

## CHAPTER 36

### DESIGN OF F-M RECEIVERS

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#### SECTION 1 : F-M RECEIVERS

(i) Comparison with A-M (ii) Aerial and r-f design (iii) Local oscillator design  
(iv) I-F amplifier (v) F-M detection and A-M rejection.

##### (i) Comparison with A-M

Many of the problems encountered in F-M receiver design are new, but the majority are similar to those in A-M receivers and yield to similar treatment to that described in Chapter 35.

The advantages conferred by F-M are not so great that an obviously improved service will necessarily be available to the F-M listener unless suitable programme material is broadcast, and unless the receiver design makes full use of the increased dynamic range, frequency response and noise reduction which are possible with F-M.

To obtain the advantages of the noise reducing properties of frequency modulation it is necessary to design the F-M receiver with considerable care. If this is not done the receiver can destroy much of the improvement obtainable. In many applications F-M does offer undeniable advantages over A-M but this is not true in every case. It should not be assumed that F-M offers a cure-all for every type of noise problem, and for some communication systems A-M may be superior to F-M (see Refs. 9 and 24).

The main differences between A-M and F-M receivers are due to the higher frequencies used for the signal, local oscillator and intermediate frequencies, the need for greater i-f bandwidth, the different type of detection required and the necessity for some form of amplitude rejection. New techniques required for the higher F-M signal frequency would be equally necessary for an A-M system operating on the same frequency.

##### (ii) Aerial and r-f design

See also Chapter 22 (Aerials) and Chapter 23 (R-F Amplifiers).

(A) The inductance of a straight non-magnetic wire with a length much greater than its diameter can be obtained (Ref. 1) from the approximate formula

$$L = 0.005 l [2.3 \log_{10} (4 l/d - 0.75)] \quad (1)$$

where  $L$  = inductance in microhenrys,

$l$  = length of wire in inches,

and  $d$  = diameter of wire in inches.

The inductance of one inch of 22 A.W.G. wire is thus 0.025  $\mu$ H, and its reactance at 100 Mc/s is 16 ohms, so that when laying out circuits for use at v-h-f every lead used must be considered as an inductance. This factor, and the reduced input admittance of valves at v-h-f, are responsible for most of the problems in F-M receiver front-end design.

(B) Since at 100 Mc/s one quarter wavelength is only about 30 inches, it is quite possible to provide a **tuned aerial system** with F-M receivers. A type commonly used with console receivers is made from 300 ohm ribbon type transmission line. The usual arrangement is approximately half a wavelength long with the two wires connected together at the ends and one side cut to take the feeder. The aerial is tacked around the cabinet, on the back or inside, or it may be placed under floor coverings.

Such aerials operate quite satisfactorily at distances up to fifteen or twenty miles from a transmitter in favourable circumstances, but they are too big for mantel models. However a "**mains aerial**" (or "line antenna") consisting of a small capacitor between one side of the mains and the hot end of an unbalanced aerial primary can be used with comparable results.

For improved results in difficult locations an **outside aerial** should be erected as a matter of course, and in all cases it is advisable to investigate the effect of slight changes in position or orientation of the aerial since they may make a large difference to the signal strength. This is due to the fact that the dimensions of the waves are comparable with those of houses, rooms and even people, so that reflections occur, and standing waves exist which vary in amplitude with small changes in location. When a receiver is operated with an inbuilt aerial on low-level signals this is strikingly illustrated by signal strength variations being caused by reflection from people moving in the vicinity of the receiver.

(C) Aerial coil design is greatly influenced by the input conductance of the first valve, but in all cases it is necessary for the layout of the stage to be such as to include as much as possible of the required tuned circuit inductance in the coil. This prohibits the use of switching within the tuned circuit, and since a typical aerial coil secondary inductance might be  $0.06 \mu\text{H}$  (less than the inductance of two inches of 22 A.W.G. wire) a good system is to mount the secondary coil across the tuning condenser (if one is used) when no a.v.c. is applied to the input stage. If a.v.c. is used an assembly can be made of the tuning condenser, tuning coil and a.v.c. by-pass. In the usual case in which a 300 ohm transmission line is used, the main function of the aerial coil is impedance matching and no appreciable voltage gain is obtained. In some cases there may even be a slight loss. See Ref. 14 for calculation of aerial coil constants, although in the example a 75 ohm transmission line is used.

