vhf-uhf
BY G.R.JESSOP, C.Eng., MIERE, G6JP

manual

RSGB

21/-

Ex-Libris

Arthur Missira

MCMXCVII
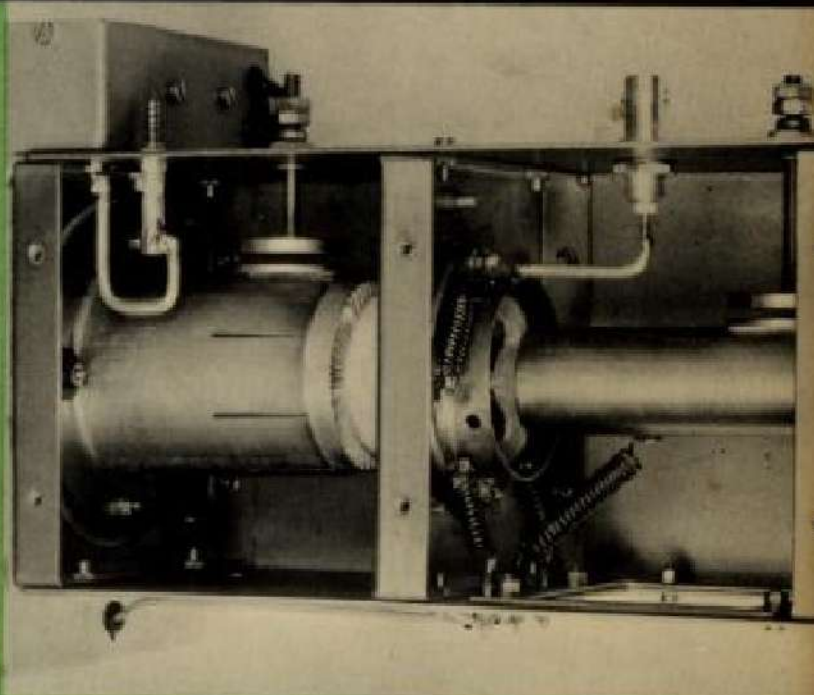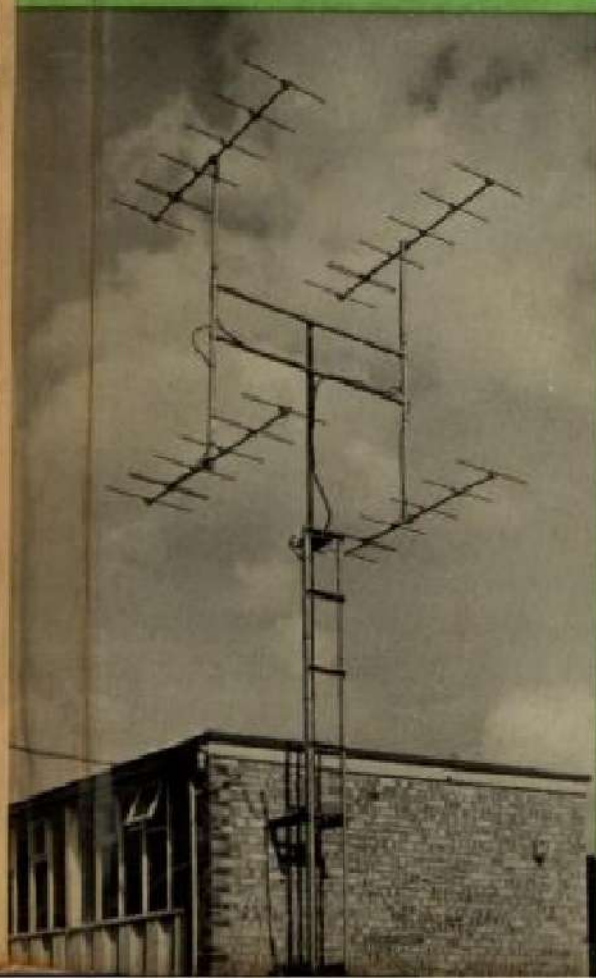
VHF – UHF MANUAL

VHF - UHF MANUAL

by G. R. Jessop, C.Eng., M.I.E.R.E.
G6JP

ACKNOWLEDGEMENTS

CONTENTS

PREFACE

The radio amateur has in recent years shown an increasing interest in communication on frequencies in the v.h.f. and u.h.f. region. Activity on 4m, 2m, 70cm and 23cm has increased rapidly and many operators from the lower frequencies are moving into this part of the radio spectrum, either as an alternative to the crowded h.f. bands or because of the great interest in the techniques involved. This very desirable increase of interest appears to appeal to both experienced and newly licensed operators. The introduction of the Class B licence for telephony only on 144 MHz and above has attracted to v.h.f. many of the younger generation of amateurs.

A great deal of the equipment in use is home constructed, partly due to the considerable developments taking place on the v.h.f. and u.h.f. bands, and because of this it is felt that a manual devoted entirely to equipment suitable for use in this part of the frequency spectrum will find a ready acceptance.

Considerable information already exists in other more general handbooks, such as the RSGB *Radio Communication Handbook* but these must of necessity give somewhat limited coverage to equipment for any particular section of the radio frequencies available to the amateur.

The purpose of this specialised manual is to give a wide range of material for each of the bands, so the constructor has a selection of designs from which to choose and to suit his own individual need, together with a suitable amount of basic information on general matters.

The author's interest in v.h.f. started in the far-off days of 5 metres when together with two other amateurs, 5CV and 6YK in West London, he set about getting going on the then new band of 56 MHz.

From these seemingly far-off days, the rapid growth of demand for equipment for frequencies above and beyond 56 MHz, which occurred during the World War II, has continued rapidly ever since. The amateur may not so easily do things that have not been done before, nevertheless there are still many refinements that were found as a better way of doing the job and quite a few of the circuit arrangements from amateur designs have found their way into modern commercial equipment.

The new components and devices now available to the amateur enable him to construct highly efficient and refined equipment for communication and to continue the study of propagation problems.

G.R.J.

INTRODUCTION

The problems of communication increase as the frequency is raised and the range of v.h.f. and u.h.f. signals is relatively short by comparison with the h.f. bands, no round-the-world contacts are possible unless aided by satellite or reflected from the moon.

Nevertheless contacts over extended ranges are often possible, up to around 1000 km and beyond under favourable propagation conditions. Highly directional aerials are readily erected in almost any area and the effective gain, and therefore the radiated power, can be very much greater than on the h.f. bands.

Before considering any of the problems that arise in this most interesting frequency region let us first look at the bands and space available for operation.

Band	Frequency Range	Maximum d.c. power input	R.f. output p.e.p. for A3A and A3J only
4 metres	70.025 - 70.7 MHz band width = 675 kHz	50 watts	133 watts
2 metres	144 - 146 MHz band width = 2000 kHz	150 watts	400 watts
70 cms	425 - 429 MHz and 432 - 450 MHz band width = 22 MHz	150 watts	400 watts
23 cms	1215 - 1325 MHz band width = 110 MHz	150 watts	400 watts
13 cms	2300 - 2450 MHz band width = 150 MHz	150 watts	400 watts

The types of emission permitted on all these frequencies are:

AMPLITUDE MODULATION

A1 Telegraphy by on-off keying, without the use of a modulating tone.
A2 Telegraphy by on-off keying of an amplitude - modulating audio frequency or by the on-off keying of the modulated emission.
A3 Telephony double sideband.
A3A Telephony single sideband, reduced carrier.
A3H Telephony single sideband, full carrier.
A3J Telephony single sideband, suppressed carrier.

FREQUENCY (OR PHASE) MODULATION

F1 Telegraphy by frequency shift keying without use of a modulating tone, one of two frequencies being emitted at any instant.
F2 Telegraphy by on-off keying of a frequency modulating tone or on-off keying of a frequency modulated emission.
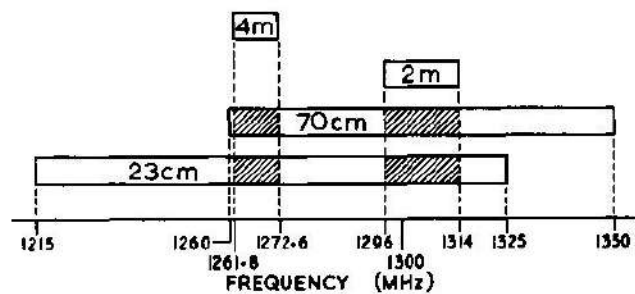F3 Telephony.

Fig. 1.1. Relationship of the v.h.f. and u.h.f. bands.

The above chart clearly shows that although the 4m band is not harmonically related to all the other bands nevertheless it is suitable for multiplying into the 70 and 23cm bands. Normally, however, this band is regarded as the odd band in this respect and most operation is derived from 2m exciters.

The 4m band has many attractive characteristics, it is, however, unfortunately not yet allocated on an international basis and only the U.K. and Rhodesia have bands in this portion of the spectrum. There are possibilities of cross-band transatlantic contacts with the U.S.A. and Canada who have the 6m band. It is also a very reasonable band for mobile operation where a standard quarter wave whip aerial is of convenient size and the average propagation characteristics ensure a satisfactory communication range.

Band Plan

With the rapidly increasing use of these v.h.f. and u.h.f. bands, by both newly licensed and experienced operators, some planning is desirable in order to

prevent undue and often unnecessary, interference occurring.

So far the activity on 23cm has not justified planning but generally since the 2m, 70 and 23cm bands are harmonically related, the fact that plans exist for 2m and 70cm almost certainly ensures that operation in the 23cm band will follow that used on the lower frequencies.

V.H.F. BAND PLANS

All v.h.f. operators are reminded of the British Isles Two Metre and Seventy Centimetre Band Plans, which are sponsored by RSGB. Observance of these plans will assist in DX working and in avoiding interference to certain frequencies in the 144-145 MHz band.

Zone	2 metres	70 cm.	
1	144.0 - 144.1	432.0 - 432.1	C.W. Only (Nationwide).
2	144.1 - 144.25	432.1 - 432.25	Berkshire, Cornwall, Devon, Dorset, Hampshire, Somerset, Wiltshire, Channel Isles.
3	144.25-144.5	432.25-432.5	Brecon, Cardiganshire, Carmarthenshire, Glamorganshire, Gloucestershire, Herefordshire, Monmouthshire, Pembrokeshire, Radnorshire, Worcestershire.
4	144.5 - 144.7	432.5 - 432.7	Kent, Surrey, Sussex.
5	144.7 - 145.1	432.7 - 433.1	Bedfordshire, Buckinghamshire, Essex, Hertfordshire, London, Middlesex.
6	145.1 - 145.3	433.1 - 433.3	Cambridgeshire, Huntingdonshire, Leicestershire, Norfolk, Northamptonshire, Oxfordshire, Rutland, Suffolk, Warwickshire.
7	145.3 - 145.5	433.3 - 433.5	Anglesey, Caernarvonshire, Cheshire, Denbighshire, Flintshire, Merionethshire, Montgomeryshire, Shropshire, Staffordshire.
	145.41	433.41	Single Sideband Spot Frequency (Nationwide).
8	145.5 - 145.8	433.5 - 433.8	Derbyshire, Lancashire, Lincolnshire, Nottinghamshire, Yorkshire.
9	145.8 - 146	433.8 - 434	All Scotland, Northern Ireland, Isle of Man, Cumberland, Co. Durham, Northumberland, Westmorland.

Two Metre Band Channels: The following frequencies in the 144-145 MHz portion of the 2 metre band are tabulated on the schedule to the Amateur (Sound) Licence to be avoided as they are allocated to aeronautical use: 144.0, 144.09, 144.18, 144.27, 144.36, 144.45, 144.54, 144.63, 144.72, 144.81, and 144.9 MHz. Remember! The safety of aircraft and human lives depend upon the interference-free use of these channels.

CHAPTER 2

PROPAGATION

On the frequencies covered by the v.h.f. and u.h.f. bands the propagation characteristics are different from those met with on the h.f. bands. Good conditions are much more likely to be directly related to the local weather prevailing at and between the stations at the time of the contact.

On the lower bands, particularly 70 MHz and to a lesser extent 145 MHz, Sporadic E occasionally provides exceptional ranges, but generally normal ranges are confined to ground wave and are much affected by the terrain near to the station. There is no "sky wave" in the usual sense as experienced on the bands below 30 MHz.

Since ground wave working is the usual condition the higher the location and the higher the aerial above ground the better. In urban areas where large steel frame buildings are frequently encountered it is often best to make use of these obstacles as large reflectors rather than direct aerials away from them. They do give a considerable screening effect.

In the following sections of this chapter many of the different forms of propagation are covered.

TROPOSPHERIC PROPAGATION OR

V.H.F. WEATHER

It is well known that there is a close connection between "the weather" and one's ability to exchange signal reports with distant friends well beyond the horizon, and it is common knowledge, too, that you stand a better chance of working DX when the pressure is high and the weather is settled, than when it is blowing half a gale. But it soon becomes equally obvious that not every anticyclone, however fine the weather, brings about a period of anomalous propagation, and that even when such conditions exist, not everyone enjoys equal benefit. Occasionally, too, a long-range contact is made when the pressure is comparatively low and it would seem likely that we cannot hope to explain what is happening in terms of atmospheric pressure alone.

What we must do, in fact, is to find out what is happening in the atmosphere between the transmitter and the receiver at a time when conditions favour long-range transmission, and for this we need a function which not only shows the meteorological processes involved but also what is happening to the radio wave as it travels along the path.

Potential Refractive Index

In two articles on "V.H.F. Weather", R.G. Flavell, G3LTP, gave a full explanation of the potential refractive index method as applied to tropospheric wave propagation.[1]

Its major advantage is that it enables us to see a vertical section of the atmosphere over a chosen point with, and this is the real crux of the matter, most of the factors that do not affect propagation removed. This is accomplished by taking the radiosonde ascent information at various heights, and adiabatically bringing these to a standard level - say 1000mb. These "parcels" of air from various heights can therefore be seen in refractive index comparison.

The method has a limitation. Since the adiabatic lapse rates are different for wet and dry air, if large vertical sections are plotted through regions of condensation, that is through cloud, this would introduce difficulties. Normally, under anticyclonic conditions, medium and high level cloud is dried out by the subsidence, and low level cloud at the point of density balance is relatively thin - between 100 - 200 feet thick - and so the regions of vertical condensation are usually quite small.

If we plot the p.r.i. as a time section at midnight and midday vertically over the chosen station for a number of days, we would find that as conditions improved, the p.r.i. lines on the time section fall and crowd together indicating the point or layer where the greatest refraction is taking place.

Alternatively, we could plot a Delta K map taking a number of radiosonde ascents at the same time from different stations and noting the p.r.i. number at the 850mb level, and subtracting this from the surface level reading. We would end up with a series of numbers which could be plotted geographically and linked by isopleths to produce a propagation map.

Radio-Refractive Index

We are all familiar with the effects of light waves passing from one transparent medium to another of different refractive index, even if our experience is limited to the sight of a stick apparently bent as it passes into a bucket of water. A mirage is another manifestation of changes in refractive index, and this example is the counterpart in terms of light waves.

At frequencies above those returned by the ionosphere the path taken by a radio wave is determined by the variations in the radio-refractive index of the air it encounters. The actual changes in magnitude which are sufficient to make all the difference between normal and abnormal conditions are extremely small, and it is customary to express radio-refractive index in "N-units" which represent the excess over unity in parts per million.

The radio-refractive index of a given sample of air may be calculated from measurements of pressure, temperature, and some function of the water-vapour content, such as vapour pressure or dew point, by the use of the expression:

$$N = \frac{77.6\,P}{T^2} + \frac{3.733 \times 10^5\,e}{T}$$

where T is the absolute temperature, P the atmospheric pressure in millibars, and e is the partial pressure of water-vapour, also in millibars.

Although this is the form of refractive index necessary for mathematical calculations of ray-bending, it is not the most suitable for graphical or pictorial analysis. One reason for this is that the normal changes of N with height tend to obscure the abnormalities which give rise to anomalous conditions.

Potential Refractive Index, K (which must not be confused with k, the effective earth radius factor), achieves this end by reducing the meteorological data to a standard pressure of 1000mb. This results in a change of temperature in accordance with a well-known law of thermodynamics (the same one which causes the end of a bicycle pump to get warm when you compress the air inside) and although the amount of water-vapour remains unchanged, most of its attributes, such as vapour pressure, dew-point and relative humidity will alter.

If these new values, together with 1000mb as the value for P, are inserted in the original expression for N, the result will be the potential refractive index, K.

A value of potential refractive index is the value of normal refractive index which a sample of air would have if it was taken adiabatically to a pressure of 1000mb. If we convert a change of atmospheric pressure to a change of height, it can be seen that air which is forced to rise will expand and cool · due to the reducing pressure - while on the other hand if it is forced to sink (i.e. anti-cyclonic subsidence) it will be compressed and its temperature will rise. These changes are termed adiabatic. That is, cooling or heating has taken place by pressure changes alone, and without loss or gain of heat by any other means.

Anticyclones

It is well known that the most prolonged and pronounced periods of anomalous propagation coincide with the presence of regions of high atmospheric pressure in the vicinity of the transmission path. At such a time the air circulation consists of light surface winds spiralling outwards from the centre of an anticyclone, and for these to be maintained air must be brought slowly down from aloft.

As the subsiding air descends, its pressure increases, producing dynamical warming (the bicycle pump effect again). Now air at, say 20,000 feet holds very little moisture even when saturated and as the capability of the air to retain water-vapour increases very rapidly with temperature, by the time the air has descended to about 5,000 feet the amount which it contains is only a small fraction of the amount which it could hold at its new temperature. Therefore, although no water-vapour has been taken away, the descending air has become very dry, compared to air normally found at that level.

The interesting point, which may now be obvious, is that the mechanism controlling the temperature, and hence, indirectly, the relative humidity or dew-point of the subsiding air, is the same one which was used to reduce the refractive index value to a standard pressure of 1000mb and this means that the air, as it decends, tends to retain its original value of K (Fig.2.1).
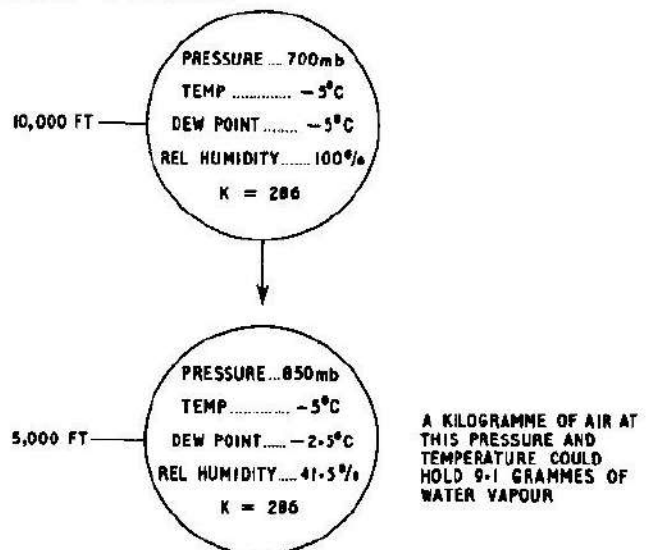

Fig.2.1. An example of changes resulting from the adiabatic descent of a sample kilogramme of air containing 3.78 grammes of water-vapour.

This is where potential refractive index scores over other derived units, because, as we shall see, large-scale downward movements of air are almost invariably involved when v.h.f. signal enhancements occur and the K value provides us with a means of identifying and following a particular part of an air mass as it undergoes adiabatic change.

Now to return to the anticyclone. The portion of the atmosphere near the ground is generally a region of turbulence due to convective currents and mechanical stirring brought about by the presence of obstructions, either natural or man-made. This causes the removal of water-vapour from the surface by evaporation, to be distributed in the first few thousand feet of air where it may ultimately show itself in the form of cloud. Rising air cools and increases in humidity by an inversion of the same reasoning already applied to descending air. Eventually then, it would reach saturation and further cooling would cause water droplets to form if condensation nuclei are present. At this stage the process is no longer reversible, but that need not concern us here.

It is quite likely therefore that the warm, dry subsiding air will find itself in close vertical proximity to the relatively cool, moist, turbulent air stirred up from the surface, and there is then a sharp transition at the subsidence boundary.
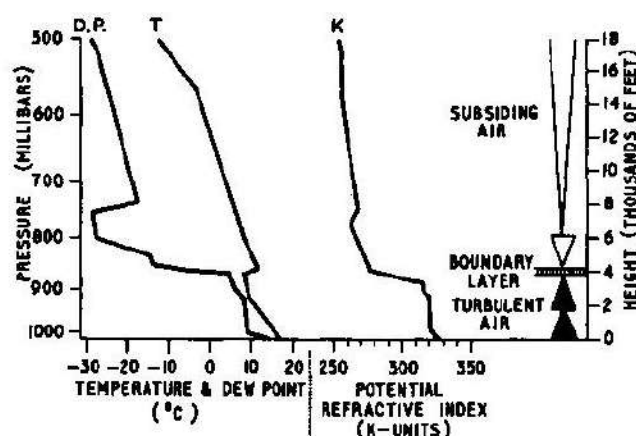

Fig. 2.2. Distribution with height of temperature, dew-point and Potential Refractive Index through the subsidence boundary layer.

In terms of K, or N for that matter, an ascent through the boundary layer is characterised by a sudden steep fall in refractive index with height, Fig. 2.2., and this is a change in such a direction as to cause abnormal bending of the radio waves back down towards the earth. Under these circumstances, a low-angle ray from the transmitter could be returned to the ground many miles beyond the normal horizon.

The Subsidence Boundary Layer

The effectiveness of the transition layer will depend upon its height, its horizontal extent and its intensity. When warm, dry, subsiding air is present aloft, the more the water-vapour which is available below, the sharper will be the transition across the boundary. Many operators will be able to testify from their own experience to the excellent results obtained during an opening when thick fog sets in during the evening, and, as another instance, the almost unlimited

supply of cool, moist air available in the lower layers over the sea accounts for the higher median signals received on over-sea paths, as compared to land paths of a similar length.

Under the meteorological conditions normally encountered near the surface, the amount of water-vapour held in suspension is the most important factor in determining the refractive index. The quantity required to saturate a given volume of air increases rapidly with increasing temperature, approximately doubling in value for each rise of 10°C. The low values of temperature usually found in the lower levels in a winter anticyclone result in relatively low values of refractive index, even when the air is saturated. This reduces the magnitude of the transition layer between the convective air below and the subsiding air above, which, in turn, reduces its effectiveness as an aid to long-range v.h.f. propagation.

In summer when the temperature is high and the air is capable of holding a large amount of water-vapour, air passing over land is normally dry (because of the large amount of water-vapour required to saturate it), again resulting in low values of refractive index. This often produces a very marked diurnal variation in propagation effects, high signals being most common in the late evenings and early mornings.

The best compromise between temperature and availability of water-vapour commonly occurs in autumn, and some of the most spectacular openings occur around September and October. Autumn anticyclones may persist for two or three weeks, and conditions will be observed to vary considerably over this period although pressure remains high.

Most Favourable Part of an Anticyclone

There is a well-known rule-of-thumb among amateurs both here and in the United States, which suggests that the most favourable conditions for v.h.f. signal propagation occur when the path lies on the western side of an anticyclone.

We have seen that subsiding air is an essential requirement, bringing air from aloft which becomes dry and warm as compared to the normal low-level air. But the mere presence of this air is not enough. To provide the steep lapse of refractive index necessary for abnormal ray-bending there must be air of high refractive index below it, and it is the transition region between them which is all-important. Now this region is the result of a slow, steady flow from above, and a turbulent or convective motion from below, the latter largely determining the height and sharpness of the boundary.

Consider the circulation around an anticyclone, which is clockwise in the Northern Hemisphere. On the Eastern side air is travelling towards the South, usually towards regions of increasing surface temperature, a situation generally leading to an increase in convection and turbulence. This raises the

boundary and, by spreading it, makes it a more gradual transition. The reverse argument applies to the Western side, where the boundary tends to become lower and more pronounced, due to the more stable conditions below. On these grounds then, one would expect the Western side to be more likely to give rise to long-range propagation, and this is borne out by observation. It needs to be emphasised though, that it is not the subsiding air which gives abnormal conditions it is the sharpness of the contrast between its refractive index and the higher values encountered near the ground which is important. For a given path length there will be an optimum height for the boundary layer, although as a general rule, the lower it descends, the more effective it will be, until a height is reached which cannot be "seen" by the lowest angle radiation from the transmitter, and the lowest angle reception at the receiver. The possibility of multi-hop propagation, particularly over paths crossing the sea, complicates any attempt to derive a hard-and-fast connection between layer-height and distance over which reception is possible for a given power, and there is frequently evidence of "skip" effects over long-distance paths.

A descending layer may ultimately reach the surface, initially over high ground, and subsequently over a wide area. When this occurs during a spectacular opening a mysterious "drop-out" results, since the refractive index gradient in the subsiding air is markedly below normal. High altitude stations may then experience the disappointment of hearing their neighbours still in contact with distant stations which to them have disappeared.

For a station located at an appreciable height above sea-level, periods of what is to everybody else abnormally good propagation may turn out to be just the reverse. Strange though it may at first seem, the cure for this is to reduce the height of the aerial system - a difficult decision to accept for anyone who has transported equipment up the side of a mountain the hard way, but nevertheless a fact. Under such circumstances, wet-and-dry bulb thermometers should reveal low relative humidity which will confirm any such suspicion regarding the cause of the apparent lack of activity on the band.

Locating the Subsidence

We are now in a position to appreciate how potential refractive index assists in an analysis of the state of the atmosphere during periods of anomalous propagation.

Along a particular transmission path during normal conditions refractive index tends to be distributed horizontally as shown in Fig 2.3. The changes of refractive index with height are inversely proportional to the spacing of the lines, which are here uniformly distributed.

Subsiding air will cause potential refractive index lines to be brought down to abnormally low levels, as discussed earlier (Fig.2.2). The lines nearer the ground tend to rise slightly and the overall effect is of a region of steep lapse at some distance above the ground. Whenever this is present signals become higher in strength over a given path, and receivable over greater distances, provided the paths are favourably placed with regard to the location of the subsiding air.

A very simple technique enables this to be placed. Consider once again Fig.2.3. If the change in K between top and bottom of the diagram is calculated for various points horizontally, the greatest differences will occur in the region of the subsiding air. If such differences are derived for a horizontal "slice" of the air above the British Isles, we can, in fact, locate the subsidence by drawing the different contours, or Delta-K lines (Fig.2.3). It is convenient to take for this purpose the difference between the K value at the surface and that at a height where the air pressure is 850mb (approximately 5,000 feet) since this information is always available without interpolation from data reported by radiosonde stations.

It has been found that paths experiencing high signals are symmetrically placed to the pattern of the corresponding Delta-K chart, and that when signals have fallen the pattern has become asymmetrical. In other words, although conditions are generally good in the example pictured, the path AB is more likely to experience signal enhancement than either of the paths AC or DE, since the former has a symmetrical distribution of Delta-K values along it.
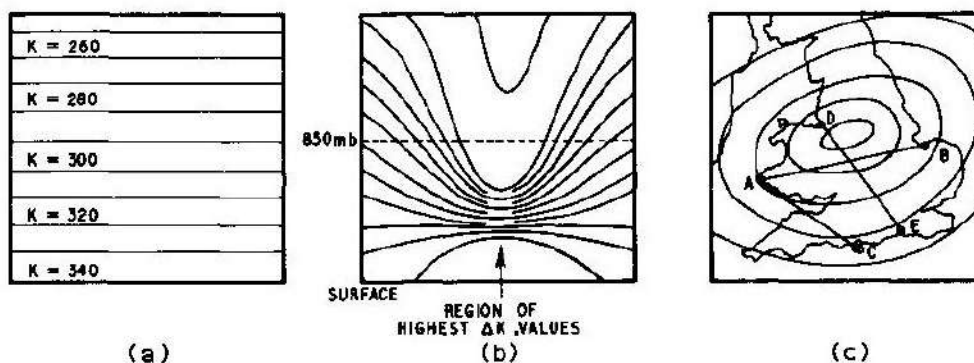

Fig.2.3. Development of the Delta-K chart. (a) Uniform distribution along path. (b) Subsiding air revealed by displacement of K Lines. (c) Contours of constant difference in K values between surface and 850 mb. Path AB, symmetrical to the pattern is more favoured for anomalous propagation than paths AC or DE.

In practice, the subsidence boundary layer does not always occur between the surface and a pressure of 850mb. It is fortunate that when it is above, the region of steepest lapse (i.e. where the lines are closest together) generally has below it a compensating slack gradient. In this case the Delta-K contour lines enclose a centre of low value instead of high, as originally considered, but the same symmetry observation is found to apply.

Calculating K

Referring to Fig.2.4 which is a skeleton outline of part of a meteorological temperature-entropy chart known as a "tephigram" on which K isopleths have been drawn, three scales are of interest in the present matter, namely, the pressure lines sloping diagonally from bottom left to top right, the temperature lines running vertically and the K lines which curve round the top left-hand corner. In the interests of clarity the values above K = 310 have been omitted although the lines themselves have been inserted.
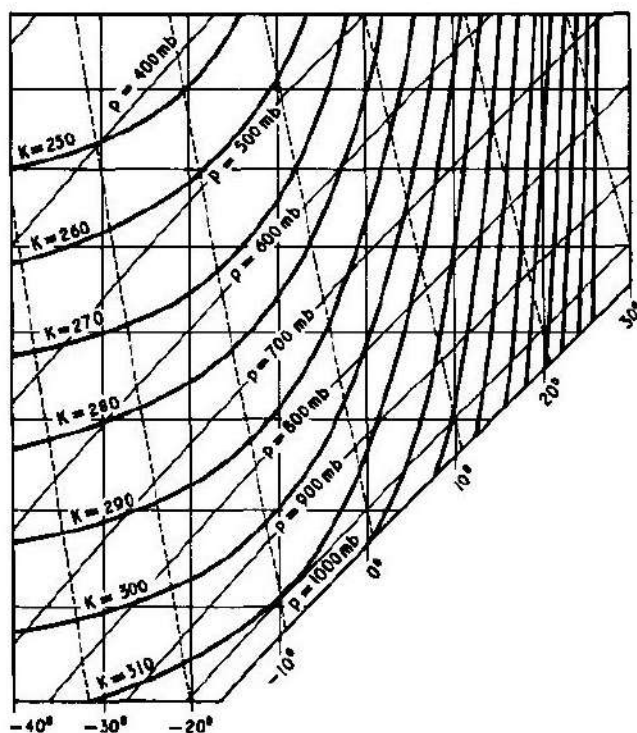

Fig.2.4. Section of tephigram for calculation of K. (Reproduced by permission of the Journal of Atmospheric and Terrestrial Physics).

To find the K value corresponding to a given combination of pressure (P), temperature (T), and Dew-point (Td)

(a)   Locate the point corresponding to (P, Td), using the appropriate skew axes.

(b)   Through this point lay, and hold steady, a transparent straight-edge parallel to the sloping dashed lines.

(c)   With a pencil in the right hand follow the pressure line to the point (P, T).

(d)   Return horizontally to the straight-edge and make a dot beside it.

(e)   Interpolate from the K curves the potential refractive index value at the dot.

As an example, where P = 1000mb. Td = 0°C and T = 10°C, then K = 302.

If required for raytracing purposes, the normal refractive index value, N, may be calculated from the K value by the use of a factor dependent on the pressure, derived from the expression:

$$K/N = (1000/P)^{0.714}$$

where P is the air pressure in millibars at the level of the observation.

The K-Profile

As has been described earlier, K values may be calculated from observations of pressure, temperature and relative humidity. A graph of K values obtained from a radiosonde ascent, plotted against height (or more commonly air-pressure, as this is more readily available) constitutes a K-profile.

Details of all the British radiosonde reports are published by the Meteorological Office in their Daily Aerological Record. With the diagram described earlier and a certain amount of practice an ascent to 500mb can be converted to Potential Refractive Index and plotted on graph paper (constituting a K-profile) within five minutes. Although a plot to this pressure level is usually worthwhile during periods of subsidence, all the features likely to affect v.h.f. propagation will appear well before 500mb, and it may be thought an advantage to terminate the work at 700mb (about 10,000 ft) as has been done with the illustrations to this article.

It is a good plan to record all the data directly on the plotting sheet at the appropriate pressure levels, as shown in Fig.2.5 1cm to ten K units and 20mb respectively are very convenient scales to use and K may be expected to range between 250 and 340 units over most of the year, occasionally extending a few units beyond either end in the summer, referring of course, to British Isles conditions.

It is hardly necessary to point out that although the published skeleton diagram can be used for K calculations, it is not sufficiently large, or detailed, for accurate work. The best plan is to transfer the K lines to a full-sized tephigram chart, as this will allow values to be read off in comfort to the nearest K unit.

The Cross-Section

An early authority on propagation, Publius Ovidius Naso (better known under his trade name, Ovid) once wrote, "Things which are not of value singly are useful collectively". Whilst it is hardly fair to regard an isolated K-profile as being of no value, it
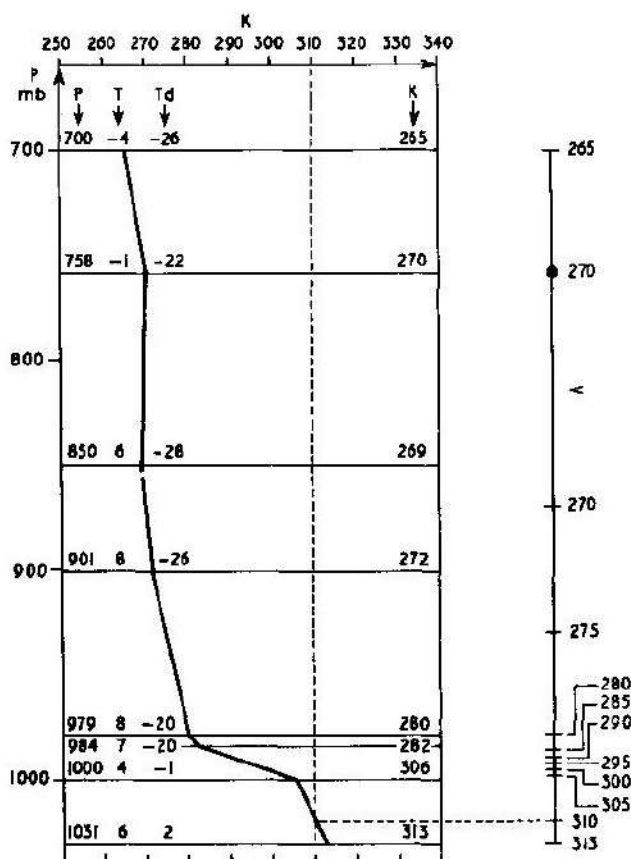
Fig.2.5. K-profile derived from radiosonde pressure (P), Temperature (T), and Dew Point (Td) data, obtained by Hemsby, Norfolk at 12.00 GMT on 3 December, 1962. The vertical line to which various values have been projected would normally form part of a cross-section or a time-section.

appropriate level of the cross-section. A surprising amount of detail results but the work is considerable.

The construction of K cross-sections is one of the most rewarding ways of investigating anomalous tropospheric propagation. If very long-range signals have been heard, at a time when nothing of interest appeared on the appropriate cross-section, it is fairly safe to presume that Sporadic E, the aurora, or some orbiting product of the space age has been the agent responsible.
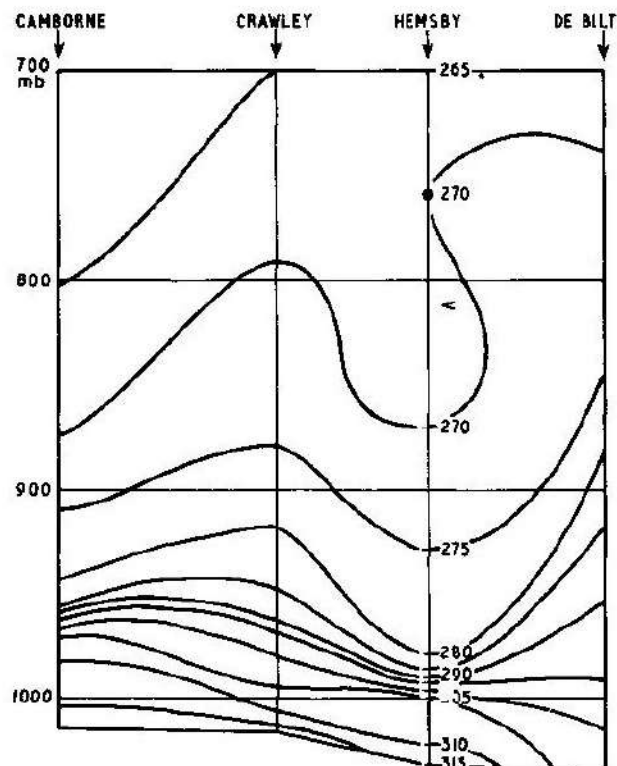

Fig.2.6. K cross-section through Hemsby, using the levels derived from the K-profile of Fig.2.5. This was a notable high signal occasion.

is certainly true to say that its usefulness is enhanced many dB when combined with other K-profiles in a cross-section.

A cross-section is used to find the vertical distribution of potential refractive index along a given path at a given time.

The way the levels at which certain K values occur can be transferred to a vertical line is shown on the side of the K-profile diagram (Fig.2.5). The line would be one of a series, spaced in proportion to the distance between selected radiosonde stations as nearly as possible along the path of the cross-section. Care is needed to identify on the vertical line turning points where the profile slope changes sign at one of the chosen values, turning points at other intermediate values, and ordinary crossing points, beyond which a rise or fall continues. These all supply useful information which helps to complete the space between stations (Fig.2.6).

When the chosen path is too far from a line of radiosonde stations for them to be representative, data must be interpolated by drawing a series of horizontal charts of the K distribution over the British Isles at 50mb (or less) intervals, and transferring the distribution along the line of the path on each to the

The Time-Section

Another way of displaying potential refractive index in the vertical is to show how it varies from profile to profile above a given place. The result is a chart very similar in appearance to the cross-section, but the horizontal scale is now not distance, but time, and its length is limited only by the interest or patience of the construction (Fig.2.7). The levels at which various K values occur on each profile are again transferred to a series of verticals on a separate sheet, in this instance evenly spaced and 2cm apart (a scale of 4cm = 24 hours).

The time-section gives an immediate indication of periods when anomalous v.h.f. and u.h.f. propagation is likely. It reflects the passage of pressure systems on the surface weather chart and, being based on the results of only one radiosonar station, is easy to maintain on a current basis if the necessary meteorological data is available.
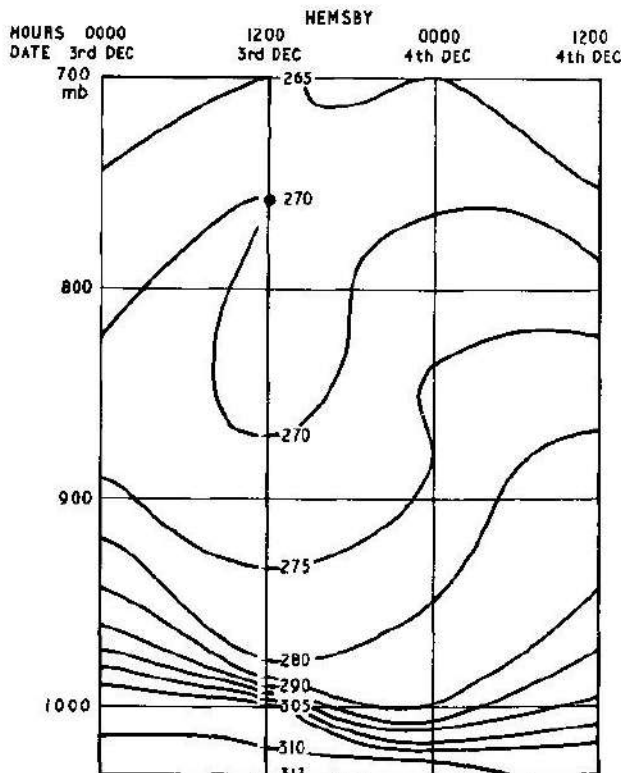
Fig. 2.7. K time-section for Hemsby, in which the profile of Fig. 2.5 is combined with others showing conditions before and after.
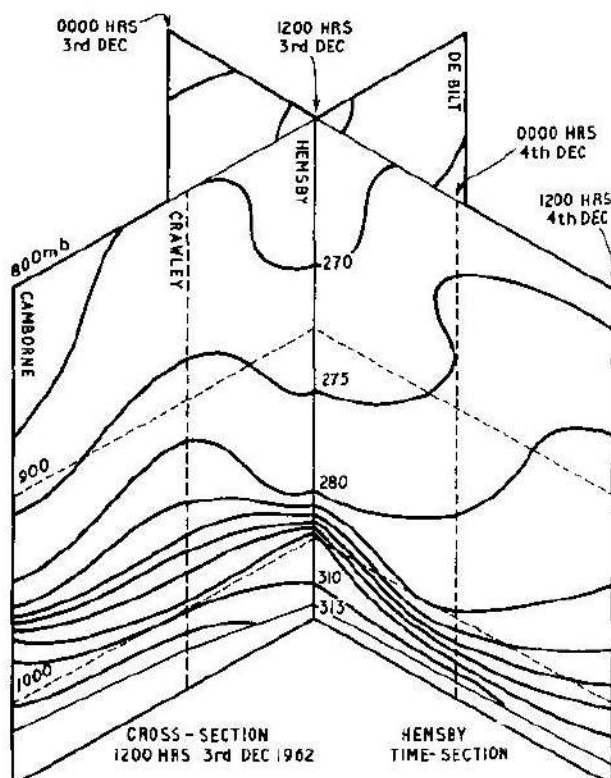

Fig. 2.8. Showing how the time-sections and cross-sections are related. Other planes containing time-sections for the other stations and cross-sections for the other times may be imagined to interlock in a similar way.

3-D with a Difference

Since any K-profile can form part of both a time-section for a given station and a cross-section through that station for a given time, it is evident that the two forms of presentation must have a lot in common. In fact, a series of cross-sections for a particular period and the individual station time-sections over that same period interlock completely as is suggested in perspective in Fig. 2.8.

Although this exercise is one which would be difficult to complete without having to retain the services of a qualified psychiatrist, the idea is an important one as it not only relates the appearance of particular features on both types of diagram but, should difficulties arise, also allows one sort of plot to be used as a guide to drawing up the other. It hereby follows that observations made relating to the interpretation of either diagram may be regarded as applying equally well to both.

Air in Motion

The most familiar demonstration of air in motion is of course, provided by the wind. But although a pen recording of the signal strength of a distant v.h.f. station frequently bears a strong resemblance to an anemometer record of wind speed, there is no evidence of a direct connection between them. There is, however, also a less apparent but very important movement in the vertical, ranging from the gentle downward flow of subsiding air associated with an anticyclone to the fierce up-currents found inside a thundercloud, which have been known to tear the wings off aircraft, and are for this reason rather unpopular with the airline companies.

It was shown earlier that air descending adiabatically (i.e. without addition or subtraction of heat) retained its K value and resulted in low values being brought down towards the ground. The converse applies initially to ascending air, but here the laws of nature demand that temperature falls as pressure decreases, a circumstance put to good use by the manufacturers of compression-type refrigerators. Eventually the temperature drops so far that it reaches the dew-point and condensation occurs, marked by the appearance of cloud. Further ascent causes more water-vapour to condense and this may result in quite large droplets being produced by coalescence, which are then likely to fall out of suspension, a phenomenon known to us in the trade as rain.

Up to the condensation level, the K value of a sample of air ascending adiabatically remains constant; beyond that the latent heat released brings about a decrease in potential refractive index with height which, however, only serves to emphasise the appearance of ascending air on the cross-section and time-sections. It is soon recognized as the trademark

of depressions, a consequence of the winds spiralling inwards having to displace the air upwards in order to prevent a dreadful pile-up from developing at the centre.

Lows and Highs

As an example let us examine a portion of the Crawley (Sussex) potential refractive index time-section for April 1962,Fig.2.9., representative of conditions over Southern England. On Easter Sunday evening signals from the West of England were received in the London area at good strength. This was the 22nd. and there is no mistaking the signs of descending air on the time-section, nor of the boundary which formed around 900mb. An anticyclone was approaching from the Atlantic.
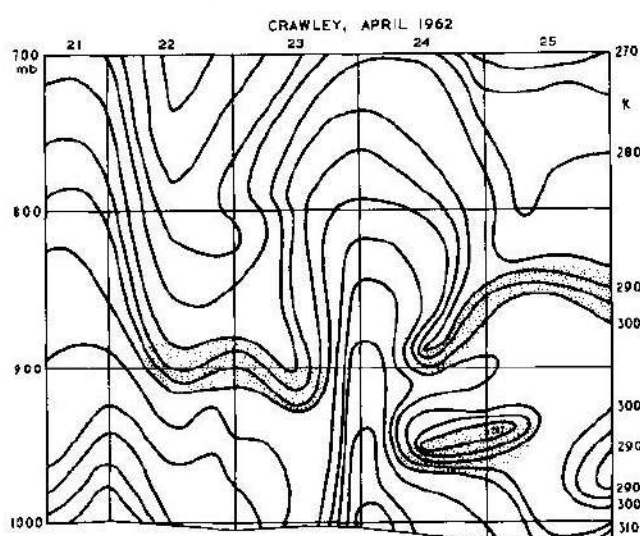

Fig.2.9. The K time-section for Crawley, Sussex, for the period April 21-25, 1962 analysed in some detail in the text. The layers of steep K gradient have been shaded.

During the evening of 24th April signals from Continental stations were received in South East England at quite good strength but subject to fading. Reports of conditions on the 23rd are conspicuously absent, but an ominous tongue of ascending air will be seen on this diagram, and this has forced the subsidence literally out of the picture. It was caused by the flank of a depression centred over North East France which quickly moved away past the Low Countries, allowing the anticyclone to return and extend across to the Continent. The double layer apparent on the 24th is very conducive to deep multi-path fading.

However, on the following evening 25th. April, conditions had improved to provide excellent contacts between stations in South East England and F, ON, PA, DL and DJ. Notice that there is a single layer again, this time just above 900mb. Amateurs in the Midlands and North shared none of this activity, but their turn was to come on 26th. April when a different

high pressure centre took over and the main region of subsiding air appeared over the UK.

RADIO REFLECTION OF V.H.F. SIGNALS BY BACK SCATTER AURORAL PROPAGATION

If we consider the chain of events which can lead up to an auroral occurrence, our starting point must be inside the sun. Something happens there, exactly where we do not know, to cause a solar flare (eruption) to take place somewhere on the disc. The ultra-violet light-rays, X-rays etc. which are emitted, result in a considerable increase of ionization in the earth's ionized layers. In fact, the lower layers are so heavily affected that long distance communication becomes impossible; absorption and relatively low level reflection being so severe. At about the same time the solar eruption causes particles to be blown out from the sun's interior due to very strong magnetic fields being created. These particles result in the phenomenon now termed the "Solar Wind". The sun's normal magnetic field can also be disturbed, and it is believed that very large electric currents flowing across the sunspot, i.e. eruption hole, result in local magnetic disturbances, which cause the emitted particles to have spiralling trajectories. These become trapped in a magnetic cone, which reaches out into space, sweeping round with the sun's rotation until this catches up with the earth, and literally sprays it with the emitted particles. The result is an aurora.
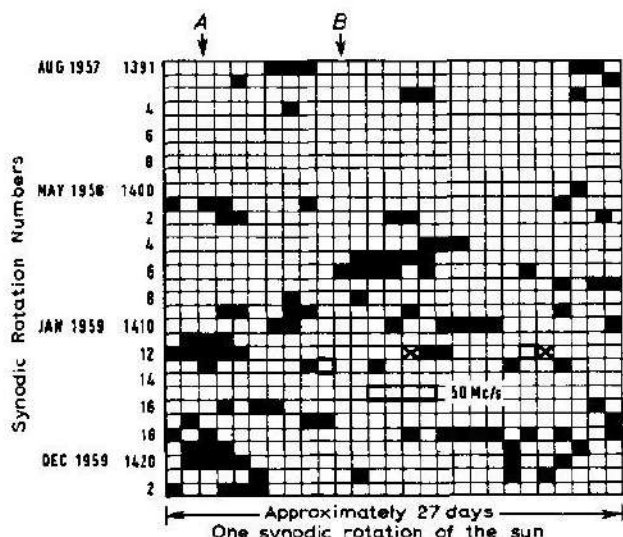
The position of a sunspot in relation to the earth is a most important factor influencing auroral and associated effects. If a flare occurs near the sun's central meridian, at a point looking at us, this can cause far greater disturbance than one, say, nearer the polar latitudes, out on the westerly limb (disc edge) which is rotating away from us.

