is not used.

The incoming signal is fed to the grid of the frequency changer via coupling coils L_2 and L_3 and capacitor C_3. L_1C_1 form an effective i.f. trap against unwanted signals close to the intermediate frequency of the receiver. The aerial is tuned by the ganged aerial tuning capacitor C_{20}.

The oscillator section of the triode-hexode frequency changing stage is connected as a tuned-grid, shunt-fed circuit.

The output of the frequency changer is at the intermediate frequency of the receiver and is amplified by the i.f. amplifier before being passed to the detector stage.

The a.f. output of the detector stage is developed across volume control R_{10} and fed via capacitor C_{15} to the grid of the triode section of V_3.

The d.c. component of the voltage developed across R_{10} is decoupled by R_9 and C_{14} and fed to the control grids of V_1 and V_2 as an a.g.c. voltage.

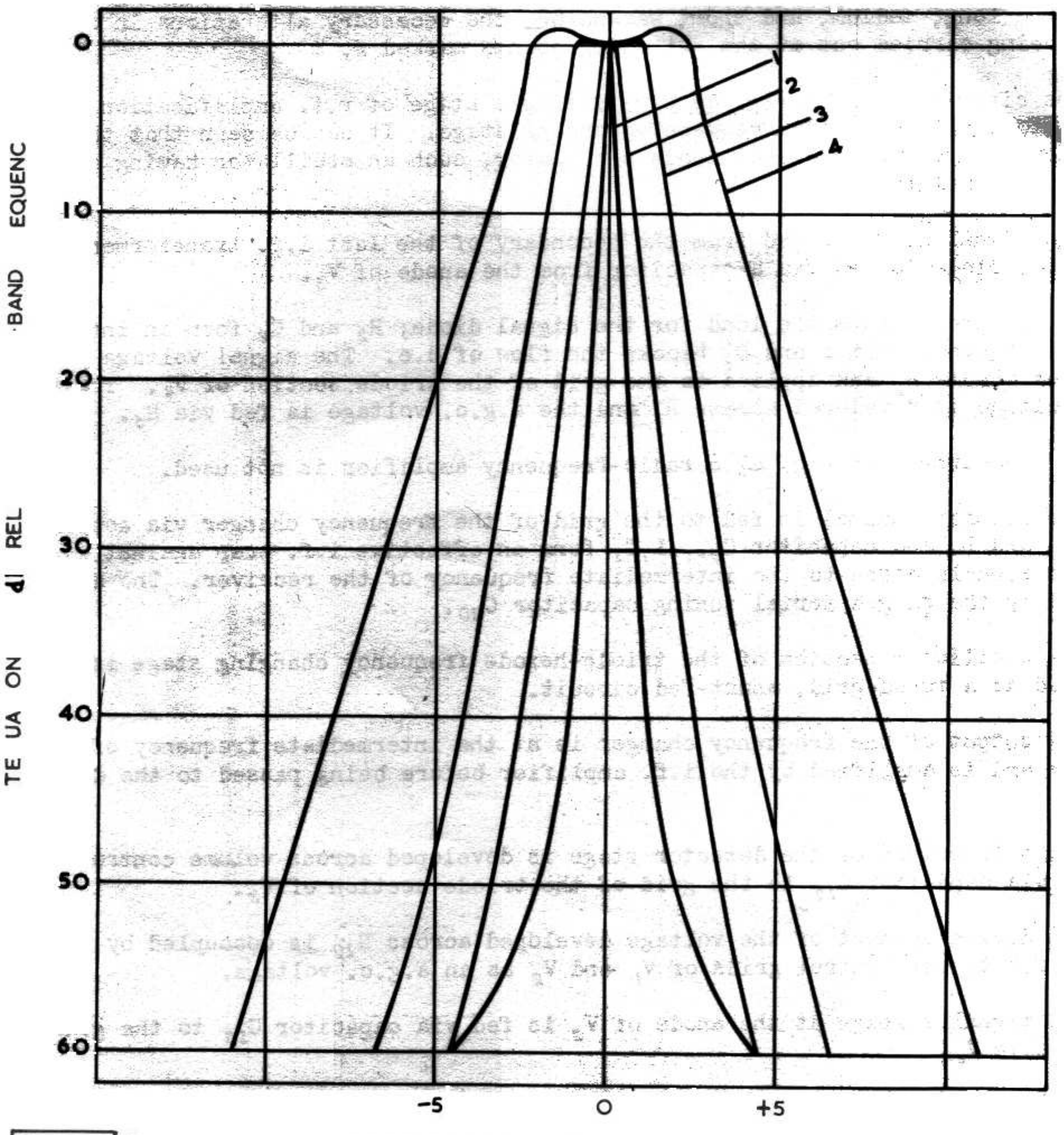
The signal voltage at the anode of V_3 is fed via capacitor C_{16} to the grid of output valve V_4.