(D) **The valve used in the first stage** has a considerable effect upon the signal-to-noise ratio of a receiver. Although the impedance\* between the grid of the input valve and ground might be as low as say 800 ohms in a typical case, reference to Chapter 23 Sect. 6 shows that the equivalent noise resistance of the valve (probably of the order of some few thousands of ohms) is effectively in series with the input circuit and can thus be the most important source of noise in the first stage.

Also the input impedance of the first valve controls the  $Q$  (and thus the gain and selectivity) of the aerial coil. It is possible to increase this input impedance—or even make it negative—by using an unbypassed cathode resistor in the r-f stage, but inductance in the cathode circuit and inter-electrode and electrode-to-ground capacitances, which are not accurately known, make it difficult to calculate the value of resistor required. A convenient experimental procedure is to couple a signal generator to the aerial primary and to measure the voltage on the grid of the r-f valve with the heater open-circuited. When power is applied to the heater, the voltage on the grid of the valve will fall owing to its reduced (hot-cathode) input impedance. A suitable value of unbypassed cathode resistor will reduce this drop to negligible proportions.

The effect of the resistor on changes in input capacitance with changing a.v.c. voltage (see Chapter 26 Sect. 7(ii) also Figs. 23.15 and 23.16) should be checked before the value of the resistor is finalized.

(E) The main problem with r-f stages is **instability**, and apart from the causes experienced at lower frequencies, impedances common to the input and output of the

\*See Chapter 2 Sect. 8 for valve admittances.

stage become particularly important. These impedances are usually inductances and at 100 Mc/s the inductance of the rotor shaft of the gang condenser becomes sufficiently important in some cases to determine the maximum gain that can be obtained from a r-f stage. This is one reason for the use of permeability-tuned signal frequency circuits, or of loaded resonant lines with a movable short-circuiting bar (Ref. 2). Other couplings can be avoided to a large extent by returning all leads from the input circuit to one point and leads from the output circuit to a different point. If two cathode leads are available in the r-f valve they should be used for this purpose.

Coupling from output to input of the r-f amplifier can be caused by inadequate internal shielding due to an impedance between the screen grid and the cathode. This is usually due to the inductance of the leads in the screen by-pass capacitor, but this inductance can be put to good effect by series tuning it to the operating frequency. By using a small (100-500  $\mu\mu\text{F}$ ) mica capacitor with leads as short as possible, perhaps  $\frac{1}{2}$  in. in all, much better by-passing is obtained than with larger capacitors. The exact frequency of series resonance is not particularly critical. In the case of pentagrid converters of the 6BE6 type satisfactory by-passing is particularly important to prevent undesired impedances from appearing in the oscillator circuit.

Owing to the impedance at signal frequencies of the B+ and filament leads it may happen that more than one B+ by-pass is necessary. Hot spots may occur in the tuning range, and self-resonant by-passes (100-500  $\mu\mu\text{F}$  mica) should be tried at various points in B+, a.v.c. and heater wiring.

A stage gain of 10 can usually be obtained from a r-f stage without undue trouble, and higher gain (at least up to 15) is possible, particularly when there is no coupling between grid and plate circuits in the rotor shaft of a gang condenser.

### (iii) Local oscillator design

See also Chapter 24.

Local oscillator stability is important in a F-M receiver because much of the noise-reducing ability of the receiver depends on the centring of the signal on the discriminator characteristic. Detuning in excess of about 25 Kc/s can noticeably affect noise reduction and audio distortion (Ref. 11). On the other hand even a technical user of a F-M receiver can make tuning errors of this order. It is necessary therefore for the local oscillator stability to be of the order of 25 Kc/s in 100 Mc/s, i.e.  $2\frac{1}{2}$  parts in 10 000, if the receiver performance is to be unaffected.