It is useful to remember that not all solar flares and associated eruptions disturb the earth's magnetic field, in fact, many flares which are responsible for short wave "Dellinger" type fadeouts never cause aurorae at all.

It appears to be important for auroral effects that the solar flare be in certain areas, usually denoted "M" regions, meaning "magnetic disturbance region" or "activity centres". Very good radio aurora can also occur without any noticeable solar flare (as it is possible for the flare to occur unobserved on the far side of the sun) and the auroral particle stream, sweeps round with the solar rotation until it catches up the earth.

The surface of the sun has no features which enables us to fix positions and, being composed of gas, it rotates at different speeds at different latitudes. An average of its rotation viewed from the earth is termed a "synodic period" meaning that looking from the

earth the same part of the sun is directed towards us once per synodal rotation (27 days approximately), so the same eruption hole, if still active, can repeat auroral disturbances for a number of synodal rotations. This has been plotted in the Synodic Plot for 1957-59, Fig.2.10, which tends to show that some parts of the sun's surface are more prone than others to cause auroral disturbance. It should be noted that the plots are not sun spots but "auroral occurrences".

Fig.2.10. A synodic plot, which shows the considerable difference in auroral occurrence in relation to the position of the Sun's disc. Note the frequency (line A) rotation 1411-1420 in contrast to line B where no repeats occurred. Point A on the Solar disc is most likely an activity region.

Now, a sunspot cycle is measured not only by the number of spots appearing at a given time, but also by their position on the disc and their magnetic polarity. Sunspots start in the higher latitudes, about 30° North and South, and slowly appear in lower latitudes as the solar cycle proceeds, ending in the equatorial regions about 7° North and South. As it takes about two years for an old cycle to end, after a new one has started, at sunspot minimum the new reversed magnetic polarity spots occur at about 30° North and South together with old cycle spots at about 7° North and South. During sunspot maxima the concentrations are about 13° North and South. There is also a tendency for the maximum number of eruptions to occur in the sun's Northern Hemisphere during one 11 year cycle. The spot magnetic polarity also changes. This really means there is a 22 year cycle.

It is interesting to note that ever since the days of the earliest astronomical records right up to the present time, large spots hardly ever occur within a small band centred on the sun's equator. This is due to the magnetic polarity changing at this point.

In the 1930's Dr. Dellinger showed that there was a correlation between the occurrence of solar flares and very sudden fadeouts of long distance shortwave communication due to very considerable ionization in the lower layers.

The "Solar Wind"

This is a term used to describe the constant outflow of ionized gas expelled by the sun. During solar outbursts (flares or spots) the "solar wind" can increase from a gentle breeze to become more like a raging hurricane, when giant streams of strongly ionized very hot gas are shot out like jets from a hose pipe. In terms of speed the ionized gas travels fast (about 20-24 hours for the sun/earth transit), but it is slow compared with the ultra-violet, X-ray, radiation and cosmic particle times of about 8 to 15 minutes for the 93 million mile journey. This is due to the ionized gas spiralling, because the sun's normal magnetic field is very distorted near its sunspots where usually an extremely strong magnetic longitudinal field exists. This grows out into a kind of cone around which the ionized gas spirals, resulting in a slow forward velocity.

As the sun rotates, the magnetically trapped ionized gas streams sweep up to and engulf the earth. However, auroral occurrence is much more likely during Spring and Autumn periods. This is because the earth's magnetic field is coupled to the solar wind and the sun's magnetic field more effectively at these times, due to the tilting of the sun's axis, i.e. it is rather like mutual coupling between coils.

There is considerable evidence to support the contention that the "axial" times are best, i.e. the periods when the sun's North or South Pole (rotation axis) is tilted towards the earth. This reaches a peak twice a year, first on 10th. April and again on 10th. October. During the "Epochs" in early December and June the Solar Equator is seen from earth with the Poles at the limbs. This is generally a very quiet auroral period. The solar wind can affect the earth's magnetic field, and in order to appreciate this fact we must first understand just how the field comes about.

The Earth's Magnetic Field

Many believe the earth's magnetic field is due to a large mass of iron at the centre of the earth which somehow or other becomes magnetized. This is far from the truth, in fact, it is doubtful whether it is possible to magnetize the earth's core at all because it is certainly above the Curie temperature of iron and could only be para-magnetic (very weak). A more reasonable explanation, and one which, as will be seen later, fits the aurora case well, is that large electric currents flow in the ionized layers (particularly the E region) and set up magnetic fields (the auroral and equatorial jets). The total effect is known in scientific circles as the "ring current" magnetic theory. There is also the "dynamo" theory which is used to explain the production of the earth's stable background magnetic field - so called because it requires the earth's rotation

for its existence, together with circulating currents in the crust. There is also the effect of surface magnetism in rocks etc., but for all practical purposes this can be ignored. Fig.2.11 illustrates the earth's magnetic field in space as the field angle of dip.

In order to measure magnetic force, a system of "absolute units" was introduced by Gauss at his laboratory in Gottingen during 1832. This, expressed in terms of "unit length, mass and time" gives the basic magnetic unit - the gamma · 10^5 gammas = 1 c.g.s. (centimetre, gramme, second) unit. Any point on the earth may thus be measured in terms of the magnetic force and is directly comparable to the basic unit laid down by Gauss.
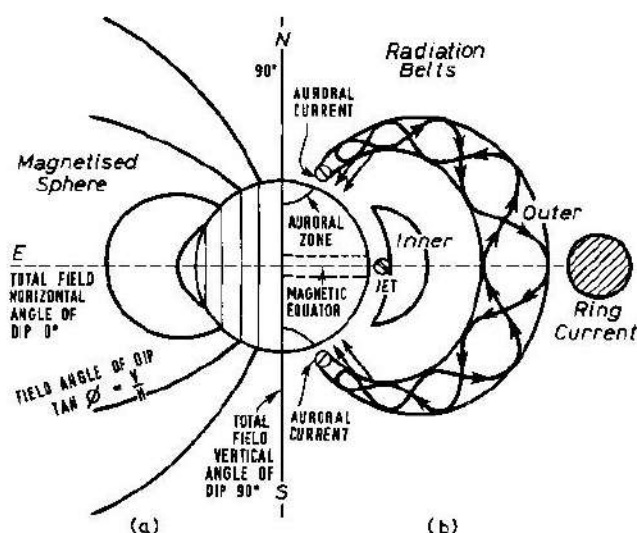
Fig.2.11. (a) Illustrates the earth's magnetic field in space as the field angle of dip. (b) Shows the relationship of the radiation belts. The outer parts are mostly charged particles which ultimately cause auroral effects. In practice there are not two distinct zones but rather a merger from one to the other.

The stable background magnetic force is listed for any recording station as a mean reading, e.g. for Lerwick in 1960 the "horizontal force" was 14,000 gammas (0.14 c.g.s. units) and "vertical force" 47,000 gammas (0.47 c.g.s. units). All the variations are relative to these basic "means". Due to the "secular" (long period) magnetic changes these "mean" values vary slowly, e.g. for Lerwick the 1960 horizontal force increased by 15 gammas and the vertical force by 28 gammas over the 1959 "means". The range between the extreme values recorded during 1960 were - "H" force 3283 gammas, "Z" force 1151 gammas. The "declination" (difference between geographic and magnetic North) at London in the year 1815 was 24° West, whereas at the present time it is only 10° West. So secular changes are caused by the earth's internal background magnetic field changing slowly with time. The precise reason for this is, at the present, not fully known.

The earth's total magnetic field (internal plus external) is measured in three directions, usually referred to as "vertical" (Z), "declination" (D) (this being the angle between the magnetic North Pole and the geographic North Pole), and "horizontal" (H). The curves shown in Fig.2.12 illustrate the daily magnetic departures from the mean variations of D-H-Z meaned for the year 1959 for Eskdalemuir, Scotland. Magnetic effects are recorded for quiet and stormy days. It is interesting to note that Z, H and D swing about a mean in relation to time e.g. in relation to the sun's position relative to a fixed point on earth. This tends to show that the electrical conductivity of the ionosphere and associated electric currents vary also, and that the daily variations in the earth's surface magnetic field are mainly due to external causes. It should also be noted that, under magnetic storm conditions, it is the extent of variations that changes, not so much general character. Here then is the clue to auroral activity - large surface magnetic changes are the result of large ionospheric electric currents, mainly in the E region. This means that strong ionization must be present to carry the current, many charged particles (e.g. electrons and protons) must be available to supply the driving power, and the sun must be able to supply the particles.
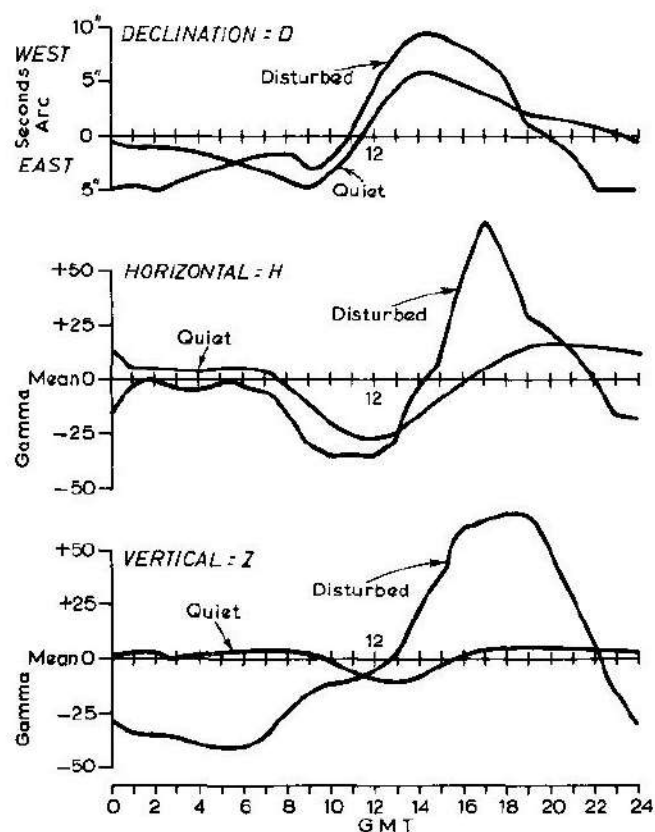

Fig.2.12. This illustrates the daily magnetic departures from the basic mean of changes in D-H-Z for the year 1959 at Eskdalemuir (Scotland) showing means of both "quiet" and "disturbed" days. Note that it is the extent of change which varies with disturbed conditions, and not so much the general character.

"K" INDICES	0	1	2	3	4	5	6	7	8	9
Lerwick, Shetland	0	10	20	40	80	140	240	400	660	1000
Eskdalemuir, Scotland	0	8	15	30	60	105	180	300	500	750
College, Alaska in Auroral Zone	0	25	50	100	200	350	600	1000	1650	2500
Kuyper, Sumatra on Equator	0	3	6	12	24	40	70	120	200	300

(a)

September 1957 — LERWICK OFF SET 10 HIGH
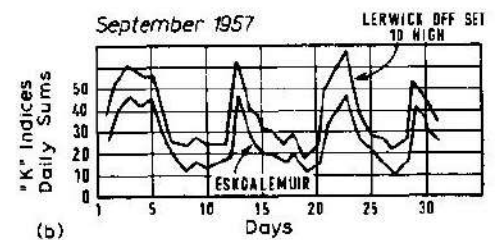
(b)

Fig.2.13. (a) Shows the comparability of K indices for different magnetic observations. A lower limit of Gamma variation in either D, H or Z for any three-hour period (00.00.03.00, 03.00.06.00 GMT etc) will give the appropriate K indices for the station. (b) Shows the comparability of K indices for observations in a given region. For the sake of clarity the charts have been displaced although these are to the same scale and show the total for each day of September 1957. This is the basis of the Kp figure. For example, 12 observatories in a region are "meaned" to give the "planetary" (Kp) figure - an indication of the magnetic activity for the region.

All magnetic variations at a station can be compared by a coefficient K, which is a measure of any range of variation out of Z, D and H added over and above the particular station's normal variation, the extent of K being defined for the particular station. This means that any figure for K 1-9 is comparable station-to-station in terms of magnetic variation from normal, even though the exact individual variables are different. This point is explained in the table shown in Fig.2.13. which lists Lerwick (Shetland), Eskdalemuir (Scotland), College (Alaska) (a station situated right in the auroral zone) and Kuyper (Sumatra), almost at the equator. The K indices are the range of the most disturbed element D, H or Z, taken over a three-hour interval. As a rough guide, to increase K by one means that the gamma variation is about doubled. Eight K figures are issued each day starting at 00.00 - 03.00, 03.00 - 06.00 GMT etc.

The importance of K indices is that they are a direct measure of the earth's magnetic field variation and, indirectly, an inference of auroral (E layer electric current) ionization conditions. As a matter of interest, during magnetic storm conditions the total auroral electric currents can reach a figure of 180,000 amps. This flows as a sheet current at around the 110 kilometre height (lower E layer) where, as will be seen later, back scatter radio aurora ionization occurs. If the K indices from a number of suitably placed observatories are compared, then the K figure can be deduced. This is a measure of the "planetary" magnetic activity.

As the solar wind, or, to be more precise, the spiralling stream of ionized particles, approaches the earth, it gets trapped by the radiation belts. These can be considered in two parts, outer auroral and inner equatorial. The outer, auroral, belts must closely follow the earth's magnetic field pattern, so there comes a point (well out in space) where the individual charged particles become influenced by the earth's magnetic field which causes them to be deflected, spiralling round to the earth's night side.

As the location of the surface magnetic poles is known it can be assumed that the strongest fields will be near to these points. In practice, however, it is not a point but rather a ring round the magnetic poles in

geomagnetic latitude 67°. These are usually referred to now as the "radiation horns" or "auroral zone horns". As the ionized particles get nearer, and so become influenced by a stronger magnetic field, the rate of deflection increases, until, finally, the particles are spiralling into the magnetic polar rings or auroral zone horns. The protons and electrons separate in much the same manner as Fleming's right or left hand rule applies to electro-magnetic theory. The earth's outer radiation belts (Van Allen) now become filled with spiralling charged particles which are trapped. If the magnetic field strength is increased enough, a point is reached when a given particle charge, velocity and spiral thread pitch, will be compressed so that the forward movement is halted, reversed and then spiralled out along a line of magnetic force (Fig.2.14). The spiralling direction remains the same but the particles shunt to and fro between the earth's magnetic polar rings with such velocity that they make a round trip in only hundredths of a second. A see-saw spiralling "dance" of the trapped particles then begins. Some may carry on the "dance" for millenia of time, but for others a chance collision, deflection of the magnetic field, insufficient energy etc., can cause their end, or, to be more specific, a change of state.
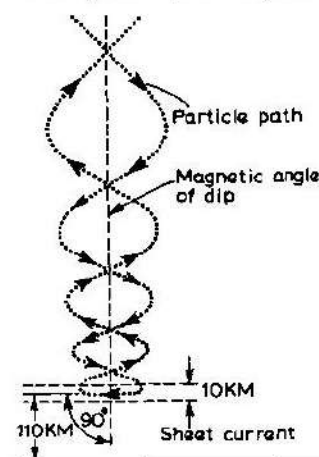
Fig.2.14. This diagram shows a typical auroral radiation belt particle path which is shunting to and fro between the northern and southern auroral zone horns. Particles reverse when the spiral is at 90° to the magnetic angle of dip irrespective of height.

2.11

Due to the spiralling, particles tend to drift eastwards or westwards, depending on the charge. Because the spiral is not circular, due to the magnetic field at the top of the spiral being weaker than at the bottom (nearer the earth at equatorial transit), this total progress tends to separate the different energy (particle) levels into layers (or shells as they are called), which encircle the earth. The USA Argos and Rainbow bomb tests were carried out partly to find out more about this process, known as the "onion skin" theory. The increase in low level E region ionization (due to the electron shells) causes low frequencies to be trapped, resulting in considerable signal increase of 10 kHz to 500 kHz signals and also atmospheric static noise. At the same time galactic noise tends to be reduced as the electron shells shut it out. This effect is most noticeable over the range 20 MHz - 40 MHz in the higher latitudes.

The importance of "optimum height" to radio aurorae must not be forgotten. This is the point where the greatest ionization and sheet current takes place, e.g. if a particle is slowed down by collision, its ability to cause ionization gets progressively less, but when penetrating into the denser atmospheric gases the chances of a collision are greater. Our best radio reflection point will, therefore, be where the highest level of ionization occurs, for good radio aurorae large quantities of particles or masses of electrons are required, and in practice this occurs only in a narrow band about 100-120 km in height, where the auroral sheet currents are flowing (E layer), virtually at the top of the atmosphere. However, before it is possible to obtain backscatter radio signals from this area, certain conditions must be met. The most essential requirement is that the earth's magnetic field (angle of dip) must be at or very near right angles (90°) to the transmitted and received wave front, at the point where the ionization is taking place. The actual area of sky from which useable ionization can be obtained is a function of frequency and location, but the peak or highest signal area will always be at the point where the wave-front is at right angles to the magnetic field at the optimum height. (Note the use of the term "magnetic field" - this is deliberate in order to take account of vertical and horizontal components).

The Field Angle of Dip

At the magnetic equator all the field is horizontal; at magnetic North or South Poles it is all vertical. At any point in between it is a result of the vertical and horizontal amounts. This is called the "angle of dip" and can be written as $\tan \phi = Z/H$. This is about 67° at London at the present time whereas at Lerwick, where the mean horizontal force is 14,000 gamma and the vertical force at 47,000 gamma, it is 72° 53'. It is this "angle of dip" down which the spiralling particles are coming that our transmitted and received wave fronts must cross at right angles at the

optimum height, namely 100-120 km. This is where the greatest level of ionization will occur. Fig.2.15 illustrates the importance of this point. There are definite limits imposed (for any station) where the required 90° optimum limit will occur, both in the North/South and East/West directions.
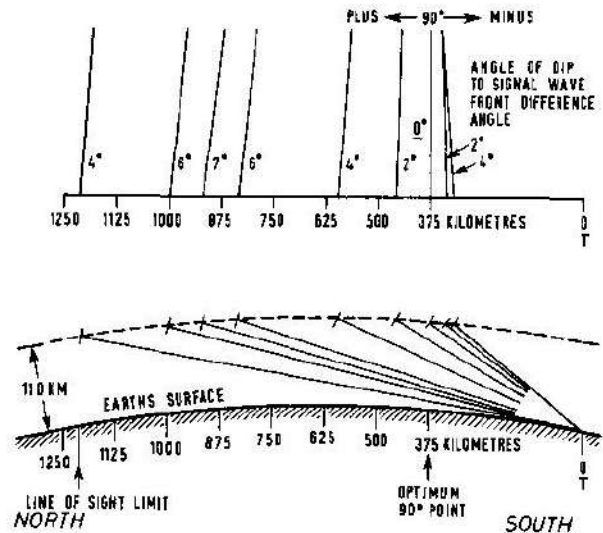

Fig.2.15. The magnetic angle of dip to signal wave front shown as a "difference angle" for a transmitted wave from point T along a geomagnetic meridian looking north, as the wave crosses the 110 km height point. Point T has been taken as London.

If consideration is given to the effect of firing a signal up, say, a geomagnetic meridian to the North, then it will be found that at a point about 375 km distant the signal would be crossing the 100-120 km height auroral sheet current high ionization point at about 90° to the field angle of dip. This will give a good reflection but as we progress farther North we shall find that the resultant angle deteriorates so that the best bistatic results will be when both stations are positioned so that they are able to use the 90° position. In practice, this means separation - East/West is preferable to North/South. If the North/South component is interposed then for one or other station the 90° angle requirement cannot be met, with resulting deterioration of signal level.

Daily Time Periods

This is fairly straightforward and fits a standard two period pattern. Fig.2.16a applies to the UK and Fig.2.16b to Sweden for two typical auroral openings. If we compare any one opening with another we will always find slight differences, both in time and character, for any one area. Some aurorae are very intense during the afternoon phase, with little or no midnight phase, or are reversed with intense midnight activity and little afternoon. Occasionally they are very intense during both the afternoon and midnight
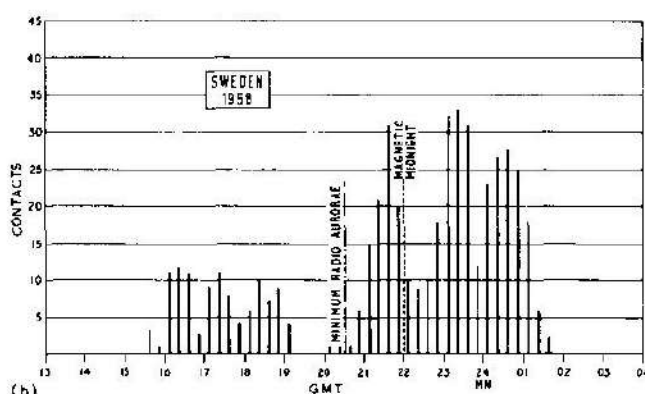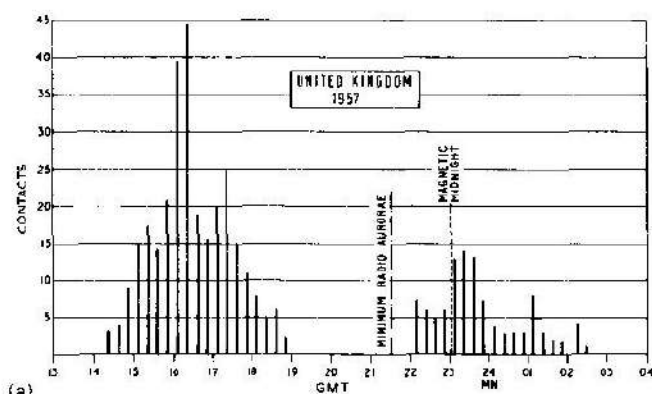
Fig.2.16. (a) This chart shows the time against number of contacts (2m) for all UK locations during 1957. (b) Time against number of contacts for a number of Swedish stations during one opening in 1958. Note the wax and wane effect.

phases. Some show a gradual build-up of activity, whereas others start very suddenly, and some last only a short period - say half an hour - while others go on for three hours at a phase.

Superimposed on the general picture is a seasonal variation which results in small time changes of the individual phases, e.g. in summer the afternoon phase tends to be slightly earlier and the midnight slightly later than at the equinoctial times, but for all practical purposes this effect can be ignored, as it has no real effect on general amateur operating.

Auroral Types

The second point which emerges from the daily time patterns is what the writer terms "wax and wane effect". This is a phenomenon apparently common to all stations in a given area. This is rather important, as many observers have been caught out by thinking that a decline in conditions was the end of an opening, when in fact it was simply a wane. An example of this is Fig.2.17 which tends to indicate that G4LX went to bed too early! The other observation of interest is the pattern of stations worked. There appears to be a definite relationship between the opening of some aurorae and the appearance of DX. In other words, an aurora tends to open with the real long range DX contacts, of up to 1200 km to Eastern or South Eastern Europe. It then declines to the more local DX, i.e. GM or near continent, PA, OZ, ON etc., from London, and closes with a flourish of the long range DX. The operating patterns of Fig.2.17 are typical examples.

Plan Field Orientation

The basic geometry necessary for bi-static auroral reflection showed how large strictly North/South distances are propagationally difficult, but if we

consider the plan component, then the picture is much more hopeful.
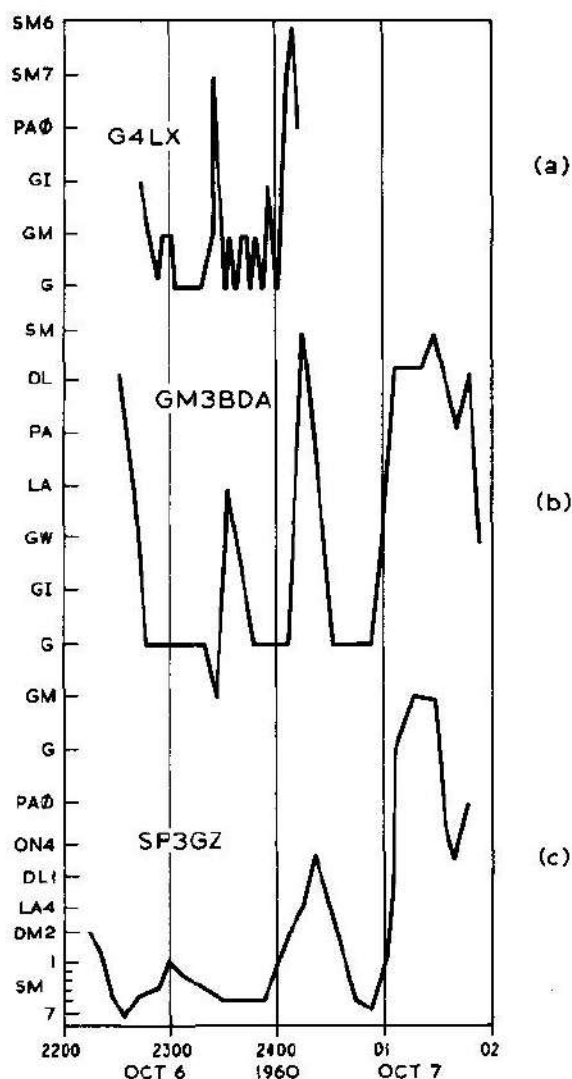

Fig.2.17. (a,b,c) Some DX patterns of the midnight phase opening of 6-7 October, 1960.

Fig.2.18 shows the plan field orientation pattern to signal wave front angle for a single station. The shaded area is where the best signal wave front to magnetic field orientation angles are, at the 110 km height. The line where the transmitted or receiver wavefront meets the earth's magnetic field angle of dip at right angles (the specular angle) is shown as 0° with plus and minus variations to this. In practice, the strongest signals will be received from the shaded area, the optimum being along the specular angle. However, we are not concerned with just our own position but with that of the other station as well. This modifies the picture somewhat in relation to those two stations.
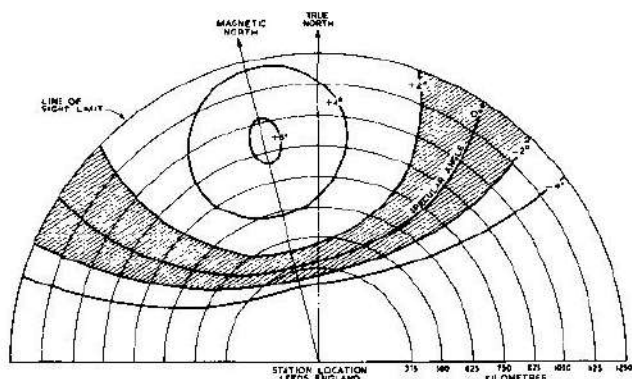
Fig.2.18. This shows the plan field orientation pattern for a single station at 110km height. It can be considered as the transmitter and receiver orientation patterns overlapped or superimposed. The shaded area is where the best orientation angles are. The specular angle line is where the transmitter or receiver wave front meets the earth's magnetic angle of dip at right angles at the 110km height, with the variations as plus and minus to this. In general practice it is not possible to use more than about plus or minus 5° from the specular angle, but it must be remembered that the ionized area extends about 20km in height which modifies the pattern somewhat.

The transmitted plan pattern has, in effect, been moved in relation to the received one, so the patterns are no longer superimposed, but rather overlap one another, the precise area of overlap being a function of each station's location. In other words, bi-static radio auroral contacts can only occur if the plan field orientation pattern of one station suitably overlaps the plan field orientation pattern of the other, and if this area of sky is ionized.

Fig.2.19 shows the overlap of the respective patterns for stations in London and Stuttgart. It is clear that contacts are possible, and if the London station beamed North-East the specular reflection line could be used over a considerable distance, thereby improving the chance of greater signal strength. The Stuttgart station's field pattern is such that it cannot better 3° to 6° off the specular line in the overlap area. The beam headings shown in Fig.2.19 were those

actually used by the respective operators, and are about the best compromise that can be made between the respective field lines with 30° beam widths. The importance of this contact is that it shows a typical "East of North" bias for the UK station's beam heading. Many UK operators expected the best signals to come from North or even magnetic North.

This is not the case, however, as the optimum beam heading depends upon the location of the two stations and their respective plan field orientation overlap. If both stations can use an area of sky, which allows specular reflection to both of them then the greatest levels will result. But if either location is changed, then the best area of sky is that which is as near the specular as possible, though with resultant deterioration of signal levels.
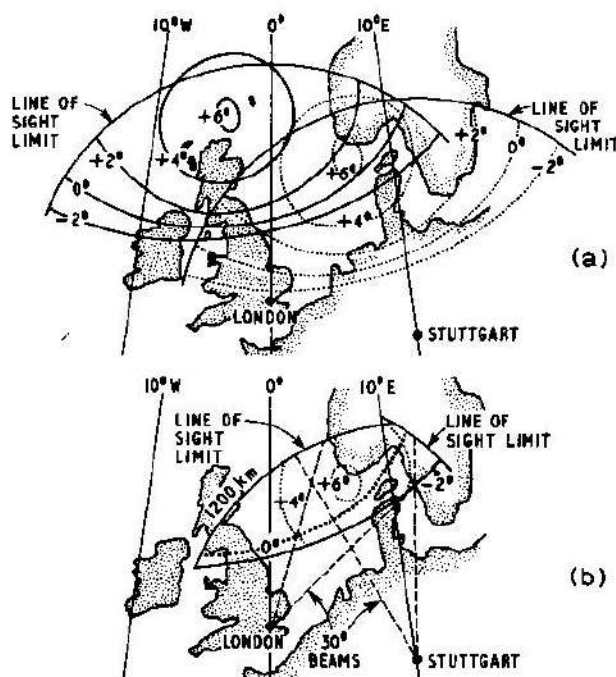

(a)

(b)

Fig.2.19. (a,b). This shows the plan field orientation patterns for stations in London and Stuttgart, the overlap area, and the respective operators' beam headings at the times of contact. The line of sight limit is taken as 1200km, beams (2m) 30° wide. The dotted line is specular reflection for the London station

It cannot be stressed too often that the opportunity to work any particular DX station, when the strong ionization is in the right place, does not last very long. If, say a London station hears a Scottish station working into Poland, it does not follow that it is the best time for him to try as the very intense ionization necessary for these contacts appears to be small in area, and, owing to the earth's rotation, will sweep over the backscatter region.

It must be understood that *the general auroral ionization region is very large in comparison with the area that can be used by any two stations owing to*

their field orientation requirements. Also, the earth's rotational and magnetic axes do not coincide, which means that the ionization area is swept from East to West, and to a lesser degree, North to South, as far as Europe is concerned. As the Kp planetary figures climb, the aurora spills out farther South anyway. This total effect therefore brings suitable reflection areas into view for any particular pairs of stations at only certain phases of the opening. However, opinions may well diverge on this point and further observations will be required before a more complete understanding of this phenomenon is possible.

The wax and wane effect could also be superimposed on this East/West, North/South drift, and tends to break up the DX pattern into geographically workable area.

Fig. 2.20. This shows how the solar "X" rays can increase the ionisation levels in the summer hemisphere due to the higher concentration.

SPORADIC "E"–"E's" AURORAL "E"

When writing about Sporadic "E" in the sense generally understood by the v.h.f. man, this usually means a super opening with very high signal levels on two metres with the opening mainly to Southern Europe. To be strictly correct, the precise mechanism responsible for the very high ionization necessary to achieve these m.u.f's is not really understood. Many theories, such as wind shear at "E" level and the soft X-ray solar radiation have been put forward, but so far all fail to give a full explanation of observed phenomena.

Time Patterns

Limited observation by amateurs and others has shown a definite pattern. Firstly, this is a "summer hemisphere" phenomenon, usually showing itself between May and September during daylight hours. At two metres Sporadic "E" is fairly rare, occurring only about 0.1% of the total "E" time, but the openings when they occur are usually very spectacular. From, say, the Midlands, Northern Italy, Yugoslavia or equivalent distances in Southern Europe, are the sort of ranges one can expect with signal levels of a very high order.

The openings, when they appear, come out of the blue and disappear just as quickly, usually lasting only a few hours. The patchiness of the ionized cloud and its movement can usually be observed as different DX areas appear in turn. It is from the UK a phenomenon extending mainly South West to South East. There is no case on record to the writer's knowledge where such a condition occurred to the North, although this could be due to lack of v.h.f. amateurs on the few islands and in Northern Norway which in the main make up the only populated areas.

The fact that all types of Sporadic "E" predominantly appear as a "summer hemisphere" phenomenon tends to support (in the writer's view) the solar soft X-ray theory for enhanced ionization at "E" level, (Fig.2.20). If we now lower the frequency from 144 MHz to 28 MHz it is possible to show daily time patterns, Fig.2.21, which clearly show the daytime preponderance, but it must be borne in mind that at 28 MHz we are dealing with what most amateurs refer to as "E'S" rather than Sporadic "E". However, in recent times with the 70 MHz band it would appear that "E's" extend up to this frequency more often than most amateurs ever supposed, and the shorter ranges 700-1000 km where "E's" first makes itself evident, has shown that Northerly working, Southern UK to Scotland and Shetland Isles is possible as well as DX TV reception from a very large area of Europe In fact, in the writer's view much of this lower frequency "E's" seems to come down from the North East and may, in fact, be born in the Arctic skies, which during summer months are solar illuminated 24 hours a day so therefore can pick up whatever soft X-rays are going. There is also overwhelming evidence that a type of "E's" which the writer prefers to call "Auroral E" accompanies Radio Aurora, but again only during the summer periods and again at the lower frequency.

"Auroral E", though mainly keeping to the daylight time pattern can occur during the early hours, but keeps very well to the Synodic Solar Rotation Pattern and is noticed when Auroral propagation is also prevalent. One theory put forward to explain this is that high velocity electrons, which are contributing to the Auroral Ionization, cause X-rays to be emitted by collision with gas molecules (at about "E" level height), and this X-ray radiation in turn causes clouds of ionization which drift down from the North. Whether "Auroral E" fits a two period Daily Time Pattern like Radio Aurora is something not yet known.

The *Scientific Studies Committee* of the RSGB are paying particular attention to many aspects of "E" phenomena, in particular the correlation of Northerly
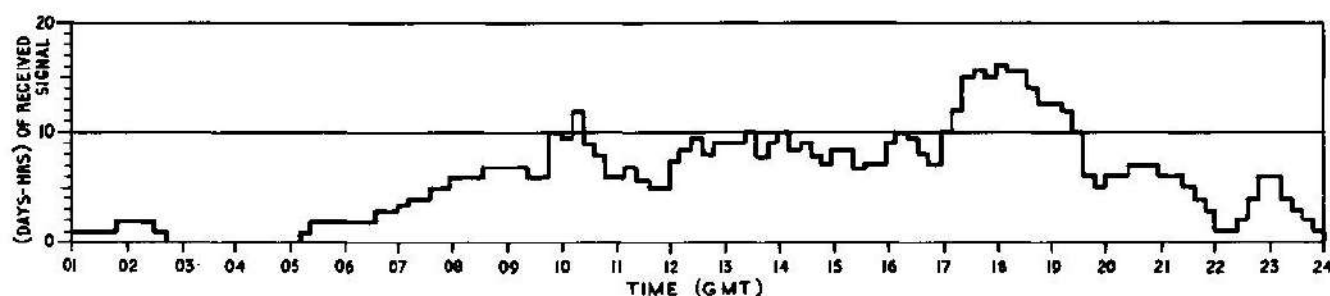
Fig.2.21. This shows the daily time patterns of GB3LER on 29005 kHz as recorded in London from May to August 1965.

and Southerly paths from various points in the UK and just over what frequency range can we expect what results. This is a field which the v.h.f. man has sadly neglected, but which may prove very profitable with a better understanding of the various phenomena.

SCATTER PROPAGATION

Both mediums, Ionospheric and Tropospheric, under certain conditions can be considered as unhomogeneous, i.e. they consist of blobs of material as distinct from a uniform media. It is this patchiness which gives rise to what is known as Scatter Propagation. True Ionospheric and Meteor Scatter differ insomuch that, in amateur circles with the limited transmitter power and antenna gain available, in practice, (at v.h.f.) only Meteor Scatter is possible. It is not strictly true to say Scatter when dealing with meteors, as the technique is a bit like Auroral, there being a relationship between the angle of the meteor trail to the wave front angles of the respective stations to obtain reasonable chance of making a contact. This is further complicated by the meteor particle size and speed as these factors govern the level of ionization which a given particle can produce, i.e. a slow burn-up means a longer trail of weaker ionization whereas a very fast entry and burn-up tends to concentrate it. Obviously particle size is important, the larger the better. In practice most particles are about the size of sand grains with a few bits of grit thrown in so in practice meteor scatter contact can only be successfully accomplished under Meteor Shower conditions i.e. such as the Perseids, the shower names being the sky star constellation points from which radiants the showers appear to come, as there is an optimum angle of meteor trail to Wave Front Angle, and showers appear to come from fixed radiants, it follows that some paths i.e. East/West are better than

others at a particular time, also the best compromise of greatest usable meteor angles is about 7° off the mid point position at "E" level height (110 km).

As the whole science of meteor scatter communication is rather involved, and the difficulties for the average amateur very formidable, a full detailed explanation of the mechanism and techniques required is beyond the scope of this manual and no one should seriously take up meteor scatter work without making a detailed study of all the relevant literature on the subject.

Tropospheric scatter is the only reliable everyday propagation method open to the v.h.f. man who desires to work skeds over rather long paths 250 miles plus. Signal levels are very weak but very consistent. Of course, it takes skill, high power, high gain aerials, first class receiver, DX ears and c.w. is essential. The technique is to aim at the other station with as low an angle of radiation as you can get. The mechanism of propagation being a Forward Scatter one, due to the varying refractive index of the multitude of blobs which make up any air mass.

If schedules are maintained with any particular DX station at say 144 MHz it will be found that a certain minimum signal is always present, with extensions of improvement, as normal tropospheric conditions improve. Of course, to get the best out of any scatter propagation circuit it is essential to have a frequency stability of an extremely high standard, both for the transmitter and receiver. The receiver calibration must be beyond doubt since very narrow receiver band widths are essential to obtain usable signal to noise ratios at the very weak signal levels encountered. In practice, band searching is not a practicable proposition for any scatter propagation, with the exception of auroral backscatter where signal levels are reasonable.

[1] VHF Weather. *RSGB Bulletin.*
    Part 1, March 1963
    Part 2, March 1964

TUNED CIRCUITS

The choice of type and construction of a tuned circuit will, of course, depend very much on the resonant frequency, and will vary from a lumped circuit (conventional inductor and capacitor) at 70, 144 and possibly 435 MHz, to linear circuit (tuned line) and on to cavity resonators.
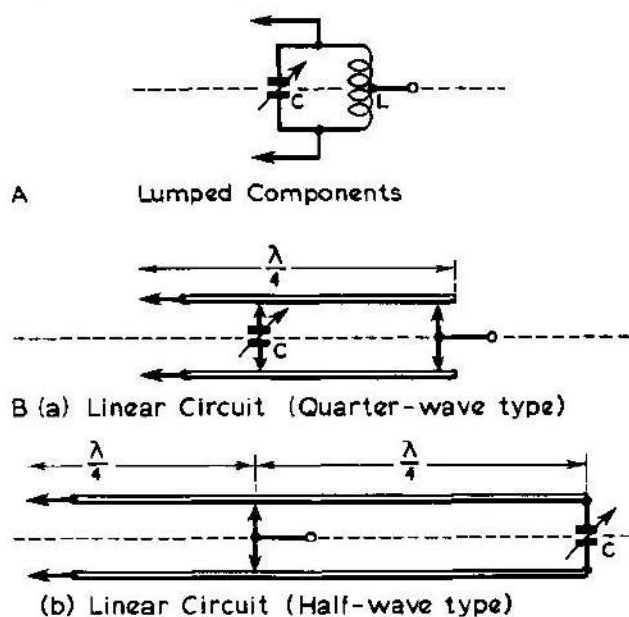
A   Lumped Components

B (a) Linear Circuit (Quarter-wave type)

(b) Linear Circuit (Half-wave type)

Fig.3.1. Three examples of tuned circuits.

Also, the choice on 144, 435 and 1296 MHz will to some extent depend on the power level at which the circuit is to be used, and if it is required for a power amplifier where efficiency is of prime importance, then the more elaborate circuit is justified. While this may be true for the two higher bands, there is little point in using a linear circuit at 144 MHz because the increase in power output resulting from higher efficiency is not really worthwhile. There are however some cases where the linear circuit is more convenient on account of better mechanical construction, this is usually the case where higher power valves or transistors are being used. In such cases quite often the output capacitance is quite high and therefore with a lumped type of circuit (coil) it would be quite difficult to get good circuit efficiency. Here a linear circuit is useful, as with in certain air cooled valves where the use of a large diameter tube for the anode line enables the proper air flow conditions to be met.

Another advantage of this type of circuit is that it can become part of the cooling system and this is of considerable importance in types where conduction cooling of the anode is employed.

In the linear type of circuit it is often an advantage to use a half wavelength instead of the more usual quarter wave, this is particularly useful when the amplifier device has a high output capacitance necessitating significant shortening of the line. In the half wave line it is usual to tune the end remote from the amplifier and to feed the h.t. in at a point approximately a quarter wavelength from the "open end". Fig.3.1 illustrates these circuits.

There are, of course, some cases when compromise is necessary between the parallel tuned lumped constant circuit and the linear type, in these cases series tuned circuits may be resorted to as means of getting an efficient tuned circuit.

Push-pull configuration has been shown in the above circuits but they are equally applicable to single ended use. In this case the circuits become half that shown, as indicated by the dotted lines.

Another form of circuit which is generally used in the higher frequencies, 400 MHz upwards, is the resonant cavity. Fig.3.2 illustrates the development from a lumped constant circuit, through the linear circuit into a cavity.
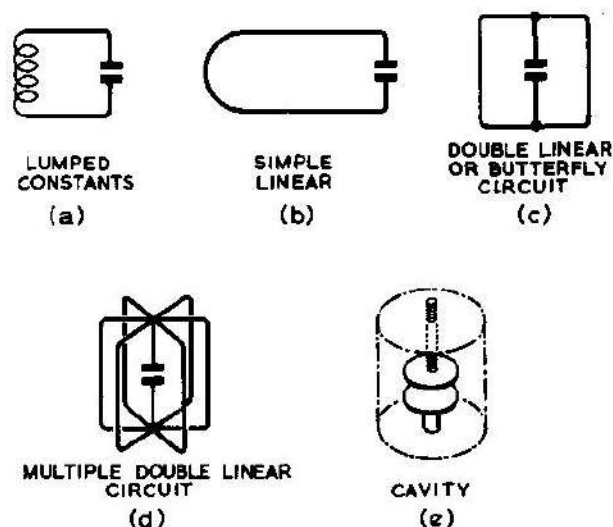
LUMPED CONSTANTS (a)   SIMPLE LINEAR (b)   DOUBLE LINEAR OR BUTTERFLY CIRCUIT (c)

MULTIPLE DOUBLE LINEAR CIRCUIT (d)   CAVITY (e)

Fig.3.2. Illustrating the development of the cavity from the original lumped constant circuit.

Another circuit which is similar to the cavity, is the trough line. This has the advantage that it is very much easier to construct and if closed in on the open side becomes in effect a square section cavity.
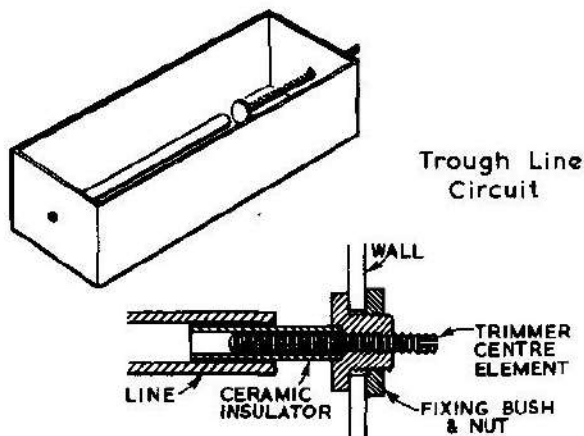

Trough Line Circuit

Fig.3.3. Trough line circuits. The tuning capacitor may be either in line with the inner conductor (resonant line) or attached to the side wall of the trough. Another convenient method is to use a tube for the line and to fit into the end of it a ceramic trimmer so that tuning is effected by adjustment of the trimmer centre element. In this case the outer element can be removed or connected to the line as required.

At the higher frequencies, conventional tuning capacitors are either inconvenient or too large mechanically, or both, and it is better to use simple disc capacitors. The size will depend on the value required, and this can be ascertained from the chart below.

Series tuned circuits, which among other applications are suitable for use where the driving source has a relatively high capacitance, such as anode output capacitance, can be used conveniently to feed a push pull or balanced output if the series tuning capacitor is of similar value to that of the source.

As shown in Fig.3.5, Co (out) should be equal to Ct. In this case resonance must be obtained by adjustment of the inductor or an additional resonating capacitor across it. The actual value of Ct will have to be equal to the hot capacitance (working condition) of the valve or transistor. It will need to be adjustable to enable it to be set for proper balance.
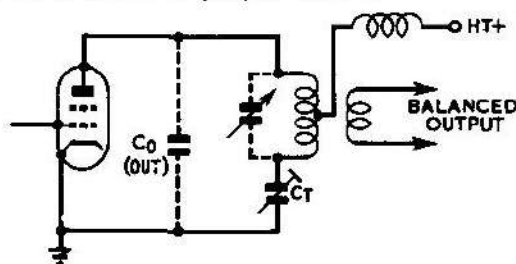
Fig.3.5. Illustrating a series tuned circuit.

Coupling to these various types of circuit should be arranged so as to transfer the maximum amount of energy either into it or out from it to the load or following stage.

Both in the lumped constants and linear circuit coupling follows conventional means, either by a suitable inductor or tapping point that will provide a satisfactory impedance match. In the case of cavity or trough line circuits the method is the same, but because of the different mechanical construction may
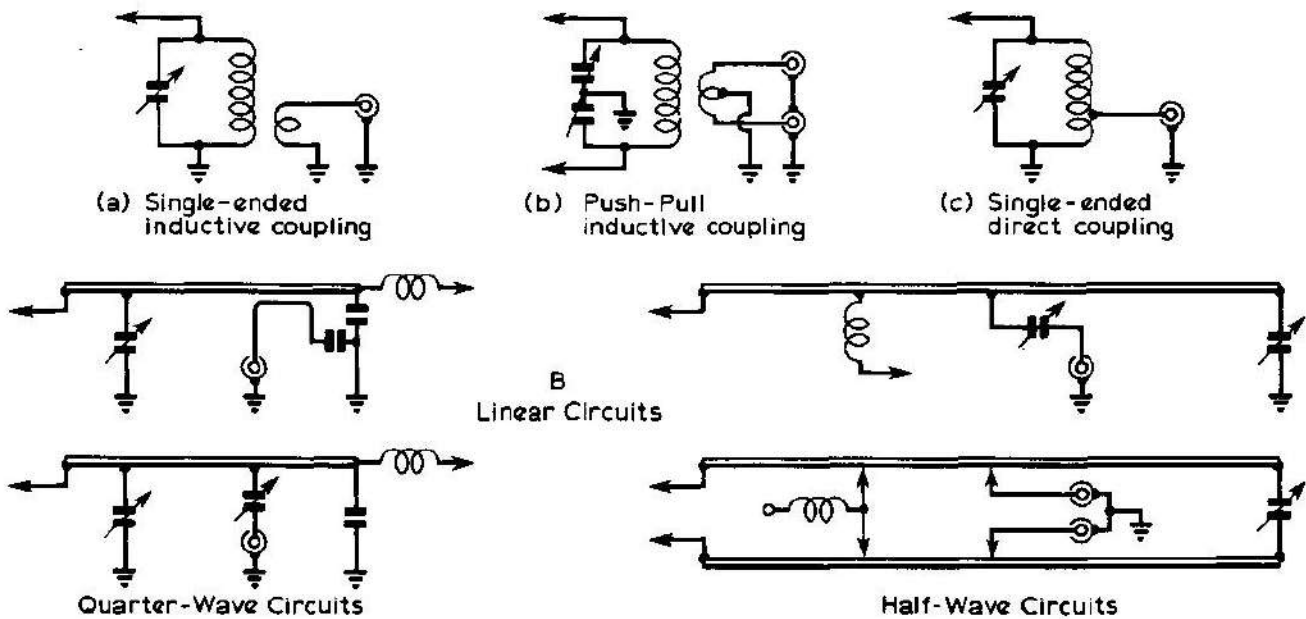

Fig.3.4. Capacitance between two parallel discs of various diameters.

$$CpF = \frac{0.244 \times area \text{ (inches)}}{spacing \text{ (inches)}}$$

(a) Single-ended inductive coupling (b) Push-Pull inductive coupling (c) Single-ended direct coupling

B
Linear Circuits

Quarter-Wave Circuits   Half-Wave Circuits

Fig.3.6. Shows methods of coupling to the various types of circuit discussed.

appear to be unusual. The diagrams in Fig.3.6 illustrate these points.