The amplified signal at the anode of V_4 is transformer coupled to the loudspeaker.

The h.t. supply for V_4 is fed directly via L_{10b}, and the h.t. for V_1, V_2, and V_3 is fed via L_{10a} and is R.C. smoothed by R_{14} and C_{22}. The h.t. supplies are fed via the centre tap on L_{10} since this reduces the hum content to a lower level.

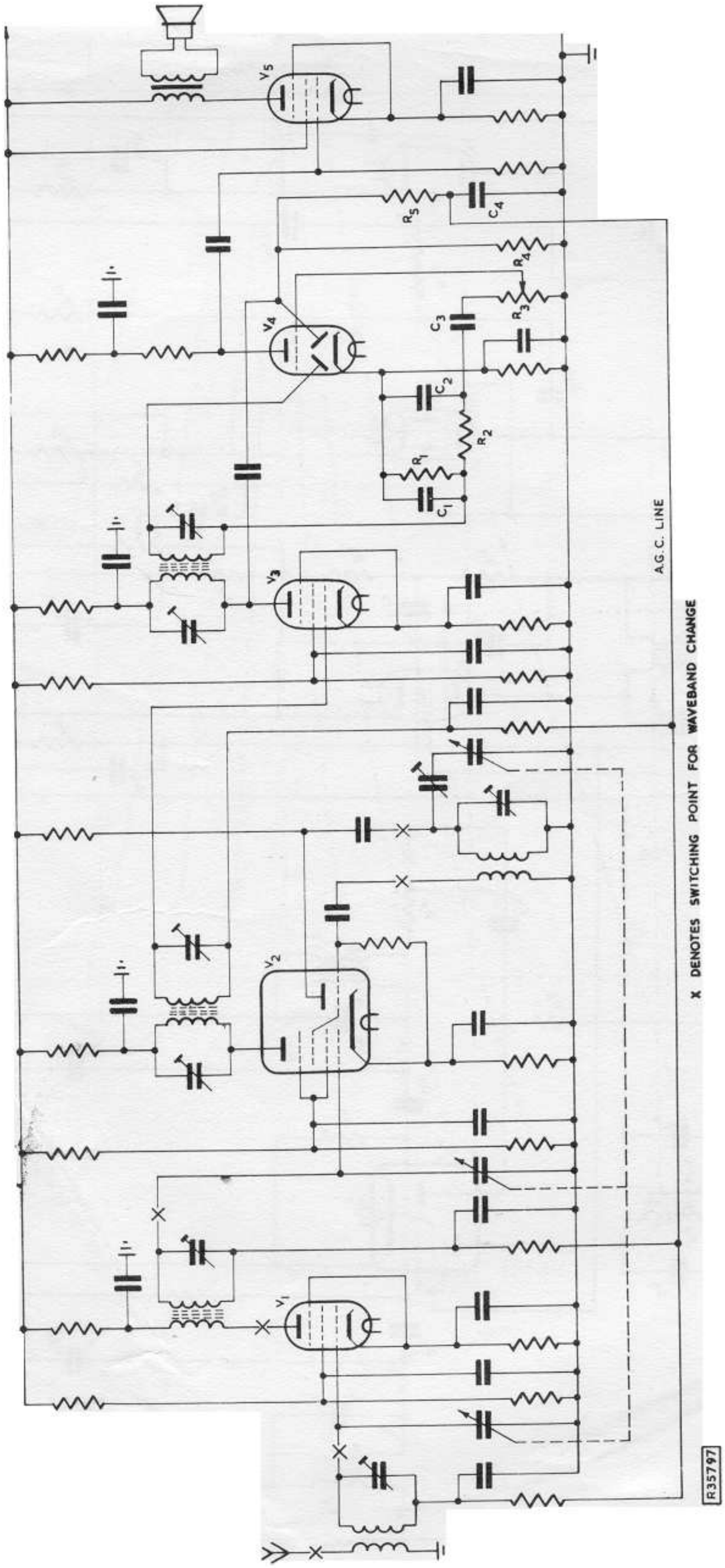
END

References: E.P. ELECTRONICS 2/2
E.P. ELECTRONICS 4/3
E.P. ELECTRONICS 2/4
E.P. ELECTRONICS 3/1



R.32936. B

Fig.