This requirement is not easy to meet and an average figure for an uncompensated oscillator is more like 8 parts in 10 000. By keeping heat sources away from the oscillator circuit and introducing compensation after all possible precautions have been taken, a long and short term stability of 2 parts in 10 000 is possible.

Additional oscillator stability can be obtained by using automatic frequency control if the local oscillator is a separate valve, and if suitable control voltages can be obtained from the detector circuit (Refs. 10 and 21). Improvements in oscillator stability of five to ten times are possible, but care must be taken to see that the a.f.c. will allow the receiver to tune from a strong local station to a weak one in an adjacent channel.

Some designs have used permeability-tuned-oscillator circuits to eliminate the temperature instability of a tuning condenser (Ref. 11). Harmonic operation has also been used and gives a reduction in frequency shift due to capacitance variations when a larger tuning capacitor is used to tune to the lower oscillator frequency.

Harmonic operation has the added advantage of greatly increasing the separation between the oscillator and signal frequencies and it thus improves oscillator frequency stability in cases where pulling is experienced between oscillator and signal frequency circuits.

For other methods of frequency stabilization see Chapter 24 Sect. 5.

Grid-tuned and plate-tuned oscillators are not commonly used in the F-M band because the feedback winding becomes comparable in size with the tuned winding.

This leads to tracking troubles and the possibility of an oscillator changing from grid tuning to plate tuning or vice versa in a particular band.

Both Colpitts and Hartley circuits are used, the Hartley (Fig. 24.4B) probably because it is widely used on the broadcast band with 6SA7 type converters. However unless the receiver has only one waveband, switching of the cathode lead is necessary which introduces undesirable impedance between the cathode and the tap on the coil and may lead to parasitic oscillation. Added to this the Hartley circuit is in any case more prone to parasitic oscillation than the Colpitts, and is also liable to change over to a Colpitts type using valve interelectrode capacitances to give a tapping point in the oscillator circuit.

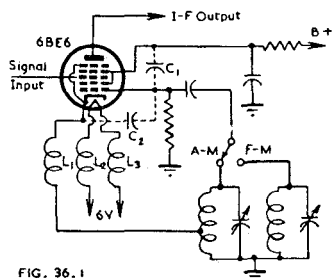


FIG. 36.1

Fig. 36.1. Simplified switching from Colpitts high frequency oscillator to Hartley medium wave oscillator.

Because of this, the Colpitts oscillator is popular in F-M receivers, and a convenient arrangement for a two band receiver is shown in Fig. 36.1 (from Ref. 16). The inductance of the choke  $L_1$  is small enough to be negligible on the broadcast band so that it is only necessary to switch one connection to the oscillator circuit to change from the F-M to the broadcast band. The ratio of internal capacitances between screen grid and cathode and between oscillator grid and cathode determines the amount of feedback, and, if desired, external capacitances can be added to modify the feedback and increase the grid current, as shown ( $C_1$  and  $C_2$ ) dotted in the circuit. However, these capacitors also have an effect on the signal frequency circuit and sensitivity and signal-to-noise ratio should be checked at the same time as the oscillator grid current. It may be necessary to compromise between maximum sensitivity, maximum signal-to-noise ratio and maximum oscillator grid current.

The chokes  $L_2$  and  $L_3$  comprise the two remaining sections of a trifilar choke made by twisting together three lengths of enamelled 28 A.W.G. copper wire and then winding 15 turns, closely-spaced, on a  $9/32$  in. former,  $\frac{1}{2}$  in. long. The sections  $L_2$  and  $L_3$  introduce impedance between the heater of the oscillator valve and ground so that variations in heater-cathode capacitance do not appear across a part of the tuned circuit. This is a recommended precaution against microphony and may improve frequency stability during the heating period (about five minutes) of the mixer.