CAVITIES AND TROUGH LINES

The diagrams show the various methods of coupling, and tuning, cavities or trough lines.

(a) Shows input and output coupling using direct connection to the tuned central line.

(b) In this case capacitative coupling is shown.

(c) This illustrates inductive coupling.

(d) Mixed capacitative and inductive coupling is shown.

In all these the central line is tuned by end capacitance and there are occasions when it is convenient to cut the line to a slightly higher frequency than that of operation and resonate by a trimming capacitance. In this case the line is connected directly to the cavity or trough at both ends. This is shown in Fig.3.7 (e) and any of the coupling methods shown in the other diagrams may be used.

In these circuits there will be considerable circulating currents on the inside surface and for this reason the various parts should be a good fit. If plating is used such as silver, it should be applied to the inside surface as plating on the outside is unimportant other than to prevent corrosion or improve appearance.

DESIGN OF TRANSMISSION LINE RESONATORS

When designing a resonator to be used as a tank circuit it is necessary to know first how long to make the lines. The resonant frequency of a capacitatively loaded shorted line, open-wire or coaxial, is given by the following well-known expression:

$$\frac{1}{2\pi f C} = Z_0 \tan \frac{2\pi L}{\lambda}$$

where f is the frequency
C is the loading capacity
λ is the wavelength
L is the line length
Z_0 is the characteristic impedance of the line

(a)   (b)   (c)   (d)   (e)
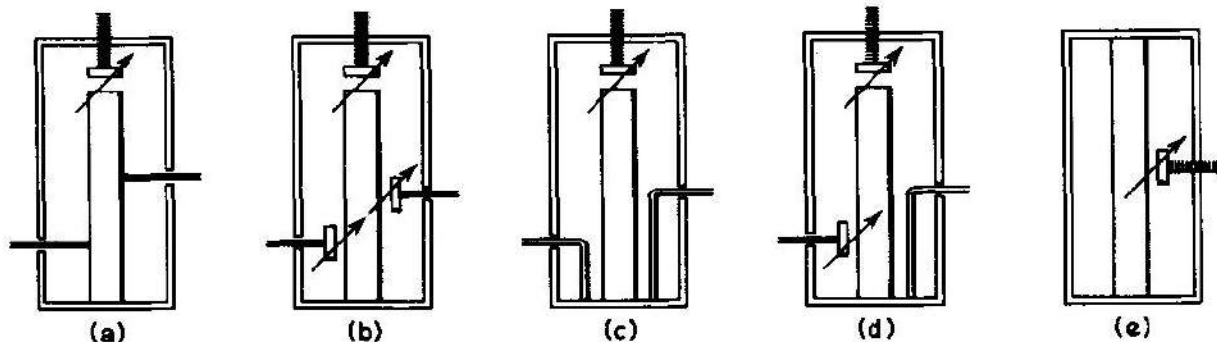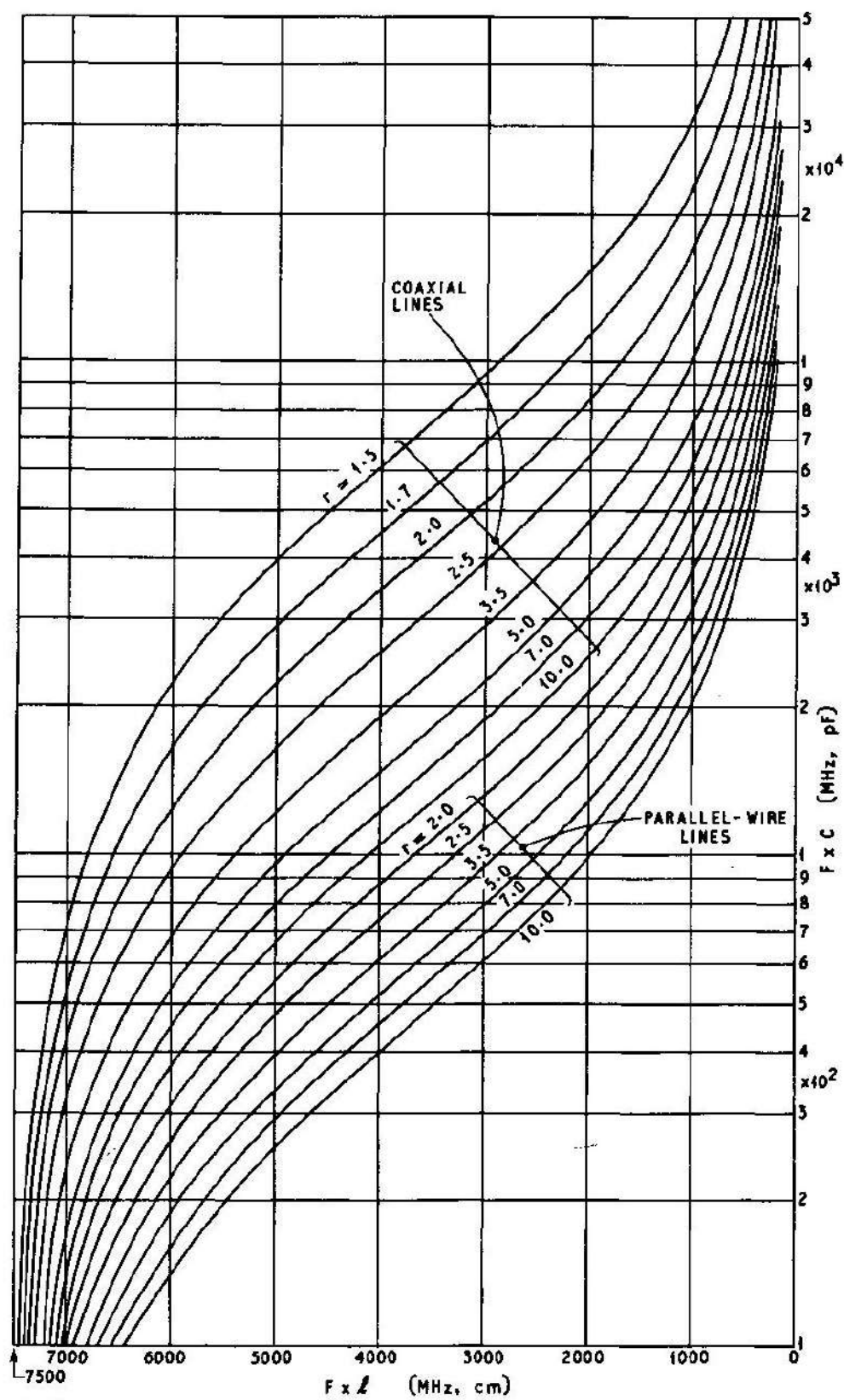
Fig.3.7. Methods of coupling to a trough line or cavity.

Fig.3.8. Resonance curves for capacitively loaded transmission line resonators.

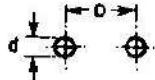The characteristic impedance is given by:

$$Z_0 = 138 \log_{10} \frac{D}{d}$$

for a coaxial line with inside radius of the outer D and outside radius of the inner conductor d

$$\text{or } Z_0 = 276 \log_{10} \frac{2D}{d}$$

for an open-wire line with conductor diameter d and centre-to-centre spacing D.

The results obtained from these expressions have been put into the form of the simple set of curves shown in Fig.3.8.

In the graphs, fL has been plotted against fC for different values of Zo, with f in MHz, C in pF and L in centimetres.

In the case of coaxial lines (the lefthand set of curves) r is the ratio of conductor diameters or radii and for openwire lines (the righthand set of curves) r is the ratio of centre-to-centre spacing to conductor diameter.

The following examples should make the use of the graphs quite clear:

Example 1

How long must a shorted parallel wire line of conductor diameter 0.3" and centre-to-centre spacing 1.5" be made to resonate at 435 MHz with an end loading capacitance of 2 pF (the approximate output capacitance, in practice, of a QQVO3-20 (6252) push-pull arrangement)?

First, work out f x C, in MHz and pF

$$fC = 435 \times 2$$
$$= 870$$
$$= 8.7 \times 10^2$$

The ratio, r, of line spacing to diameter is:

$$r = \frac{1.5}{0.3} = 5.0$$

Then, using the curve marked parallel-wire lines r = 5.0 project upwards from 8.7 x 10² on the horizontal "f x C" scale to the graph and project across from the point on the graph so found to the vertical "f x L" scale, obtaining:

$$fL = 2800$$
$$\text{therefore, } L = \frac{2800}{435} = 6.45\text{cm approximately}$$

The anode pins would obviously absorb quite a good deal of this line length but, if the lines were made 6cm long, with an adjustable shorting bar they would be certain to be long enough.

Example 2

A transmission line consisting of a pair of 10 s.w.g. copper wires spaced 1" apart and 10cm long is to be used as part of the anode tank circuit of a QQVO6-40A (5894) p.a. at 145 MHz. How much extra capacitance must be added at the valve end of the line to accomplish this?

For a pair of wires approximately ¼" in diameter spaced 1", r is about 8. Also f x L is equal to 145 x 10

i.e. 1450. Estimating the position of the r = 8 curve for a parallel wire line between r = 10 and r = 7, f x C is found to be about 1.55 x 10³, i.e. 1550. Hence the total capacity C required is given by:

$$145 \times C = 1550$$
$$C = 1550 \div 145$$
$$= 10.7\text{pF}.$$

Now the output capacitance of a QQVO6-40(5894) push-pull stage is around 4 pF in practice, so about 7 pF is required in addition. A 25 + 25 pF split stator capacitor should therefore be quite satisfactory giving 12 to 15 pF extra at maximum capacity.

Example 3

A coaxial line with outer and inner radii of 5.0 and 2.0cm, respectively, is to be used as the resonant tank circuit (short circuited at one end of course) for a 4X150A or 4CX250B power amplifier on the 70cm amateur band. What length of line is required?

In this case:

$$f \times C = 435 \times 4.6$$
$$= 2001$$

using the r = 2.5 curve for coaxial lines,

$$f \times L = 4620$$
Hence
$$L = 4620 \div 435$$
$$= 10.6\text{cm approximately}$$

This length would include the length of the anode and cooler of the 4X150A of course but, as in Example 1, a line 10cm long would be certain to be long enough, especially as the output capacity used in the calculations is that quoted by the manufacturers for the valve, the effective capacity being somewhat greater in practical circuits. A shorting bridge or disc capacitor should be used for tuning the line to resonance.

Designing for Maximum Unloaded Q

The anode circuit efficiency is given by:

$$\text{Efficiency (per cent)} = \frac{\text{unloaded Q} - \text{loaded Q}}{\text{unloaded Q}} \times 100$$

It is obvious that the highest possible unloaded Q is needed to get the greatest anode circuit efficiency. The Q is greater for radial and coaxial resonators than for comparable parallel wire circuits and the former types should always be used where possible. It should perhaps be explained that unloaded Q is the Q of the anode circuit with the valve in position and all voltages and drive power applied, but with no load coupled up to it. The loaded Q is, of course, that measured when the load is correctly coupled to the anode circuit.

The losses due to radiation which are likely with open line circuits are appreciable and screening should be used where the maximum efficiency is required. In fitting such screening, care must be taken to ensure that there is an adequate flow of air to cool the valve envelope, and that the screen is not too close to the tuned lines.
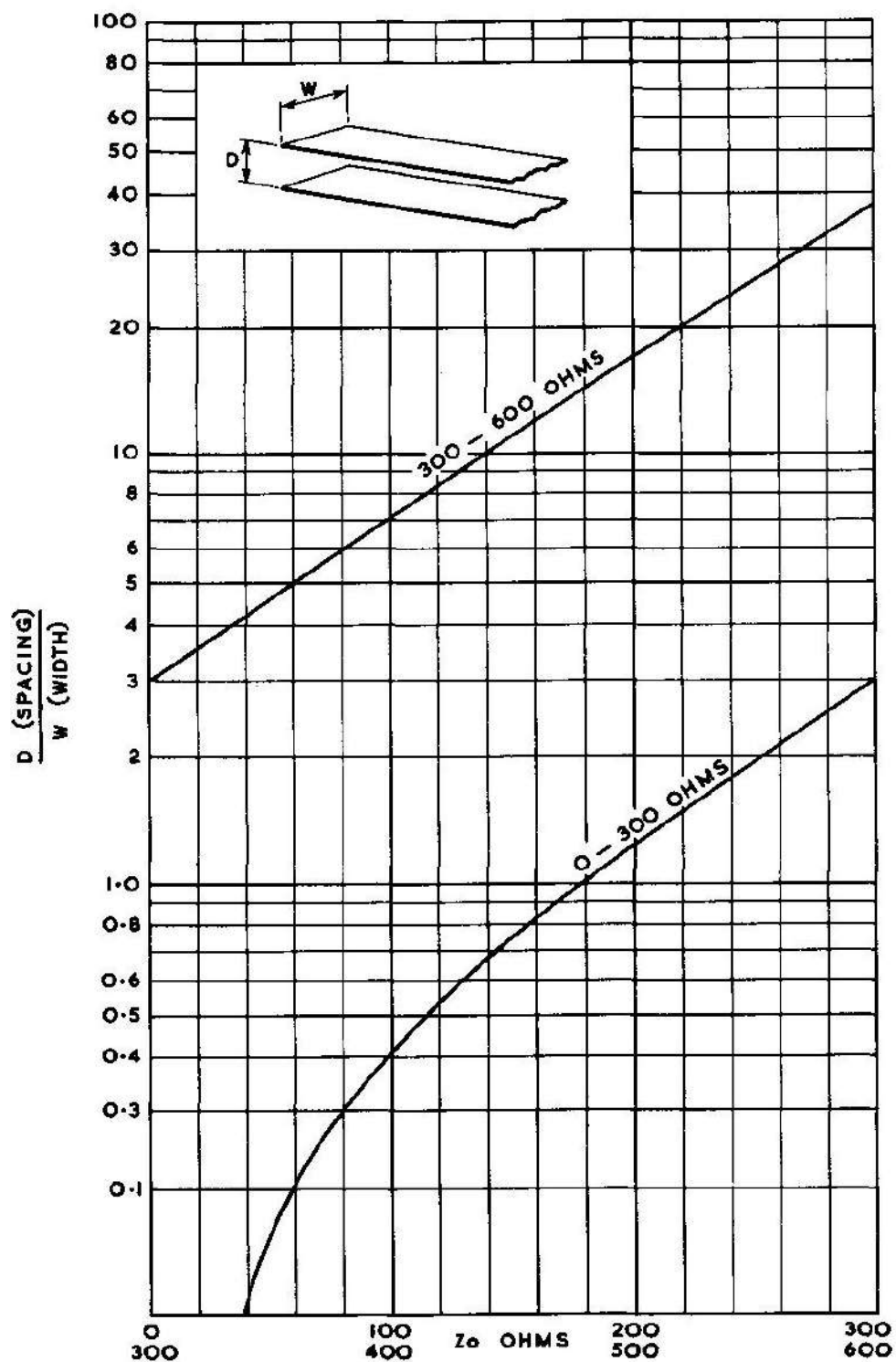
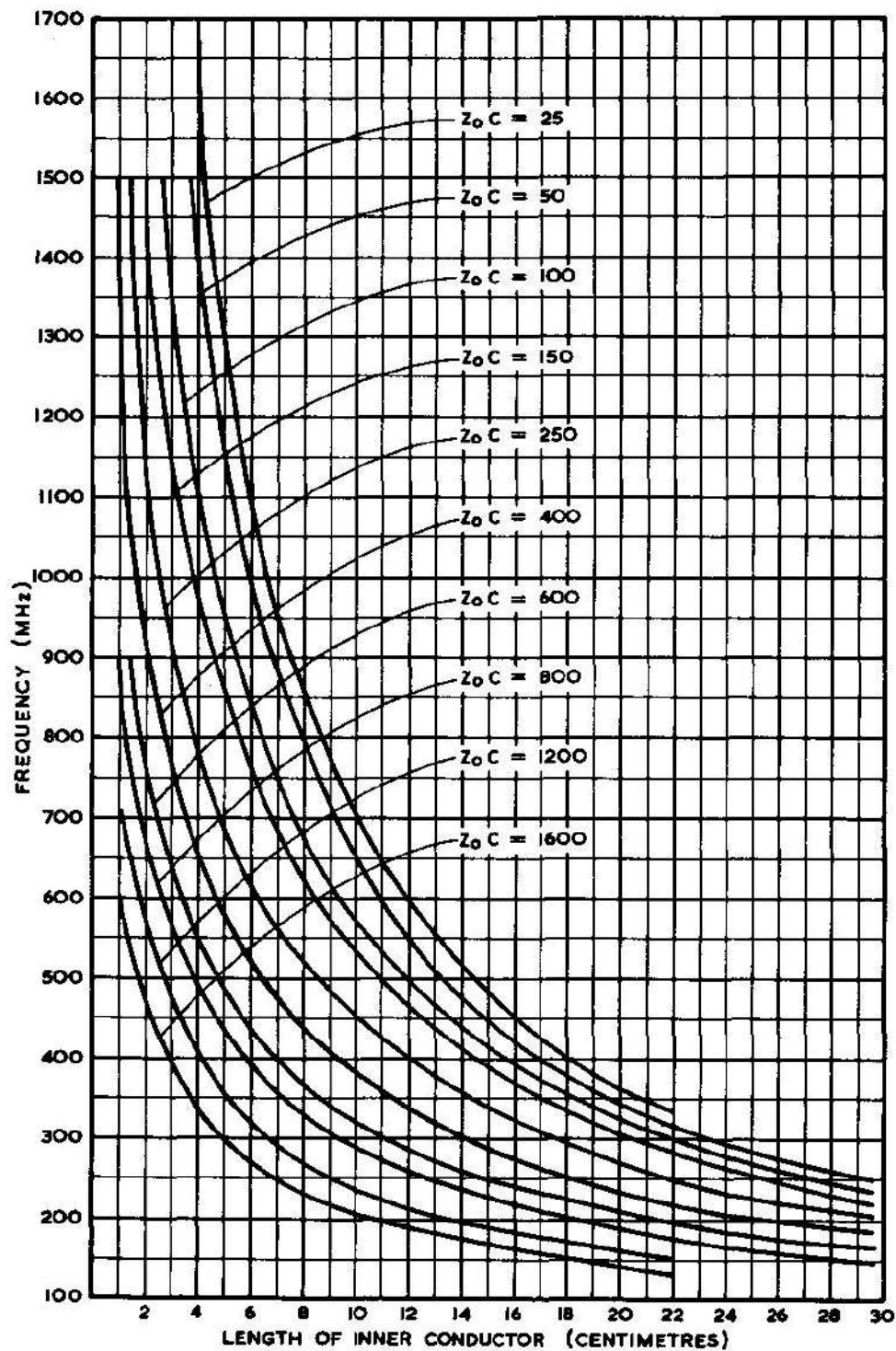*Fig.3.9. Characteristic impedance of balanced strip transmission line.*

Fig. 3.10. *Chart plotting frequency against length of inner line for various values of the characteristic impedance multiplied by the total capacitance. C is in pF and Zo in ohms.*

HELICAL RESONATORS.

This is a high Q circuit which in effect combines the lumped constants circuit with the cavity. It has not been used to any extent in amateur applications but offers mechanical advantages which justify its consideration, especially for filters. It consists of an inductor in a closed screen as shown in Fig.3.11 and is suitable for use at all frequencies in the v.h.f. and u.h.f. bands.
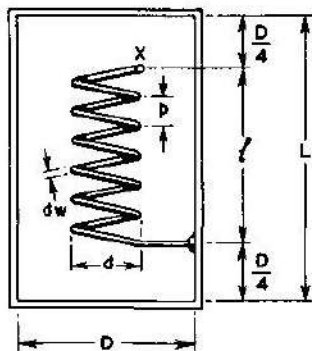
Fig.3.11. Illustrating a helical resonator.

The screen may be either circular or square cross section, allowance being made for the latter. The unloaded Q = 50Df½ where D is the diameter of the screen, but for a square screen use 1.2D″ f is the resonant frequency in MHz.

The number of turns required $n = \dfrac{1900}{fD}$ turns

Pitch (spacing between turns) $p = \dfrac{D^2 f}{2300}$ inches

Impedance characteristic Zo $= \dfrac{98000}{fD}$ ohms

These formulae assume that $\dfrac{d}{D} = 0.55$ and $\dfrac{l}{d} = 1.5$

From the general arrangement it will be seen that the helix is terminated at the lower end, on the side of the screen and the free end (X) is normally connected to a low loss tuning capacitor.

The actual value of the tuning capacitor will depend on the operating frequency and for the range over which it is required to resonate the circuits. Probably the most suitable would be one of the p.t.f.e. neutralising capacitors or other concentric type.
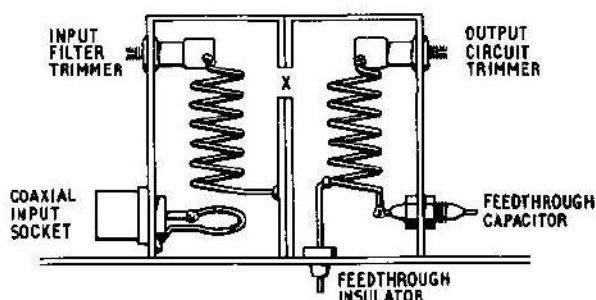
Fig.3.12. A typical arrangement of helical resonators.

The equivalent electrical length of the circuit should be about 95% of a free space length, and the actual conductor length is in the region of 28% of the free space length. Stray capacitance is approximately equal to 0.15D pF.
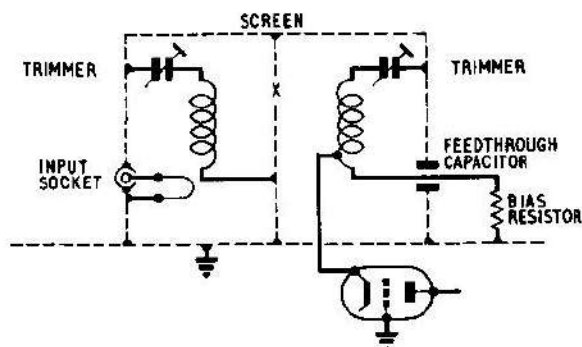

Fig.3.13. The equivalent circuit. This arrangement shows two tuned circuits as the input filter and input circuit to a grounded grid amplifier. Coupling between the filter and input circuits is provided by an aperture in the screen (X). Alternatively, a normal coupling method can be used, such as a link, taps or probe.

Design examples:

(a) 70 MHz Filter. Fixed Frequency
Screen 1¾″ x 1¾″ (inside)
No. of turns 14
Diam of helix 0.9″
Wire of helix 16 s.w.g.
Helix length 2″ without end trimmer and with silver plating on inner surface and helix, an unloaded Q of between 850 and 1000 can be obtained.

(b) 148-174 MHz Tunable Resonator (suitable for 144 MHz with slight increase of trimmer capacitance)
Screen 1″ x 1″ (inside)
Resonant frequency with end trimmer: 240 MHz.
No. of turns 7
Diameter of helix 5/8″
Length of helix 1″
Capacity to resonate at 174 MHz 1.4 pF
                              148 MHz 2.4 pF
Unloaded Q approximately 750

The high unloaded Q of these resonators require that precautions are taken to avoid the introduction of losses. Where a former is needed for the helix it must be of low loss material but it is better to utilise air spacing wherever possible.

Silver plating of the inside of the screen and the helix is desirable, certainly above 100 MHz. Silver plating of adequate thickness would give a Q about 3% higher than plain copper.

The lower termination of the helix should be as short as possible and fixed to the side of the screen. As shown in Fig.3.11 the coil should be D/4 from each end of the screen.

As mentioned earlier, coupling to or from the resonator can be by any of the usual means, such as tap, loop, probe or aperture. When loop coupling is used it should be approximately perpendicular to the axis of the helix and relatively close to the bottom turn of the helix, preferably parallel to it.

Material for Tuned Circuits

In the case of simple tuned circuits, copper wire is normally used, either bare or enamelled and is usually quite adequate. There is little advantage in silver plating except at the highest frequencies or as a protective surface. The largest convenient size should be used since in most cases the inductor will be self supporting and should therefore be mechanically stable. Tinned copper wire is not satisfactory due to its high r.f. resistivity.

Linear Circuits

These may be made from copper or brass, aluminium can be used when there are weight considerations.

As these metals are subject to corrosion in moist atmosphere, silver plating (gold in special cases) should be used as a protective coating, although there is little advantage from an efficiency point of view. Chromium plating should not be used for r.f. circuits.

Cavities and Trough Lines

Here again copper or brass are the most suitable materials and plated as indicated above, any joints should be soldered or at least close fitting, since significant currents flow around the inside of these units and resistive losses must be kept to a minimum.

Corrosion

As mentioned earlier copper and brass will readily corrode in moist atmosphere and a protective coating, usually silver plating, should be used.

In connection with corrosion, it is important to avoid the use as far as possible of dissimilar metals, because, in the presence of moisture (water vapour) electrolytic corrosion can be quite extensive, for example brass screws used to fix aluminium will show considerable corrosion after only a few weeks in the average atmosphere. It is advisable to use plated fixing screws where possible but stainless steel are also quite suitable.

Further notes on corrosion will be found later in this Manual under the heading *Material for Elements, page 9.35*.
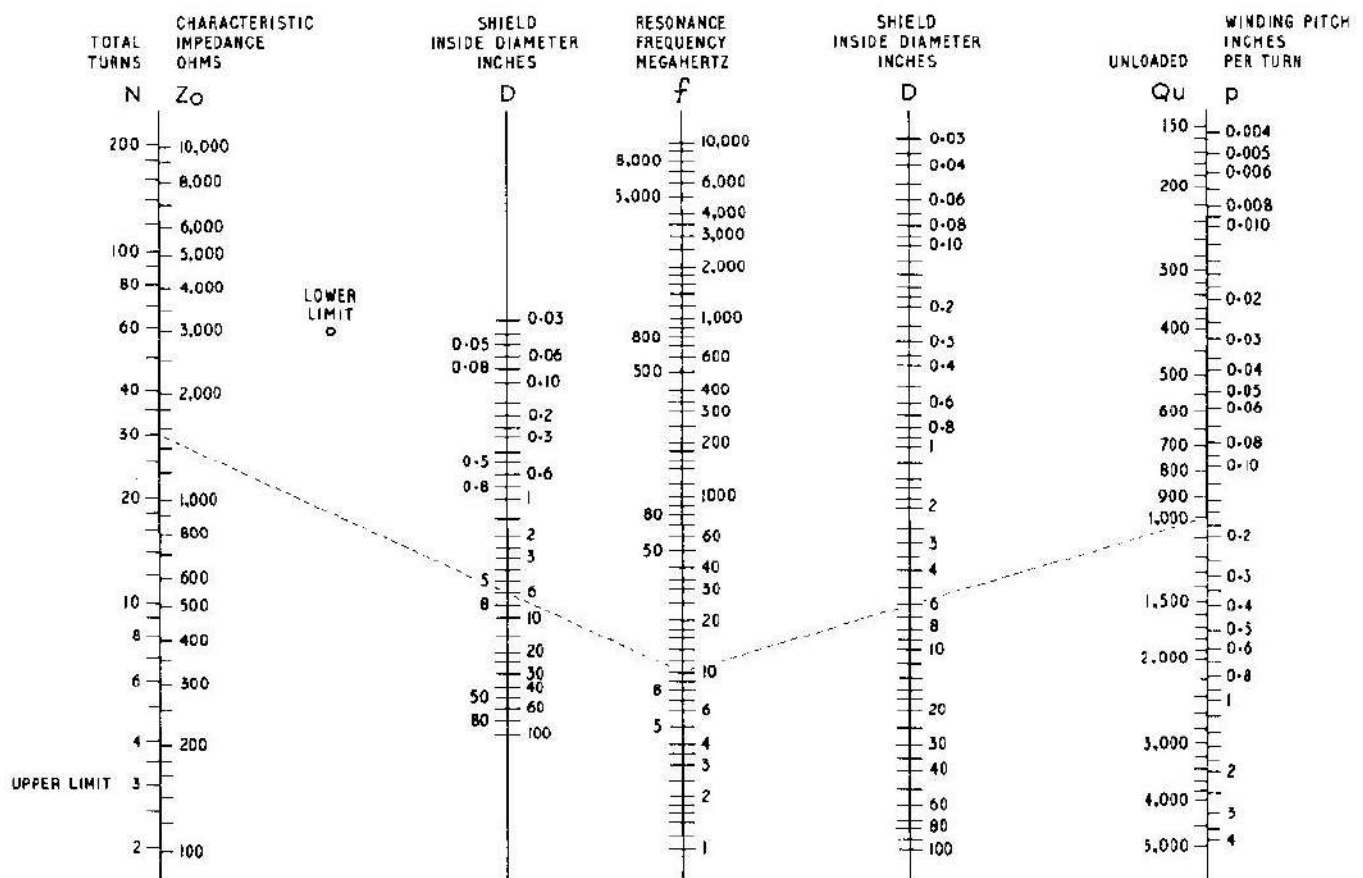
Fig.3.14. Design Chart for quarter-wave helical resonators. Lines indicate example.

FILTERS

Operation in the v.h.f. and u.h.f. region where the amateur bands are situated amongst the TV and FM broadcast bands, will almost inevitably cause some interference. The relationship between these bands is illustrated in the chart of Fig. 4.1.
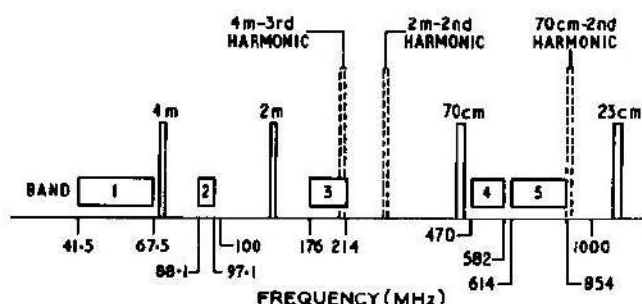

Fig.4.1. Chart illustrating relationship between the TV bands and harmonics from the amateur v.h.f. and u.h.f. bands.

The approach to overcoming this problem will have to be rather different from that normally applied to h.f. band equipment where often the matter can be resolved by use of a low pass filter at the transmitter and a high pass filter in the receiver input.

The most satisfactory method is to use some form of band pass filter suitable for each of the bands concerned. The filter may be either a multi element type with a relatively wide band, say 5 to 7% of the frequency being used, or a narrow band high Q strip line type which is tuned to the frequency in use.

The former is the more elegant approach but the design, and especially the setting up, of such a filter without fairly elaborate test equipment is not easy if the insertion loss is to be kept down to a maximum of 1dB.

The high Q strip or tube line filter on the other hand is readily made and tuned but it will also have to be adjusted for any appreciable change in operating frequency, this can be a problem when a v.f.o is used.

Either method will provide a very considerable measure of protection against out of band radiation such as harmonics and spurious responses from transistor and single sideband transmitters.
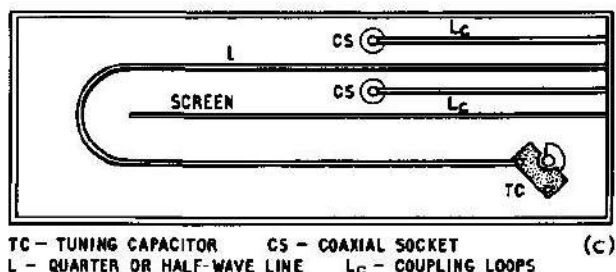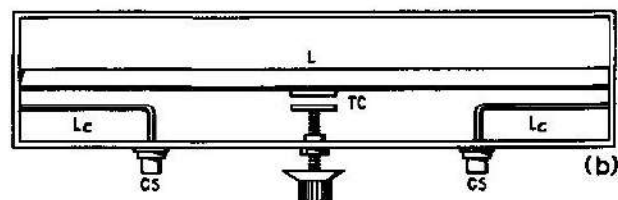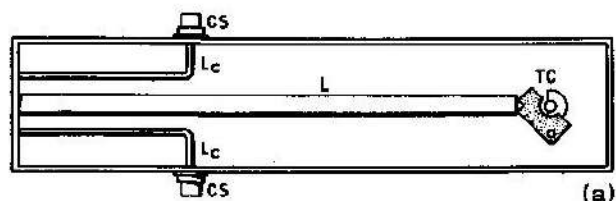

TC – TUNING CAPACITOR   CS – COAXIAL SOCKET
L – QUARTER OR HALF-WAVE LINE   Lc – COUPLING LOOPS

Fig.4.2. Typical forms of high Q filter. (a) Quarter-wave line with tuning capacitor at one end. It is suitable for any of the bands, but at 4m is likely to be inconvenient. (b) Half-wave line with centre tuning capacitor. Particularly suitable for 70 and 23 cm. (c) Quarter-wave line folded into U shape with a screen between the legs of the U line. May be suitable for 4m.

Low power Band Pass Filter

A simple band pass filter can be made by over coupling two pairs of tuned circuits and then coupling them together through a suitably small capacitor. A screen is fitted between the pairs to reduce stray coupling.
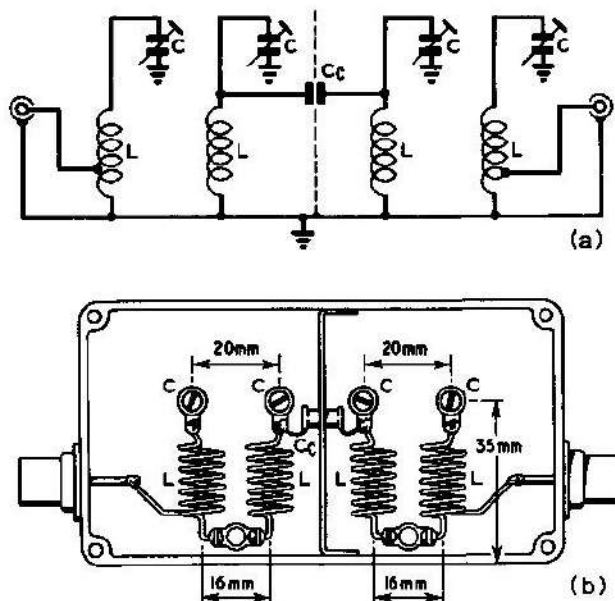
(a)

(b)

Fig.4.3. (a) Filter circuit. Cc = 0.5pF, C = 6.0pF max., L = 6½ turns 18 s.w.g. ⅛" dia. spaced one turn. Tap at one turn up on end coils. As can be seen from the layout (b) the whole unit can be built into a small diecast box. In the prototype, concentric ceramic trimmers were used but there is space for small air trimmers which are necessary if high power operation is required. In any case air trimmers are desirable to avoid heating of the dielectric and to avoid breakdown due to the high voltage which will develop in a high Q circuit.
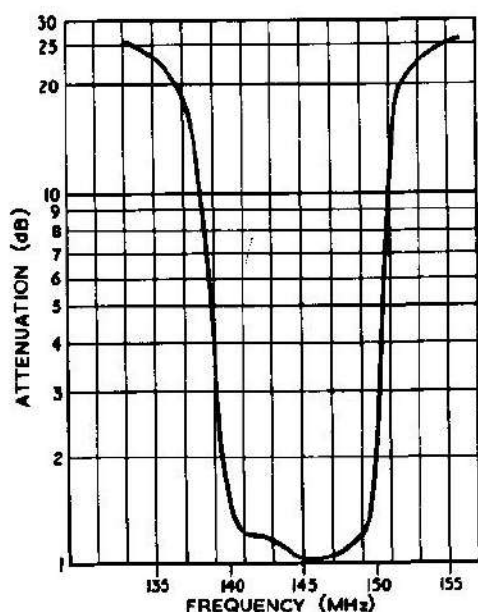
Fig.4.4. Performance curve of the filter described.

A 70cm Helical Resonator Filter

A practical band pass filter for 70 cm using two helical resonators coupled by iris in the separating screen. It has an insertion loss of less than 1dB with more than 20dB attenuation outside the pass band.
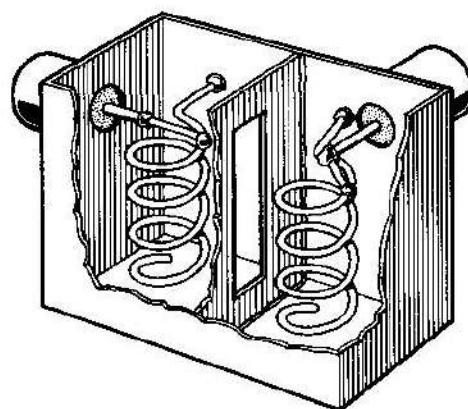
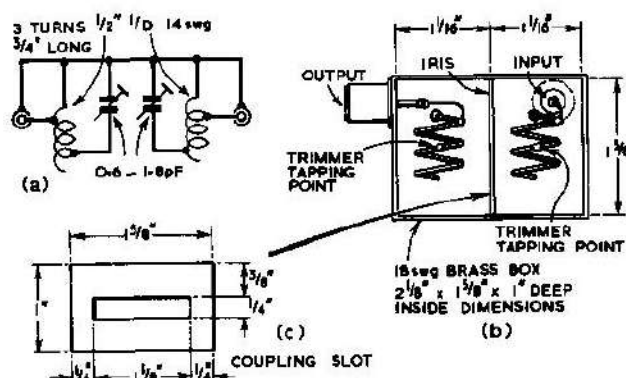

Fig.4.5. General arrangement of the 70cm helical filter.

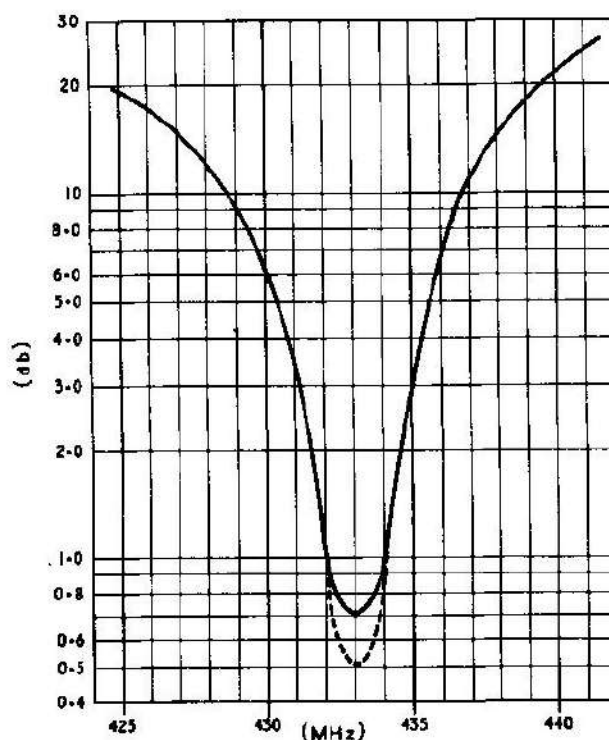

Fig.4.6. Layout of the 70cm. filter.

Fig.4.7. Performance curve of the 70cm. filter.

High Q Filters

The construction of this type of filter is relatively straightforward but the actual method will to some extent depend on the facilities available for doing the mechanical work.

For general purposes the box type structures are the easiest to fabricate and provided good joints are made to the end plates, by screwing together or soldering, a satisfactory high Q will be obtained.

The form of the circuit will generally be of near quarter-wave long strip or tube tuned at one end or, for the higher frequencies, half-wave lines tuned at the centre. In the case of the lowest frequency band (4m) the length of the filter may be inconveniently long and in this case the line may be folded provided the two legs are screened from one another.
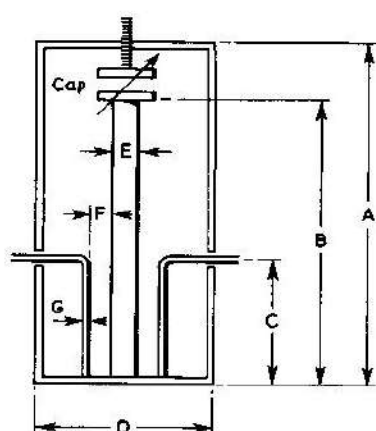

Fig.4.8. Showing a high Q filter. The end capacitor may be a disc or conventional type, except at 70cm or 23cm when a disc must be used.

Dimension (inches)	4m	2m	2m	2m	70cm
A	39	19	14	12	6½
B	36	17	12½	10	6
C	16	8	3½	3	1¼
D	3	3	4	2½	1½
E	½	½	7/8	½	½
F	¼	¼	16	1/8	1/8
G (s.w.g.)	12/14	12/14	14	14	14
Cap.(pF)	15	15	15	15	7.5

In Fig.4.2 are shown several forms of filter, the two straight types may be made up either in square or round cross section, but the folded type (c) must be restricted to a rectangular cross section.

The inner conductor (L) may be of rod, tube or strip, the latter being most satisfactory for the folded type.

The length of the inner conductor may be between 60% and 90% of a resonant quarter wave but for half wave line filters, it should be between 80% and 90%.

There is a good deal of flexibility in these filters and provided the inner conductor is properly resonated by its tuning capacitor, and the outside box or tube is properly connected with low resistance joints, then no difficulty should be encountered.

The coupling loops should be of equal size since no change in impedance is required when using a filter between the transmitter and the aerial and with careful adjustment the insertion loss should not exceed 1dB.

HIGH POWER BAND PASS FILTERS

Filters for use with transmitters of output powers up to 100-120 watts can readily be constructed using strip line tuned circuits.

The arrangement is of three shortened strip lines tuned by capacitors at the top end, input and output coupling is through small fixed capacitors to the top end of the outer lines, the third (central) line is free and couples the input and output circuits.

The strip lines are made of 1″ wide copper and in order to get these large units into the box without over coupling they are set at an angle of 45 degrees (incidentally this angle will be provided naturally if the standard type of trimmer is used with the centres given). Fixing of the line to the end wall of the box is provided by OBA brass cheesehead bolts with saw cut opened to allow the line to be brazed centrally to the end.

The top end of the line is bent to make contact with both the capacitor stator pillars, which are soldered directly to the line after assembly — a fairly large soldering iron will be needed for this.

Input and output capacitors are taken directly from the connectors to the top of the respective line. Setting up is straightforward provided an output power measuring device of some form is available. Initially C2 should be set near maximum capacity, this and the C1's are then adjusted for maximum output, taking care to keep the C1's value similar. If this is not done an asymmetrical response will result.

In the 432 MHz filter there is a response at 288 MHz, this is some 13dB down on the wanted frequency but it should be borne in mind in cases where this frequency is likely to appear in the transmitter output such as in the case of a varactor tripler directly feeding the aerial.

From the curve of the 144 MHz filter the insertion loss (at the band centre) is less than 1dB. On actual power measurement this was found to be a loss of approximately 10 per cent or 0.5dB which is a small price to pay for keeping your harmonics under control.

Although the 2m unit was intended for this band only, it is also suitable with the present mechanical details for use up to about 175 MHz.
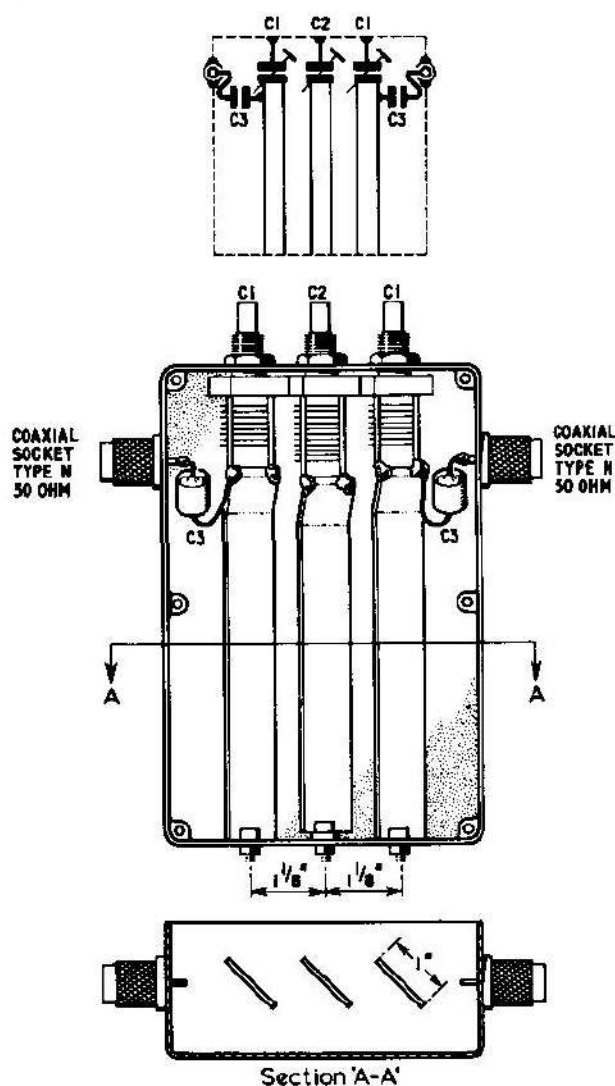

Fig. 4.9. Mechanical layout and electrical circuit of the 144 MHz filter. For 432 MHz a smaller, 4¾" x 3¾" box is used.

	144 MHz	432 MHz
Line length, inches	6¼*	2½
Line size, inches	1 x ¼	1 x ½
Line centres, inches	1⅛	1¼
C1 pF (Jackson C804)	50	50
C2 pF (Jackson C804)	60	15–0.045" spacing
C3 pF	4.4 (see text)	5
Insertion loss, dB	0.6	1.2
Bandwidth, MHz	6	10
Out of band attenuation, dB	26	26
VSWR – 50 ohms	1.1	1.1
Connectors	Type N	Type N

* Centre line approx. ¾" shorter to allow for rib in cast box and longer capacitor.
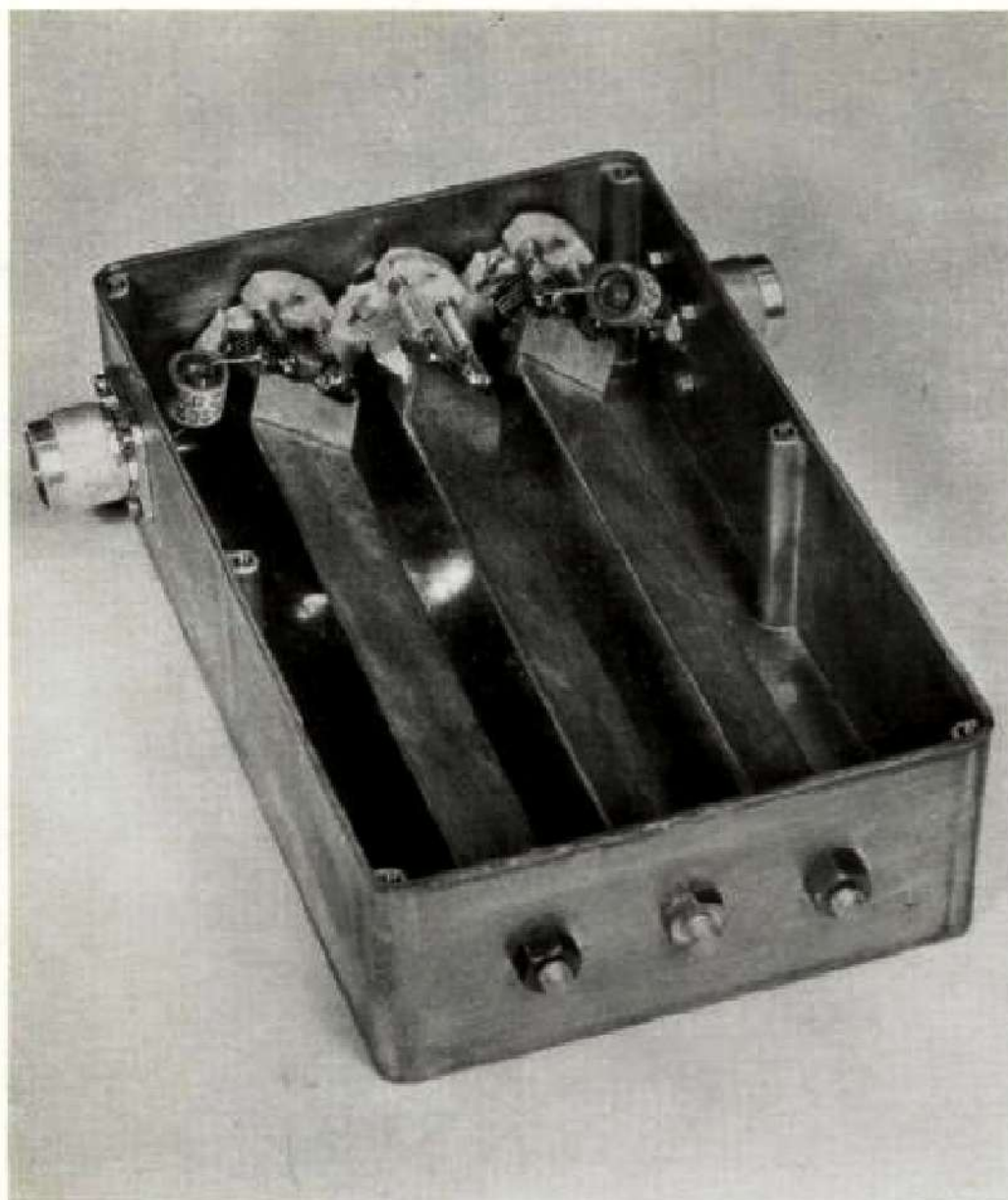
Fig. 4.10. Layout of the strip line filter.