V5
V4
V3
V2
V1
R5
C4
R4
R3
C3
C2
R2
R1
C1
A.G.C. LINE
X DENOTES SWITCHING POINT FOR WAVEBAND CHANGE
R35797

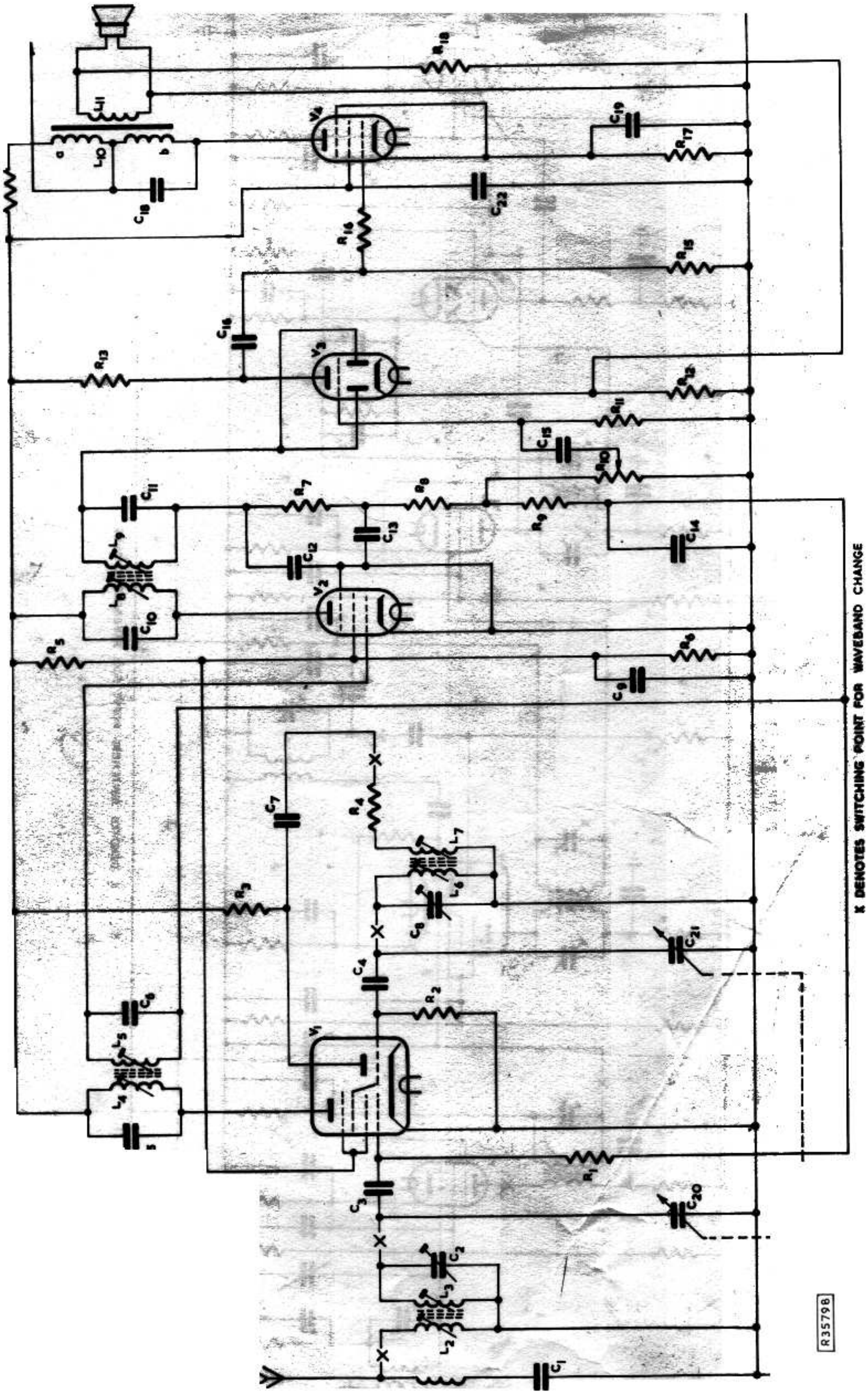


Fig 47