Care is needed in the local oscillator design to avoid microphony and modulation hum since any a-f variation in oscillator frequency results in an equivalent output from the frequency discriminator. A check on modulation hum produced by the local oscillator can be obtained by comparing the hum in the receiver output when intermediate frequency and signal frequency inputs are fed into the converter grid.

#### (iv) I-F amplifier

See also Chapter 26.

When the maximum  $Q$  allowable by bandwidth considerations is used in a F-M transformer the gain obtained with the types of valves currently used approaches the maximum gain possible without an undesirable amount of regeneration. Usually there are three valves amplifying at the intermediate frequency and the over-all gain is greater than 10 000 times, so that regeneration problems are common.

A convenient method of tracing regeneration is described in Ref. 1. It consists of feeding the i-f signal to the hot side of the plate circuit of the second last i-f amplifier by means of a very small capacitor ( $3 \mu\mu\text{F}$  or less), retrimming the circuit, and converting the detector to an A-M type by open-circuiting one diode and replacing it with an appropriate capacitor to keep the circuit aligned. When A-M signals are fed into the amplifier the output will be due to the generator and to any regenerative effects from earlier stages which may be present. If the receiver output is decreased

by short-circuiting the input of the second last i-f amplifier regeneration exists and various parts of the receiver can be short-circuited to trace it.

The method has the advantage of making it possible to deal with regenerative signals alone in the first stage instead of regenerative plus desired signals as when tracing regeneration by more conventional means. In addition the A-M signal generator can be set to various frequencies in the passband of the i-f amplifier to trace regeneration which may be more severe at some frequency other than the centre of the pass-band.

When regeneration occurs due to plate-to-grid capacitance in a single valve the circuit of Fig. 36.2A (Ref. 5) provides economical neutralization as illustrated in Fig. 35.2B. Phase shifts and stray inductances complicate the circuit sufficiently to make the calculation of  $C_n$  difficult but an average value is about  $3000 \mu\mu\text{F}$  and correct neutralizing is obtained when, using the tracing method just outlined, no change in output occurs as the input transformer is tuned through resonance by  $C_2$ .

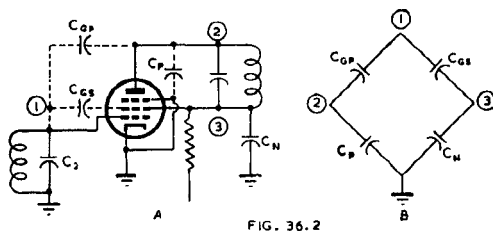


Fig. 36.2. Circuit for i-f amplifier neutralization (A) Actual circuit (B) Capacitance rearranged to show bridge (Ref. 5).

**Series resonance** can be obtained in i-f by-passes by allowing half-inch leads with typical  $0.01 \mu\text{F}$  paper capacitors. Larger values may give less by-passing and once the self-resonant frequency is passed the effective series impedance increases more rapidly than the frequency at which the by-pass is used.

The i-f amplifier design can be responsible for poor noise rejection on the part of the receiver unless adequate stability is provided against the effects of time, temperature, moisture and varying signal strength. The detuning of one or more i-f tuned circuits from the centre point of the discriminator characteristic will result in decreased noise rejection whether or not a limiter is used (Ref. 3).

**Stability versus time characteristics** are usually satisfactory when good trimmers are used, but compression type mica trimmers may give trouble, particularly when the effects of vibration in transit between factory and the user's home are considered.

**The effects of temperature and humidity** can be offset by suitable impregnation methods for coils and the use of low temperature coefficient condensers to tune the transformers. Silvered-mica types are satisfactory as are most air-dielectric trimmers.

**Frequency instability with varying signal inputs** applies to individual circuits rather than to the i-f amplifier as a whole and can be caused by a decrease in regeneration as gain is reduced by a.v.c. application or by valve input capacitance changes due to a.v.c. applied to i-f amplifiers or to sudden changes in the amplitude of the applied signal in the case of limiters. One method of overcoming the latter trouble is to use resistance coupling between the last i-f amplifier and the limiter (Ref. 21), although this can only be done at the expense of gain and selectivity.