The original dimension (C₃) is for 50 ohms line, for use with 75 ohm line the value of this component should be reduced to 3.2 — 3.4 pF. However, the insertion loss is in fact sufficiently low for its use in feeders of either impedance. The capacitors C_3 should preferably be made adjustable and standard *Philips* type trimmers of 2-8 pF are suitable. They should be about ¼" mesh for 50 ohms.

This filter will also be of benefit to the receiver, when this is subject to adjacent high strength signals such as from tv transmitters.

The centre line with C_2 set at maximum tunes to about 144 MHz. Hence when the input and output lines are tuned to resonance with it the overall response is quite sharp and is about 3dB down at ±1 MHz. However as C_2 is decreased in value the higher frequency cut-off moves upwards and a double humped response develops. With C_2 at about 10% unmeshed there is an almost square shaped response 2 MHz wide with a slight dip in the middle of around 0.1dB.
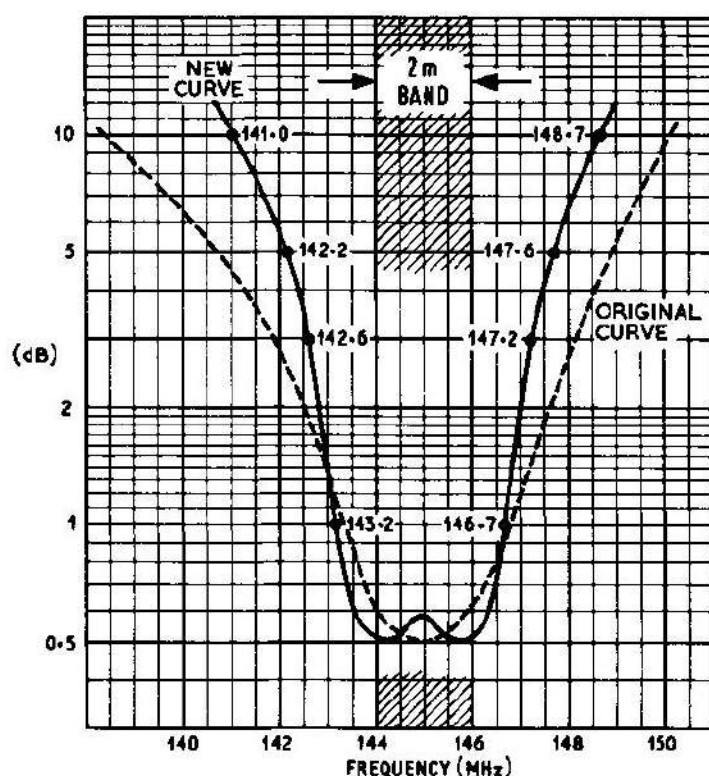
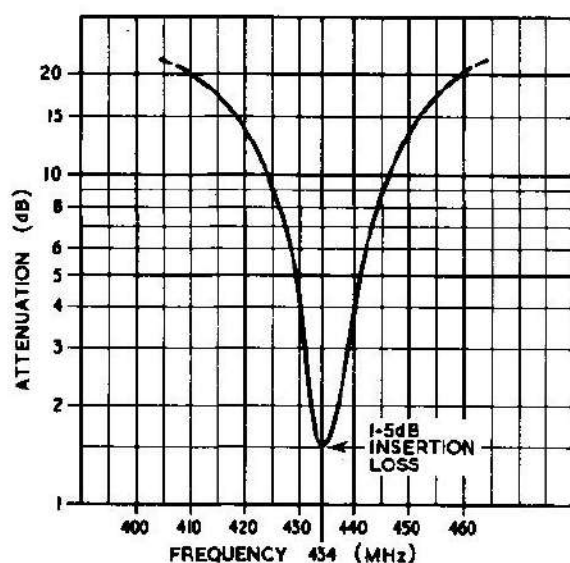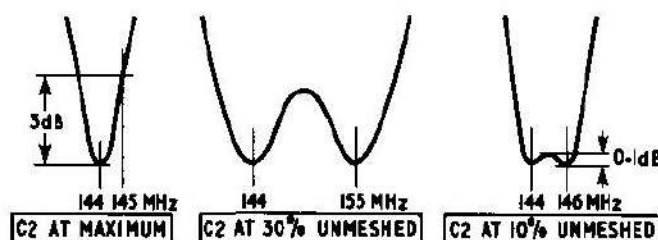Fig. 4.11. *Performance curve of the 144 MHz strip line filter.*

Fig. 4.12. *Performance curve of the 432 MHz strip line filter.*

Alignment procedure

Set the coupling capacitors (C_3, *Philips* 2-8pF trimmers) meshed by $\frac{1}{8}$" and C_2 at 10% unmeshed. Apply a signal and peak output by adjustment of C_1. Slightly readjust coupling capacitors in step retuning C_1 until maximum power transfer is achieved. Replace and fix the box lid before the final adjustment of the capacitors C_1. If transmitter output at several frequencies in the band is available, make the adjustments at the centre of the wanted range. If after this the higher frequencies show reduced output, reduce the capacity of C_2 slightly. If however, there is some loss at the lower frequencies, retune both C_1 to optimise the lower frequencies, and adjust C_2 to maintain output at the higher wanted frequency.

COOLING

Cooling of valves and semiconductors is important and the makers' recommendation should be observed if satisfactory performance is to be obtained.

Temperature measurement of valves is relatively easy but this is by no means a simple matter in the case of semiconductors particularly as they are sensitive to many forms of overload. Often heat sinks become cumbersome and the situation can be significantly improved by use of a cooling fan to improve the heat exchange.

Materials with higher thermal conductivity are the most suitable for heat sinks. It is however important to ensure that there is no significant contact potential between the device and the heat sink material otherwise electrolytic corrosion will occur. Coolers made of metal other than copper or aluminium will need to be of greater thickness to give the same performance.

The thermal efficiency of a heat sink may be improved when the surface is roughened and covered with matt black paint, such as lamp black, or anodising. Care must be taken to ensure that the area in contact with a semiconductor is smooth and free from burrs and other irregularities.

The use of a silicone grease may be helpful in the contact area before assembly, this will reduce the thermal resistivity and assist in preventing corrosion.

The characteristics of suitable materials are shown in the table below:

Metal	Relative Resistivity	Thermal Expansion
Aluminium	1.64	25.5×10^{-4}
Brass (typical)	3.9	18.9×10^{-4}
Copper	1.0	16.7×10^{-4}
Silver	0.95	19.5×10^{-4}
Gold	1.416	14.3×10^{-4}
Tin	8.72	26.9×10^{-4}
Nickel	7.8	12.8×10^{-4}

CHAPTER 5

RECEIVERS

RECEIVER REQUIREMENTS

The receiver needed for satisfactory reception of v.h.f. and u.h.f. signals is generally similar to that used for the h.f. bands. The most usual arrangement is to use a communications band receiver as a tunable i.f. and to put a crystal controlled converter ahead of it. By this means a high gain receiving system can be set up, the choice of tuning range to suit the converter can be anywhere within the frequency range of the receiver. It is, of course, possible to arrange for the local oscillator in the converter to be tunable and to use a fixed setting of the main receiver, but this raises problems of frequency stability and repeatability of the converter oscillator.
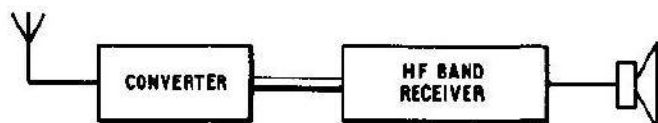
Fig. 5.1. Block diagram showing a converter and receiver.

Satisfactory tunable oscillators can, of course, be made but they need to be very robustly constructed and should be at a suitably lower frequency to that of the required final frequency in order to give the best chance of obtaining the necessary stability. Oscillators anywhere between 8 and 72 MHz have been successfully built and used and the final frequency arrived at by use of frequency multipliers. In this respect oscillators for receiver local oscillator or transmitter control are the same except for the actual frequency and perhaps the power output.

Details of various forms of variable frequency and crystal controlled oscillators are given under the heading *Oscillators* later in this chapter.

When a general coverage h.f. bands receiver or a good multiband broadcast receiver is used as a tunable i.f. amplifier, the frequency of the main i.f. amplifier will be 455/465 kHz or lower and since most of the bands we are concerned with are 2 or more megahertz wide (4m band only 675 kHz) image signals of the wanted signal will occur at twice the i.f. and since

these will be only about 900 kHz (455 x 2 or 465 x 2 or less) from the signal frequency they will be within the tuning range of the receiver.

Such signals will, of course, be worse from local transmitters and can be a considerable nuisance, as well as the unwanted signals which occur at half the intermediate frequency. A preferred i.f. amplifier frequency will be one which certainly eliminates the image responses and the 1.6 MHz region appears very suitable, especially as crystal filters for single side band operation are now available.

The use of 1.6 MHz as the i.f. is suitable for bands 2 MHz wide but will suffer similar disadvantages as 455 kHz, when related to the 70cm band, even if only the harmonically related section to the 2m band is used, i.e. 432 MHz to 438 MHz, 6 MHz wide.

From this it is obvious that for these higher frequency bands an ideal receiver would have to be made for each different band. It is therefore necessary to compromise and if 1.6 MHz is adopted both the 4m and 2m bands will be clear of image responses, (it could equally be used for 10m) and such a system would at least be better than one at the lower frequency. As increasing the i.f., although giving an improvement in the above respect, reduces the selectivity that can be obtained from a standard transformer, and it is necessary to increase the number of tuned circuits to recover the selectivity. This can conveniently be done by using two transformers in place of one, arranged back to back.

Another point to be remembered is that the gain of a single stage at 1.6 MHz will be lower than one at 455 kHz and therefore, to obtain the same overall gain from the i.f. amplifier it may be necessary to use three stages instead of two.

Selective filters such as mentioned earlier, can be used for single side band, but if these are not used, a Q Multiplier instead can give a real improvement.

As already mentioned, a receiver for v.h.f. and u.h.f. will, since the bands are of different width, be a compromise. The actual frequency used for the main i.f. amplifier will depend on the image frequencies likely to be met and if it is essential to avoid these then there is no alternative but to use a separate system for each band.

However, some form of compromise is possible which offers some advantages over the general use of a standard general coverage receiver where an i.f. of

455/465 is used, and this consists of an i.f. amplifier operating at 1.6 MHz. At this frequency suitable voltage amplification can be obtained with three stages and suitable components are readily available. Also, crystal filters are not too difficult to fabricate and the use of a Q Multiplier will enable adequate selectivity to be obtained.

The use of this frequency will give freedom from image signals for the 2m band and this is particularly important with the tendency for s.s.b. operation to increase in this band. It seems almost inevitable that this form of communication will grow quite rapidly with the use of transverters as a means of using h.f. band equipment on this band.

Any receiver built specially for amateur use in this higher frequency region should, as with any modern receiver, be capable of receiving all the usual forms of voice modulation:

> Amplitude modulation
> Frequency or phase modulation
> Single sideband amplitude modulation

The increased complexity of the inclusion of a triple detector is justified, and should be a feature of all new equipment and this particularly applies to the correct and adequate detection of frequency modulated signals. There is no doubt that one of the reasons why n.b.f.m. has not achieved the popularity it deserves, is the simple fact that most receivers are not fitted with an adequate detector and operators have resorted to compromise for detection. It is recommended that all users of n.b.f.m. for transmission should also add a suitable detector to their present or new receiving system.

The block diagram of the improved v.h.f. - u.h.f. receiver, shown in Fig.5.2 gives the essential constituent parts of a suitable system. All the various units need some description and these will be dealt with individually.

Front End Units

The units may be either built-in or plug-in and the actual requirements will vary considerably from one operator to another. Before deciding the details of these units, it is important to decide the tunable first i.f. section.

Assuming a 1.6 MHz main i.f. amplifier and the use of a tuning range of 2 MHz (± 1 MHz about the mean frequency), this can be at more or less any frequency the constructor wishes. It is desirable that this should be at a frequency at which breakthrough is not troublesome, frequencies between 2-4 MHz and 28-30 MHz have been used without trouble, but at the lower frequencies there is likely to be a greater chance of breakthrough than at higher frequencies.

At the same time oscillator stability is better at the lower frequencies and therefore rather greater care in this respect is needed when using a higher frequency.

There is no reason why the tunable first i.f. should not be made to cover one of the popular h.f. bands if this is desired, but since the tuning range has to be 2 MHz this really is only suitable for use at 28 MHz. Nevertheless, with a good slide rule type tuning dial and suitable h.f. band converters, the arrangement can be usefully used for the lower frequencies, but owing to the narrow frequency bands the amount of band spread on the dial will be limited to 85% on 10m, 22.5% on 15m and 17.5% on 20m.

If the tunable first i.f. is likely to be operated without a converter on one of the h.f. bands, it will be necessary to incorporate one stage of amplification and it may be desirable either to add a preselector, or band pass filter to provide adequate selectivity when operating as a straight forward receiver.

When the tunable i.f. is low it is preferable to use double conversion and this should be done using a single crystal to avoid inter-oscillator spurious signals. This use of double conversion is necessary to separate adequately the signal and oscillator frequencies and is particularly appropriate for 2m, 70cm and 23cm converters.

At a higher frequency first i.f. such as 24-26 MHz or 28-30 MHz the need for double conversion does not arise for 2m converters but is desirable for 70cm and 23cm.
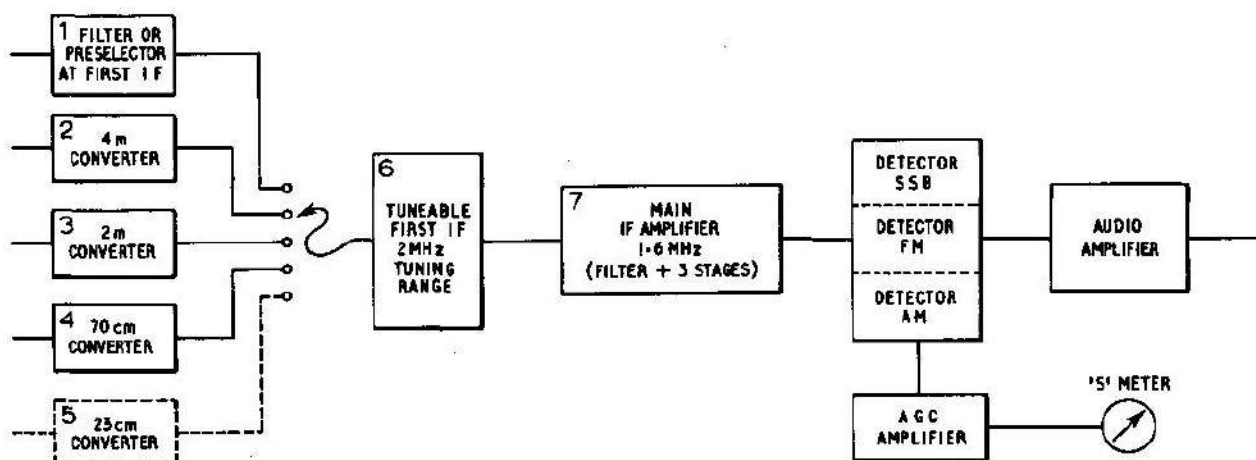

Fig.5.2. Block diagram of improved v.h.f. - u.h.f. receiving system.

The use of Field Effect transistors for front end converters is particularly attractive because they can be made completely plug-in self contained units if desired, each with its own built-in battery. Alternatively, a suitable power source can of course be provided from the main unit.

When a number of converters are to be built into the main unit, it is important to suitably screen the selector switch to ensure that there is a minimum pick up of outside signals within the tunable first i.f. range.

In selecting a crystal for the local oscillator in a converter it is essential to ensure that the fundamental and harmonic frequencies do not appear in the pass band of the tunable first i.f. section. Ideally the crystal frequency should be above the highest frequency to which the tuning will reach, alternatively a crystal close to the lower frequency will be found satisfactory.

Typical crystal frequencies for a tunable i.f. of 24-26 MHz:

Band	Band centre frequency	Local oscillator	Crystal frequency	Multi-plication
4m	70.3 MHz	45.3	22.65 or 15.1	2 or 3
2m	145 MHz	120	30 or 10	4 or 12
70cm	435 MHz	410	20.5 or 22.777	20 or 18
23cm	1305 MHz	1280	20	64

The Main Receiver

Comprising the tunable first i.f. main i.f. amplifier, detectors and audio amplifier.

This may be conventional with the exception of the inclusion of a suitable filter supported by a Q multiplier as required, together with detectors for the three forms of voice transmissions. An additional refinement which is very desirable is amplified a.g.c. This addition is only a small increase in complexity and is well worth including.

The question of whether this "fixed" part of the equipment should be based on valves or transistors is to some extent dependent on personal preference. In the i.f. amplifier stages using transistors gain will be much lower than where valves are used.

Field Effect transistors however, largely overcome these disadvantages and are a satisfactory alternative to the valve. The design detailed is based on valves and is given as a typical arrangement which has proved to be very satisfactory, translation into a semi-conductor design is fairly straight forward.

If a specialised receiver is to be built for the harmonically related 70cm and 23cm bands, it should be based on the improved receiver with a 1.6 MHz main i.f. amplifier and the three forms of detector.

The communication part of the bands is in general confined to ± 1 MHz and therefore the tunable first i.f. ahead of the main i.f. amplifier will only need to cover a 2 MHz swing. By careful choice the whole of the three band converters can be arranged to

operate from a single crystal thus avoiding any chance of inter-oscillator spurious beat frequencies. Naturally with the precise frequency requirements the number of crystal frequencies that may be used are very limited.

Details of crystal frequencies suitable for use in converters:

Tunable 1st i.f. MHz	17	25	29.8	37
Crystal frequency	32	24†	57.6	36†
2m Multiplier times	X4	X5	X2	X3
Frequency	128	120	115.2	108
70cm Multiplier times	X3 X3	X3 X4	X5	X4 X2
Frequency	288	288	288	288
23cm Multiplier times	X4*	X4*	X4*	X4*
Frequency	1152	1152	1152	1152

†These frequencies will prevent reception at the bottom band edge, local signals will occur at 144, 432 and 1296 MHz.

*In addition to that required for the 70cm mixer.

Fig.5.3 illustrates the principles involved in this arrangement and the table gives details of suitable crystals and the local oscillator frequencies for each of the three converters.
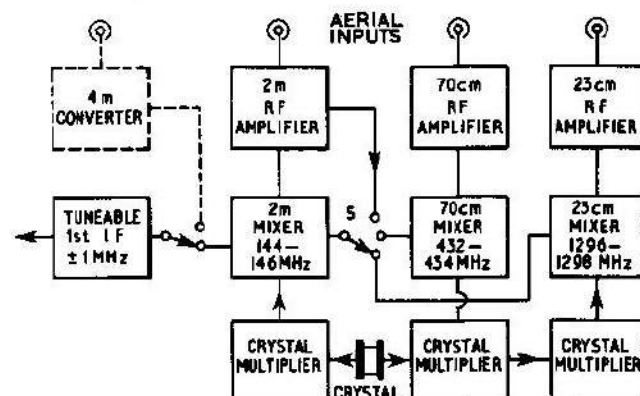

Fig.5.3. Illustrating the principle of a 3 band receiver using a single crystal.

The 2m mixer and its oscillator are used on 2 metres in the conventional manner. On the two higher frequency bands the output from the relevant mixer is fed into the 2m conversion before passing on into the tunable first i.f. stage. The selector switch S should be of a type which does not introduce any serious mismatch or noise into the 2m mixer stage.

This component may be arranged by using diodes for the circuit switching, so avoiding any problem being caused by use of a mechanically operated switch, see page 5.62.

Where reception on 4m is needed, a separate self contained converter for this band can be introduced as indicated in the block diagram.

Undoubtedly the best approach to the somewhat elaborate multi-band v.h.f./u.h.f. front end, is to use Field Effect transistors for all the signal frequency amplifier and mixer stages. Bi-polar type transistors

and/or varactor diodes are suitable for use in the oscillator/multiplier stages. In this manner the whole unit can be built into a relatively small space, which may be either a plug-in unit or a built-in section of the whole receiver.

Details of the various circuits involved for each of the individual stages can be obtained from the various units described elsewhere in this Manual.

Multi-Purpose Detector for NBFM, SSB/CW or AM

Probably one of the major reasons for most operators not providing proper detection for n.b.f.m. is the rather difficult matter of fitting one or other of the classical detectors into an existing receiver.

The small transformers required for the Foster-Seely or ratio detectors are not generally available for the frequencies normally used in i.f. amplifiers.

It is therefore not proposed in this Manual to deal with these methods of detection, but rather to offer a circuit which is capable of being used for all the usual methods of modulation, amplitude, narrow band frequency and single sideband.

Fig. 5.4. Diagram showing detector circuit for receiving n.b.f.m., s.s.b., c.w. and a.m.

In this circuit the valve is a 6BN6, a special type of valve, known as a gated beam valve, with a pentode characteristic, and two control grids No. 1 and No. 3 and is used both as a pentode and a diode.

In the position for n.b.f.m. (1) the valve is operating normally but with no b.f.o. the tuned circuit in the No. 3 grid (quadrature coil) is tuned for resonance at the intermediate amplifier frequency and is then trimmed for best audio output.

In position 2 for s.s.b. or c.w., the conditions are the same but the b.f.o. is switched on. In both these positions the valve has considerable gain, whereas in position 3 all h.t. supplies are switched off and the valve operates as a diode. VR1 is provided as a gain equalisation control and VR2 is adjusted for best a.m. output.

If a.m. is not required, the full gain of the valve can be used and VR1 would become the audio gain control; the only switch then required would be to turn b.f.o. on or off. It is probable that one or other of the dual control pentodes other than the type shown, such as the 6AS6, 6DT6A or EH90 can be used in a similar circuit.

The basic circuit, the quadrature grid demodulator, has been used for many years in television receivers where f.m. sound is used, for it is a self limiting device. But it is important to have adequate i.f. gain ahead of the detector to ensure its proper working. The quadrature coil in the grid No. 3 may be any convenient tuned circuit of the same frequency as that of the i.f. amplifier.

Tunable I.F. Amplifier suitable for V.H.F.

A circuit of a tunable i.f. amplifier is given in Fig. 5.5. The r.f. amplifier TR1 uses an OC170 in the grounded base configuration. The d.c. stabilization uses an emitter resistor returned to a positive bias, which is varied to adjust the r.f. gain. Silicon diodes CR4 and CR5 in Fig. 5.5 are fitted across L2 to protect TR1 from high power signals.

The mixer, TR2, operates in the common emitter mode and is perfectly stable. It is emitter coupled to the local oscillator TR3. With injection from the oscillator collector to mixer base, it is impossible to approach the desired oscillator injection level, resulting in poor sensitivity. It will be noted that the collectors of TR1 and TR2 are taken to the hot end of their respective coils, no advantage being found in tapping them down.

Capacitor C16 in the local oscillator is a negative temperature co-efficient type to compensate for changes in C17 and L8 with varying ambient temperature. Zener stabilization at 4.7V is used for the oscillator h.t. If a 6V supply were used the Zener would have to pass a high current when the battery was new, to make sure that it would still stabilize when the supply is down to 5V. Consequently a 7.5V h.t. line is used, the battery being discarded when it is down to about 6 volts.
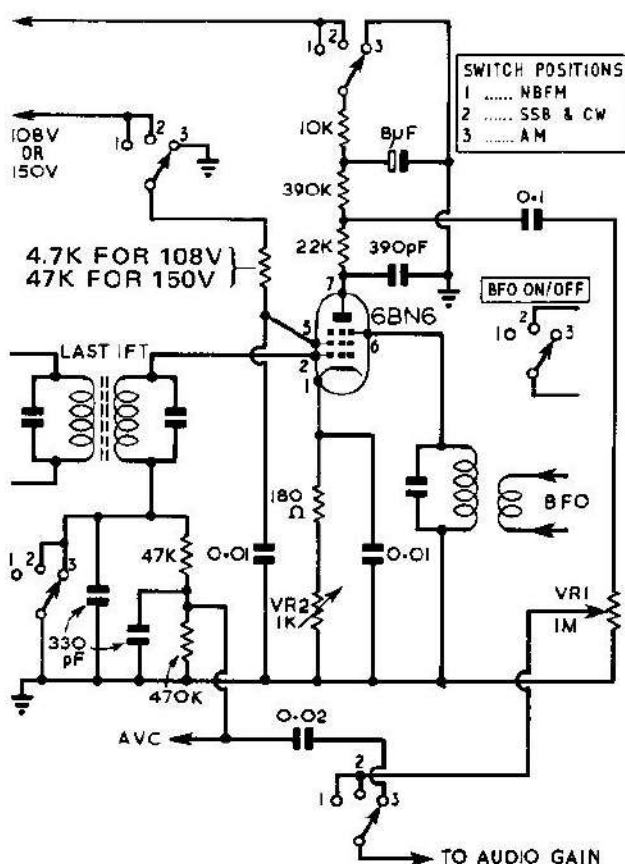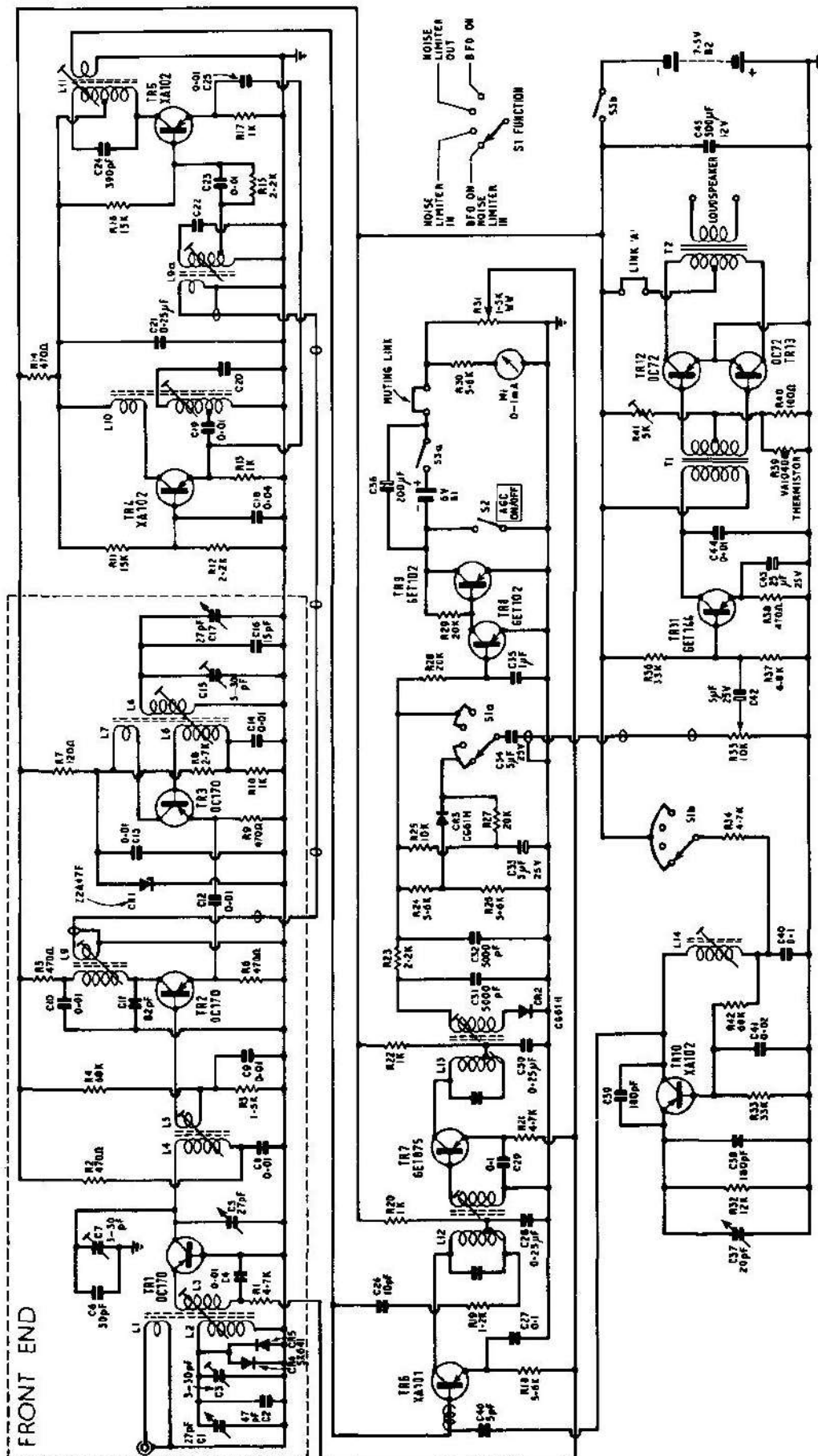
Fig. 5.5. The complete tunable i.f. with a.g.c. and audio amplifiers. C1, 5 and 17 are a three gang capacitor from an RF27 unit, with vanes removed to provide 27pF for each section: C4, 8, 9, 10, 12, 13, 14, 31 and 32 are 30 volts working disc ceramics; C2, 6, 26, 38 and 39 are silvered mica; C16 is a 15pF N750 ceramic; C18, 19, 21, 23, 25, 27, 28, 29, 30, 35, 40, 41 and 44 are metallized paper; C33, 34, 42 and 43 are 25 V wkg electrolytic and C45 is a 12 V wkg electrolytic R31, 1.5 K ohm wirewound variable; R35, 10 K ohm carbon, with S3; R39 is a Mullard Thermistor type Va1010 and R41 is a 5 K ohm preset miniature potentiometer. All fixed resistors are Erie type 9,5 per cent tolerance. M1 is a 1mA panel meter. Suggested alternatives to certain of the original transistors are: TR4-5-6-10, OC45; TR7 and 11, AC126; TR7 OC82. B1 is a 6 volt Vidor VT1, and B2 is 7.5 volts using a VT1 and a V11 in series.

Tunable I.F. Coil Table

Coil	Turns	Pins Used	Coil	Turns	Pins Used
L1	2	1.4	L5	2	1.2
L2	7	3.2	L6	1½	1.2
L3	1	5.6	L7	2½	1.6
L4	7	3.2	L8	7	3.4

L1-8 are wound with 24 s.w.g. wire on ¼" diam. Aladdin formers with core and short can.

L9, L9a, medium wave aerial coil with emitter tap and low impedance input winding.

L10, medium wave oscillator coil.

L11, 12, 470 kHz i.f.t.

L13, 470 kHz i.f.t. (second detector type)

L14, 470 kHz i.f.t. (for valve use) with one winding and all capacitors removed.

T1, T2 interstage and output transformers.
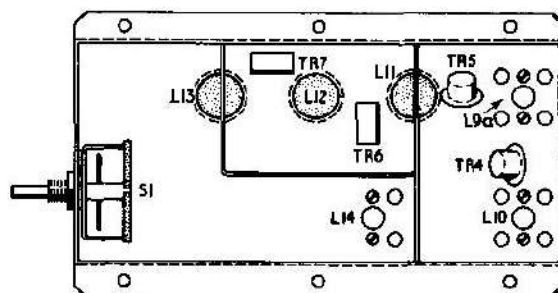
Fig. 5.5A. Under chassis layout of the i.f. amplifier.

The I.F. Amplifier Unit

The first mixer is coupled to the second mixer TR5 by 75ohm cable. L9 and L9a have 75ohm coupling windings. Top coupling via a 5 pF capacitor may also be used, the cable capacity appearing across one of the tuned circuits. This method was found equally effective and is an alternative if 1.6 MHz coils are not available with 75ohm input windings.

The second mixer TR5 and local oscillator TR4 are quite conventional separate stages being preferred to prevent frequency pulling on strong signals. Stabilized h.t. is not necessary for TR4. The i.f. amplifiers TR6 and TR7 are also straight forward; the second stage using a GET 875 not requiring neutralization. These stages are biased by taking the emitters to a positive bias, i.e. the gain control line, but note that neither mixer is taken to this control line.

The detector and noise limiter CR2, CR3, is a type used successfully in several valve receivers and is scaled down for the semiconductor impedances. The noise limiter is very good, without an excessive insertion loss. The noise limiter switch SI is combined with the b.f.o. switch to save panel space.

The b.f.o. circuit (TR10) was chosen because a tapped coil is not required. B.f.o. injection is by a 5 pF capacitor, C46, from TR10 collector to TR6 base. The end of the capacitor at TR6 base is not connected directly to the base, but the lead is wrapped around the base lead to provide an effective capacity rather less than the 5 pF. If the coupling is too great, pulling of the b.f.o. is liable with strong signals, therefore the output from the b.f.o. is lower than the desired amount, thus requiring low r.f. gain when copying s.s.b. Alternatively, a circuit similar to that of TR4 may be used.

A.G.C. Amplifier

When no signal is being applied to the receiver, TR9, which is connected in series with the negative lead of the bias battery BI, is turned on by R29, and in this condition nearly all BI voltage appears across the r.f. gain control R31.

When a signal is received, some of the d.c. component at the detector, CR2, is passed via R28 to TR8. This negative going voltage begins to turn TR8 on, hence removing drive from TR9 base, so that TR9 is no longer heavily conducting and bias voltage is dropped across it, thus reducing the voltage across R31. Capacitor C35 decouples the a.g.c. system from the detector. The meter, with its multiplier, R30, monitors the a.g.c. voltage, giving "S" meter indication and with the a.g.c. off, this meter reads the bias battery voltage. When a strong signal is being received, the voltage on R31 slider is virtually constant whatever the position of the slider.

The receiver is muted by disabling the bias supply, and with this system the receiver is completely silent when muted.

It should be noted that if lower gain transistors than those specified for TR8 and TR9 are used, R29 and R28 will have to be reduced in value, placing more load on the detector.

Front-end Construction

The layout of the front-end is clearly seen in the underside diagram. It is built on a 6" x 4" 16 s.w.g. tinplate chassis, which is fitted with sides and screen of 20 s.w.g. tinplate. The tuning dial is an Eddystone type 598, fitted with an extra epicyclic reduction unit so that a slow 60:1 reduction is available.

I.F. Amplifier Construction

The i.f. amplifier is built on a sub-chassis 6" x 2¾" x 1½" deep of 20 s.w.g. tinplate, and a screen is soldered across the width to make a 2¼" long compartment for TR4 and TR5. An L-shaped screen forms a 2¼" long compartment for TR6 and TR7. The b.f.o. is rather cramped as is the noise limiter, and it was originally intended to fit the a.f. gain control R35 in the i.f. amplifier, but again space was short.

The chassis has ¼" flanges at each side and if these were turned inwards, instead of out, and fitted with captive nuts, then the chassis could be 3" wide. This would still allow room for the loudspeaker magnet.
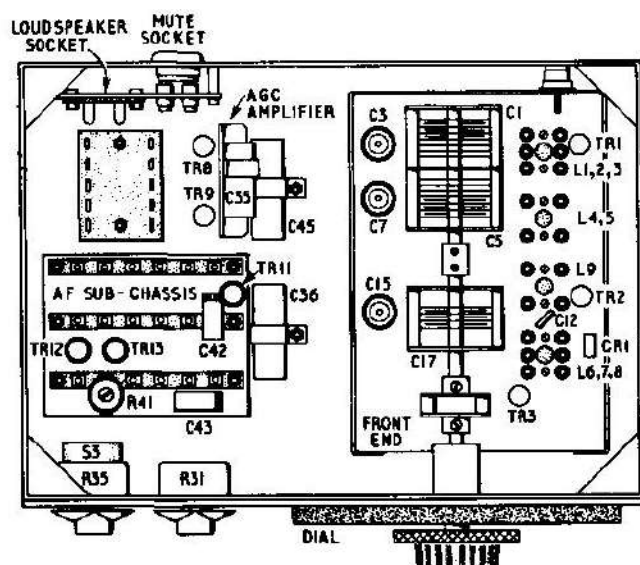
Fig. 5.5B. Under chassis layout of the receiver.

The connecting leads to the i.f. amplifier are made sufficiently long to enable the chassis to be withdrawn from the rear, although the lead to the b.f.o. pitch control C37 has to be unsoldered. The a.g.c. amplifer is built on a small bracket. TR8 and TR9 being held in small cable cleats. The audio amplifier is contained on an 18 s.w.g. plate. A maximum undistorted output of 220 milliwatts is obtained for 4mV input.

General Construction

The receiver is constructed as a number of sub-units. The main chassis has a cut-out to enable the front-end unit to be fitted from underneath with the tuning coils protruding through the cut-out.

The i.f. amplifier is built on a sub-chassis, bolted on top at the right hand side of the main chassis. The wiring to the i.f. amplifier is made into a cableform of sufficient length to enable the sub-chassis to be withdrawn from the rear.

The audio stages and the a.g.c. amplifier are both built on plates which locate on pillars underneath the main chassis and underneath the i.f. amplifier.

The main chassis is a standard 10" x 7" x 2½" item to which is bolted the ½" thick aluminium front panel, 6" high and 10" wide. The side plates are of ¾" Tufnol, drilled and tapped 4BA on the edges to enable them to be bolted to the chassis and panel. A 16 s.w.g. aluminium base plate with turned up flanges is fitted to the main chassis, and the flanges of this base plate are equipped with 6BA captive nuts. All the "hot" wiring is completely screened by this method of construction, which is desirable to prevent pick up of spurious signals in the tunable range of the receiver.

The batteries occupy the space above the tuning gang, the centre line of the batteries also being the centre line of the tuning gang. The batteries are held in place by a strap.

The bias battery voltage is indicated by the "S" meter, as explained in the description of a.g.c. system, and this battery is fitted nearest the front panel. The collector supply battery is then at the rear from which position it can be checked easily.

Alignment Procedure

Before switching on for the first time, set R41 to maximum resistance. Open link "A" and insert a milliammeter to monitor TR12 and TR13 collector current. Switch on and set R41 such that the total collector current of TR12 and TR13 is 8mA. Switch off and re-connect link "A".

Feed in 470 kHz to L9a, and adjust L13, L12, L11 for maximum output. If these transformers are a long way off tune, however, it may be necessary to inject the 470 kHz signal, via a 0.1 µF capacitor, to the base of the preceding transistor.

Feed in 1.6 MHz to L9a and tune 10 until the signal is found, then adjust L9a for maximum output. L10 operates on the high side, i.e. at 2.07 MHz, and this can be verified by checking that the image is at approximately 2.5 MHz

With the tuning gang set to maximum capacity, feed in a 26.4 MHz signal to L1. Adjust L8 to tune in this signal, then peak the signal with L4 and L2. Set the tuning gang to minimum capacity, re-set the signal generator to 30 MHz and adjust C25 to tune in this signal, and then peak the signal with C7 and C3.

Repeat the above procedure, for the oscillator only, i.e. L8 at 24.4 MHz and C15 at 30 MHz (dial readings) until the band edges are set.

The mixer and r.f. circuits are then peaked up, which is best done at frequencies inside the band edges, e.g. 26.7 MHz and 29.7 MHz. Repeat these adjustments until no further improvement can be obtained.

The oscillator works on the low side, and this should be verified by image frequency measurement.

Performance

Sensitivity at 28 MHz, Less than 1μV for 6dB signal to noise ratio.
Image at 30 MHz, 40dB down.

OSCILLATORS

In both receiver (or converter) and transmitter a local oscillator is required, and its performance in either is of considerable importance. Its stability is of prime importance because it will almost invariably be used with several stages of multiplication to reach the working frequency.

Until recently fixed frequency crystal oscillators have been used for operation in the v.h.f. and u.h.f. region, but there is now a significant change towards some form of variable or tunable oscillator. Band planning has to a large extent been successful and there is not the same need to operate on a common frequency as there is on s.s.b. or h.f. bands.

However, s.s.b. operation is naturally increasing on these higher frequency bands and common frequency working is being practised. In the main the equipment is based on h.f. band apparatus with some form of transmitter converter.

Before deciding the form of oscillator to be used, it is as well to consider the advantages and disadvantages of the fixed oscillator compared with a variable or tunable unit.

Crystal Controlled Oscillator

Advantages	Disadvantages
Absence of controls other than selector switch	Frequency choice must be such as to avoid spurious signals in the pass band.
Clean note and freedom from modulation.	There are likely to be more stages needed to reach the final frequency.
Negligible short term or warm up drift.	Inability to move frequency when interference from another station is present
Reliable frequency location in the band.	

VXO Variable Crystal Oscillator

Advantages	Disadvantages
Variable but limited tuning range.	More complex and requires more space and a mechanically good dial for resetting accuracy. Similar number of multiplier stages needed to reach final frequency.
Clean note if tuning range not too great.	
Short term and warm-up drift can be maintained. Reliable frequency location in the band.	

Variable Frequency Oscillators

| Advantages	Disadvantages
Direct frequency calibration possible.	Stability, often difficult to obtain long term drift to low value
Free choice of frequency tuning range to suit requirements.	Sensitive to ambient temperature and ventilation conditions.
When high frequency used fewer frequency multiplier stages needed to reach final frequency.	Care needed to ensure clean note free from unwanted modulation.
	Valve or transistor change necessitates recalibration.

TYPES OF CRYSTAL OSCILLATOR

There are very many suitable circuits for crystal oscillators but those especially suitable for generating harmonics are normally required in this application for local oscillator in a converter or control of a v.h.f. or u.h.f. transmitter. Usually crystals in the range 6-12 MHz are used and output frequencies from the multipliers to suit the band in use. In general all the popular circuits give substantial output, from a single valve, of 3rd or 5th harmonic.

The three classic crystal oscillators are the Miller, Colpitts and Pierce of which the last two are now much more widely used. All these oscillators use the crystal in its parallel mode. A digest of the wide variety of oscillators appears in Chapter 4 of *Amateur Radio Techniques*.

Squier Oscillator

This circuit is suitable for use with crystals of the fundamental type. Feedback between grid and anode
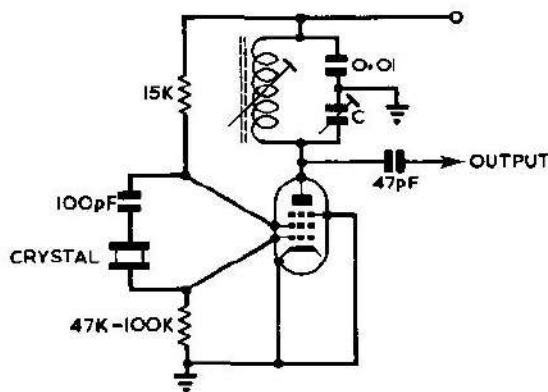
Fig. 5.6. A MODIFIED PIERCE OSCILLATOR
This circuit will give substantial output up to about the 9th harmonic but it is generally used with the anode circuit LC tuned to either the 3rd or 5th harmonic, the latter with its smaller output is suitable for use in converters rather than transmitters. The valve should be a high slope straight μ characteristic type such as used for television receiver i.f. amplifiers.
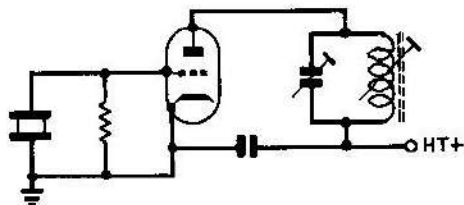
Fig.5.7. MILLER – *typical frequency stability of about 25 ppm can be obtained, with a relatively high output. The need for a tuned circuit is a disadvantage. Suitable for use with crystals up to 20 MHz.*
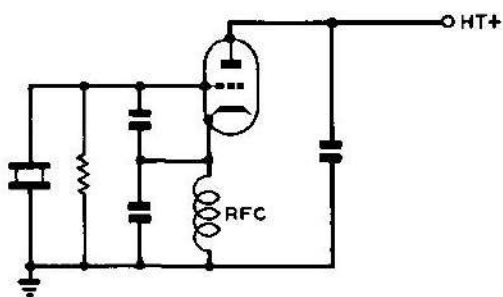
Fig.5.8. COLPITTS – *although the output of this oscillator is lower than with the Miller circuit, the frequency stability is better, generally around 10 ppm.*
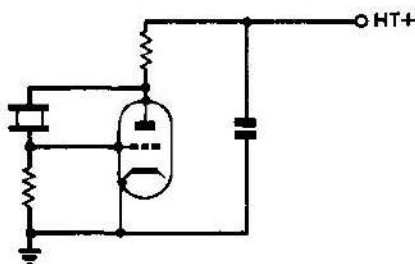
Fig.5.9. PIERCE – *no tuning is needed in the fundamental form and suitable for crystals up to 20 MHz. Switching crystals presents no particular problem.*

is by direct inductive coupling, the amount of which is determined by the position of the h.t. feed point on the inductor L.

The amount of feedback is increased as the tap is moved away from the grid end. When overtone crystals are used the amount of feedback needed is smaller than with fundamental types. The inductor L is tuned to the overtone frequency (harmonic) required. Care must be exercised to ensure that there is not too much feedback, otherwise the crystal will lose control, in addition the output frequency may not be an exact harmonic of the crystal frequency.
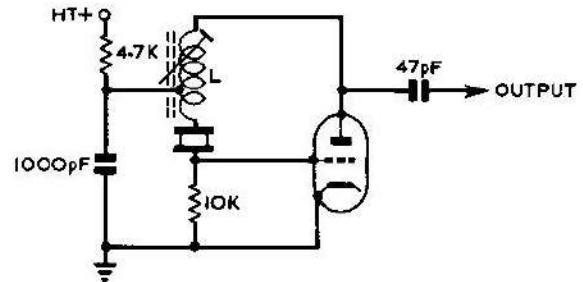
Fig.5.10. SQUIER – *some care is needed in the adjustment of this circuit.*

Butler Oscillator (Modified)

This circuit is suitable for use with overtone crystals, consisting of a grounded grid amplifier (V1a) and a cathode follower (V1b). The anode circuit of the amplifier is tuned to the required harmonic frequency usually second or third, outputs at higher orders are generally too small for transmitter use. If output only at the overtone frequency is required, this may be obtained by inductive coupling to L1, and the anode of the cathode follower should be earthed to r.f. and the tuned circuit L2 C2 omitted. The resistors used in the two cathode circuits should be matched, increasing the value will increase the output, but the stability will deteriorate if this is taken too far while reducing the value will reduce the output until oscillation ceases. A compromise value such as that shown should be used.
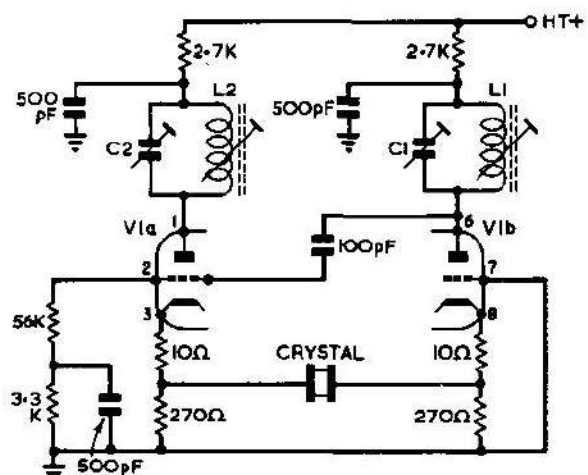
Fig.5.11. BUTLER – *for use with overtone crystals.*

Transistor Overtone Oscillator

As with valves there are many suitable circuits which may be used either for local oscillators in a converter or control of a transmitter.

The circuit shown is typical and with the details given it is suitable for a 144 MHz converter when a 28-30 MHz tunable i.f. is used. It may also be used with an h.f. band s.s.b. transmitter with output in the 28-30 MHz to a converter for use on 144 MHz.
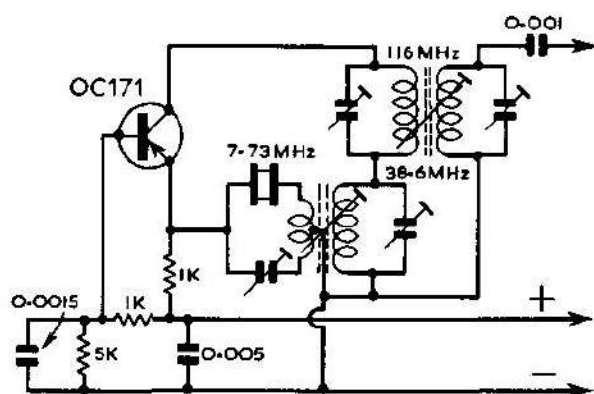
Fig. 5.12. A typical transistor oscillator.

VXO - VARIABLE FREQUENCY CRYSTAL CONTROLLED OSCILLATORS

In recent years the use of circuits which enable a crystal to be pulled from its fundamental frequency has been actively developed and several reliable circuits are now being extensively used, either with a single crystal or with a bank of switched crystals to cover the whole band.

In most v.h.f. or u.h.f. bands, the crystal used is usually of relatively low frequency such as 6-12 MHz and there is a fairly high degree of frequency multiplication to reach the operating frequency.

Typical frequency shifts of up to 200-300 kHz at 144 MHz are common, but better stability will result if the range is limited to about 100 kHz. Larger or smaller frequency shifts will become available on higher or lower final frequencies.

If for example, it is desired to cover the whole of the 144 MHz band, 144-146 MHz, then a simple switched bank of crystals together with a frequency shifting circuit will provide a satisfactory means of achieving this, with probably 10 crystals. The actual number will depend on what degree of stability and clearness of note is desired.

There are a large number of arrangements for moving the crystal frequency. The correct values of the frequency shifting components will vary and are best made to suit the particular crystal or crystals being used.
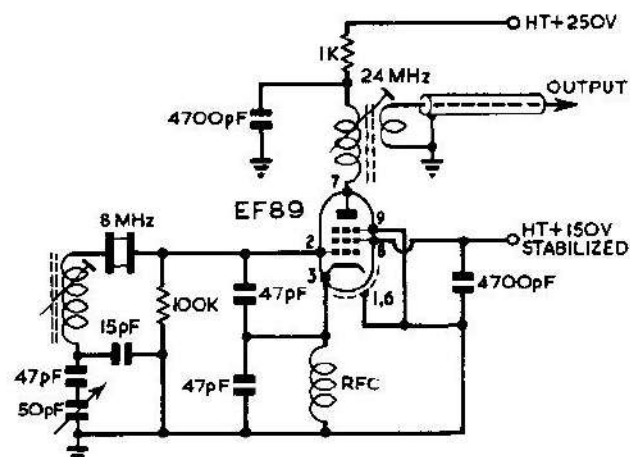
Fig. 5.13. Simple method of varying the frequency of a crystal oscillator.

The object should be to keep signal quality clean, this as mentioned earlier almost inevitably means restricting the frequency variation to around 100 kHz. These oscillators find considerable application for net working or in association with s.s.b. transmitters.
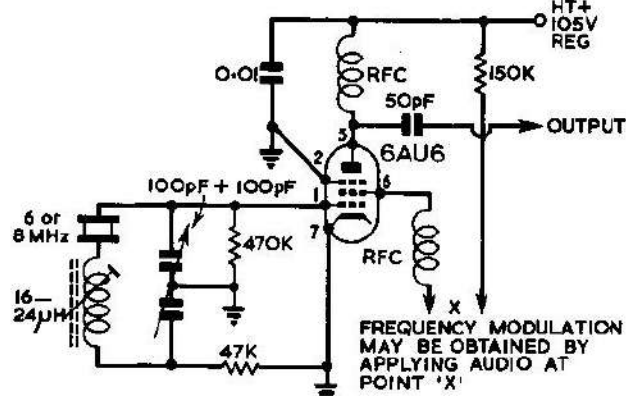
Fig. 5.14. Alternative arrangement for variations of frequency.

Where wider frequency ranges are required a normal variable frequency oscillator should be used and this is best operated at a relatively low frequency followed by the required frequency multiplication to the final frequency.

The Kaliatron 72 MHz oscillator offers another approach to this problem and in this, as with all variable (tunable) oscillators, rigid mechanical construction is essential together with adequate temperature compensation to counteract thermal effects caused by the active components in the valve or semiconductor and also changes of ambient temperature.

The indiscriminate use of variable frequency oscillators should be discouraged and close adherence to the band plans will make for the best use of the frequencies available.

A Kaliatron Oscillator for 72 MHz.

The fundamental frequency of operation is 72 to 73 MHz and is suitable for fully covering the 2m band (144-146 MHz).
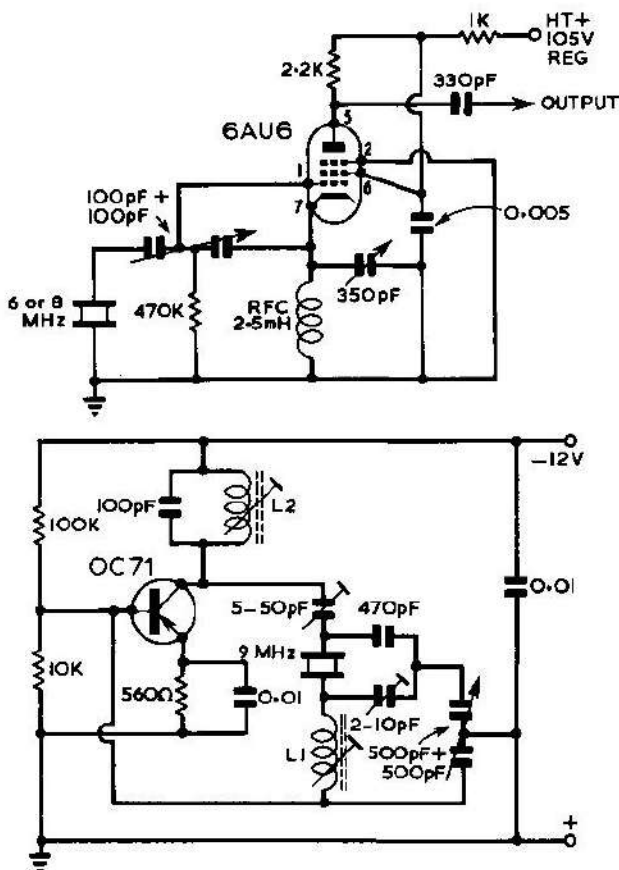
Fig. 5.15. *Two further circuits suitable for giving a change in frequency.*

The tuned circuit of the oscillator consists of the hairpin shaped inductor, L, and the capacitors, C1, the frequency setting component and C2 which covers the tuning range, in this case 1 MHz (72-73 MHz).
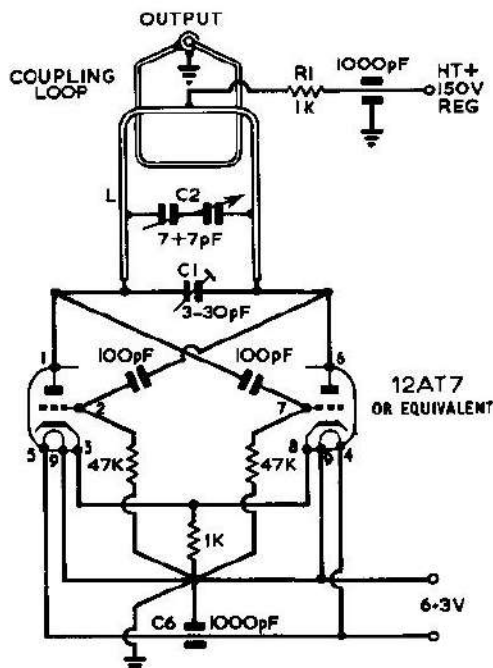

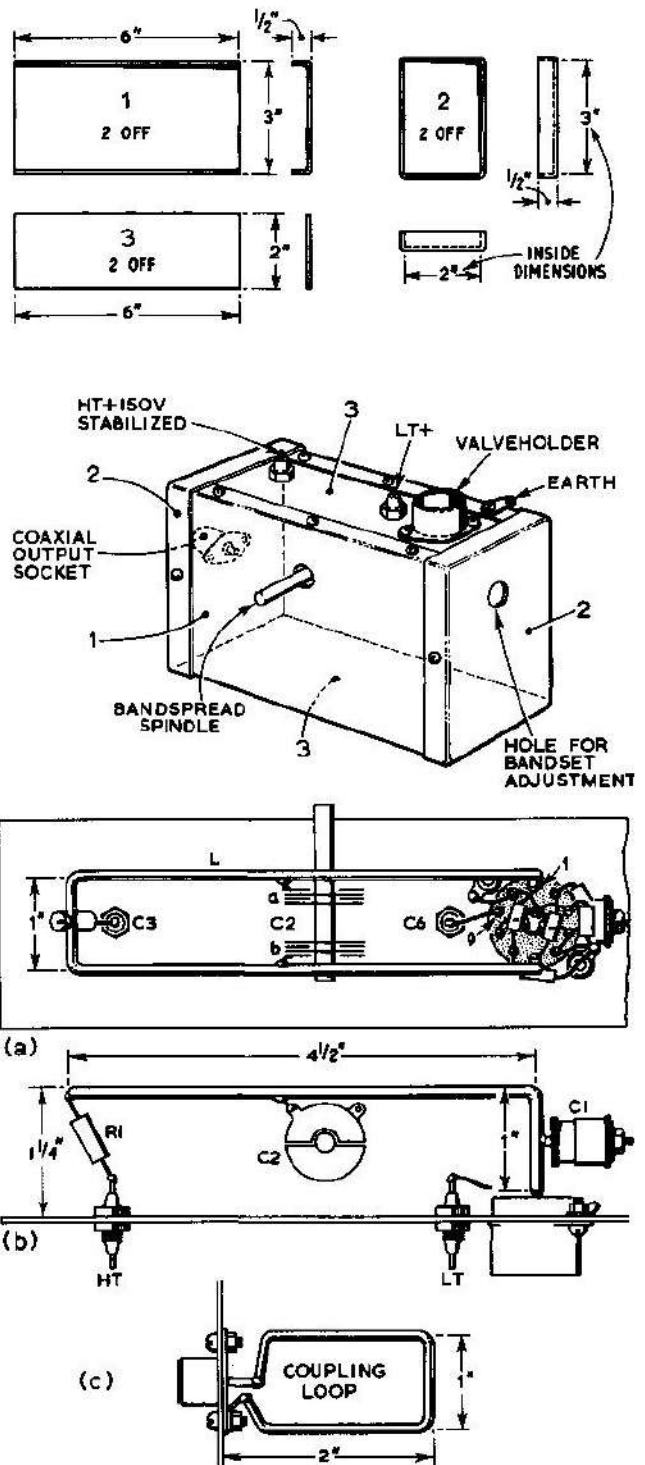Fig. 5.16. *The Kaliatron variable frequency oscillator for 72 MHz.*

Fig. 5.17. *Details of the construction of the copper or brass box. The thickness of the material may lie between 16 and 22 s.w.g. Diagram (a) shows a view of the underside of the top plate of the box, and (b) a side elevation. The coupling loop (c) is mounted approximately ½" underneath the tuned circuit.*

The inductor is formed from 10 s.w.g. copper wire to the dimensions shown in the mechanical detail drawings and is rigidly supported by connection to the valve socket, tuning capacitor and h.t. feed resistor at

the centre. The actual position of C2 should be located at a point on the circuit so that the frequency range is limited to the 72-73 MHz, this limit must be strictly observed to prevent inadvertent out of band operation.

The output coupling loop is mounted ½" horizontally below the tuned circuit at the h.t. feed end. Coupling should be as loose as possible consistent with obtaining the required output to drive the following stage which should be a buffer amplifier isolation stage. A valve such as a CV 138 or 6AM6 is a suitable amplifier.

NOISE

In the v.h.f. and u.h.f. region, the factors which affect the receiver performance are different from those encountered on the h.f. bands. It is necessary to pay considerable attention to noise reduction within the equipment itself since as the frequency is increased from around 100 MHz, noise generated in valves, semiconductors, resistors, inductors and leads becomes of increasing importance.

The lowest of the bands which are considered in the Manual may be regarded as about the transition frequency where the change over from standard methods to more refined techniques becomes necessary.

This of course does not mean that a sudden change over from the conventional circuit arrangements to linear or cavities are necessary, these do not really become necessary until frequencies around 400/500 MHz are reached.

Attention to small details become increasingly important - lead lengths and self inductance of leads apart from contributing noise may seriously affect the circuit constants. In order to make use of radio signals it is necessary to amplify them without at the same time introducing any more noise than is absolutely necessary. It is of prime importance to achieve as high a signal/noise performance as possible. To do this it is important to appreciate the factors which cause noise.

Generally the sources of noise can be divided into:-

(a) External Noise   Atmospherics
Thermal noise from the earth
Solar and cosmic noise
Manmade impulse noise (e.g. ignition)

(b) Internal Noise   Resistive noise
Shot noise
Flicker noise
Partition noise
Transit time and induced grid noise
Thermal effects in transistors

From these it is clear that little can be done in respect of external noise, except in the case of the manmade impulse and other spark noises. Effective noise cancelling systems can be devised and details of these are given in Chapters 4 and 16 of the RSGB *Radio Communication Handbook*.

On the other hand, by design, a considerable influence can be made on the internal noise (noise generated within the equipment), but it is important to know how this arises. The various types are briefly described below.

Resistive Noise

This type of noise is to be found in all resistors, together with the resistive elements of inductance and capacitance. The actual level of resistive noise depends on the type and construction of the resistor.

Even a so-called "ideal" resistor of the metallic type will generate noise. This noise is created by the random movement of free electrons within the molecular structure of the resistive element, and since the "activity" of these electrons is dependent on temperature, the level of noise will vary with changes in temperature.

Some qualification of the foregoing temperature v. noise statement is however needed. Most resistors operate at a temperature in the region of 17 deg.C. or 290 deg.K unless special cooling is undertaken. Since it is only at Absolute Zero - 273 deg.K below 0 deg.C - that the motion of the random electrons ceases, and the resistor would become "silent" it will be appreciated that, over the normal working temperature range, any changes in temperature will be proportionately very small when compared to the elevation above absolute zero. For this reason, the inherent noise level within average resistors does not change very appreciably over the range of temperatures encountered under normal working conditions.

Carbon resistors, which are almost universally employed in equipment are likely to be as much as ten times more noisy than metallic types, and this fact
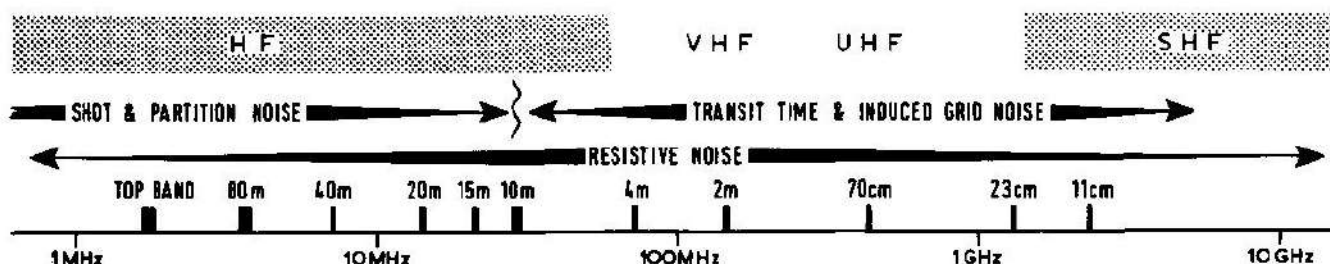

Fig. 5.18. Chart illustrating the relationship between types of noise and frequency.

should be borne in mind when producing items such as high gain microphone amplifiers.

As a matter of interest, specially designed resistors operated at high and/or low temperatures, are available for use as noise sources for equipment evaluation.

Shot Noise

Shot noise arises from the random emission of electrons from a hot cathode within a valve structure.

There are two classifications of emission from the cathode or filament of a valve, one being designated "saturated emission" and the other "non-saturated emission". In the former, and considering a simple diode with a pure tungsten filament, the anode current will be directly controlled by the temperature of the filament and when this is at its permitted maximum, the anode will "take up" all the electrons available from the filament. There are no "spares" drifting around.

Under non-saturated emission conditions, there is much more emission available from the cathode/filament, than that which is being drawn by the anode. This gives rise to a space charge between the filament and the anode, and this has the effect of smoothing out some of the fluctuations in anode current caused by the shot effect. Most valves employed in equipment are operated under space charge conditions i.e. non-saturated emission - and the anode current fluctuations are much less than in the case of the saturated emission diode used in noise generators.

Flicker Noise

This takes the form of large amplitude pulses, and is produced in some types of valves at low frequencies, being, generally, most troublesome in the frequency range 100 Hz - 50 kHz.

In high gain audio amplifiers, the presence of this noise form is most detrimental to satisfactory operation, and has led to the development of special "low noise" valves for this particular application. The EF86 and its associated types are examples.

Partition Noise

In multigrid valves, the division of the total cathode current between the anode and the various other electrodes are subject to the fluctuations caused by shot effects. In turn, these additional electrodes will, by a process similar to modulation, cause increased random fluctuations in the anode current.

From this it will be appreciated that the fewer the electrodes in a valve the better, and moreover, why it is that a triode will always have a much lower noise factor than a pentode or other multi-electrode valve.

Transit Time Conductance and Induced Grid Noise

The random fluctuations in the electron stream emitted from the hot cathode - see shot noise - in passing through the grid structure on the way to the anode, will induce a noise voltage on to the grid from the electrostatic charge carried by the electrons themselves.

The magnitude of this will depend on frequency. At low and medium frequencies, this noise will be self-neutralized as the voltage induced by the electrons approaching the grid will be cancelled by an equal and opposite voltage induced by those receding from the grid. At higher frequencies, however, 20 MHz and upwards, the actual time taken by an electron to pass from the cathode to anode will be an appreciable fraction of the operating cycle, the fraction increasing with frequency. This transit time, as it is known, results in a difference in phase between the electrons approaching the grid, and those receding from it. The result is that not all the induced voltage is cancelled.

As the frequency is raised, the magnitude of the current induced in the grid also rises. This approximates to a resistive thermal noise generator operating at a relatively high temperature. This effect is normally termed transit time conductance. The value of which is proportional to the square of the operating frequency.

In order to assess the merit of a valve for receiver purposes two parameters are often quoted: (a) Equivalent Noise Resistance and (b) Noise Factor.

Equivalent Noise Resistance

This is defined as an ideal resistance which, when maintained at a normal operating temperature, would, if placed in the grid of a noiseless valve, produce anode current fluctuations equal to shot and partition noise of an actual valve of similar characteristics.

This method of assessment has been used for many years on the continent, but it is only really of value in the frequency range 50 kHz to 20 MHz. It is of little or no value in equating a valve for suitability in v.h.f. or u.h.f. applications.

The equivalent noise resistance for various types of valves can be calculated as follows:
(i) Triode (Shot Noise only)

$$R = \frac{2.5}{G_m} \text{ ohms}$$

(ii) Pentode or Tetrode (Shot and Partition Noise only)

$$R_{eq} = \frac{I_a}{I_a + I_{g2}} \times \left(\frac{2.5}{gm} + \frac{20 \, I_{g2}}{gm^2} \right) \text{ ohms}$$

(iii) Triode Mixer (Shot noise only)

$$R_{eq} = \frac{4.0}{gc} \text{ ohms}$$

(iv) Pentode or Multigrid Mixer (Shot and Partition Noise only)

$$R_{eq} = \frac{I_a}{I_a + I_{g2}} \left(\frac{4.0}{gc} + \frac{20 \, I_{g2}}{gc^2} \right) \text{ohms}$$

In the foregoing formulae, the value of I_a and I_{g2} is expressed in amps, while that for Gm and Ge are in amps/volt.

From these calculations, it will be seen that the lower the equivalent noise resistance, then the better the valve. While this is true at frequencies where the induced grid noise is not troublesome, and while valves with a high Gm will produce the lowest equivalent noise resistance, this does not automatically make them the most suitable for every application.

Noise Factor

There is more or less a standard method of expressing the performance of a valve intended for service as an r.f. amplifier on frequencies above 20 MHz; this is quoted as its noise factor. Assuming that the first stage has adequate gain, this can also be used to state the performance of receivers and converters.

The measurement of noise factor is carried out using a diode noise source such as the A2087 whose noise performance can be calculated from:

$$F = \frac{e}{2kT} I_d R_s$$

where
- e = the electron charge = 1.59×10^{-19} coulomb
- k = Boltzmann's constant = 1.372×10^{-23} joules per deg.K
- T = Temperature of source resistance in deg.K
- I_d = noise diode anode current in amperes
- R_s = Source resistance in ohms

In cases where the normal operating temperature is of the order of 17deg.C - 290deg.K - the formula can be simplified to

$$F = 20 \, I_d R_s \text{ as a ratio}$$

or $\quad F = 10 \log (20 \, I_d R_s)$ in decibels

Noise diodes are available for frequencies up to about 500 MHz or so, but when measurements are made at or near the maximum frequency of the noise source, some precautions are necessary to avoid producing optimistic results, and corrections are often needed. For most purposes, however, the diode noise generator is an entirely reliable and stable source producing repeatable results, and when used up to about 150 MHz little or no correction is needed provided that a suitable diode is employed.

The circuit of a diode valve noise generator is given in Fig.10.40. This will perform the tests required on most amateur equipment for frequencies above 20 MHz, and possesses a high degree of repeatability over long periods. It should be noted that if a CV2398 diode is used in place of the A2087, then the filament supply must be increased to 6 volts.

For comparative tests, and when repeatability over a short period is adequate, then the simpler germanium diode noise generator given in Fig. 10.43 will be found quite satisfactory.

Full details of the construction and use of noise generators may be found in Chapter 19 of the RSGB *Radio Communication Handbook*.

At frequencies of 500 MHz and above, it is usual to find that an inert gas discharge tube is employed as the noise source. Some special noise diodes however can be used up to 1000 MHz but corrections for transit time errors must be made and it is for this reason that the gas discharge tube usually takes over.

In a gas discharge tube, the positive column of gas emits electromagnetic radiation, and this can be readily coupled into either a co-axial line or wave guide circuit.

Transit Time Conductance and Induced Grid Noise

Above 20 MHz the induced grid noise effects predominate, imposing a shunt conductance across the input circuit of an amplifier. This shunt conductance, Ge, can be calculated as follows:

$$G_e = \frac{gm \cdot 5 \cdot (af)^2}{V_1 \cdot 10^{15}} \left(1 + \frac{3.3 \, b/a}{1 + (Va/V)^{1/2}} \right) \text{(A/V)}$$

This relationship assumes the following:

(a) The valve geometry is planar.
(b) The initial velocities of the electrons leaving the cathode surface are zero.
(c) The emission is space charge limited.
(d) The grid plane can be considered as equipotential.
(e) The μ value is large.
(f) The signal voltages on grid and anode are small.
(g) The transit angles through the cathode-to-grid and grid-to-anode spaces are small.

where:
- $V1 = 5.69 \times 10^3 \times a^{4/3} \times J_c^{2/3}$ (volts).........
- gm = mutual conductance of the valve under the given operating conditions (A/V)
- a is the grid cathode spacing (cm)
- b is the grid anode spacing (cm)
- f is the frequency in hertz
- Va is the anode voltage (V)
- J_c is the cathode current density of the valve under the given operating conditions (A/cm2).

Hence $\quad G_e \propto f^2$

The transit time shunt conductance Ge should not be confused with the shunt conductance placed across the input terminals of an amplifier containing a valve, either in the grounded cathode or grounded grid connection.

In the case of the grounded cathode connection, shown in Fig.5.19, the interaction between the valve and the cathode lead inductance produces a

conductance Gc, placed across the amplifier input, of value given by:

$$G_c = gm (4\pi^2 f^2) \, L_c \, cg\text{-}k \text{ (A/V)}$$

where:

gm is the mutual conductance of the valve in A/V.

f is the operating frequency in hertz

L_c is the cathode lead inductance in henries.

cg-k is the grid-cathode capacitance in farads.

here again, $G_c \propto f^2$

G_c does not contribute any noise to the amplifier but it does serve to apply extra damping to the input of the amplifier, which is not usually desirable.

In the case of the grounded grid connection, a shunt conductance term G_{in}^- is applied across the input, taking into account the valve, the transit time conductance G_e and the load applied to the output of the valve circuit.

G_{in} is given by:

$$G_{in} = G_e + \frac{gm}{1 + 1/(ra \, G_L)}$$

where gm is the valve mutual conductance.

G_L is the external load shunt conductance

r_a-1 is the valve anode conductance

To minimise the noise factor of an amplifier, the effects of the equivalent noise resistance and transit time conductance, considered as noise sources, must be made as low as possible at the operating frequency; always bearing in mind the importance of sufficient gain to eliminate the effects of second stage noise contributions.

Noise in Transistors

Noise in transistors can be broadly divided into three classes, which are similar to those affecting valves, these are:- Flicker noise,
Thermal noise and
Shot noise

Flicker noise occurs at low frequencies and is generally increased as the frequency is decreased, it therefore is only likely to concern us in audio amplifiers such as microphone and speech amplifiers in transmitters or the i.f. and audio stages of a receiver.

Thermal noise is caused by the random motion of charges within the device. This motion gives rise to electrical power which is proportional to absolute temperature and the band width. The noise voltage (rms^2) which appears across the device terminals is equal to

$$4kTRf_{bw} \text{ volts}^2$$

where

k = Boltzmann's constant = 1.38×10^{-23} joule /K.deg.

T = Absolute temperature in K.deg.

R = Resistance in ohms.

f_{bw} = frequency bandwidth in hertz.

Note, that this is very similar to the noise factor of a valve.

Shot Noise occurs under conditions of current flow. The current flow may be caused by a field, as in the case in a valve or by diffusion as in the case of transistors. It is caused by the random nature of the arrival of the charges. If these charges arrived uniformly, a single frequency would be generated. The shot noise energy associated with a stream of charges (carriers) - the d.c. current - is proportional to the charge of an electron, the d.c. current flowing and the band width.

The transistor itself has in effect three self-contained noise generators within the device:
(a) Shot noise in the emitter base junction,
(b) Thermal noise in the base resistance, and
(c) Shot noise in the collector-base junction.

In spite of these problems the modern transistor and especially the Field Effect type will in general provide a lower noise r.f. amplifier or receiver front end than valves, but care must be taken to prevent accidental overload occurring and so far no adequately fast protective device has been developed.

VALVES SUITABLE FOR R.F. AMPLIFIERS

As indicated from the previous section on the causes of noise, one of the factors, that due to partition, virtually forces the triode as the only suitable type design. Pentodes are too noisy for low noise amplifiers, except possibly on 70 MHz where they may be acceptable, but even at this frequency the triode will show some benefit. At 144 MHz and above they are essential. Some recent transistors have good low noise performance and those of the Field Effect type are especially useful,their very much higher input impedance and freedom from cross modulation effects make them the first choice for new equipment.

Ordinary bi-polar transistors, although some have good noise factors, are very low input impedance devices and have to be well tapped down on the tuned circuit. In addition they generally also suffer from cross modulation effects, are sensitive to r.f. input and can be easily damaged.

The choice then really is between specially designed valves and Field Effect transistors and in the following details those suitable are given, together with circuits for normal low noise transistors, which in any case can be successfully used in mobile and portable equipment.

As mentioned earlier the signal-noise ratio is the most important factor for any receiving equipment and it is most desirable that the noise factor is measured. Equipment to do this is quite simple, and either a saturated thermonic diode, from which the noise output can be calculated, or a noise source using a semiconductor diode, which can be used for comparative tests, should become part of the serious amateurs' test equipment. Details of a noise generator are given later in this Manual. Apart from actual measurements if an increase in noise can be detected

when the aerial is connected the converter has an acceptable noise factor.

Although the noise performance of an r.f. amplifier is very largely controlled by the valve or transistor employed, the circuit itself should be as good as it can be made.

The tuned circuit should have as high Q as possible and both the aerial and the amplifier correctly matched, so that as much energy as possible is transferred from the aerial to the amplifier itself.

As will be found from noise generator tests, the best noise factor will not coincide with the maximum signal gain, but preliminary tuning of the circuit for maximum signal output should be done initially, and then varied as shown by the noise measurements.

The three conventional circuit configurations are

grounded cathode, grounded grid and cascode, their relative advantages and disadvantages are set out below.

Valves most suitable for these applications have high mutual conductance and as low capacitance as possible. In assessing the mutual conductance it is important to relate this to the standing anode current and it is therefore in terms of mA/V per milliamp of anode current.

In an r.f. amplifier, whichever circuit arrangement is used, the input and output circuits should be screened from one another. Such a screen is best fitted across the valve socket and should be a reasonably close fit to the profile of the socket. Careful positioning will enable the maximum isolation between circuits without interfering with the wiring to the socket connections.

CIRCUIT	ADVANTAGE	DISADVANTAGE	APPLICATION
Grounded Cathode	High power gain. Low input conductance.	Requires neutralising.	Narrow band amplifiers up to 300 MHz.
Grounded Grid	Wide band characteristic very stable. Low input conductance.	Low power gain.	Wide band amplifiers above 150 MHz.
Cascode	Stable high power gain. Low input and output conductance.	Requires two stages one of which requires neutralising.	Narrow band amplifiers up to about 300/400 MHz.

(a) Inductively Neutralized   (b) Capacity Bridge Neutralized   (c) Grounded Grid
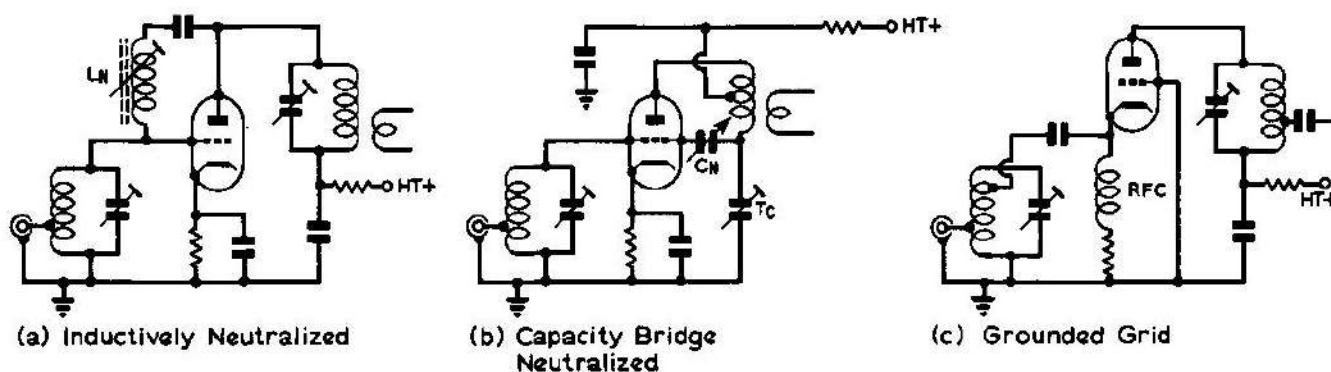
Fig. 5.19. Three types of r.f. amplifier.

As mentioned earlier, triodes operating with grounded cathode need to be neutralised to prevent self oscillation due to the relatively high anode to grid feed back capacitance. There are two normal methods by which this can be achieved:-

(a) Inductive neutralising (Fig. 5.19a)
(b) Capacity bridge (Fig. 5.19b)

The latter is generally rather simpler to arrange and adjust. In the circuit of Fig.5.19(b) the anode tuning capacitor TC should have an operating value as near as possible to that of the valve's anode to all other electrodes capacity. In cases where the anode circuit is

parallel tuned it may be necessary to introduce a balancing capacitor, in addition to the tuning capacitor.

Valves for use as grounded cathode operation at v.h.f. and u.h.f. are normally fitted with a number of cathode leads, in order to lessen the series inductance of the cathode connection. Grounded grid r.f. amplifiers have a distinct advantage in obtaining stability without neutralising due to the much reduced capacitance between the anode and cathode when the grid is operating as an earthed screen. Input and output circuit isolation is of course necessary as for

neutralised amplifiers. Valves intended primarily for use under these conditions are either fitted with a number of leads to the grid, or are of the disc seal type of construction, thus enabling a very low inductance connection to the grid. There is, however, some difficulty in matching the input to the low impedance of the valve which is generally in the region of about 100 ohms. As indicated in Fig.5.19(c) the valve may be tapped down the input tuning circuit which will often give a worthwhile improvement in performance. In simple amplifiers the aerial coaxial feeder may be connected directly to the valve as far as r.f. is concerned, i.e. through a suitable isolating capacitor to enable cathode current bias to be applied to the valve.
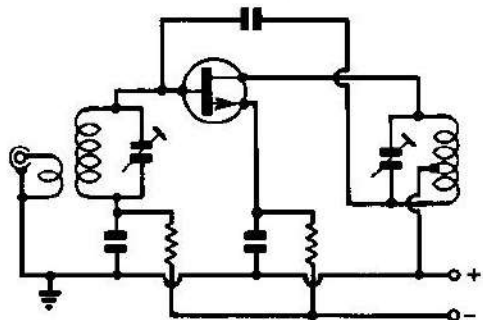
Fig. 5.20. Showing a FET neutralised amplifier.

In Fig.5.20 an example of the amplifier given in Fig.5.19(a) is shown using a Field Effect transistor in place of the valve. The higher input impedance of this type of semiconductor enables them to be substituted for the valve without a great deal of re-designing of the tuned circuit.

Standard types of semiconductors intended for r.f. amplifier service at these frequencies can, of course, be used, but it will be necessary to match them into the tuned circuits by tapping on to the inductor, usually well down from the high potential end.

Cascode Amplifiers
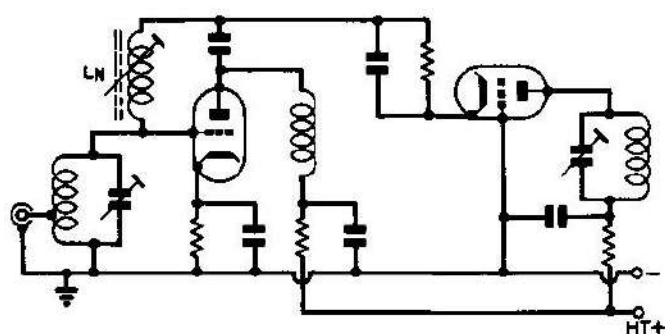
Fig. 5.21. Showing a cascode amplifier.

In Figs. 5.21 and 5.22 are shown typical cascode amplifiers. In this amplifier arrangement the first stage operates as an impedance matching device feeding into a grounded grid following amplifier. The first stage must be neutralised, usually as shown by the inductive

method although capacitative arrangements have been used in some designs.
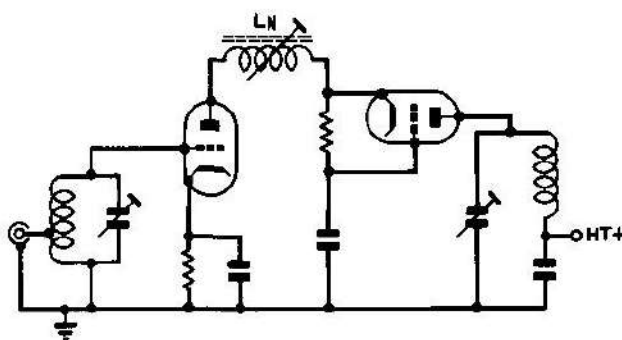
Fig. 5.22. A simplified or series cascode.
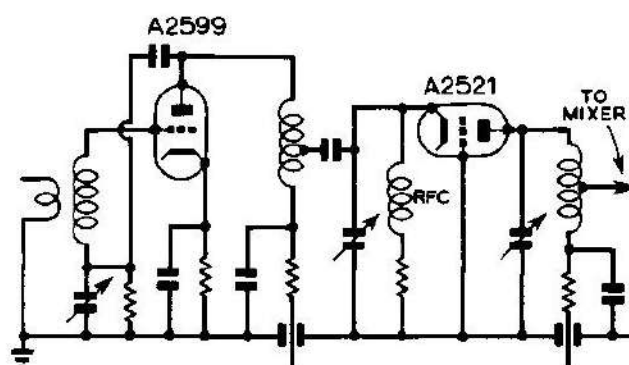
Fig. 5.23. The Author's form of modified cascode amplifier.

In the original form, which may be either a.c. coupled, Fig. 5.21, or series d.c. coupled, Fig. 5.22, the first stage has low gain due to the very heavy damping of the output circuit by the following stage. Some improvement in gain can be arranged by the method shown in Fig. 5.23 in which the first stage is capacitatively neutralised and the second stage is tapped down the anode circuit of the first stage, thus permitting a significant increase in gain and improved selectivity. The neutralisation in this arrangement needs to be more complete than in the original form, in order to preserve the stability of the stage with the reduced damping applied to the output circuit.

TRANSISTORS FOR PRE-AMPLIFIERS AND CONVERTERS

Transistors of high performance and low noise with satisfactory stage gain are freely available and suitable for most receiver pre-amplifiers or converters. Equally they are suitable for the local oscillator and multiplier use.

Generally, unlike valves, few types are entirely interchangeable and the list of available suitable types is very extensive, the typical practical designs given later quote the particular type used but equivalent performance types may, of course, be substituted.
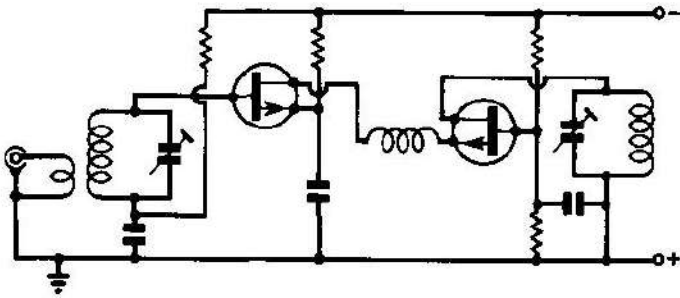
Fig.5.24A. A FET cascode.

Designs are also given for Field Effect transistors, these types having significant advantages over the bi-polar types especially in so far as they, like valves, are voltage operated devices and are of substantially similar input impedance, whereas the bi-polar transistor is essentially a current operated device of low impedance.

One of the major problems of the bi-polar type transistor is that they are rather transparent, i.e. they will accept and amplify unwanted signals often far removed from the tuned circuit resonant frequency; this is due largely to the lack of isolation between the input and output circuits. Field Effect types show a marked improvement in this respect.

When the device is of low impedance it is essential to tap the feed points down on the tuned circuit in order to obtain a reasonable "Q", and hence a sufficiently narrow band width. As mentioned earlier Field Effect types being higher impedance devices can be operated almost the same as a valve and in many cases typical valve type components are quite suitable for use with these devices.

Noise figures of semiconductors are now significantly lower than those available from valves, but generally, they are easily damaged by overloads which valves would withstand without trouble. It is important therefore, to ensure that there is adequate protection against over-load particularly where there is any chance of poor isolation in a transmit-receive relay.

Mast head pre-amplifiers, provided suitable over-load protection is provided, enable superior receiver performance to be obtained, and sufficient gain is readily available to more than overcome the noise contributed by the feeder cable. Complete pre-amplifiers with a small long life battery can be installed with or without switching the battery supply and the life of suitable batteries at the low current drain is likely to exceed 12 months and renewal can be included in the annual overhaul of the aerial.

Diode Multipliers
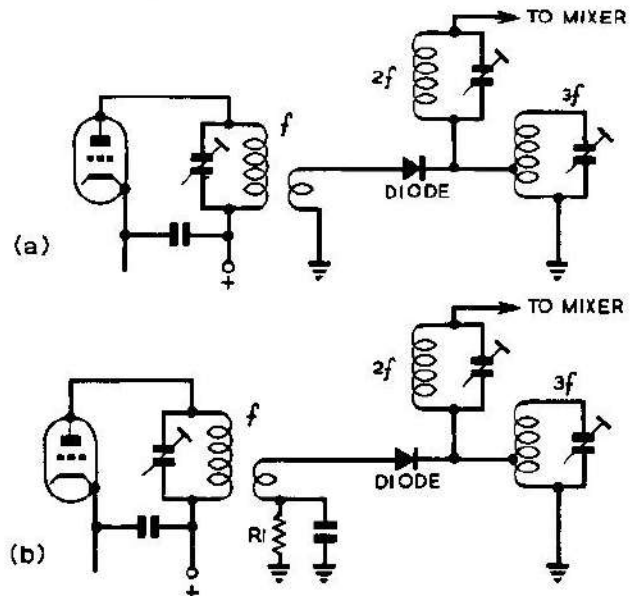
Fig.5.24. Two types of diode multiplying circuits.

The small "crystal" diode, such as the GEX66, or 1N82, offer an attractive means of frequency multiplication in converters, especially on the higher frequencies, such as for 70 and 23 cm. Like their transmitting counterpart, the varactor diode, and unlike the valve, it is necessary to do something about the unwanted harmonics. In the case of a valve a simple tuned circuit at the required frequency is usually sufficiently selective to reduce other harmonics to a
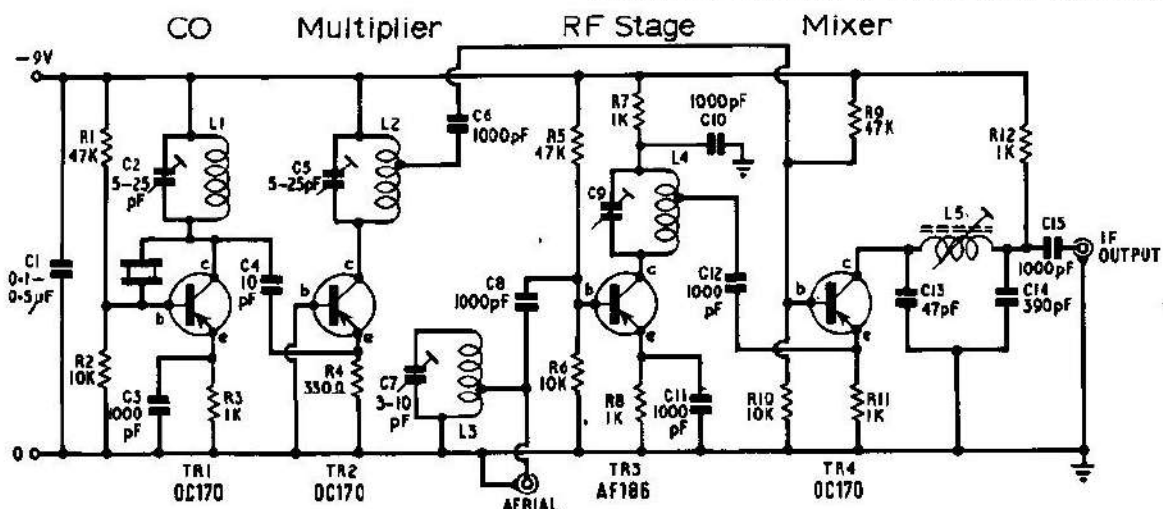

Fig.5.25. A practical 2m converter using bi-polar transistors. Coil details appear on page 5.19

low enough value not to be troublesome. Whereas in a diode multiplier used as a tripler, a trap circuit for the second harmonic is needed to prevent unwanted response to signals which when mixed with the second harmonic fall into the passband of the receiving system. In spite of this disadvantage, their use as the final stage of frequency multiplication is a great convenience, no power supply, other than r.f. is required and this enables them to be wired directly in the circuit. As they are low impedance devices they can conveniently be driven by a simple loop coupled to the previous tuned circuit and a typical arrangement is shown in Fig.5.24(a).

In Fig.5.24(b) some improvement in performance can be obtained by setting the diode bias to its most suitable value by choice of the resistor R1. This will vary depending on the diode and the frequency multiplication factor. As indicated in both the diagrams the diode should be tapped well down the tuned circuit.

A CONVERTER USING BI-POLAR TRANSISTORS FOR 2M or 4M

In this converter the local oscillator - multiplier is provided by the two OC170 v.h.f. transistors. The crystal should preferably be of a high frequency, so that the multiplier is then operated at reasonable efficiency, which means that the stage multiplication is limited to three or four times the oscillator frequency.

The actual frequency will depend on the i.f. required and if this is 4-6 MHz then a 35 MHz crystal with a four times multiplier will be suitable, or a 44 MHz crystal with a three times multiplier giving 132 MHz would be suitable for an i.f. of 12-14 MHz.

In the circuit, L1 and C2 are resonated to the crystal frequency, and L2 and C5 are tuned to the required final frequency, either three or four times that of the crystal frequency. Output from this circuit is taken to the mixer base via the isolating capacitor C6.

The third transistor, an AF185 or other low noise v.h.f. type, is operated as the r.f. amplifier. The circuit of this stage may, of course, be replaced by the alternative neutralised amplifier previously described. The input tap, from the inductor, one turn up from the earthy end of the coil, is common with the aerial tap and goes via the isolating capacitor C8 to the base of the AF186.

The output from the collector tuned circuit is also taken from a tap near the low r.f. potential end of the inductor and this is fed into the emitter of the fourth (mixer) transistor.

Alternative Pre-Amplifiers for 2M and 4M

Both the circuits shown are similar and the component values are the same in both, except that in Fig. 5.27 the aerial input is coupled to the input circuit L1 by a coupling coil. This is necessary so that bridge neutralisation can be achieved. For simplicity Fig.5.26 is preferred but when correctly adjusted the neutralised amplifier will give a superior performance. As mentioned previously it is necessary to tap the input to the transistor well down the input inductance. The output impedance (collector) is sufficiently high to feed in at the top of the tuned circuit.
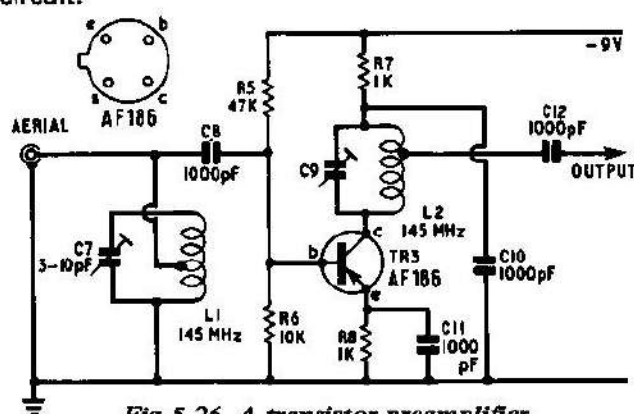

Fig.5.26. A transistor preamplifier.

2M CONVERTER COIL DATA

INDUCTOR	NO. OF TURNS	WIRE SIZE	DIAM.
L1	15	26 s.w.g.	¼" former for crystal 44 MHz.
L2	6	18 s.w.g.	¼" self supporting tapped 1 turn for 132 MHz.
L3	6	18 s.w.g.	¼" self supporting tapped 1 turn.
L4	6	18 s.w.g.	¼" self supporting tapped 1 turn.
L5	55	30 s.w.g.	¼" closewound for i.f. = 4-6 MHz.
	25	30 s.w.g.	¼" closewound for i.f. = 12-14 MHz.

4M CONVERTER COIL DATA

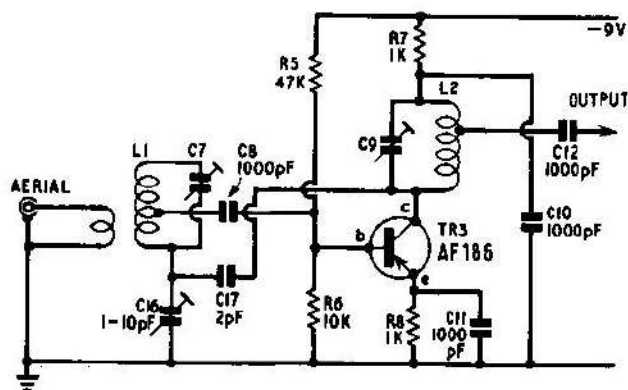| INDUCTOR	NO. OF TURNS	WIRE SIZE	DIAM.
L1	25	26 s.w.g.	¼" former for crystal 16.25 MHz.
L2	11	18 s.w.g.	¼" self supporting tapped 1 turn for 65 MHz
L3	10	18 s.w.g.	¼" self supporting tapped 1 turn.
L4	14	18 s.w.g.	¼" self supporting tapped 2 turns.
L5	50	30 s.w.g.	¼" close wound for i.f. = 5.1-5.7 MHz.

Fig. 5.27. An alternative circuit to that in Fig. 5.26.

The position of the tap on the output circuit will depend on the following stage and this is most likely to be a low impedance input coupling, normally fed by an aerial or it may be the input to a transistor mixer. In these cases the tap should again be well down towards the low potential end of the inductance. Alternatively, if this is to feed into a valve mixer, then the tap may be nearer the high potential end of the circuit. In any case the points of both the input and output circuits should be adjusted for optimum results.