When impulse noise is present the detuning with each pulse may result in phase distortion of the signal which will no longer be centred on a symmetrical pass-band. The noise itself will also be heard and under extreme conditions the effect of a limiter can be to increase noise, as the increase due to mistuning more than offsets any other effect of the limiter (Ref. 3).

In the absence of noise, detuning due to a.v.c. application decreases the amplitude rejection of the receiver and may increase phase distortion.

On the other hand, some detuning due to a.v.c. might be preferable to the amount of detuning caused by overloading in the grid circuits on strong signals. Resistance-capacitance networks having suitable time constants are often used to provide additional self bias when the signal strength is high. This system is usually preferred to normal a.v.c. application.

To minimize detuning effects the i-f amplifier must be non-regenerative and the capacitors on the secondary—at least—of limiter input i-f transformers should be increased as much as the consequent reduction in gain will allow. If the primary can conveniently be wound with a different number of turns from the secondary the primary inductance may be increased when the secondary inductance is reduced and gain will be unchanged if  $\sqrt{L_p L_s Q_p Q_s}$  remains constant. This should not be carried to the point where detuning of the primary circuit becomes noticeable either when valves are changed or due to differences in wiring capacitances between different chassis.

In addition, an undue increase in  $L_p$  can cause the grid-to-plate gain to become excessive, resulting in instability.

The same precautions can be taken with the input circuits of stages to which a.v.c. is applied and in addition an unbypassed cathode resistor can be used to reduce input capacitance changes. Figures 23.15 and 23.16 show the effect of varying values of unbypassed cathode resistors on the change in input capacitance of type 6AC7 and 6AB7 valves. For type 6BA6 a common value of resistor is 68 ohms.

### (v) F-M detection and A-M rejection

There are three main methods of performing the functions of F-M detection and A-M rejection in commercial F-M receivers, the limiter-discriminator combination, the ratio detector and the locked oscillator. See Chapter 27 Sect. 2 for design details and Ref. 8 (from which some of the following information is taken) for a consideration of the capabilities of each. The effect of the type of detector used on over-all receiver performance can be summarized as follows.

(a) Assuming reasonably high a-f sensitivity, the use of a **ratio detector** can save one valve from the number required for a receiver using a limiter and discriminator or a locked oscillator.

(b) The downward A-M handling capability of the ratio detector is less than that of the other two types. For the ratio detector an average figure is about 60% for medium and large inputs whereas the other types rapidly approach 100% once their threshold input is passed. Serious distortion occurs in a ratio detector if the downward amplitude modulation is excessive since this stops the diodes from conducting (see Chapter 27 Sect. 2).

(c) The ratio detector gives some A-M rejection at very small inputs whereas the limiter discriminator combination has a threshold (the beginning of limiting) below which there is no A-M rejection. The locked oscillator also has a threshold—the input required to lock the oscillator. As a result, interstation noise levels on a receiver using a ratio detector are noticeably low for a given sensitivity.

For medium and high inputs all three types reduce amplitude modulation to a small percentage of the original, with the proviso in the case of the ratio detector that the downward amplitude modulation must not exceed about 60%.

An amplitude modulated signal tuned to the mid-point of a balanced discriminator will not give any a-f output, and in such a case the A-M rejection of the discriminator alone is 100%. However, if the signal is not centred on the discriminator characteristic there will be some a-f output.

Even 100% A-M rejection will not remove noise from a F-M carrier. If the carrier is considered as a rotating vector and at a particular instant a noise carrier of twice the amplitude and with a 180° phase difference is added to it, the resulting carrier will be unchanged in amplitude but rotated through 180° by the noise. Thus, phase modulation (and so equivalent frequency modulation) of the desired carrier has occurred and noise must appear in the output of the detector. Since the amplitude is unchanged there will be no limiting. In practice, both amplitude and frequency modulation of a desired carrier occur, and although the receiver may have 100% A-M rejection the effect of the frequency modulation cannot be eliminated.