Suitable coil sizes are:-

2M	L1	6 turns	18 s.w.g.	¼ diam.	tapped	1 turn
	L2	6 turns	18 s.w.g.	¼ diam.	tapped	1 turn
4M	L1	10 turns	18 s.w.g.	¼ diam.	tapped	1 turn
	L2	14 turns	18 s.w.g.	¼ diam.	tapped	2 turns

The output circuit of the mixer, a pi circuit consisting of L5, C13 and C14, will provide a good match to the input of the following receiver. In this case only the inductor is made adjustable, which is simpler than tuning this circuit by the capacitors. Some adjustment of the aerial trimmer in the main receiver may be necessary to obtain equal performance over the whole of the 2m band. This is not likely to be necessary on 4m where the band width is less.
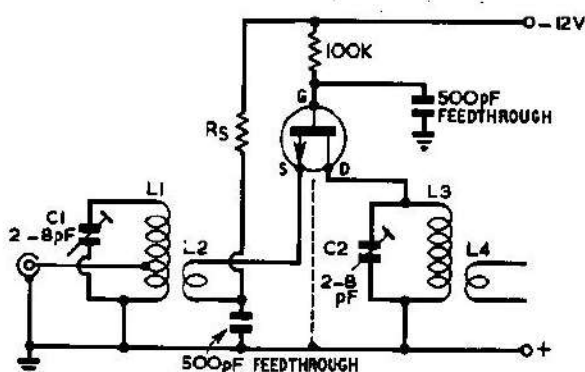
Fig. 5.28. An FET pre-amplifier for 144 MHz. L1, L3 6 turns ¼" dia. L2, L4 1 turn coupling coil. C1, C2 2-8 pF trimmers. Rs, the value should be suitable for the transistor used. For a TIS34 the current should be 4mA.

The power requirements for this converter is 6mA at 9 volts. If operation is to be from a 12 volt car battery, then a 500 ohms resistor should be connected in series with the negative line.
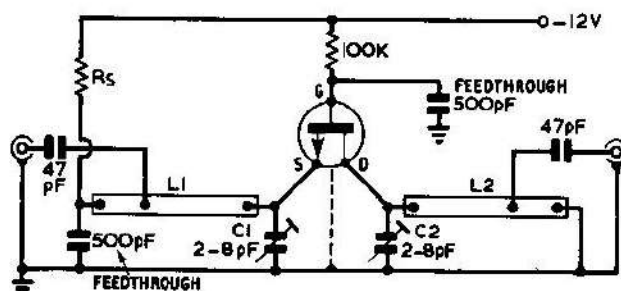
Fig. 5.29. An FET pre-amplifier for 432 MHz. L1, L2 quarter-wave line circuit tuned by C1 and C2. C1, C2 2-8 pF trimmers. Rs adjusted as for the 144 MHz pre-amplifier.

A Neutralised 6CW4 Amplifier for 144 MHz.

In this pre-amplifier the valve is used in the standard grounded cathode mode with bridge neutralising, and apart from its intrinsic value as a useful device, it also serves to illustrate the principle of the method of neutralisation of a triode r.f. amplifier.
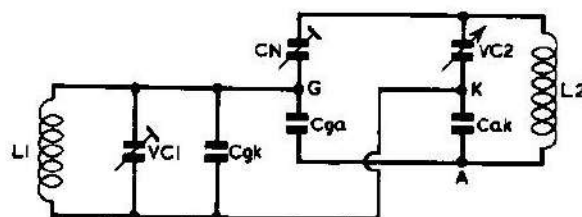
Fig. 5.30. Equivalent circuit. Cgk = capacity between grid and cathode. Cak = capacity between anode and cathode. Cga = capacity between grid and anode. Other components marked as in practical circuit.
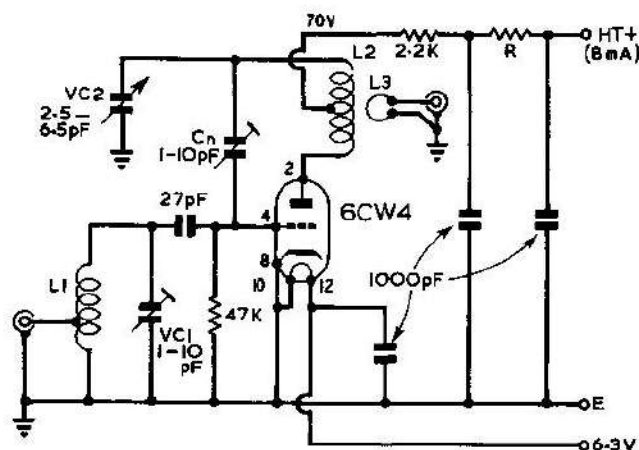

Fig. 5.31. A practical pre-amplifier. L1 5 turns 20 s.w.g. tapped at 1¾ turns ¼" dia. ¾" long. L2 8 turns 16 s.w.g. tapped 4½ turns from anode end. ¾" dia. 1" long. L3 1 turn link coil of insulated wire fitted around centre of L2. R adjusted to suit h.t. for an anode current of 8mA.

The equivalent circuit shows clearly the bridge configuration of the amplifier, the bridge being formed by Cn, Cga, VC2 and Cgk and it will be balanced when Cn/Cga = VC2/Cak. Under this connection feedback from anode to grid will theoretically be zero. The practical amplifier is constructed in a diecast box and the component positions together with the inter circuit screening is shown in the layout diagram. A noise factor of approximately 3dB is obtainable with this amplifier at 145 MHz, optimum adjustment of noise can only be achieved by use of a noise generator.
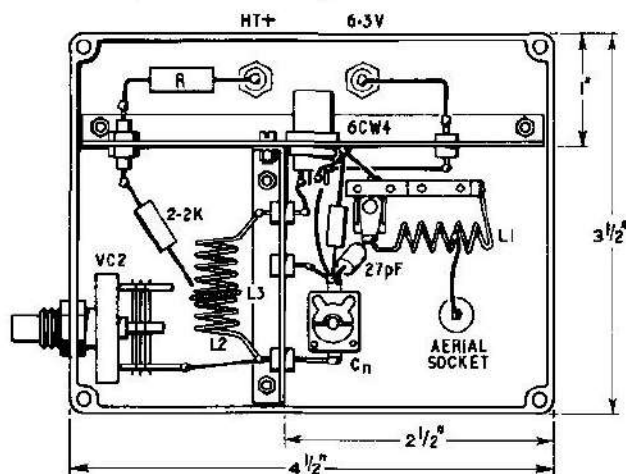
Fig. 5.32. Drawing showing layout. The pre-amplifier is built inside an Eddystone box 3½" x 4½". The output socket is mounted immediately below L3.

70cm PRE-AMPLIFIERS

Grounded Grid Using 8058 Nuvistor
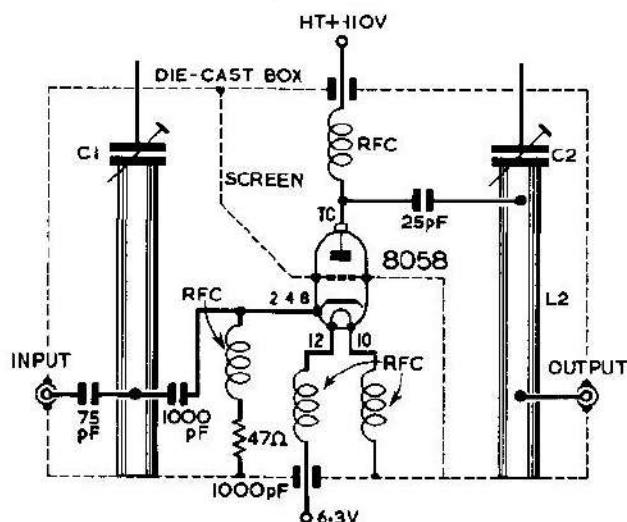
Fig. 5.33. Diagram of 8058 70cm pre-amplifier. R.f. choke is 12 t. 26 s.w.g. ⅛" diameter. L1 and L2 are ½" dia. Brass or copper tube ¼" bore. C1 and C2 are 2BA screws with low loss insulation fixed over thread to prevent short circuit between trimmer screw and body of the line. As an alternative the method described under tuned circuits could be used. Dimensions are given in Fig. 5.34.

MATERIAL – 20 swg
COPPER SHEET
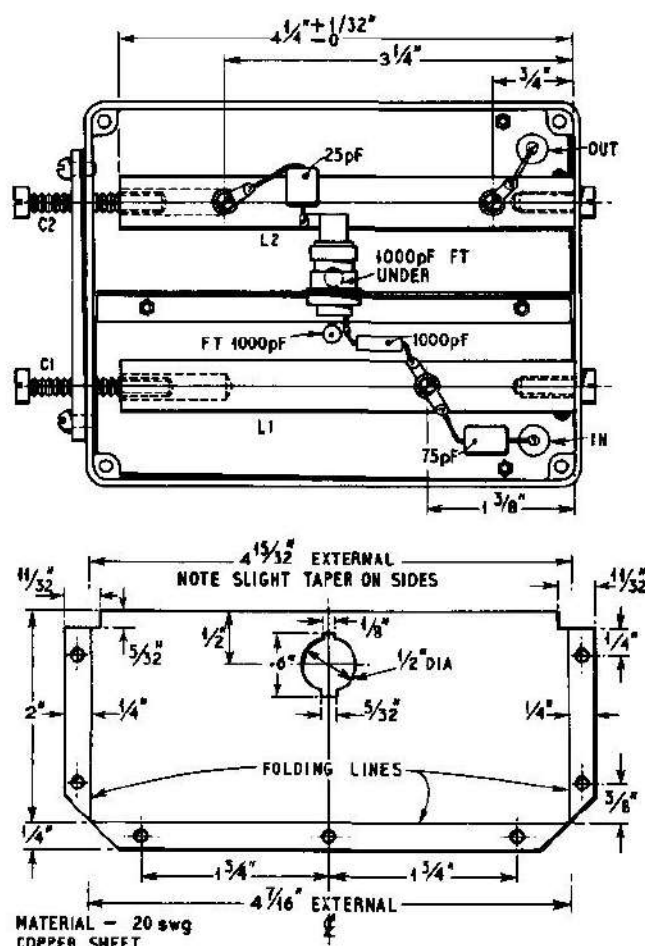
Fig. 5.34. Showing the main components and dimensions of lines of the 70cm pre-amplifier. The screen dimensions and drilling details are also given.

As mentioned earlier grounded grid connection is suitable for r.f. amplifiers at frequencies above about 150 MHz, and valves specially designed for this type of use can be expected to give a better performance than those of the usual design for grounded cathode, connected for grounded grid operation, because of the lower grid lead inductance.

The 8058 fulfils the requirement very well and can be expected to give a good performance, with little trouble provided the circuit arrangement is as shown with input and output screened from one another by the screen on which the valve is mounted. The circuit is drawn to show the parts located in the input section with the surrounding box shown as the overall enclosure. The input and output tuned circuits are lines tuned to resonance by the end capacitors. The aerial, cathode and anode and output connections are tapped on to these lines at appropriate points. The actual position of the tapping used should be found by using a noise generator.

A Neutralised 6CW4 Nuvistor

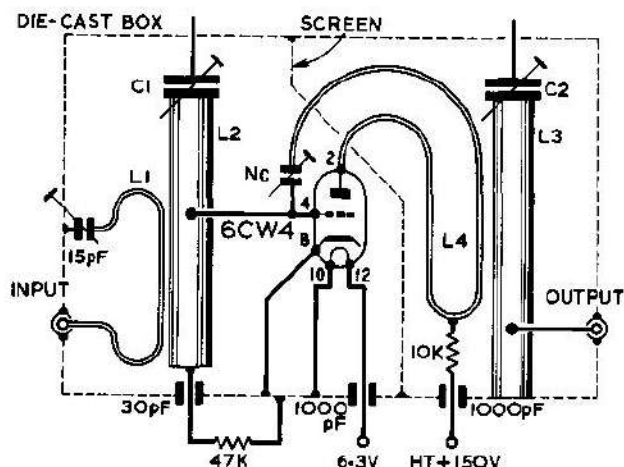In this arrangement, the grounded cathode 6CW4 valve is neutralised in a bridge circuit where NC should

5.21

Fig. 5.35. A 6CW4 pre-amplifier. Details of L1 and L4 are given in Fig. 5.36. L2 and L3 are ½" dia. brass rod or tube counter bored as shown.
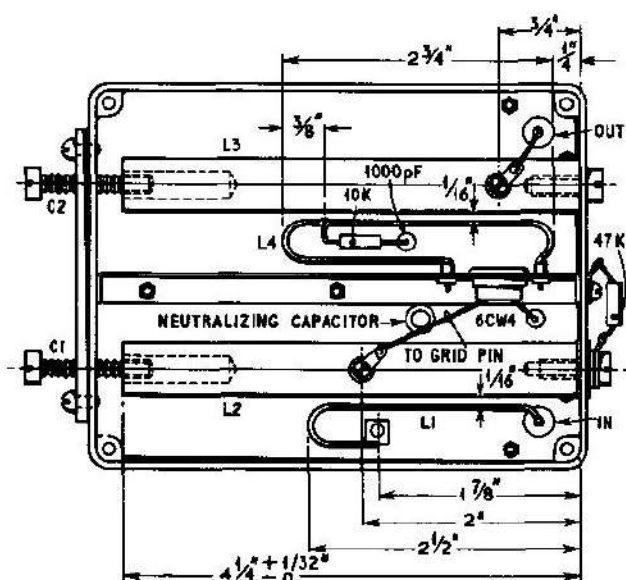

Fig. 5.36. Layout of lines and screen of 6CW4 pre-amplifier. L1 and L4 are of 16 s.w.g. and are of shape and dimensions shown.
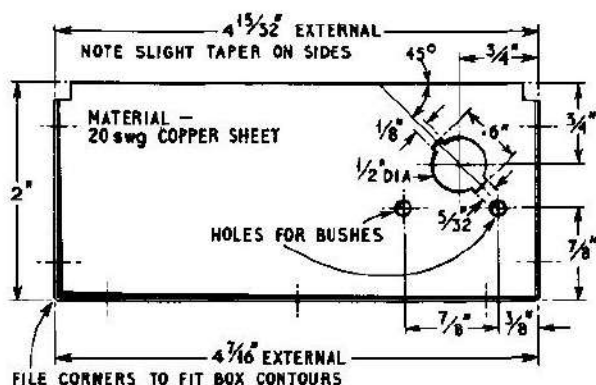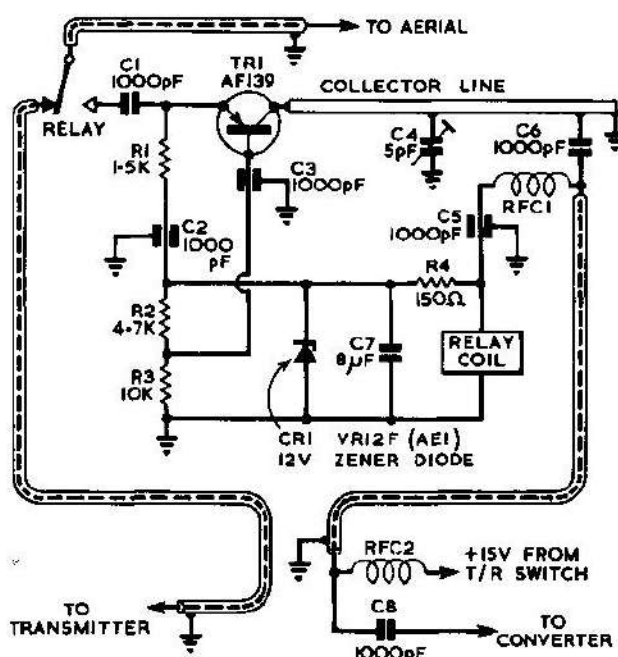

Fig. 5.37. Dimensions and drilling details of central screen upon which the valve is mounted.

be adjusted to a value equal to the capacitance of the valve when hot. Both the input and anode circuit are inductively coupled to the High Q linear circuits while the output is connected to a tap on L2. The general design is the same as for the grounded grid pre-amplifier where the encasing box is divided into input and output sections by the screen which is fitted centrally into the box. The valve mounting position is, however, changed to provide suitable layout for the tuned anode circuit. Both the input and anode circuit positions should be adjusted for optimum performance. The best noise factor can only be adequately assessed using a noise generator and details of these are given in Chapter 10. The trimmer capacitors tuning the line circuit can take any form which does not signficantly lower the unloaded Q. Tuning this neutralised type of amplifier is less straightforward than with the grounded grid type because it has to be neutralised. The line circuits should be adjusted for maximum signal and then the neutralising capacitor should be adjusted for minimum, both adjustments should be made with a strong local signal but without h.t. on the valve.

Masthead Pre-Amplifier for 70cm

Fig. 5.38. Circuit of masthead pre-amplifier showing aerial changeover relay and method of supplying operating voltage.

The use of a pre-amplifier at the masthead will enable the realisation of the best reception of which the equipment is capable, noise introduced by the cable being largely overcome together with a reduction of the signal attenuation caused by the cable itself. Although the circuit given is for 70cm the same method is equally applicable to the other bands, especially the higher frequencies where cable losses

and noise contribution increase rapidly with frequency. Care is needed in the mechanical layout to ensure that there is a minimum pick-up by the transistor when the aerial switched to transmit. It should be noted that the relay is connected so that its normal unactivated position is for transmit and that for reception it is "switched-on." This allows the aerial to be used even if there is a fault in the pre-amplifier. With care a transmitter operating at full power, 150 watts, should not damage the transistor, this may be achieved by careful test and adjustment of the cable length connecting the relay to the pre-amplifier input socket. The base current should be checked and can be reduced to about 1 µA with a 4" cable length.
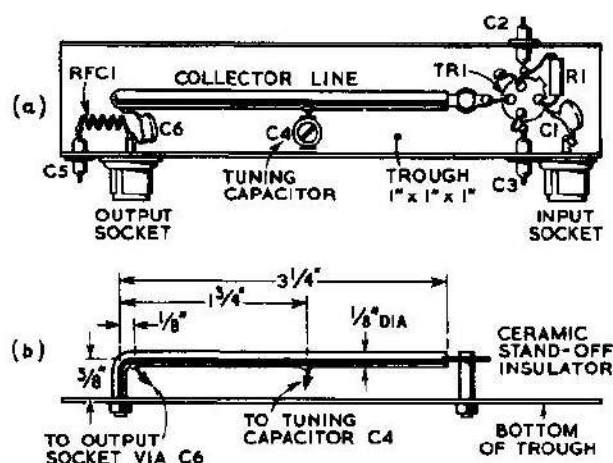
Fig. 5.39. Illustrating the general arrangement and layout of the masthead pre-amplifier.
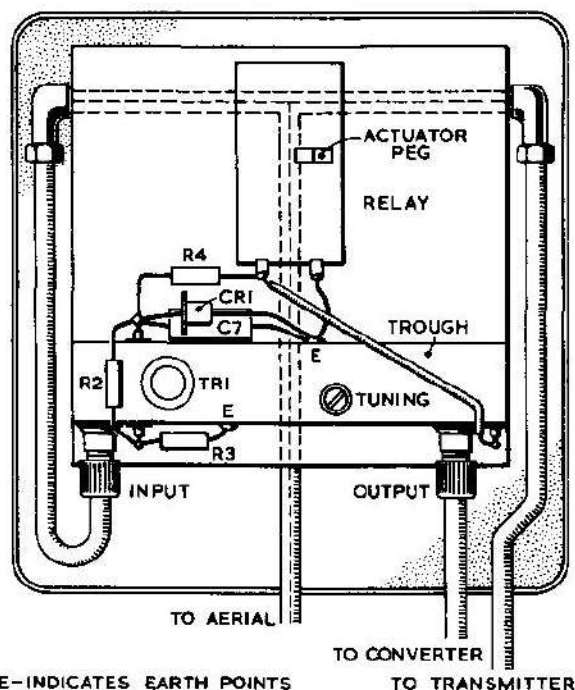
Fig. 5.40. Shows a typical arrangement of the pre-amplifier in its weatherproof box.

TUNNEL DIODE AMPLIFIERS

Tunnel diodes have been available for some years. They have been talked about for even longer, but the average amateur has very little idea of the potentiality of these devices; and this in spite of the fact that tunnel diode circuits are simple (at least on paper) and that results are very easy to achieve.
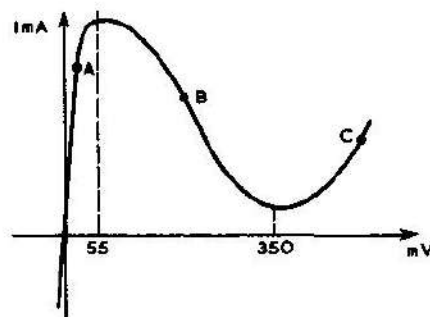
Fig. 5.41. Current plot against applied voltage to a tunnel diode.

One of the factors which has the largest effect on semiconductor action is impurity content. If the impurity content in the semiconductors forming a junction diode is reduced, the reverse voltage capability of the diode is increased. Equally, if the impurity content is increased, the reverse voltage possible is reduced. If, however, the impurity content is made sufficiently high, around 2×10^{19} atoms per cc in Germanium, the reverse breakdown voltage is reduced to zero and the diode becomes almost an ordinary conductor in the reverse direction. This degree of "doping" is said to make the semiconductor "degenerate".

The fact that the reverse breakdown voltage drops to zero is not by any means all that happens. If degenerate p and n semiconductor materials are brought together under very carefully controlled manufacturing conditions to make an extremely abrupt junction, of the order of 150 A° in thickness, the forward characteristic is also affected. Drawing a graph of current against voltage, one obtains a curve similar to Fig.5.41. Starting from zero volts across the diode, the current at first increases more or less
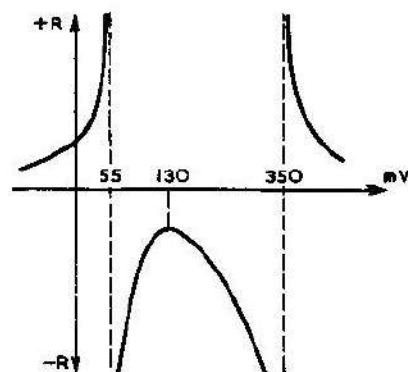

Fig. 5.42. A.C. resistance plotted against voltage.

5.23

linearly in the forward direction. At about 55mV, the current levels off and then starts to decrease until at 350mV (for germanium) it reaches a minimum and again starts to climb - more as one would have expected from a semiconductor diode.

Looking at this graph more closely, the slope of the curve at any point is a measure of the diode a.c. resistance, and this can again be plotted against voltage as in Fig. 5.42. The really interesting part of this graph is the central portion where the curve is negative. By definition, any ordinary resistor dissipates power when current flows through it. It follows then, that a negative resistance will generate power; in fact, current flowing into it will be out of phase with current leaving it. This is not a resistance in the d.c. sense: it is an a.c. resistance such as the anode resistance of a valve; a negative incremental resistance. Put this negative resistance in series with or in parallel with a load and the possibility of power gain will become more evident. Take the series case first. If a voltage V
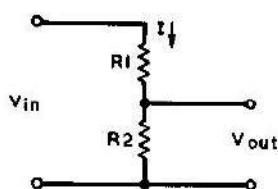
Fig. 5.43. Positive and negative resistances in series.

is applied to R1, and R2 in Fig.5.43 and the voltage across R2 is measured, it will be found to be IR2, where I is the current due to V through both resistors. V equals I(R1 + R2), and therefore the "gain" that is,

$$\frac{V \text{ out}}{V \text{ in}} = \frac{R_2}{R_1 + R_2}$$

and, if R2 is negative, the result can be greater than I. It can even approach infinity if the two Rs are equal. The parallel case is just as obvious though in this case it is better to work with conductances (reciprocal resistances) as in Fig.5.44. If sources and load
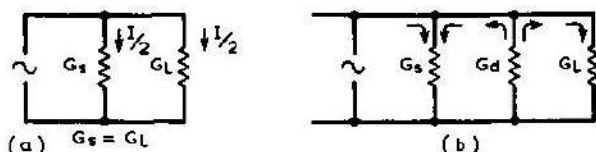

Fig. 5.44. (a) Positive conductances in parallel. (b) Positive source and load conductances in parallel with a negative conductance.

conductances are equal then the source current splits equally between the two halves of the circuit, the power in the load is $\frac{I^2}{4gL}$ and the power gain cannot be greater than 1. If the load is paralleled by a negative conductance, then current in the section will be out of phase with the driving current. That is, the negative conductance acts as an additional source and supplies additional current to the circuit. Load current can

now be greater than $\frac{I}{2}$ and power gain exceeds 1. The power gain will be seen to be

$$\frac{P \text{ out}}{P \text{ in}} = \frac{4gs \, gL}{(gs + gL - gD)^2}$$

which can again approach infinity, as can be seen by making the source and load conductances together equal the diode conductance.
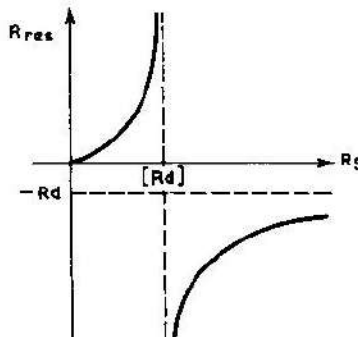

Fig. 5.45. Resultant resistance for negative (diode) and positive resistances in parallel.

Drawing a curve for the parallel case, plotting "resultant" impedance against source impedance, all other things being equal, leads to the odd-looking curve in Fig.5.45. As the source resistance is increased towards the value of the diode resistance, the resultant increases and shoots off towards infinity. That is, any current that may be flowing through the source is equally flowing through the resultant, giving rise to a much higher voltage than would appear across the source. Increase the source resistance above that of the diode and the resultant immediately reappears from negative infinity and drops towards a value equal to the diode negative resistance. But, and this is important, the resultant here is always negative, and the device will oscillate (or switch).
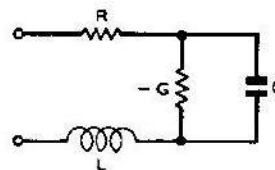

Fig.5.46. Equivalent circuit of tunnel diode for a.c.

There are many good explanations of its physical operation, but these are by no means necessary for an understanding of its circuit function. In purely practical terms, the diode is a negative conductance in parallel with a small capacitance, with the whole lot in series with an inductance and some residual resistance of the more normal kind (see Fig.5.46).

Total impedance across the terminals:

$$Z_{tot} = j\omega L + \frac{1}{j\omega C + (-g)} + R \quad \ldots \ldots (1)$$

$$= j\omega L - \frac{j\omega C + g}{(\omega C)^2 + g^2} + R$$

Equating real and imaginary parts:

$$Z_r = \frac{-g}{(\omega C)^2 + g^2} + R \quad \dots\dots\dots\dots \text{ (2a)}$$

$$Z_j = j\omega \left[L - \frac{C}{(\omega C)^2 + g^2} \right] \quad \dots\dots\dots \text{ (2b)}$$

At frequencies we can call "resistive cut-off" and "self-resonant" respectively, these will become zero:

$$\text{i.e. } R = \frac{g}{(\omega C)^2 + g^2} \text{ or } f_r = \frac{g}{2\pi C} \sqrt{\frac{1}{Rg}} = 1 \dots \text{ (3a)}$$

$$\text{and } L = \frac{C}{(\omega C)^2 + g^2} \text{ or } f_j = \frac{1}{2\pi} \sqrt{\frac{1}{LC} - \left(\frac{g}{c}\right)^2} \text{ (3b)}$$

The implication of the two equations we have ended with are quite simple. The resistive cut-off frequency, fr, is the frequency above which the diode will not amplify because its negative resistance has been effectively reduced to zero, and the self-resonant frequency fj, decides the normal maximum frequency of oscillation. The published characteristics of the IN2940 diode are:

g = 6.6 mmho

C = 5 pF nominal

L = 1muH with leads clipped very short

R = 1.5 ohms,

which gives fr as 2100 MHz, and fj as 1300 MHz. So there is obviously some u.h.f. possibility in these devices.

Looking a little more closely at the two equations and remembering that L, C and R can be added externally, if fj is lower than fr, then the device will want to oscillate. If it is higher, then it will amplify, so:

$$\frac{1}{2\pi} \sqrt{\frac{1}{LC} - \left(\frac{g}{c}\right)^2} > \frac{g}{2\pi C} \sqrt{\frac{1}{Rg} - 1}$$

$$\therefore L_g < RC$$

$$\text{or } R > Lg/C \quad \dots\dots\dots\dots \text{ (4)}$$

Also for real values of f_r $\frac{1}{Rg} > 1$

$$\frac{1}{g} > R \dots \text{ (5)}$$

Combining these two results gives :

$$\frac{1}{g} > R > Lg/C \quad \dots\dots \text{ (6)}$$

and any successful amplifier must satisfy these conditions. Actual gain is determined by matching source and load conductances to that of the diode.
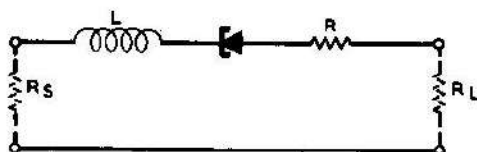

Fig. 5.47. A.c. circuit of series amplifier.

Take a practical case: a series amplifier for 145 MHz, see Fig. 5.47. Here the source and load resistances are cascaded and the total of sources and load resistance should again approach the value of diode resistance. The diode with the highest available value of negative resistance (at the inflection point) is the IN2939, and it is 150 ohms. This is only just twice 75 ohms - a usual aerial and receiver impedance value so, for the sake of argument, we will assume that impedances of the order of 40 ohms are available (actually it would not matter, as will be explained later), and that the shunt capacity across the diode is 5 pF. Considering the external source and load impedances as part of the diode residual impedance R, we can calculate the total permissible value of R from equation 3a:

$$R = \frac{g}{(\omega C)^2 + g^2} =$$

$$\frac{6.6 \times 10^{-3}}{(2\pi \times 145 \times 10^6 \times 5 \times 10^{-12})^2 + (6.6 \times 10^{-3})^2}$$

$$= 103 \text{ ohms}$$

impedances external to the diode will account for 80 ohms, the diode for 1.5 ohms, leaving 21.5 ohms to be supplied in the form of a non-inductive resistor. Now for the series inductance. Lg/C must be just less than R (at all frequencies), and working this out gives a figure of 78 mμH, of which up to 12 mμH can be accounted for in the diode leads.
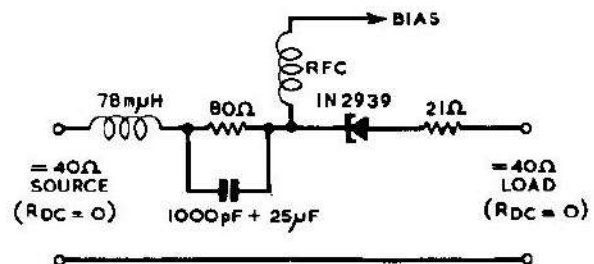

Fig. 5.48. A practical tunnel diode amplifier circuit for 145 MHz.

So the design is simply 40 ohms source and load, 21.5 ohms series resistance and 0.078 μH series inductance. If the source and load also happen to have a d.c. resistance of 40 ohms each, biassing conditions will have been met, or the diode can be supplied as shown. This kind of amplifier can give a steady 30dB gain at 145 MHz with no trouble at all, which can be increased to over 40dB at the expense of bandwidth by increasing L towards its limit of 78 mμH.

It may, of course, be said that the aerial and receiver impedances are not known accurately enough. This really does not matter much; a commercially built aerial will usually be near its design impedance, and providing it is matched to its cable, one can assume that the impedance at the bottom end is what one thinks it is. In this design there is a total of 103

ohms to use so if the aerial is of 75 ohms impedance, 28 ohms would have to be found elsewhere in the receiver. It would be just a matter of tapping down on the receiver input coil until a suitable point was found. The only possible cause of trouble would be if the impedances were substantially higher at some other frequency, which could cause oscillation.

Now a parallel amplifier. At u.h.f. the parallel amplifier is by far the easier to manipulate (the series amplifier becomes rather awkward as will be seen if figures for a higher frequency are tried in the equations above), and is conveniently in the form of a loaded quarter-wave coaxial line. Several designs have been published, most of which are impractical as far as the amateur is concerned. One beautiful design uses two diodes across a strip-line coupled hybrid to give a gain of up to 10dB - with a noise factor of 1.9dB over a bandwidth of from 270 to 625 MHz but the simple unit to be described is made with no more tools than a fine hacksaw, a drill, some taps and dies, a precision file and a screwdriver, costing not more than 10s, excluding the diode.

Designing A Typical Amplifier

The centre frequency of this unit is to be 435 MHz (a wavelength of 68.98cm). Tunnel diodes usually have a capacitance of around 5 to 10 pF, and this varies between samples, so for a parallel amplifier of reasonable length, a low impedance trough line would appear the best. Using commercially available ¾ "equal brass channel of 16 s.w.g., with a ¼" lid and a ¼" diameter centre rod, the line impedance is getting on for 60 ohms. With a line length of 5 cm, the Z_OC product is 800, so about 13 pF will be needed to tune to resonance. This gives a choice of diodes. The IN2939 and IN2940 have a design capacity of about 5 pF; the IN2969A has about 8 pF. All three have the necessary frequency capabilities. However, the conductance of the first two is 6.6 mmhos (150 ohms), which might be difficult to match, while the IN2969A is 16 mmho (63 ohms) which is more suitable since the source and load conductances are in parallel.

The IN2969A diode will want to see an admittance of 16 mmhos at its end of the line. This will be made up of two parts: that of the aerial and that of the receiver, both transformed by their respective positions on the line. From consideration of noise, the receiver output should be "undermatched," so its tapping point will be further down the line than the aerial point. We shall allow a bit over 5 mmhos for the receiver and 10 mmhos for the aerial as seen at the diode. If each has an actual conductance of 13.3 mmhos (75 ohms impedance) it is now easy to calculate where the tapping points should be. The line admittance at any point is directly proportional to the cotangent of the phase angle at that point (this means that the line impedance follows a tangent curve), but for a length of line so short in relation to a quarter
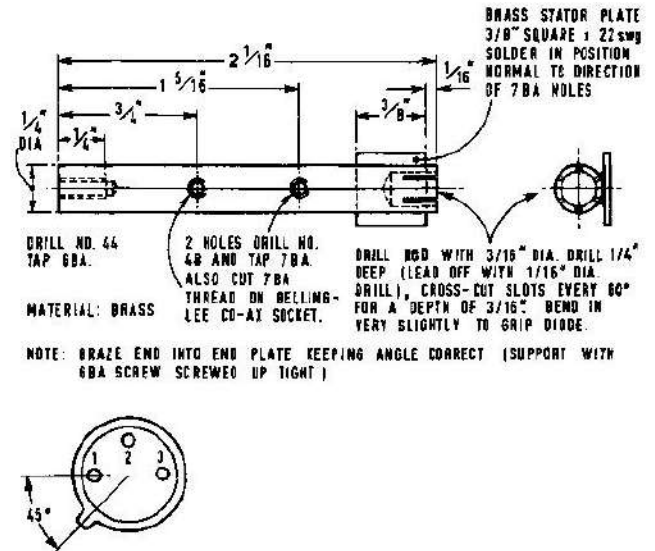
Fig. 5.49. (a) Details of the ¼" diam. centre line. (b) A TO-18 tunnel diode base. Pins 1 and 2 are the positive electrode, connected together internally. Cut these leads to ¼" and clip both to the feed-through capacitor. Pin 3, the negative electrode is connected internally to the housing and the lead should be clipped short.
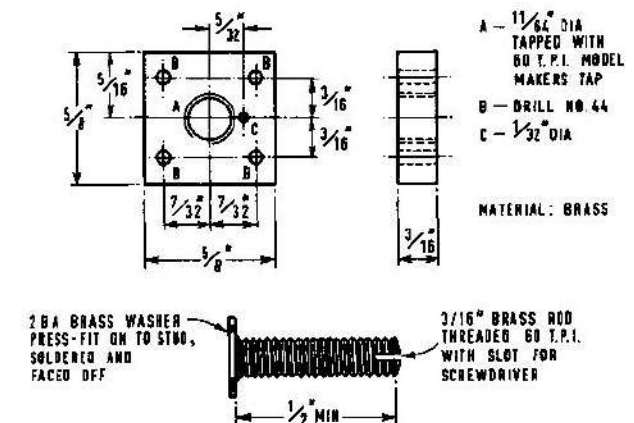

Fig. 5.50. (a) The tuning capacitor side plate, formed from ⅛" x ³⁄₁₆" brass strip. The rotor shaft spring clip is ½" of 22 s.w.g. piano wire shaped to a 'U' form, and bent at one end to fit in the ¹⁄₃₂" diam. hole C. The other end clips under one of the 8BA screws; either top or bottom left. (b) The capacitor rotor assembly.

wavelength, the change can be taken as approximately linear. The receiver point will be at $\frac{5}{13.3} \times 5\text{cm} = 1.9\text{cm}$.

and the aerial at $\frac{10}{13.3}$ 5cm = 3.8cm.

That completes the theory. We now have a 60 ohms trough line made of ¼" square channel and ¼" rod centre conductor, tapped at 1.9 and 3.8cm from the shorted end, and with a IN2969A diode doing all the work.

Constructional Details

The materials required consist of 3" of square section 16 s.w.g. ¼" brass channel, lengths of ⅛" x ¹⁄₁₆" and ⅛" x ½" brass strip, 3" of ¼" brass rod, and a few inches of ³⁄₁₆" brass rod. Also needed are a 2 BA brass washer and a 1000 pF feed-through condenser, plus, of course, screws etc. Drilling details are given in the diagrams. To tap 8 BA threads, use a
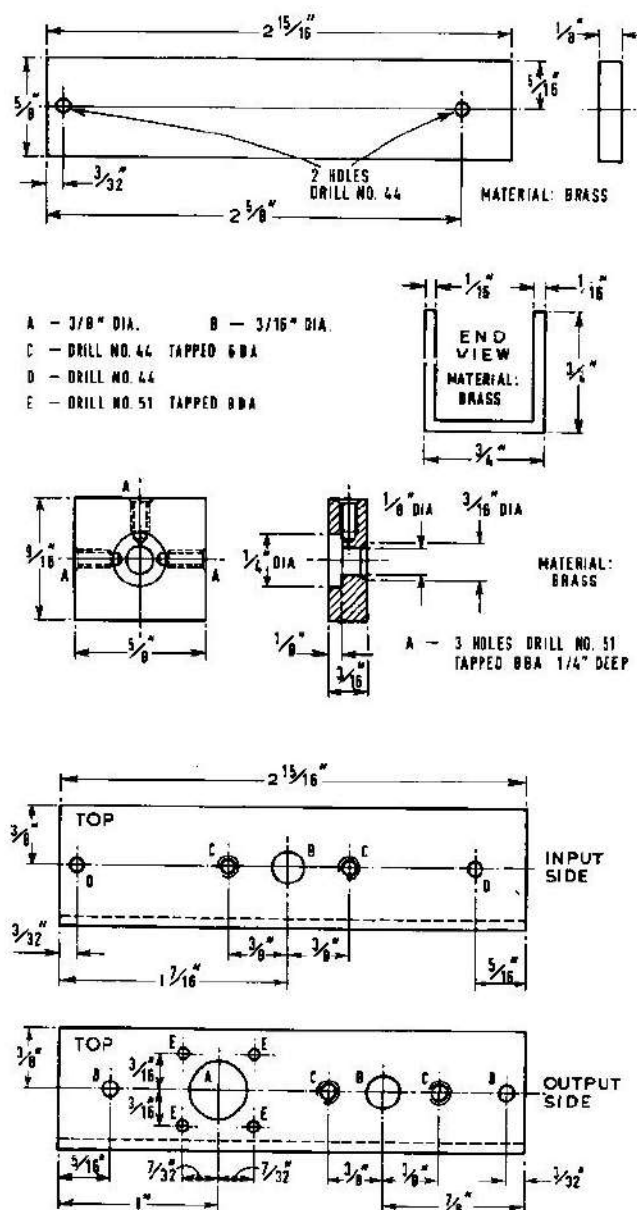
number 51 drill, for 7 BA a number 48, for 6 BA a number 44 and for the ¼" x 60 t..i. model engineer's tap, an ¹⁵⁄₆₄" drill. A number 44 drill can be used for a hole to clear 8 BA.

The centre rod is cut to 5.2cm and will be recessed in its end plate by 1mm. To make the socket for the diode at the end of the rod, first drill with a number 44 for a short distance, making certain that it is dead centre. Then re-drill with a ³⁄₁₆" drill to a depth of ¼", then make three cross-cut slots with a saw and clean up the edges.

The tuning condenser is a little more difficult. Cut a ¼" square section from the strip and after marking out the centre, drill a ¼" hole and tap with the ¹⁄₁₆" x 60 t.p.i. taper tap. The four mounting holes are drilled as shown. Drill also a ¹⁄₃₂" hole a little to one side of the central threaded hole to take the locking clip, made of 22 s.w.g. piano wire. Now tap an inch section of the ³⁄₁₆" rod with the 60 t.p.i. die, saw off the end, clean up with a fine file and carefully make a cross-cut for screwdriver adjustment. Tap the 2 BA washer with the 60 t.p.i. tap, fit on, braze (do not use too much solder) and file flat. With a little care it is quite possible to get a good-as-perfect right-angled fit, which is very necessary. The stator is made of a ¼" flat square of 16 s.w.g. brass brazed on to the end of the centre rod - be careful that its plane is exactly at right angles to that of the two 7 BA holes further down the line.

When all the parts are ready and clean, fit them together and braze the centre rod assembly and channel, but not with the sockets in position. All the other parts screw on. Finish off with a fine file and fit on the sockets and tuning condenser assembly. Having soldered in the feed-through condenser on the end plate fit the tunnel diode into its socket on the rod, clip lead 3 very short, then leads 1 and 2 to about ¼" pinch together and fit into bypass condenser and screw everything together.

Fig.5.51. (a) The tunnel diode amplifier line main body, cut from ¼" x ¼" equal sided brass channel. (b) Centre line end support plate which should be brazed into position in trough. The dimensions and drilling for the capacitor (feed-through) plate are the same, with the exception that the centre hole is ¹⁵⁄₆₄" diam. drilled right through. The 1000 pF feed-through capacitor, preferably the type having a slot in one end to accept the diode leads, is soldered into the hole. This end plate must not be soldered into the trough. (c) Cover.

Power Supply

Tunnel diodes work at very low voltages, and also, as they are majority current devices, the junction cross-sectional area is very very small (about 0.0001" diameter). They will not stand any overload.

An absolute maximum of 10mA, which represents about 4mW, is quoted for the IN2969A, and this really is an absolute maximum to keep the diode intact. Actually, however, with the voltages in use, there is not much danger of passing too much current unless the polarity is accidentally reversed. A supply is therefore needed to give up to 350mV, and it should be apparent from our equations that the diode bias supply must also be of equal or lower impedance than the diode, which is 63 ohms in this case. If it is not, the diode will simply switch itself permanently (or, bi-stably, depending on circuit inductance) into a high or low voltage state (points C or A in Fig. 5.41). To
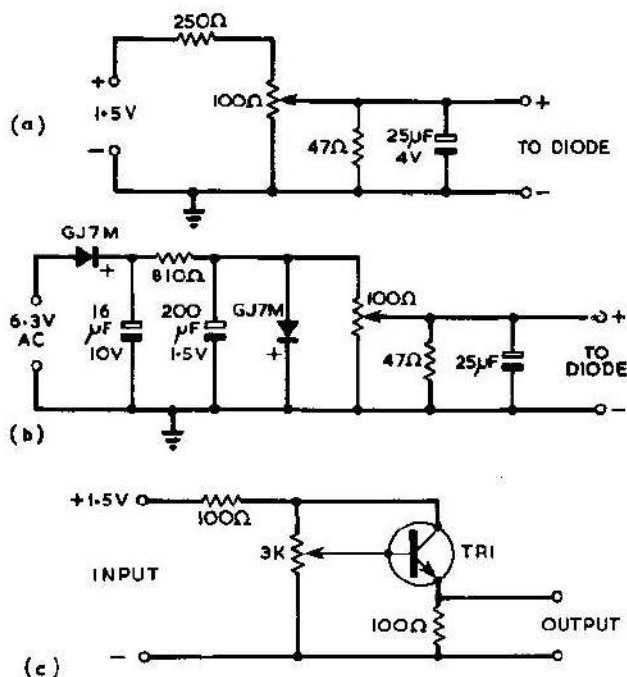
Fig.5.52. (a) A simple battery operated bias supply. (b) An a.c. fed bias supply using a forward biased diode to stabilize the voltage. (c) A series regulated transistor supply which was used in the prototype amplifier TR1 may be any n-p-n transistor having a reasonable gain. A 25 µF electrolytic capacitor must be connected across the output close to the 1000 pF feed-through capacitor, see Fig.5.51.

get stable operation at point B, the bias supply load line must cross the curve at one point only; in other words, be of lower impedance than the diode. It may also be noticed that stray inductance, even when by-passed by 1000 pF (remember $L<Rg/C$), could prove troublesome. For this reason use a 25µF decoupling electrolytic immediately across the 1000 pF bypass. The bias supply can be obtained in many ways: from batteries or other voltage sources via Zener diodes, from forward biased junction diodes or even series regulating n-p-n- transistors. The last method has the slight advantages of lower current drain from the battery and a higher value variable resistor for control - low resistance carbon track potentiometers are rather difficult to obtain. Against this must be set increased cost and complexity. Still, providing that, in this case, the tunnel diode sees a source of no more than 60 ohms, all will be well as far as biasing is concerned. Drain is very low, at the optimum setting the diode will use only about 1mA and maybe the bias supply can afford to lose a few more to get the required low impedance.

Alignment

Alignment is simple, using a sensitive voltmeter to measure the voltage across the diode, increase this to about 200mV. Plug in aerial and receiver and search for a medium strength signal. Tune up the amplifier

whilst gradually reducing the bias. The signal will increase considerably in strength without, in most cases, more increase in background noise. In all probability the unit will oscillate before 130mV (the point of maximum conductance) is reached; this only means that aerial and receiver impedances are not quite what they were thought to be. This does not matter since fairly high gain will be available before this occurs as the diode, on some part of its characteristic, can match anything provided it is of lower conductance than the diode maximum. It will thus oscillate when it over-matches. If the gain goes through a maximum and then drops off, the aerial and receiver impedances are less than 75 ohms. However, even with 50 ohms source and load, the maximum available gain would still be of the order of 13dB.

Noise, Gain and Other Points

The tunnel diode negative resistance shows shot noise as does any resistance, but the noise temperature is of the order of 300 deg.K only, which is much better than either a valve or a transistor, though not as good as a parametric amplifier or a maser. The noise is also frequency dependent to the extent that as one approaches the cut-off frequency, the noise figure gets worse. However, even at a frequency of $1/\sqrt{2}$ of fr, the noise figure is still only 6dB. At medium frequencies (in the 400 MHz range), a noise figure between 3 and 4dB can quite easily be obtained.

For the particular circuit configuration shown, the calculated gain bandwidth product is around 300×10^6 hertz and this appears to agree quite well with results obtained in its use as does the calculated noise figure mentioned above. One can literally choose the value of gain to suit one's own convenience, the limit being set by what bandwidth is required and the difficulty in holding it stable with very high values. 30dB is possibly an effective limit.

As mentioned before, tunnel diodes do not like being overloaded in any way. Up to a point they have a built-in a.g.c. action (this follows from Fig. 5.41), and this non-linearity can produce some most curious spurious responses from out of band local TV transmitters, but it will not cope with a transmitter feeding it with a few watts: both input and output sides must be well shielded from strong r.f. fields, and it is a good idea to place a 75 ohm dummy load across the input when the aerial is removed. Be careful!