(d) The output of a ratio detector is proportional to the average amplitude of its input so that some form of a.v.c. is desirable. With the other two types output is almost independent of input above the threshold level, although some form of a.v.c. is often used to prevent overloading of i-f stages and the consequent detuning.

(e) With each type there is a side tuning response when the signal is tuned to the skirt of the i-f selectivity curve. However in the case of the ratio detector this is at a much lower level than the main response and tuning is not a great deal more difficult than with an A-M receiver.

**The limiter discriminator circuit** gives distorted side responses at a level at least equal to the correct tuning position and tuning is not easy for non-technical users, unless a suitable tuning indicator is incorporated in the receiver.

**The locked oscillator circuit** may produce severe distortion as the signal crosses the locking threshold in the side tuning positions, or stations may tune with a definite "plop" as the oscillator locks.

If a.f.c. is used it is possible to arrange for the receiver to be pulled rapidly through the side responses by the a.f.c., thus minimizing their effect.

In general, the saving of a valve and the desirable tuning characteristic of the ratio detector make it attractive. However more care is needed in design to obtain correct balance between the two diode circuits for maximum noise rejection, and the limited downward modulation handling capability means that the selectivity of the i-f amplifier must be closely controlled in production. If this is not done, excessive selectivity due to decreased coupling may introduce more amplitude modulation than can be handled by the ratio detector (even without the possible addition of noise) resulting in severe distortion.

**The standard de-emphasis characteristic** required has a falling characteristic with increasing frequency and is equivalent to that provided by a simple circuit having a time constant of 75 microseconds. The components used to filter i-f voltages from the detector circuit are given suitable values to provide this de-emphasis. Fig. 15.1 gives the standard curve.

**Valves** designed especially for limiting and F-M detection are available, namely the gated beam discriminator type 6BN6 and the Philips " $\phi$ -detector" type EQ80.

In the 6BN6 two grids are operated in a quadrature phase relationship at centre frequency and the phase of one grid voltage is caused to vary with respect to the other about the quadrature point as the frequency is varied. The quadrature voltage is developed by space charge coupling to a special grid to which a parallel tuned circuit is connected. Limiting is obtained by controlling the current of an electron beam by an apertured slot located in the beam in a region of high current density.

Typical performance of the 6BN6 (Refs. 25, 26) is that with a F-M intermediate frequency of 10.7 Mc/s the output voltage for 75 Kc/s deviation is about 4.5 volts r.m.s. Maximum A-M rejection with 30% A-M and 30% F-M modulation applied simultaneously is about 35 db. The A-M rejection with respect to signal input varies but at least 20 db of A-M rejection is obtained with 1 volt input and from 15 to 30 db at signals above 1.25 volt r.m.s. An a-f output of 3.7 volts with less than 1% distortion can be obtained using accelerator and plate supply voltages of 60 and 80 volts. Higher voltages and correspondingly higher plate loads will give an a-f output up to 15 volts r.m.s. for 75 Kc/s deviation which is sufficient to drive the output valve directly.

The EQ80 (Ref. 27, from which the following information is taken) contains seven grids of which the second, fourth and sixth grids are screen grids and the seventh is a suppressor grid. To each of the control grids a voltage of at least 8 V r.m.s. is applied from an i-f transformer. The mean value of the anode current is a function of the phase shift between the two control voltages whilst the phase shift is a function of the frequency deviation. Both functions are approximately linear when the phase shift has a sweep between 60° and 120°.

The amplitude of the anode current is not dependent upon the magnitude of the control voltages (provided they are greater than 8 V) so that the valve also acts as a limiter. The  $\phi$ -detector gives an a-f output of about 20 to 25 volts. The first grid can be arranged to cut off the cathode current if the control voltages are not large enough, so that interstation noise is suppressed.

## SECTION 2 : F-M/A-M RECEIVERS

(i) R-F section (ii) I-F amplifier (iii) General considerations.

### (i) R-F section

The valve complement and other major features of a combination F-M/A-M receiver are determined by the F-M requirements almost exclusively, and the A-M receiver operates with a suitable selection of the parts already in use.

There are of course some additions, and the A-M tuning device is one. The two tuning systems may use separate gang condensers or separate permeability tuning arrangements or a combination of the two. There are two well-known systems of tuning the two bands with one control, one using a tuning condenser with A-M and F-M sections on the same rotor shaft, and the other with a cam on the rotor shaft of a normal A-M type gang, the cam operating slugs in the F-M coils.

On the F-M band, tuning should be accurate within about 20 Kc/s in a band of 20 Mc/s—one part in one thousand—which is comparable with a required A-M setting accuracy of about 1 Kc/s in a band of about 1000 Kc/s. Thus a tuning rate which is satisfactory for the A-M broadcast band should be suitable for the F-M band.

Much of the ingenuity in combined F-M/A-M design is directed towards the elimination of switching, particularly in hot circuits. This is not so much to save the cost of the switching but to avoid the losses in leads to switch contacts and in the additional stray capacitances which are introduced.

One example of this can be seen in Fig. 36.1 where one switch section is saved by converting the oscillator from a Hartley on A-M to a Colpitts on the F-M band. Fig. 36.3 (taken from Ref. 10) shows a system of combining short wave A-M input with F-M band input without switching and without undue loading of one circuit by the other.

Untuned r-f amplification also saves switching, and choke coupling can conveniently be used or a mixture of choke feed for the r-f amplifier on the F-M band and resistance feed on the A-M band. In this case there is of course a considerable loss in image rejection compared with that obtainable with tuned r-f amplification.

Other systems which have been used include a circuit in which the r-f stage is used only on the F-M band with an A-M loop switched to the converter grid for A-M reception, and a more complicated one (Ref. 4) in which the F-M signal is converted by a 12BE6 to the 10.7 Mc/s intermediate frequency and is then amplified by a second 12BE6 operating as an i-f amplifier. On A-M the second 12BE6 becomes the converter by comparatively simple switching, and an A-M loop is connected to its grid while the F-M second i-f amplifier becomes the single A-M i-f amplifier. The first 12BE6 is not used in A-M reception.

### (ii) I-F amplifier

As in the r-f section of the receiver, it is important to avoid switching hot circuits in the i-f amplifier, and the fixed tuning and large frequency difference between 10.7 Mc/s and 455 Kc/s fortunately makes a simple series connection of the two i-f's quite satisfactory (see Chapter 26 Sect. 3).

One trouble experienced with this system is that when operating on A-M, spurious responses (e.g. oscillator harmonics) can set up large signals across the 10.7 Mc/s tuned circuits and may overload later i-f amplifiers, interfere with a.v.c. arrangements

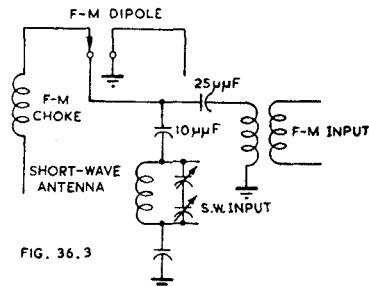


FIG. 36.3

Fig. 36.3. Combined F-M band and A-M short wave band input circuit (Ref. 10).