Other than this they are very useful and reliable little devices which work with the minimum of fuss provided a few simple precautions are observed (i.e. series inductance and bias impedance, stray r.f. etc). It is quite possible to extend the operation of a parallel amplifier to 1290 MHz with an S-band diode, where its noise figure would still make it worthwhile. Their main disadvantage that they cannot readily be cascaded, is of no consequence to the amateur.
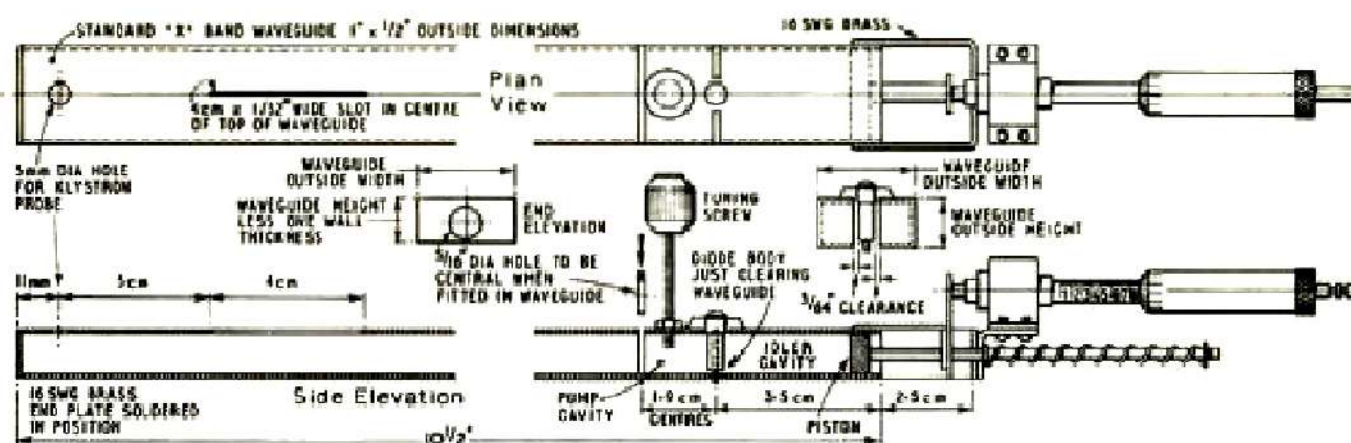
Fig. 5.53. The waveguide assembly, incorporating the attenuator slot, pump cavity and idler cavity with their adjusting mechanism. The waveguide is standard WG 16 (RG-52/U), with internal dimensions of 0.9" x 0.4"., and external measurements of 1" x 0.5".

MULTIBAND PARAMETRIC AMPLIFIER

Even though transistors, and especially FETS, can claim remarkably low noise factors at u.h.f. the parametric amplifier can still hold its own in the face of this competition. The improvement is becoming marginal, and equilibrium has been attained at 144 MHz, but the better u.h.f. noise factor is still worth having, notably to those perfectionists who will not rest until they have developed their equipment to the ultimate.

Circuit

As in a superhet receiver, a parametric amplifier relies on a form of frequency changing for its operation. When changing frequency in the normal way, one can either use a tunable oscillator and a fixed i.f. or a fixed oscillator and variable i.f. In parametric practice, the oscillator is termed the pump oscillator or pump source, and the sum or difference frequency produced by mixing this pump frequency with the signal frequency is known as the idler frequency. Generally the higher the pump oscillator frequency, the lower will be the noise output. A figure of six to seven times the signal frequency is suggested as a minimum, and therefore if operation from 144 to 1300 MHz is required, the pump range will have to extend from 864 to 7800 MHz. This could be modified a little if 144 MHz facilities are dispensed with, as transistors can offer the same performance at this frequency. Alternatively, a fixed pump frequency of say 7800 MHz could be used, implying a varying idler within the range 7944 to 9100 MHz. In practice, a 10 GHz klystron would be used for convenience, producing either 10,144 to 11,300 MHz or 9856 to 8700 MHz depending on whether the sum or difference is chosen.

It should, of course, be appreciated that a half-wave signal circuit will be required for each band; these are most conveniently designed to be detachable from the waveguide.
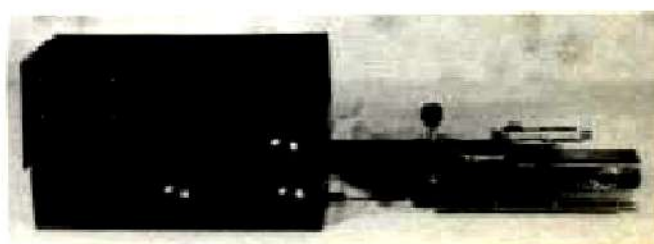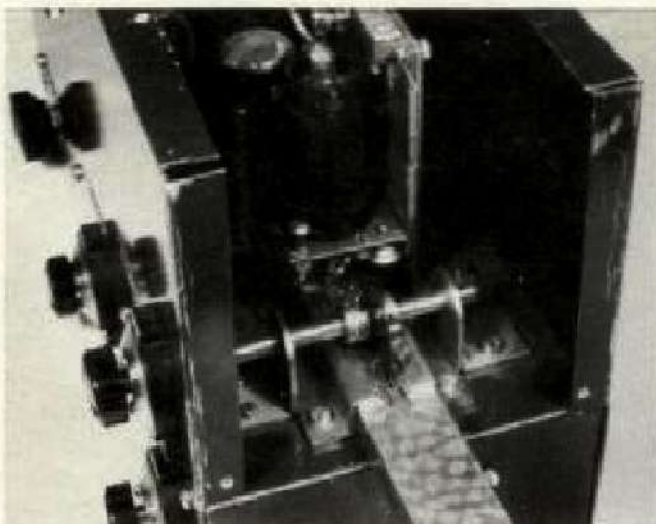

Fig. 5.53.A. Photo of a parametric amplifier.

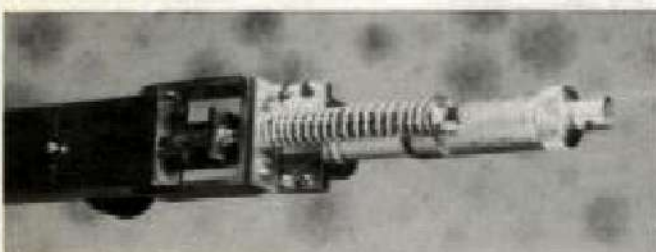## Components

The expensive parts of a parametric amplifier are the Varactor diode and attenuator, a Microwave Associates MA450 and Ferranti ZC20C. Construction of the attenuator, Fig.5.54, is from stiff card soaked in colloidal graphite which after being dried is doped with clear cellulose and mounted on the shaft by cementing between two shaft couplers.
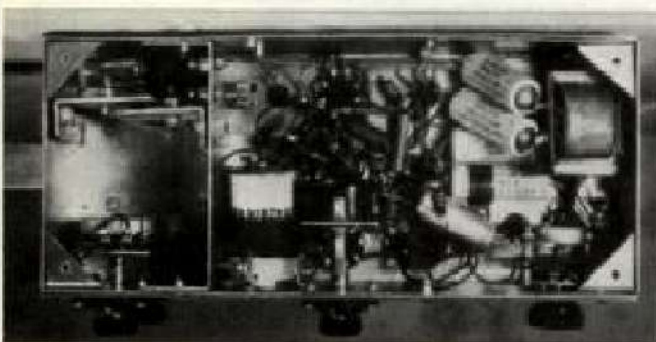
The klystron power supply chassis is a standard 12" x 5" x 2½" with sides to prevent stray draughts affecting klystron stability. The waveguide is mounted solidly on to the power unit chassis as no movement is permissible. The klystron mount was removed from a surplus radar marker unit but an alternative can be made by drilling out pin 4 of an international octal vale-holder to clear the klystron probe, and bending a suitable mount from ½" sheet metal. The metal probe of the klystron should be insulated with p.v.c. sleeving and the probe mounted with the metal parallel to the inside of the waveguide.

The signal circuit, Fig. 5.57, presents no particular problems, but it should be noted that the line carries the diode bias and is not in direct contact with the trough. The contact for the diode tip is made by removing the centre from a Belling-Lee coax socket. The waveguide is positioned and firmly clamped to the top of the appropriate signal tank, where again no mechanical movement is permissible.

Tuning Up

After checking the power unit wiring, Fig.5.58, switch on, allow a 5 minutes warm-up period and adjust the h.t. to 300 V. Allow a further ½ hour and recheck the h.t. Insert a IN21 or similar diode into diode socket and connect a v.t.v.m. (10 V d.c. or less) to the bias line after removing the connector. Withdraw the attenuator from the waveguide. Adjust the pump cavity tuning, klystron external tuning and repeller voltage for maximum reading. Depending on the diode this can be up to 7 volts, but the value is not important. Replace the diode with the Varactor and repeat the tuning procedure, noting that the pump cavity will need retuning and the voltage output will be very low, usually less than 1 volt. Insert the piston fully into the cavity and slowly withdraw it, when at some point the diode voltage should show an increase, indicating oscillation at or near signal frequency. Now is the time to connect a converter, receiver and aerial. It is advisable to use a 6dB pad between the amplifier and converter. A very rough note should be heard; if not, continue unscrewing the piston until this note is heard. Next, increase the amount of attenuation until this oscillation stops. Connect the bias line and tune in a signal, or connect a signal generator, and adjust the trough for resonance. Decrease attenuation until oscillation almost starts, and adjust the idler until it does; then increase the attenuation sufficiently to stop the oscillation (thus reducing pump power), retune the idler and readjust the trough tuning. A systematic tuning procedure should be carried out always with the aim of reducing pump power. Having resonated the trough, the diode bias can be increased, while watching the "S" meter, retuning the trough and reducing pump power. If when the bias reaches 0 V the gain is still rising, reverse the diode (if possible) or battery and carry on with the tuning procedure, not forgetting the input and output loops, repeller voltage and pump cavity adjustments. When the amplifier is producing just below 30dB gain (five S units) it will be found that all further adjustment provides no more gain, as this is the practical limit and any attempt to produce more gain will result in instability. Normally, if the amplifier is operated at between 20 and 25dB gain level, adjustment holds for weeks on end, and retuning is only required for moves of more than 250 kHz. Contrary to popular belief, this takes no longer than 5 seconds, adjustments being confined to trough and idler.
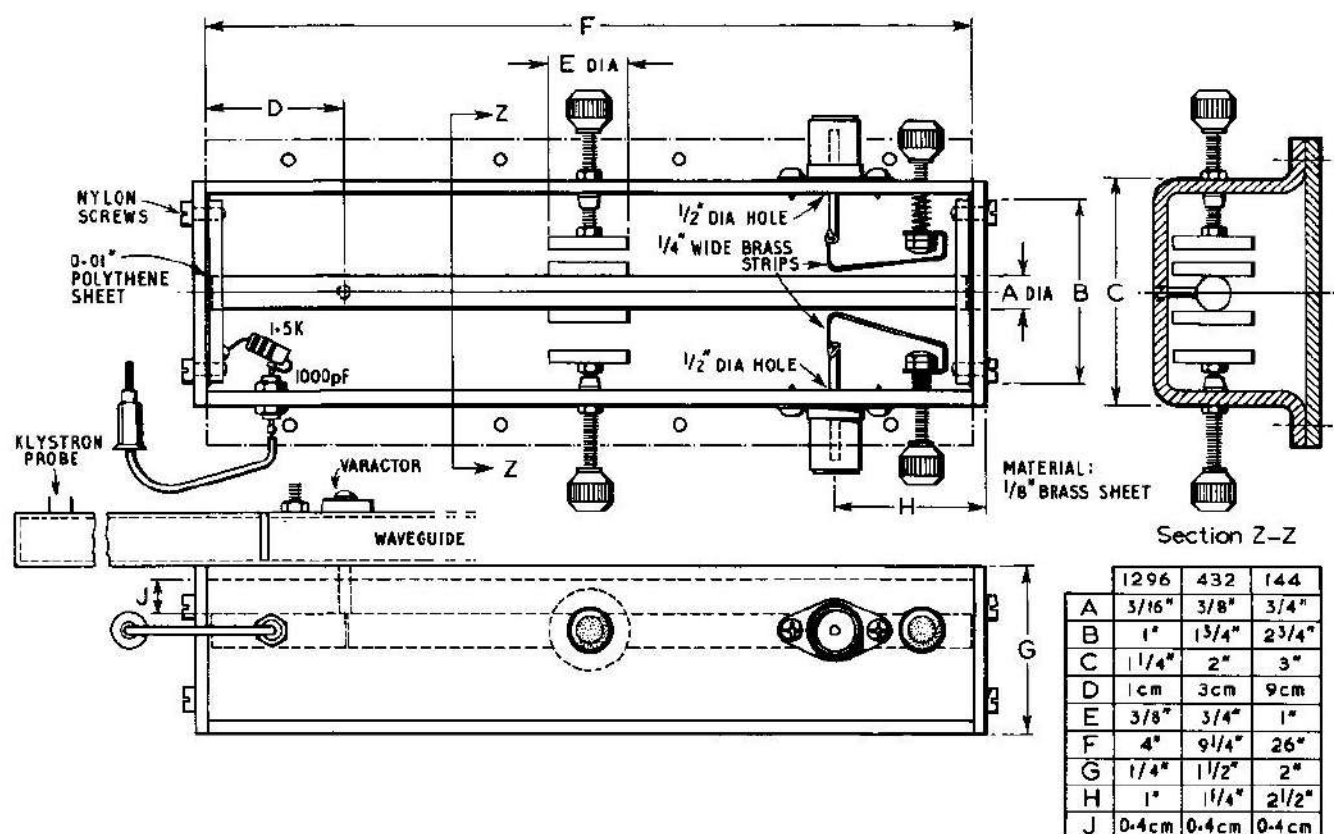

Fig. 5.54. The attenuator and Klystron mount at one end of the power supply chassis.

Fig. 5.55. An end view of the idler components. The varactor tip can be seen protruding through a hole in the waveguide wall.

Fig. 5.56. The underside of the power unit. In the right-hand compartment two GJ5Ms are mounted on the isolated sheet heat shunts, the remaining two diodes being bolted direct to the chassis.

Machining the waveguide will possibly be the most difficult operation. However, by carefully grinding a hacksaw blade to ¼" width and mounting the saw on a woodworker's guide the appropriate cuts can be made. Carefully remove burrs from the inside of the waveguide and taper file the edge of the outside so that solder can run and make a sound joint with the brass inserts which are cut and shaped from ⅛" brass sheet.

	1296	432	144
A	3/16"	3/8"	3/4"
B	1"	1¾"	2¾"
C	1¼"	2"	3"
D	1cm	3cm	9cm
E	3/8"	3/4"	1"
F	4"	9¼"	26"
G	¼"	1½"	2"
H	1"	1¼"	2½"
J	0·4cm	0·4cm	0·4cm

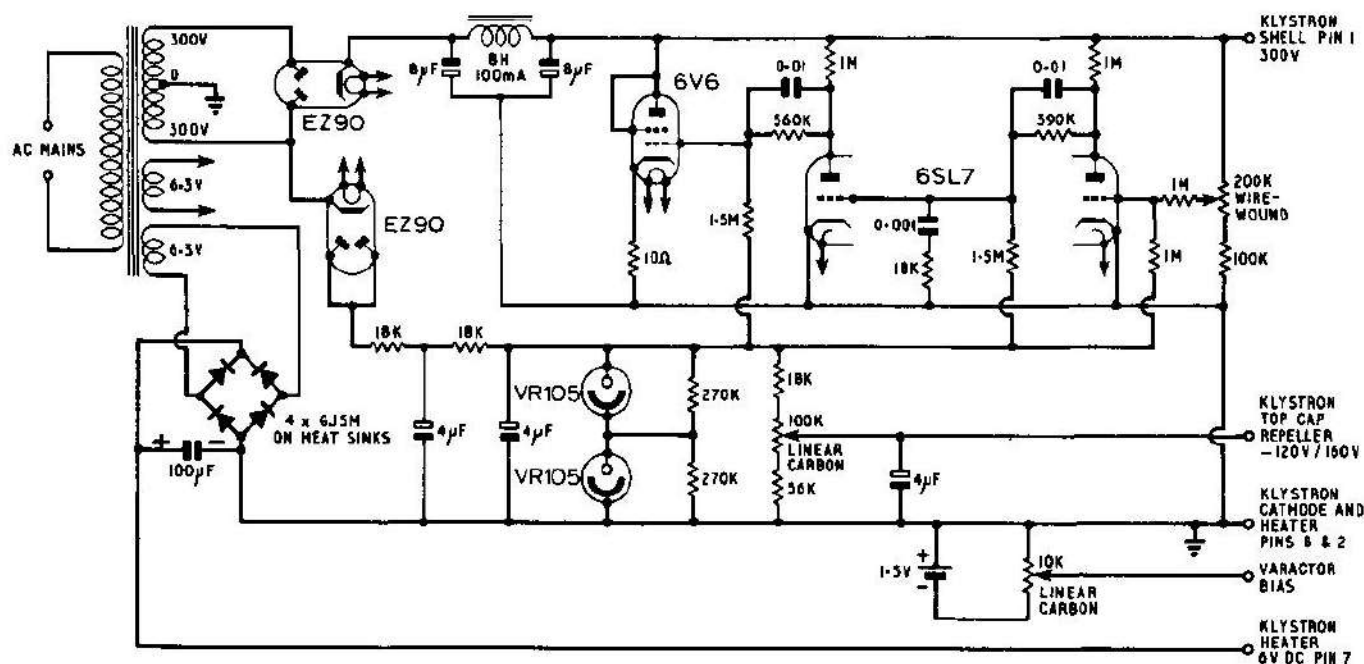*Fig. 5.57. The signal circuit layout, with dimensions for 144 MHz, 432 MHz and 1296 MHz.*

Fig. 5.58. The parametric amplifier pump oscillator power supply suitable for a 723 A/B Klystron. Remember, when constructing the klystron holder and operating the equipment, that the case of the klystron is 300 volts positive with respect to chassis.

5.31

Varactor Diodes

Initially a ZC20C was used and this performed well at 570 MHz, producing pictures over the 140 mile path from London (BBC-2). At this frequency it had a measured gain of over 25dB, but at 432 MHz however, no gain could be obtained from this device, although we know of at least one ZC20C that performs admirably on 70cm. Consequently, an MA 450 was obtained, and this is currently providing over 25dB of stable gain on 432 MHz, although very little gain can be obtained at over 500 MHz. This does not prove conclusively that the ZC20C is poor at higher frequencies but only that the techniques employed by the author would not produce the desired results. Therefore, if possible, a selection of Varactors is highly desirable to obtain the best results.

70 MHz CONVERTER WITH NEUTRALISED R.F. STAGE

This converter has a high performance and can be relied upon to provide as low noise result as can usefully be used in most urban areas.

The r.f. stage is a capacity neutralised 6CW4, inductively coupled to the 12AT7 mixer and with the local oscillator also inductively coupled to the mixer. For this converter, an 8575 kHz FT 243 crystal is used, the oscillator anode circuit being tuned to the fourth harmonic of the crystal frequency. The output of this stage is capacity coupled to the grid of the EF91 double stage, the anode of which is tuned to a frequency of 68.6 MHz, thus producing an i.f. of 1.5 - 2.1 MHz.

The circuit is shown in Fig.5.59 and mechanical details in Fig.5.59A. These diagrams are self-explanatory and little more need be said. Coil details are given in the table.

L1,   8 turns 18 s.w.g. enamel wound on ¼ in. mandrel, length 1 in. tapped 3 turns from earth end, air spaced.

L2,   13 turns 18 s.w.g. enamel wound on ¼ in. mandrel, length 1¼ in., tapped 5½ turns from CN, air spaced.

L3,   3½ turns 18 s.w.g. enamel wound on ¼ in. mandrel, length ⅞ in., air spaced.

L4,   2 turns 18 s.w.g. enamel wound on ¼ in. mandrel and placed between L2 and L3.
      Note: L2, L3 and L4 are all on the same axis.

L5,   Maxi-Q i.f. transformer type IFT, type IFT 11/1.6. Both windings in series and capacitors removed.

L6,   2 turns 22 s.w.g. p.v.c. covered tinned copper wire wound at h.t. end of L7.

L7,   10 turns 26 s.w.g. enamel wound on a ¼ in. by 1 in. former, slug tuned (Aladdin type with can).

L8,   10 turns 26 s.w.g. enamel wound on a ¼ in. by 1in. former slug tuned (Aladdin type with can).

RFC1,   2.5 mH r.f. choke.

RFC2,   51in. 18 s.w.g. enamel wound on ¼ in. mandrel, close spaced.

RFC3,   1 mH r.f. choke.

To align the converter, disconnect the h.t. supply from V1, remove the cans from L7 and L8 and first adjust L8 for maximum indication on the r.f. checking meter (Fig. 5.60). Then adjust L7 in a similar manner.
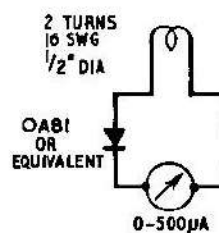

Fig.5.60. An r.f. checking device.

The tuning range of each coil is such that only the required harmonic should be selected but it is wise to check the actual frequencies with an absorption wavemeter or g.d.o. if one is available. The cans should then be replaced on L7 and L8, and the two cores readjusted for maximum r.f. output from the coupling link L4. The output should then be connected to the communications receiver, tuned to 1.8 MHz L5 will be approximately correct. Adjust C3 for maximum hiss; two positions will be found, the one with the smaller capacity being the correct one. A strong signal is then required (from a local transmitter or signal generator) fed to the aerial socket. Adjust C1 and C2 followed by C3 and L5 for maximum output then adjust Cn, the neutralizing capacitor, with an insulated screwdriver, for minimum output. (In practice this is usually found to be near the minimum capacity of Cn) H.t. may then be reconnected to V1 and initial adjustments are complete. If there is a tendency for oscillation, Cn is incorrectly adjusted; the minimum position referred to is very sharp and to a certain extent adjustment is inter-dependent with C2 and C1. No difficulty should be experienced, however, in obtaining the correct adjustment. For final adjustments a noise generator is desirable but if this is not available, C1,C2,C3 and L5 (also L7 and L8) should be adjusted for maximum output on a local signal. The capacity of C1 should then be increased slightly so that the tuned circuit is detuned towards 70 MHz and the output just drops. This is near to the optimum position for the best signal-to-noise ratio. In practice a noise factor of better than 2.5dB should be obtained.
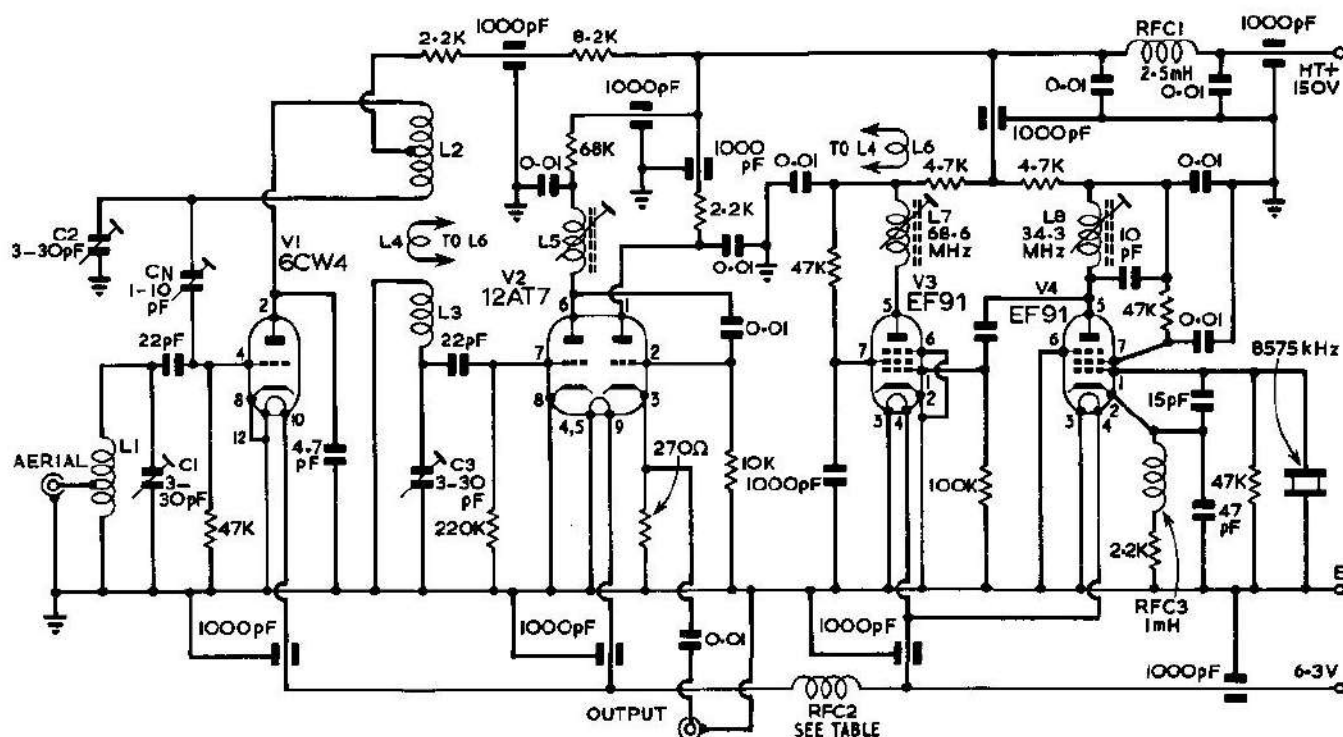
Fig. 5.59. Circuit diagram of the 4 metre converter. Coil sizes are given in the table on page 5.32.

NOTE....... HOLES 'Z' ARE 2BA CLEARANCE AND ARE FOR SUPPORT PILLARS FOR COILS OR PHILLIPS TRIMMERS
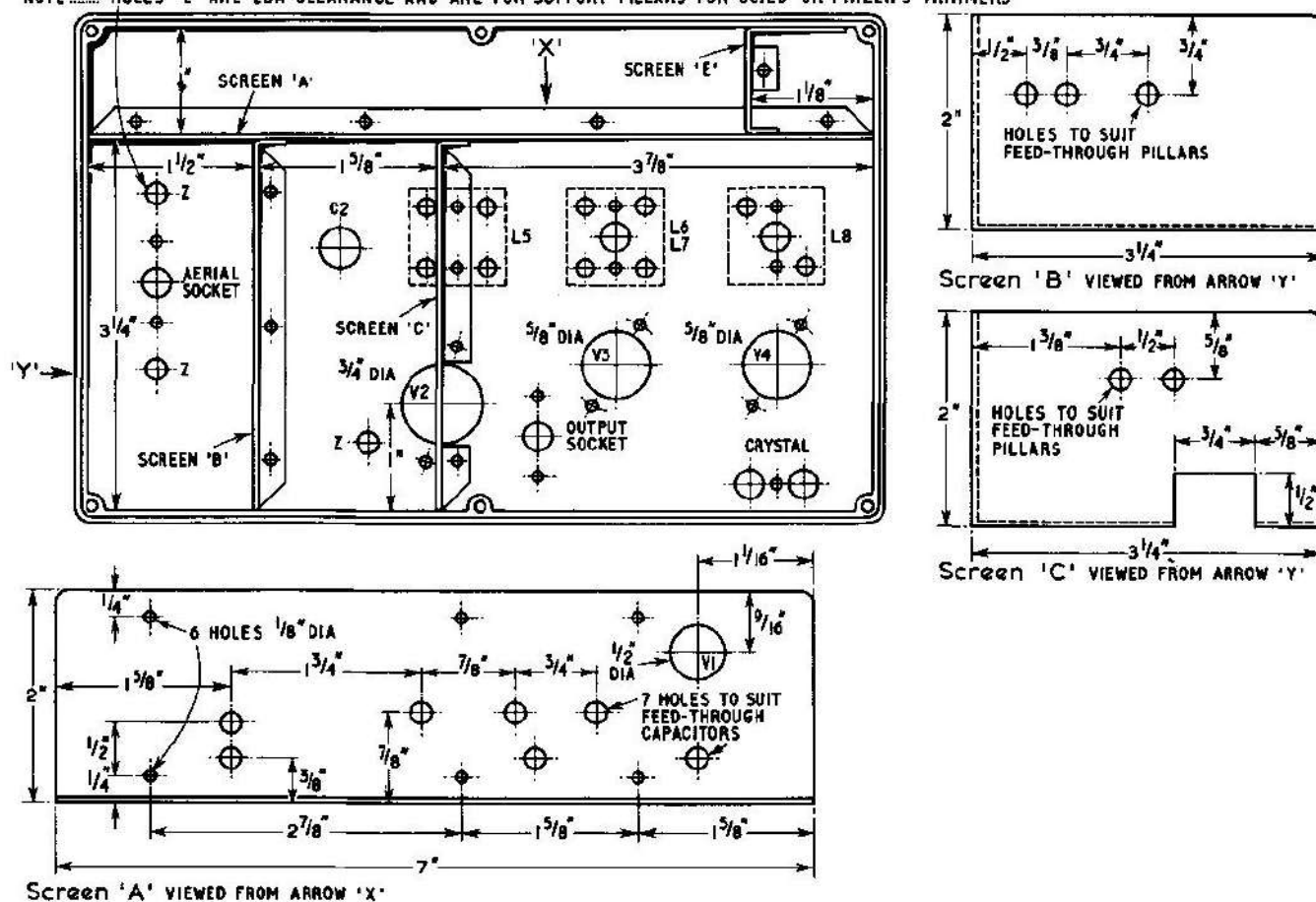
Fig. 5.59A. Mechanical layout of the 4 metre converter.

144 MHz CONVERTER WITH NEUTRALISED R.F. STAGE

In this converter, the circuit follows the same general arrangement used for the 70 MHz converter already described. The local oscillator, in this case is followed by two multiplier stages to reach the required frequency. For the 144 MHz converter, a 7100 kHz FT243 crystal is used but the oscillator anode circuit is tuned to the fifth harmonic of the crystal. The subsequent 12AT7 double triode is connected in cascade and doubling in each half produces a final frequency of 142 MHz which results in an i.f. of 2-4 MHz.

The alignment of this converter is similar to that already given for the 70 MHz converter, except that the communications receiver is set to 3 MHz corresponding to a signal frequency of 145 MHz. (If GB3VHF on 144.5 MHz is used as an alignment signal, the receiver should be set to 2.5 MHz). First, power is applied and h.t. disconnected from V1. The can is removed from L9 and the slug is adjusted for maximum r.f. indication followed by C5, r.f. indication being observed at L8 and L7 respectively and then C4. Again, the tuning range of each tuned circuit is such that only the required harmonic should be selected but it is wise to check the actual frequencies with an absorption wavemeter or g.d.o. if available. L5 which is very flat in its tuning, should be centred on 3 MHz. C1, C2 and C3 are then adjusted for maximum output on a strong local signal. Cn is adjusted for minimum output with an insulated screwdriver, the correct setting being very critical. H.t. should be restored to V1 and the converter is then ready for use. Final adjustment should again, if possible, be carried out with a noise generator. The aim should be to obtain a noise factor of better than 3dB.

L1,   4 turns 18 s.w.g. enamel wound on ¼ in. mandrel, length ¼ in., tapped 1½ turns from earthy end, air spaced.

L2,   8 turns 16 s.w.g. enamel wound on ⁷⁄₁₆ in. mandrel, length 1 in., tapped 3½ turns from CN, air spaced.

L3,   4 turns 16 s.w.g. enamel wound on ⁷⁄₁₆ in. mandrel, length ⅜ in., air spaced.

L4,   1 turn 16 s.w.g. enamel wound on ⁷⁄₁₆ in. mandrel, inter-wound at earthy end of L3, air spaced.

L5,   32 s.w.g. enamel wire wound on a ¼ in. by 2½ in. former (Aladdin type with can), 2 layers 1½ in. long and 1 layer ¾ in. long, winding from bottom of coil former up for 1½ in. and back to bottom, then up for ¾ in. Tissue paper inter-leaving is used and the winding secured with polystyrene cement. Tuning by two slugs.

L6,   as for L4 but mounted adjacent to the h.t. end of L7, air spaced.

L7,   3½ turns 16 s.w.g. enamel wound on ⁷⁄₁₆ in. mandrel, length ¾ in., air spaced.

L8,   7 turns 16 s.w.g. enamel wound on ⁷⁄₁₆ in. mandrel, length 1 in., air spaced.

L9,   10 turns 26 s.w.g. enamel wound on a ¼ in. by 1 in. former slug tuned (Aladdin type with can).

RFC1,   2.5 mH r.f. choke.

RFC2,   26 in. 18 s.w.g. enamel close wound on ¼ in. mandrel, self-supporting, air spaced.

RFC3,   1 mH r.f. choke.

NUVISTOR CONVERTER FOR 70 MHz

The r.f. stage employs a 6CW4 Nuvistor in a conventional inductively neutralized configuration which arrangement, when correctly neutralized, accounts for the excellent noise factor of the converter. The anode circuit of this valve forms one of the in-line circuits that may be seen from the arrangement of L3, L4 and L5 on the drawing in Fig.5.63B.

The mixer employs a 7587 Nuvistor in a straightforward arrangement, a small fixed bias being developed across the cathode resistor. The crystal controlled injection oscillator uses a 7586 Nuvistor in a somewhat unusual circuit configuration. Feedback to sustain oscillation is provided by a circuit tuned to the fundamental frequency fitted in the cathode circuit of the valve. Oscillation is independent of the anode circuit tuning thus the arrangement displays a high order of stability. The crystal is 20.09 MHz and produces an i.f. in the range 9.65 - 10.325 MHz for the band 70.7 - 70.025 MHz.

Construction

The converter can be built into a 3½" x 4½" diecast box. The actual chassis employs a perforated chassis plate cut down to a size which fits inside the lid of the diecast box. In the final assembly, this chassis is mounted to the lid by 6BA bolts and stand-off pillars. This form of sandwich construction results in all the h.t. and l.t. wiring being contained within a screened area, and allows this wiring to be run to the required point by the shortest possible route without it passing through any of the signal circuit compartments. Supplies to the various compartments are taken through p.f.t.e. feedthroughs pushed through the nearest convenient hole in the chassis plate.

Each of the circuits employed in the converter, that is the r.f. stage, mixer stage, and local oscillator, are contained within screened compartments constructed from 0.025" sheet copper. The screens are bolted to the chassis plate and then soldered together.
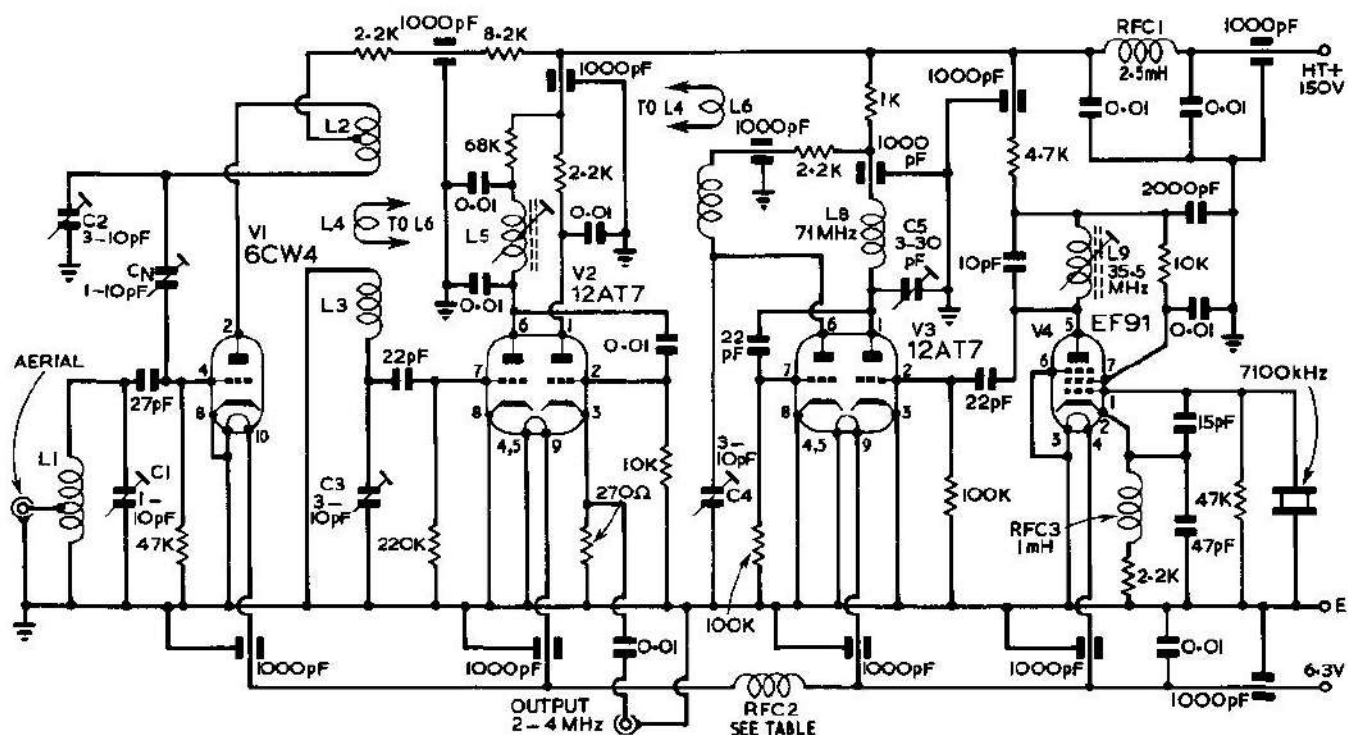
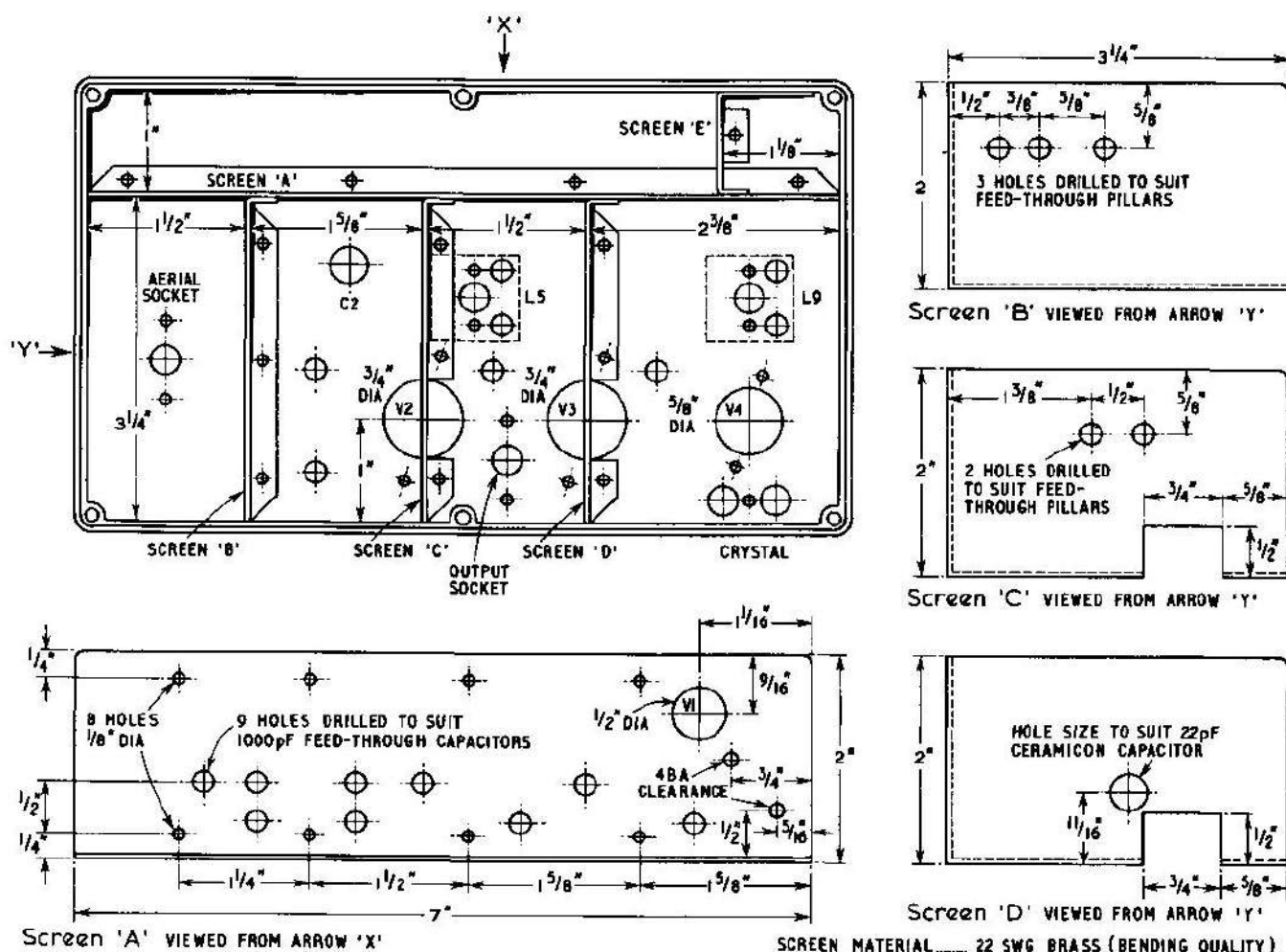Fig. 5.61. Complete circuit of a neutralised 6CW4 converter for 144 MHz.

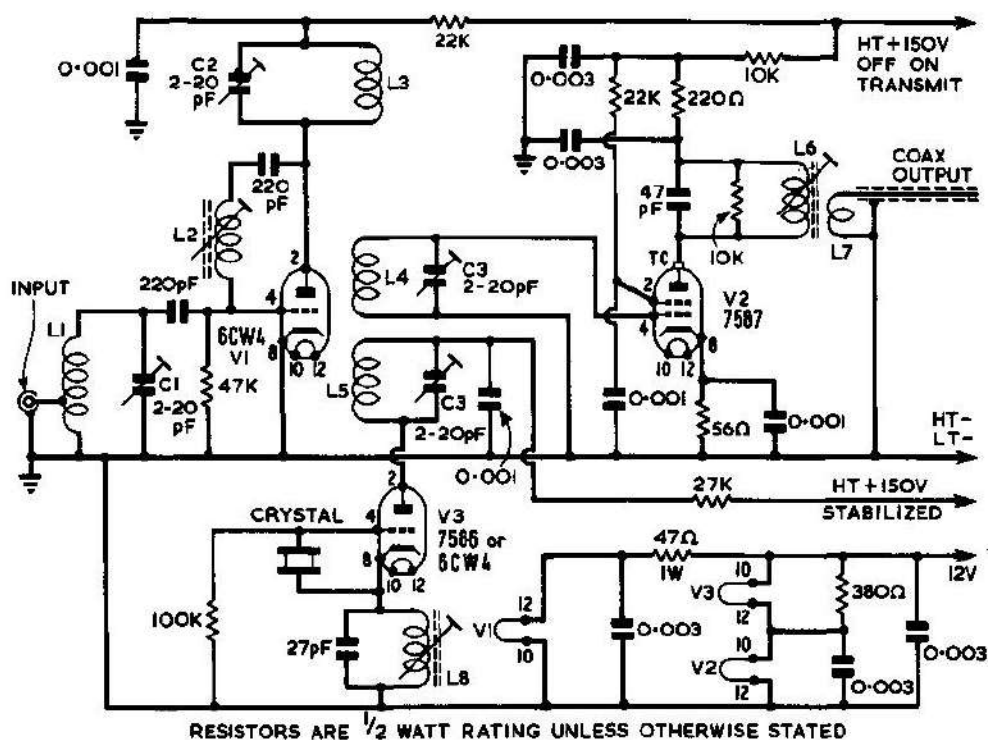Fig. 5.62. Showing layout of the principal components.

Fig. 5.63A. *Circuit of the Nuvistor converter, C1, 2, 3 and 4 are 2–20 pF variable capacitors, and all 0.001 and 0.003 uF capacitors are disc ceramics. Resistors are ½ watt rating except where indicated. L1, 8 turns, tapped at 2½ turns from earthy end, 22 s. w.g., spaced wire diam., ½" diam. former, ½" leads; L2, 27 turns, 28 s. w.g., close wound, 0.3" diam. slug tuned former; L3, 9 turns, 22 s. w.g. spaced wire diam. ½" diam. former, 1" leads; L4, 7 turns, 22 s. w.g., spaced wire diam. ½" diam. former 1" leads; L5, 6 turns, 22 s. w.g., spaced wire diam., ½" diam. former, 1" leads, L6, 34 turns, 28 s. w.g., close wound, 0.3" diam. slug-tuned former; L7, 4 turns, small diam. p.v.c. covered wire, wound over earthy end of L6; L8, 12 turns, 28 s. w.g. close wound, 0.3" diam. slug-tuned former.*
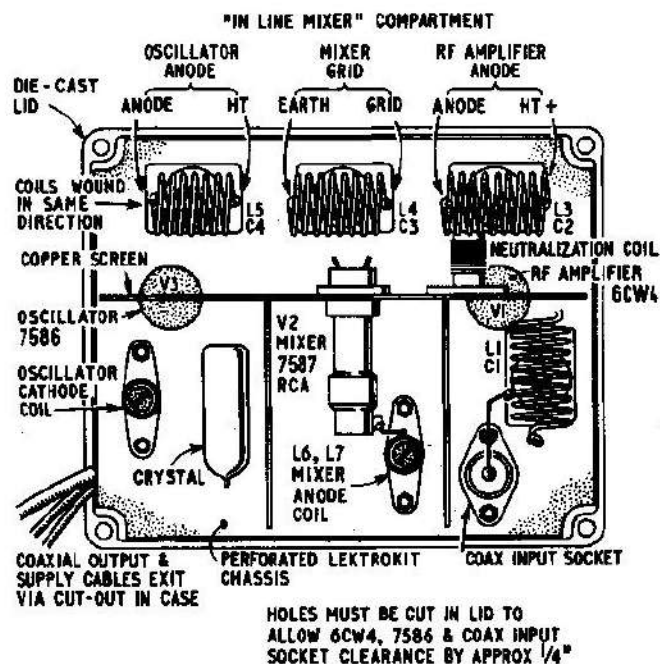
Fig. 5.63B. *Mechanical layout of the 70 MHz Nuvistor converter.*

As the setting up of the converter is carried out with it fully screened inside its box, holes have to be drilled to enable C1, C2, C3, C4, L2, L6 and L8 to be adjusted from outside.

With regard to the neutralizing of the r.f. stage, it should be noted that the neutralizing coil is situated in the mixer compartment, and that it is positioned at right angles to the other coils.

To a large extent, the performance of the converter depends on the mixer circuits. As will be seen from the illustration, the r.f. anode, mixer grid, and oscillator anode coils are all wound in the same sense. It is essential that they are connected as shown. The spacing between the coils is as follows: r.f. anode to mixer grid ¼", mixer grid to oscillator anode ¼ – ½". These spacings should not be treated as absolute, but are a starting figure prior to the adjustments described under the heading *Setting-up*. The coils may be constructed as self-supporting, but the use of a ½" diameter paxolin tube 3½" long inserted into the coils during the setting up is a help in ensuring that they are all on the same axis.

Setting Up and Alignment

The simplest method of adjustment is to align the circuits approximately to the correct frequency by the

use of a g.d.o. Care should be taken to ensure that the g.d.o. is only coupled to the circuit being adjusted, and all valves in position so that their stray capacities are taken into account, with all tuning capacitors set for maximum, initially.

With the converter out of its case, tune the cathode of the oscillator circuit by adjustment of the core of L8 to either the crystal frequency, or if an overtone crystal is being employed, to the desired overtone, and then tune the oscillator anode circuit by means of C4 to the final injection frequency.

Tune the r.f. grid circuit, C1, the r.f. anode circuit, C2, and the mixer grid, C3 to 70 MHz. Unsolder one end of the 10K ohms damping resistor across the tuned circuit in the anode of the mixer, and resonate this circuit to the centre of the i.f. passband. Reconnect the 10K ohms resistor. Set the core of the neutralising coil, L2, half way into the winding.

Place a short across L5. Connect a high resistance volt-meter 20,000 ohms/volt or better, across the 27K ohms resistor in the h.t. supply to the oscillator anode. Connect the power supplies. Tune the core of L8 while watching the voltmeter. It will be found that the voltage will fall slowly to a minimum, and then rise rapidly to a maximum. Set the core of L8 so that the voltage is some 20 per cent above the minimum

voltage on the "slow" tuning side of the core. Disconnect the power, remove the short from L5, seal L8 and fit the converter into its box.

Final neutralisation of the r.f. amplifier is best accomplished on a very strong signal. Care has to be taken to ensure that all the signal to the converter is indeed coming from the aerial, and that none of it is being picked up by the converter itself. For this reason, a g.d.o. is not suitable for this operation.

After having found, or arranged, such a strong signal, and tuned it in, the h.t. should be disconnected from the r.f. stage of the converter. Now adjust the core of L2 to the point where this signal is at its minimum. This adjustment is not very critical, but is essential for optimum performance.