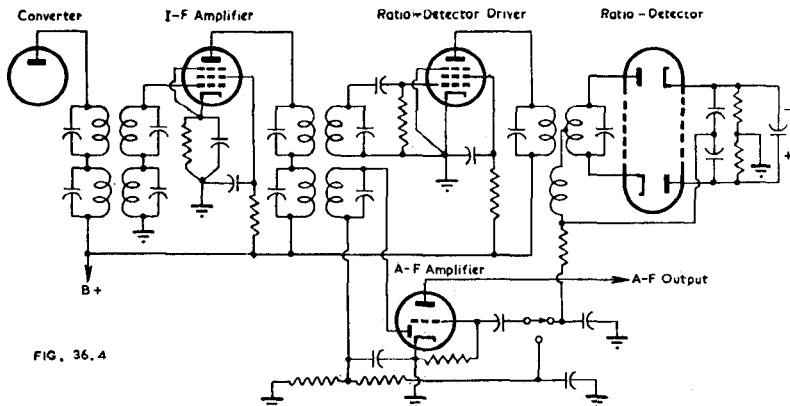


FIG. 36.4

Fig. 36.4. Combined F-M A-M i-f channel using one A-M i-f amplifier and two F-M i-f amplifiers.

or cause other trouble depending on the circuit of the subsequent sections of the receiver. To minimize this trouble it is common practice to switch the converter output to the appropriate (10.7 Mc/s or 455 Kc/s) winding. In such a case the secondary and all other i-f windings (except the output windings) would be series connected. Separate converters, each connected only to its own i-f, have also been used. This eliminates the spurious response trouble, allows very simple switching between bands (only the converter B supply need be switched) and allows more flexibility in the F-M oscillator design.

When using a ratio detector, two i-f amplifiers are needed, and with a limiter-discriminator combination an extra valve is required. As against this the i-f amplifier for an A-M receiver needs only one valve, particularly if of the high slope type commonly used as F-M i-f amplifiers.

One method of obtaining an extra F-M i-f amplifier was mentioned in sub-section (i) the case in which the A-M converter was switched to carry out this function. However, the more usual arrangement is shown in Fig. 36.4 for the case of ratio detector F-M receiver. A combined detector of the triple-diode-triode type could be used to save one valve, but the main point is that the last A-M i-f is not connected in series with the F-M winding, to avoid the necessity for switching a high level i-f circuit.

F-M detection can be carried out by tuning the F-M signal to the skirt of the i-f response and applying the resultant amplitude and frequency modulated wave to a diode. The amount of A-M introduced can be high (see Fig. 26.6) so that a few times greater detection sensitivity is available than from a conventional discriminator. This is obtained at the expense of distortion, the degree of which is dependent on the degree of non-linearity of the i-f skirt characteristic, and the circuit also gives two tuning points. Tuning must be for minimum distortion which is not necessarily obtained at the same tuning point as maximum volume or maximum signal-to-noise ratio. An additional disadvantage is that the circuit has no inherent amplitude rejection properties, as has a discriminator, although a limiter could be used.

### (iii) General considerations

The problem of providing comparable a-f output on F-M and A-M has different aspects depending on the type of detection used in each case

With a ratio detector the a-f output is approximately proportional to the r-f input so that some form of a.v.c. is needed, as with an A-M diode detector. The direct voltage generated by the detector provides an excessive amount of control when applied to the grid of the i-f amplifier unless high a-f gain is provided or unless delay is used. The circuit of Fig. 27.38B can be used for delay if a diode is available.

However "partial limiting" as discussed in Chapter 27 Sect. 2(iv)C provides a satisfactory a.v.c. characteristic and at the same time removes the need for any other means of A-M detection since if suitable constants are used this is carried out satisfactorily in the control grid circuit of the driver of the ratio detector. The direct voltage available from this circuit can also be used as an a.v.c. voltage for previous stages (on A-M or F-M).

If about one third of the a-f voltage developed in the "partial limiter" grid circuit is used on A-M, and the full ratio detector output on F-M, similar a-f output will be obtained from the receiver over a range of input signals with each system.

A small discriminator bandwidth (some commercial designs have a width of only 200 Kc/s) will increase the a-f output, which would allow a larger fraction of the A-M output to be used, and increase the output available from small signals with each system.

In the case of a limiter-discriminator combination having A-M taken from the input to the limiter stage, the a-f output is high and almost constant on F-M but is much less with small inputs on A-M. A poor A-M a.v.c. characteristic which allows the a-f output to increase rapidly can provide approximately equal F-M and A-M output when reasonably large A-M signals are received.

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