Once the above adjustments have been completed, the r.f. and mixer stages can be peaked up on an identified weak 4m signal. In these adjustments, C1 is tuned for maximum output, and the spacing between L3 and L4 varied to give best signal-to-noise ratio. With the adjustments to L3 and L4, if the spacing is made too great, the circuits will become quite sharply tuned, and the gain will drop. Naturally when L3/L4 are adjusted, C3 and C4 will require peaking, but the object is to still retain the broad character of the resonant point.
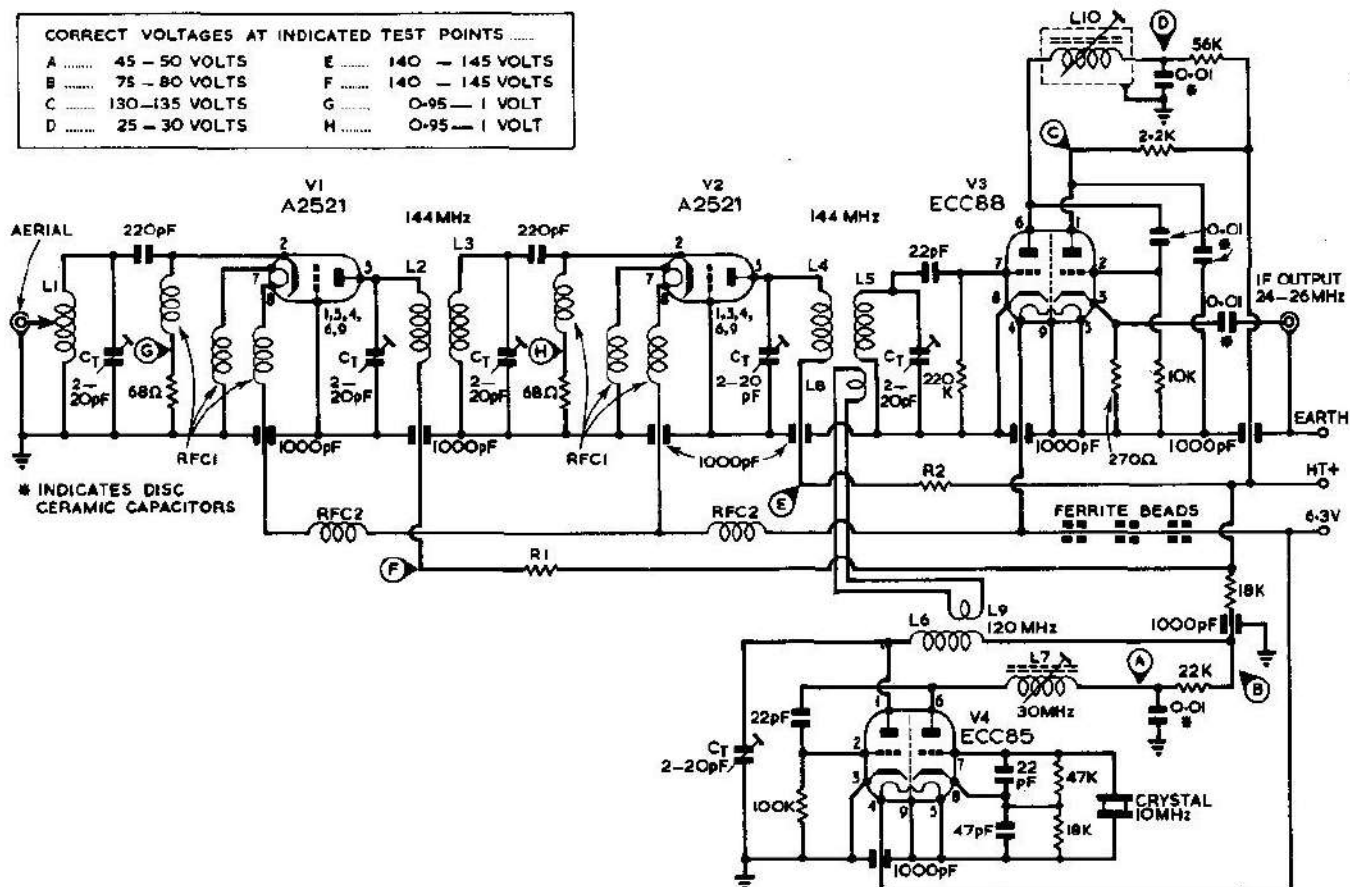

Fig.5.63. Circuit of converter with two grounded grid r.f. amplifiers. CT, 2—20 pF concentric trimmers; R1, 2, selected to pass 15mA; RFC1, 19", 26 s.w.g. wire on ¼" former; RFC2, 19" 20 s.w.g. p.v.c. wire on ¼" former; V1, 2, A2521; V3, ECC88; V4, ECC85; Crystal 10 MHz.

A CONVERTER WITH TWO GROUNDED GRID R.F. AMPLIFIERS FOR 144 MHz

In a grounded grid amplifier, although the noise performance and gain are satisfactory, there are some circumstances where the selectivity provided by one amplifier is inadequate, and therefore two stages are necessary to achieve a desirable level.

The gain with two grounded grid stages will be greater than with two similar valves connected in the cascode arrangement, the signal to noise ratio will be no worse, and in addition the likelihood of cross modulation in the r.f. stages will be reduced.

The circuit of this converter is shown in Fig. 5.63 and the coil details given in the table below.

Coil	Turns	Wire	Spacing	I.D.	
L1	4	18e	wire dia.	7/16	adjust tap for optimum noise factor.
L2	4½	18e	wire dia.	7/16	
L3	4	18e	wire dia.	7/16	spaced 1/16 to 1/8" apart.
L4	4½	18e	wire dia.	7/16	
L5	4	18e	wire dia.	7/16	
L6	5	18e	wire dia.	7/16	
L7	19	18e	cls. snd.	19/32	Neosid F804 with dust core.
L8	3	20pvc	cls. wnd	¼	⅛ to ¼ in. from earthy end of L5
L9	2	20 pvc	cls. wnd	½	wound over last turn of L6
L10	30	26e	cls. wnd	7.5 mm	Neosid former with core and can.

As an aid to fault tracing various test points are marked on the circuit diagram, and the correct voltages to be found at these points are given below:

A	45 - 50 volts		E	140 - 145 volts	
B	75 - 80 volts		F	140 - 145 volts	
C	130 - 135 volts		G	0.95 - 1 volt	
D	25 - 30 volts		G	0.95 - 1 volt	

Readings should be taken with a very high resistance voltmeter, or preferably a valve voltmeter.

It is, of course, important that the aerial is properly matched to achieve the best noise factor and this should be carried out preferably using a noise generator. If the succeeding stages are properly tuned, freedom from breakthrough can be expected.

In this converter the mixer/cathode follower uses a high slope low noise, double triode which can be more easily driven from a small local oscillator signal, so that the noise which is contributed by the local oscillator/multiplier is kept to a minimum.

Construction

The construction of this converter is detailed in the component layout drawings whilst the interstage
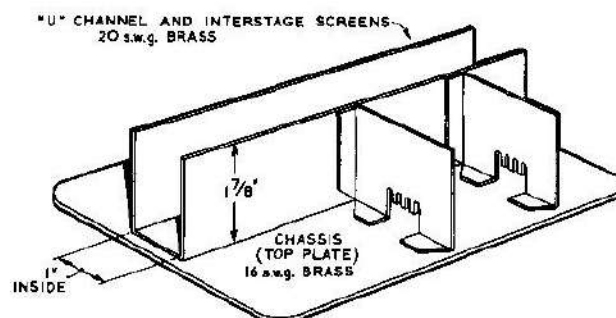

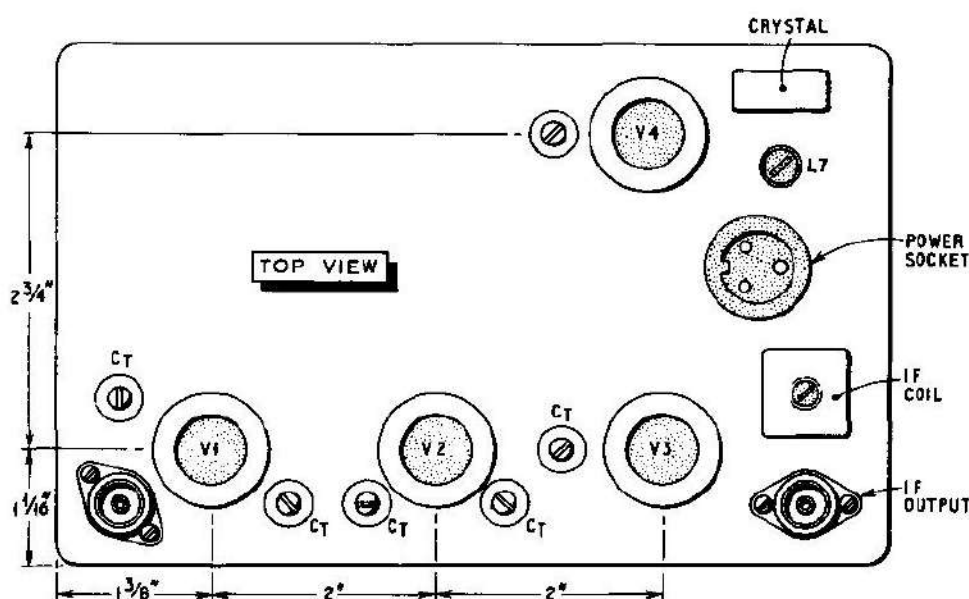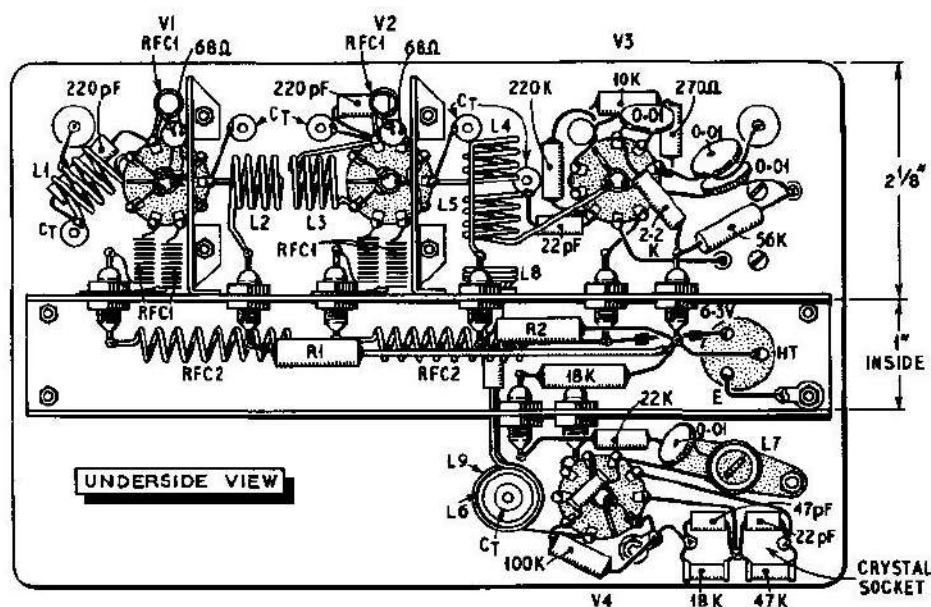Fig. 5.64. Detail of the chassis and screening arrangement.

Fig. 5.65. Top view of chassis plate, showing layout.

Fig. 5.66. Underside view of chassis, illustrating layout and major wiring.

screening is shown in Fig. 5.64. As can be seen, the mechanical arrangement has been designed so that the power supplies are isolated from the r.f. circuits, and the input circuits of each r.f. amplifier are screened from one another. The local oscillator crystal is shown as 10 MHz. It can, of course, be 30 MHz or any other frequency to suit a different i.f. as required, although for freedom from breakthrough from strong low frequency signals the specified i.f. of 24-26 MHz is recommended. If a different i.f. is used then the inductor L10 will have to be altered to suit the new frequency chosen.

Although A2521/6CR4 valves are used in this design there is no reason why other types, such as EC88, 6AM4, EC8010 which are intended for grounded grid (multiple grid connections) use should not be used. The noise factor will, of course, depend on the actual valve type employed.

With a change of valve the layout may have to be changed also, as, for example, if EC88 valves are chosen then the whole unit would have to be reversed (a mirror image) to preserve the short leads and proper component layout.
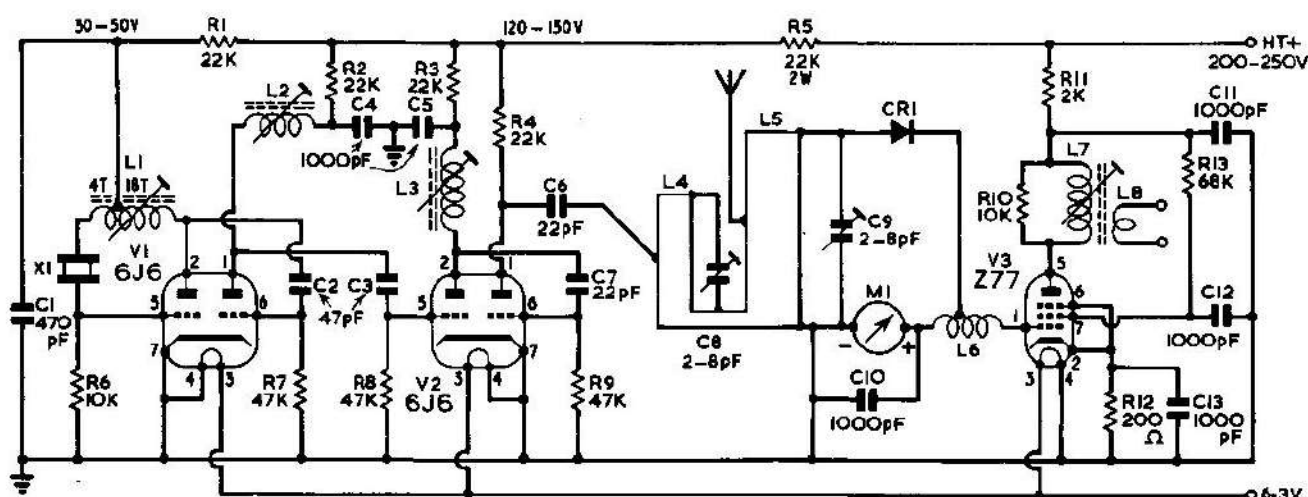
Fig. 5.67. Circuit of simple 432 MHz converter. C1, 4, 5, 6, 10, 11, 12, 13 are midget mica. C2, 3, 7 ceramicons. C8, 9, Philips trimmers 3–8 pF CR1, silicon diode, L1, 22 turns 22 s.w.g. close wound tapped at 4 turns, L2, 6 turns 20 s.w.g. close wound. L3, 2 turns 20 s.w.g. close wound. L4, L5, see Fig. 5.68. L6, 22 turns 32 s.w.g. close wound, centre tapped. L7, 37 turns, 32 s.w.g. close wound. L8, 4 turns 32 s.w.g. wound at earthy end of L7. All formers are ⅜" dia. Aladdin, with h.f. cores. Crystal is 35 MHz overtone.

70cm CONVERTER

This simple and easily constructed converter will give a performance which will enable the newcomer to get the feel of the band. It is constructed largely of the type of component usually available for the h.f. bands.
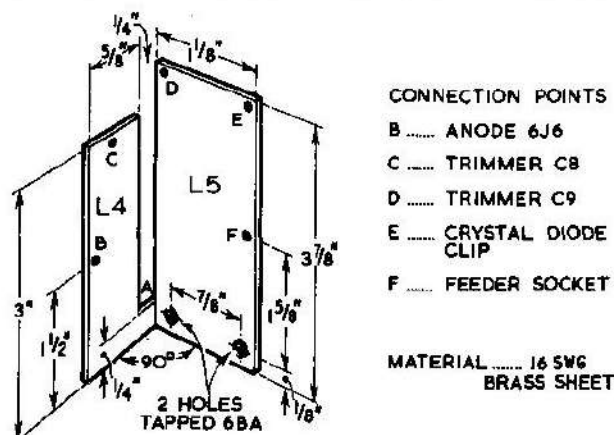

CONNECTION POINTS

B ...... ANODE 6J6

C ...... TRIMMER C8

D ...... TRIMMER C9

E ...... CRYSTAL DIODE CLIP

F ...... FEEDER SOCKET

MATERIAL ...... 16 SWG BRASS SHEET

Fig. 5.68. Dimensions of strip line. L4 and L5 mounted ¾" from chassis.

The valves used are freely available at low cost, the only u.h.f. components are in the slab type inductors used for L4 and L5. These are illustrated in the drawing in Fig. 5.68. The diode (CR1) should be a low noise silicon type such as the CV102, CV103 or equivalent. Although the crystal oscillator and multiplier stages are given as 6J6 valves, any of the double triodes such as the 12AT7 and ECC85 may be substituted but some minor adjustment to the various inductors may be necessary, notably L3 and tapping point on L4. The crystal oscillator chain for an i.f. of 12-14 MHz (L7) may be based on a crystal frequency of either 7.777 or 35 MHz.
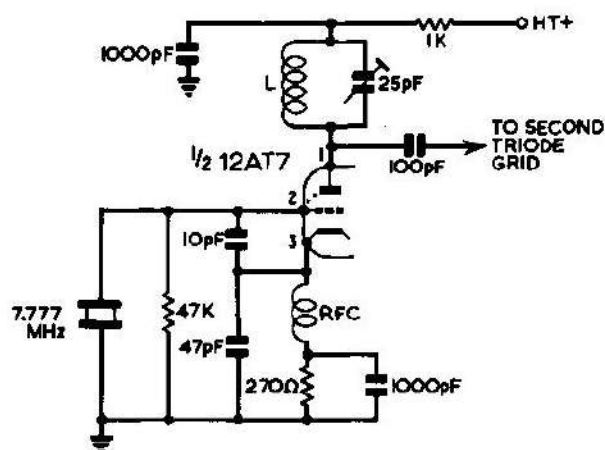
Fig. 5.69. An alternative crystal oscillator using the Colpitts circuit.

Alternative crystal oscillator circuits may be used by those who prefer the Colpitts or Butler oscillators, to that of the Squier shown in the diagram of Fig. 5.67.

In the case of the Colpitts circuit, Fig. 5.69 using a 7.777 MHz crystal, two 12AT7 valves would be needed. The anode circuit of the first triode would be tuned to 23 MHz (actually 23.331 MHz), that of the second to three times this frequency which would be almost 70 MHz. The two following anode circuits would then be tuned to 140 MHz and 420 MHz respectively.
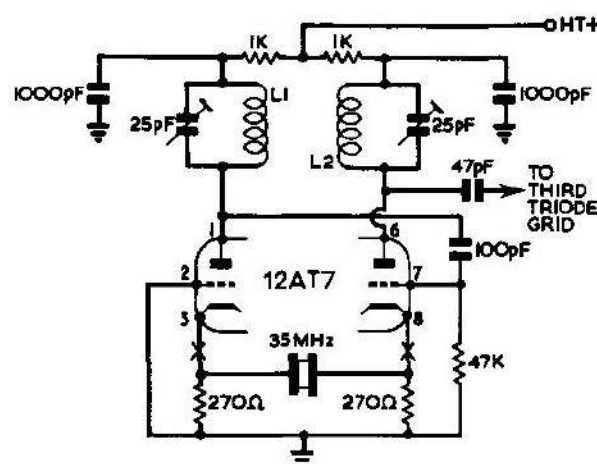

Fig. 5.70. A further alternative oscillator using the Butler circuit and a 35 MHz crystal.

A Butler oscillator circuit using a 35 MHz crystal has a considerable advantage over the others because it is easier to avoid tuning the multiplier stages to the wrong frequency. Two 12AT7 valves are used in the circuit, the first section of which is shown in Fig. 5.70. In this case the anode circuit of the first stage is tuned to the crystal frequency, while that of the second is tuned to 70 MHz. The succeeding stages, like those following the Colpitts oscillator, are tuned to 140 MHz and 420 MHz.

The use of a conventional dip oscillator will enable correct adjustment up to about 200 MHz, but the final multiplier will need to be aligned to its proper frequency using an absorption wavemeter or the dip oscillator described in Chapter 10.

A SIMPLE 70cm CONVERTER USING BI-POLAR TRANSISTORS

In this converter, which is of straightforward design, only freely available and cheap transistors are used. The crystal operates at 35 MHz and is followed by three frequency multiplier stages, the first triples to 105 MHz and the other two each double to reach the final frequency of 420 MHz. In the mixer stage a GM290 transistor is used, and is capacitively coupled to the pre-amplifier and the local oscillator. The output from the mixer is fed directly into an i.f. amplifier at 12–14 MHz. The pre-amplifier, another GM290, is operated in grounded base mode with a
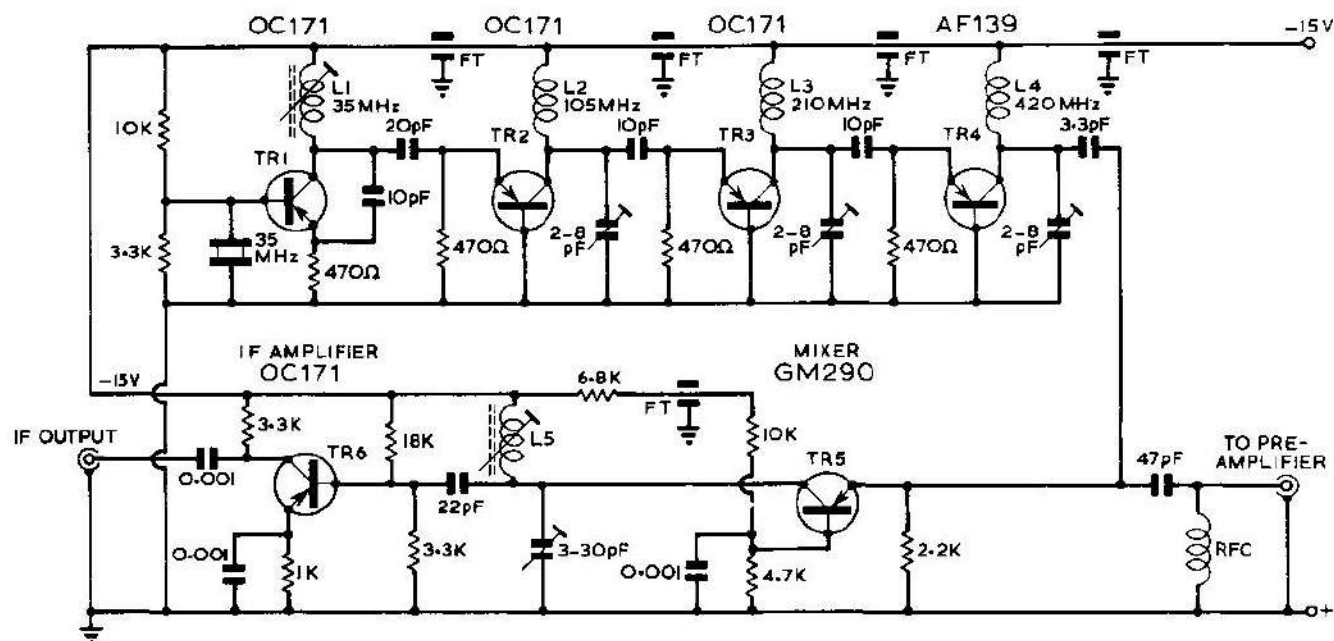
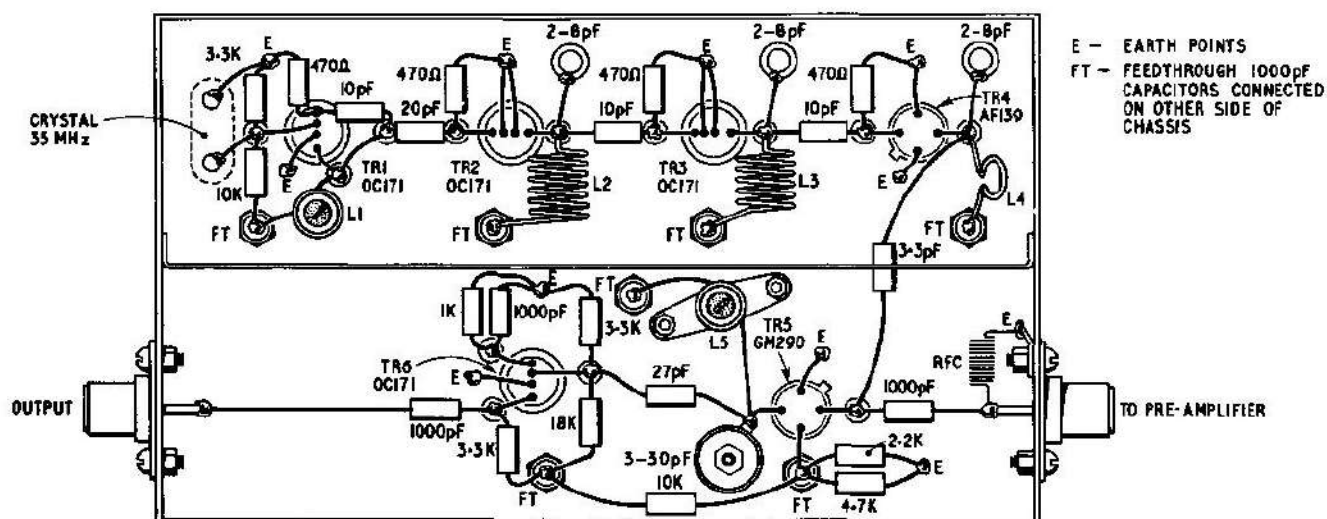Fig. 5.71. Circuit of mixer and oscillator chain for 70cm converter.

Fig. 5.72. Layout of circuit shown in Fig. 5.71. + are earth points. Ft are feed-through capacitors connected together on other side of panel. TR1, 2, 3, 6 are OC171. TR4 is AF139. TR6 is GM290.
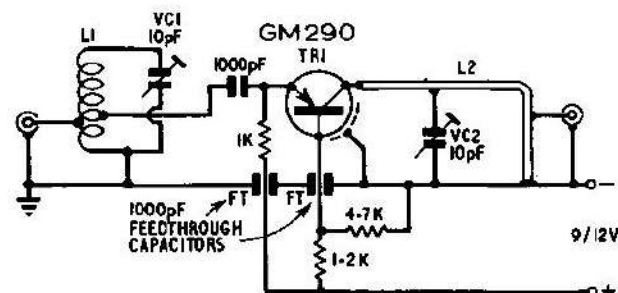

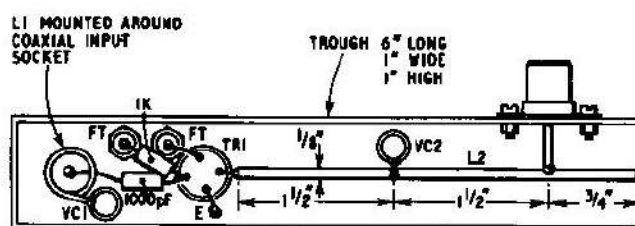Fig. 5.73. Circuit of pre-amplifier. Ft are 1000 pF feed-through capacitors. L1 is 1½ turns ¼" dia.

Fig. 5.74. Trough line layout of circuit given in Fig. 5.73.

high Q collector circuit, and is shown in Fig.5.73 while the mechanical construction is given in the drawing in Fig.5.74.

Coil Data for Oscillator chain and i.f. Amplifier

*L1	15 turns	¼" diam.		20 s.w.g.
L2	7 turns	⅜" diam.		20 s.w.g.
L3	5 turns	⅜" diam.		20 s.w.g.
L4	1 turn	¼" diam.		18 s.w.g.
*L5	30 turns	¼" diam.		28 s.w.g.

*slug tuned

FET CONVERTERS FOR 70 MHz, 144 MHz and 432 MHz

FETs and the performance of the Converters

Although Field Effect transistors have in·fact been in existence for many years, they have only comparatively recently come into general use. This was due mainly to problems in manufacturing suitable devices which have now been overcome.

FET's have "source" and "drain" electrodes which correspond roughly to the emitter and collector of normal transistors. However, the operation of the third electrode, the "gate", is rather unlike that of a transistor base as its control over the source-drain current is achieved without the injection of current carriers, by a purely electric-field effect. Consequently, the input impedance, between source and gate, can be quite large and the design of r.f. input circuits is correspondingly eased. Operation is, in fact, very much like that of a normal hot-cathode valve.

However, the FET has some operational advantage over valves for r.f. amplifier use, apart from the obvious ones of lower power consumption and satisfactory operation at low h.t. rail voltages. The form of curvature of the mutual characteristics of FET's is such that most strong-signal effects are less than those shown by equivalent valve stages (and far less than for stages using conventional transistors). This is particularly true for frequency-changer stages, for which FET's may be almost ideal devices. It should also be true that the minimum achievable noise-levels of stages using devices operating at or near room-temperature are lower than those using devices with heated, emissive cathodes.

These theoretical expectations seem to have been realised for the most part in the converters described, for the strong-signal effects are negligible, in practice, even when several 100mV signals are fed into the input simultaneously. The noise performance is also good: a noise-factor of better than 3dB will be realized on either band, provided that NF of the i.f. tuner into which the converter is fed is not itself worse than about 10dB. This latter proviso is a result of the converter gain having been held to a low value to minimize overall strong-signal effects. With the component values shown, operation is satisfactory, incidentally, at input supply voltages varying from 10 to 14 volts, but if operation from a 9 volt battery is desired, the following component modifications should be made:
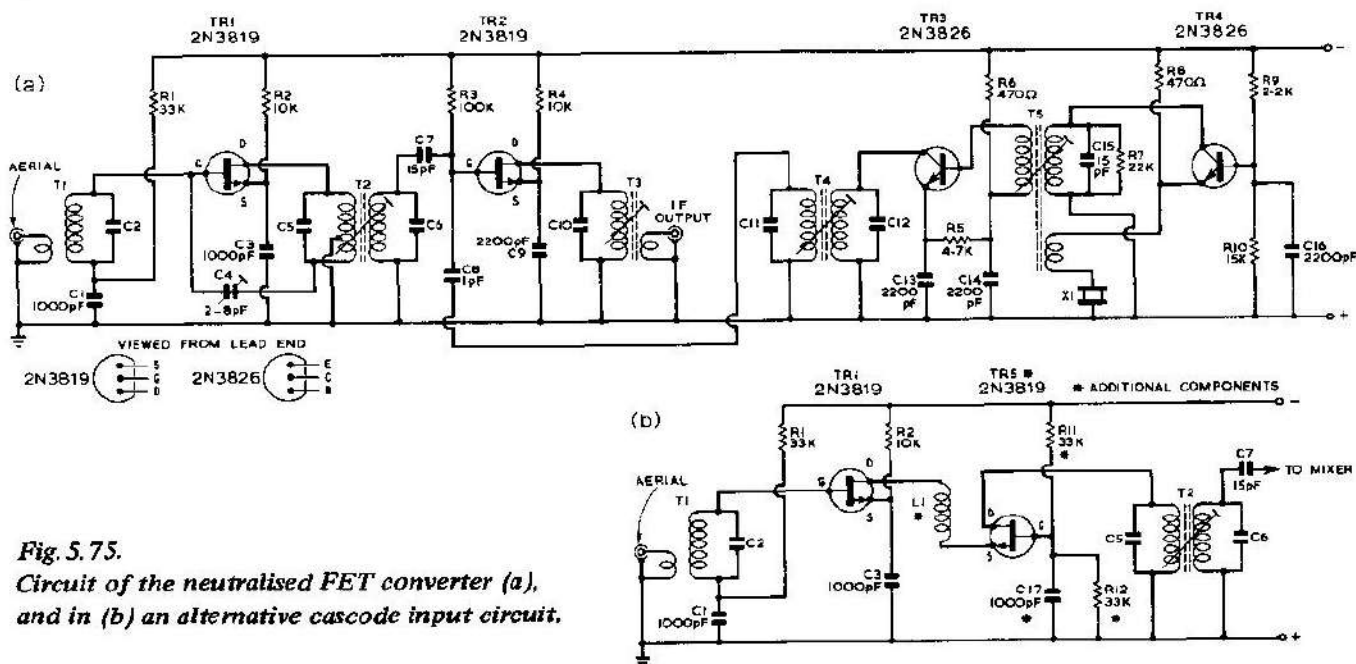
Change R2 to 6.8K ohms (10K ohms)
R4 to 6.8K ohms (10K ohms)
R10 to 10K ohms (15K ohms)
The original values are in brackets.

THE TWO AND FOUR METRE CONVERTERS

Fig. 5.75.
Circuit of the neutralised FET converter (a),
and in (b) an alternative cascode input circuit.

Coil Details

Four Metres

L1   30 turns 36 s.w.g. enam., close wound on 100 K ohms ¼ watt resistor (required for cascode only)

T1   Secondary: 8 turns, 22 s.w.g. enam., self-supporting (wound on ¾ in. mandrel)
Primary: 3 turns, 22 s.w.g. enam., pushed into "cold" end of secondary

T2   Primary: 6 turns, tapped 2 turns from earthy end, 22 s.w.g. enam., close wound on 0.3 in. Aladdin former
Secondary: 6 turns, 22 s.w.g. enam., close wound nearer chassis, spaced ¼ in. from primary

T3   See Table 1

T4   Primary: 4 turns, 22 s.w.g. enam., close wound near chassis on 0.3 in. Aladdin former
Secondary: 4 turns, 22 s.w.g. enam., close wound, spaced ¼ in. from primary

T5   Main: 11 turns, 28 s.w.g. enam., close wound on 0.3 in. Aladdin former
Emitter: 1 turn, 28 s.w.g. enam., overwound at "cold" end of main winding
Output: 2 turns, 28 s.w.g. enam., overwound at centre of main winding

Two Metres

L1   10 turns, 28 s.w.g. enam., close wound on 100 K ohms ¼ watt resistor (required for cascode only)

T1   Secondary: 4 turns, 22 s.w.g. enam., self-supporting (wound on ¾ in. mandrel)
Primary: 2 turns, 22 s.w.g. enam., pushed into "cold" end of secondary

T2   Primary: 5 turns, tapped 2 turns from earthy end, 22 s.w.g. enam., close wound on 0.3 in. Aladdin former (ignore tap for cascode)
Secondary: 5 turns, 22 s.w.g. enam., close wound nearer chassis, spaced ¼₆ in. from primary

T3   See Table

T4, 5   As for 4m

Components List

C1, 3	1000 pF disc ceramic
C2	4.7 pF ceramicon (4m), not required for 2m
C4	2-8 pF miniature trimmer, Plessey type 7365 (not required for cascode)
C5	10 pF ceramicon (4m), 2.2 pF ceramicon (2m)
C6	10 pF ceramicon (4m), not required for 2m
C7	15 pF ceramicon
C8	1 pF ceramicon
C9, 13, 14, 16	2200 pF disc ceramic
C10	22 pF ceramicon (4m), 6.8 pF ceramicon (2m)
C11, 12	33 pF ceramicon (4m), 6.8 pF ceramicon (2m)
C15	15 pF ceramicon
R1	33 K ohms
R2, 4	10 K ohms (10-14 volts), 6.8 K ohms (9 volts)
R3	100 K ohms
R5	4.7 K ohms
R6, 8	470 ohms
R7	22 K ohms
R9	2.2 K ohms
R10	15 K ohms (10-14 volts), 10 K ohms (9 volts)
T2, 4, 5	0.3 in diam. nylon Aladdin former, with iron dust core
T3	0.3 in. diam. nylon Aladdin former, with iron dust core, for i.f.s above 16 MHz
Neosid pot core on Aladdin former for i.f.s below 16 MHz
TR1, 2	2N3819 FET (Texas Instruments Ltd.)
TR3, 4	2N3826 (Texas Instruments Ltd.)
X1	HC-6/U crystal and holder, as required.

The following additional components are required for the cascode modifications:

C17	1000 pF disc ceramic
R11, 12	33 K ohms
L1	100 K ohms (see Coil Details)
TR5	2N3819 (Texas Instruments Ltd.)

Both converters are almost identical in form. A common source, neutralized r.f. stage, using a 2N3819 n-channel FET (Texas Instruments Ltd) feeds, via double tuned circuits, into a common-source 2N3819 mixer (Fig. 5.75). For those anxious to avoid even the simple procedure associated with neutralizing the r.f. stage a cascode variant is available for both bands, which does not appreciably alter the performances in any respect but does require another 2N3819. The local oscillator chain uses two silicon n-p-n planar transistors, type 2N3826 (Texas Instruments Ltd), one as an oscillator with an HC-6/U overtone crystal, operating in the 30 to 35 MHz range, the other as a multiplier, doubling for 4m and quadrupling for 2m, with double tuned circuits feeding into the mixer.

The standard output i.f. for both converters is in the region of 3 MHz but design data is given for i.f.'s between 2 and 30 MHz for the 2m converter and between 2 and 10 MHz for the 4m. An Aladdin former, with enclosing Neosid pot core, is used to accommodate the output transformer for i.f.'s up to 16 MHz, but above this frequency, the Aladdin former only is required. To alter the i.f., it is only necessary to change the i.f. transformer coil and overtone crystal and perform a small amount of re-peaking.
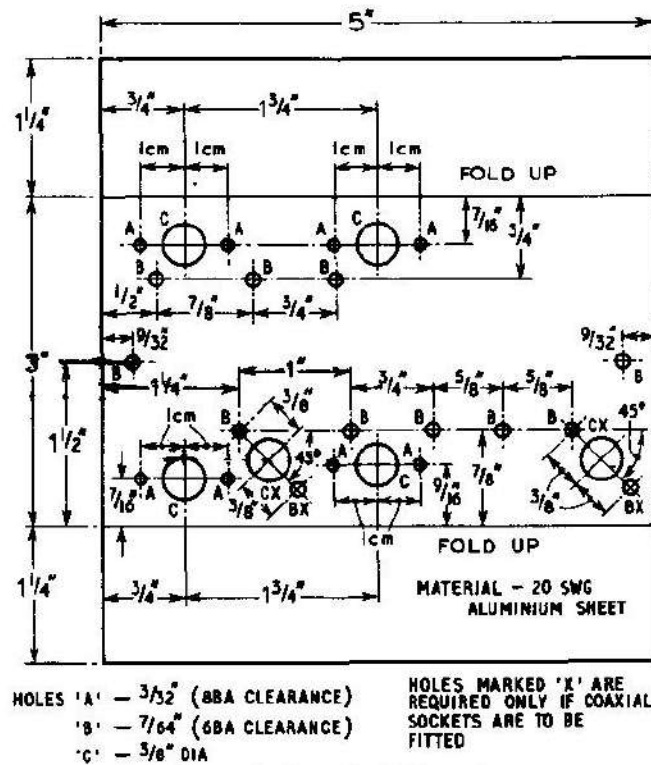
Construction

HOLES 'A' — $^3/32$" (8BA CLEARANCE)
'B' — $^7/64$" (6BA CLEARANCE)
'C' — $^3/8$" DIA

HOLES MARKED 'X' ARE
REQUIRED ONLY IF COAXIAL
SOCKETS ARE TO BE
FITTED

Fig. 5.76. Chassis drilling plan.

The converters are wired on a piece of Lektrokit board, using wiring pins pushed into the holes to support most components, including the crystal and the transistors, see Fig. 5.77. A standard, square piece of Lektrokit laminate board may be cut into three, each part of which may then be used to build a converter, although two fixing holes must be drilled appropriately in the centre piece.

The board is mounted inside a simple, trough section, 20 s.w.g. aluminium chassis, using two 6BA screws and nuts, standing the board ¼" away from the chassis. The four coil formers, i.e. those associated with the oscillator and multiplier stages, together with the r.f. mixer interstage coil and the i.f. transformer are all mounted on the chassis see Fig. 5.77. The aerial coil is air-spaced and is supported by wiring pins on the board. As the converters are fundamentally intended for incorporation into a complete v.h.f. receiver, the aerial and i.f. terminations are, like the ground and input negative connections, made to wiring pins but, if required, two coaxial sockets may be mounted on the chassis, after drilling an additional four holes. No extra shielding is required anywhere, the need for it being avoided by proper layout and coil positioning.

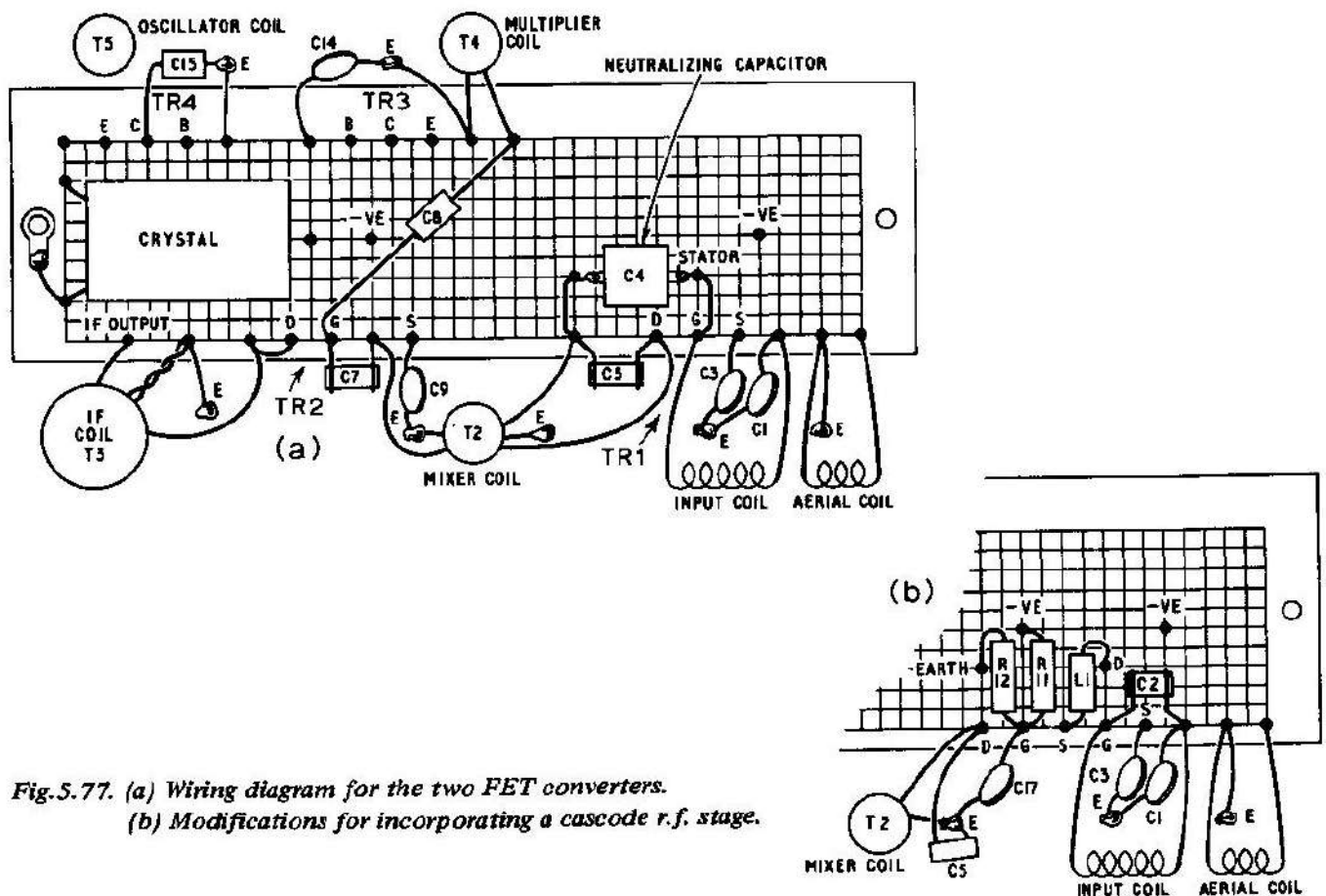The drilling plan for the metalwork for the converters is shown in Fig. 5.76.

Fig. 5.77. (a) Wiring diagram for the two FET converters.
(b) Modifications for incorporating a cascode r.f. stage.

Wiring pins should be pushed into the Lektrokit board from the underside so that about twice as much protrudes above the board as is left below. The best type of pin to use is that with a shoulder and longitudinal splines (Vero Electronics Type 2143). Otherwise, Lektrokit LK3011 pins may be used. The finished converter looks neater if some wiring is done underneath the board, prior to mounting it in the chassis. In particular, the h.t. wires and the earth wire from the crystal clip may be so inserted. If the cascode r.f. stage is to be used, the earth connection to the bottom of R2 may also be made under the board with advantage.

Mounting the Lektrokit board in the chassis requires the use of two 6BA x ½" ch.hd. screws and nuts, spacing the board ¾" away from the chassis. The crystal holder is fixed to the board by soldering the connections to two wiring pins. The crystal is inserted and is then clamped in position by bending a 6BA tag carefully over it, after the tag has been soldered to the earthed wiring pin provided.

The coils for T2 and T4 are produced by close-winding 22 s.w.g. enamelled wire on the shank of a ¼" drill (wind one more turn than is required). If the wire is then pulled off carefully, it may be sprung, by rotation, on to the Aladdin former. Note that the earthed end of the secondary of T2 is nearer to the chassis. The primary is situated at the outer end of the former, with its "cold" end nearer to the secondary. The two coils are spaced by ¾". The earthed end of T4's primary is close to the chassis and the earthed end of the secondary (which is situated right at the other end of the former) is nearer to the primary. The cores for both T2 and T4 are prepared by cutting a standard core in half, and they are locked in position, inside the former by means of narrow strips of polythene sheet.

The main winding of the oscillator coil T5 is close wound near the chassis end of the Aladdin former, its earthed end being close to the chassis. A layer of adhesive Melinex tape facilitates putting the other windings on top, the feedback coil at the cold end and the output winding near the centre of the main winding.

The i.f. transformer T3 should have each half of its tuned winding wound in two adjacent coil former sections, leaving one section available for the output-coupling secondary.

Lining-up and Operation

Getting the converters going should present no difficulties. The output i.f. tuned circuit will not normally require peaking for i.f.'s below 10 MHz. The crystal oscillator will only oscillate strongly on the correct overtone frequency and probably will not oscillate at all unlocked, although some crystals having exceptionally high shunt capacitances may cause it to do so. If a receiver in the 35 MHz region is not available, correct operation of the oscillator may

readily be checked by using the standard trick of searching for a spurious response to the oscillator on a v.h.f. receiver or even one covering the upper h.f. range. Needless to say, if the oscillator does not seem to work at all, try reversing the sense of the feedback winding.

An r.f. detector loop and diode and a set of calibrated absorption wavemeters will be found useful when tuning up the multiplier coils. However, if these are not available, the 1 pF oscillator injection capacitor lead may be temporarily moved to the multiplier collector and the multiplier coil tuned to give a noise peak. Then the injection capacitor should be reconnected to the secondary winding of the multiplier coil which may itself then be aligned, and the primary re-peaked.

It is now necessary to resonate the mixer gate circuit. For 2m i.f.'s below 6 MHz, and for 4m i.f.'s below 3 MHz, it may be found difficult to distinguish between the noise peaks due to operation above and below the local oscillator frequency. A very pronounced dip in noise between noise peaks probably indicates too much oscillator injection. This may be checked by measuring the mixer source current (0 to 10 voltmeter across R4). This current should not increase by even a few per cent when the local oscillator is switched on. If it does, reduce the coupling between the two parts of T4. Operation of the mixer at too high a level of oscillator injection will produce very poor strong signal performance.

With a 75 ohm load resistor (or an aerial) plugged into the input socket, it should now be found possible to tune the primary T2 to resonance at signal frequency, at the same time adjusting the neutralizing capacitor for best stability, in the case of the single, earthed-source, r.f. stage. The input circuit may then be set up in the usual manner for valve r.f. amplifiers and the entire converter finally peaked using a signal generator or external signals.

FET 70cm CONVERTER

An un-neutralised common-gate r.f. stage, using a 2N3819 (Texas) is inductively coupled into a common-source first mixer, also a 2N3819 (Fig.5.78). Then follows a further mixer employing a 2N3819 also in common-source or, in the alternative design already mentioned, a common-emitter 2N3826 silicon planar (Texas), with local-oscillator injection at the crystal frequency itself.

Double tuned circuits, with mutual inductance coupling, are employed between the r.f. stage drain and first mixer gate. The aerial is fed into the source of the common-gate r.f. stage via a circuit which is the exact dual of a conventional pi-coupler, with two shunt inductors and a series capacitor. This has the advantage that it tends to simulate a high-pass rather than a low-pass filter, at frequencies far outside the

matching band, and this helps to reduce possible i.f. interference. The only slight disadvantage relative to the more normal pi-coupler is that mutual coupling between the two coils must be kept fairly low.

Double tuned circuits, at the first i.f. in the 36 MHz region are used to feed the second mixer stage. The second i.f. extends from 2.0 to 4.0 MHz, for 432.0 to 434.0 MHz coverage, using a crystal of 33.077 MHz. The output i.f. may, of course, be put anywhere in the range of up to about 15 MHz by appropriate choice of the crystal and by making modifications to the first (T2) and second (T3) i.f. transformers. The changes to be made to T2 are relatively minor and are easily estimated. Sample designs for T3, suitable for several different i.f.'s may be found in the i.f. transformer table, 2m section. The formula for calculating the crystal frequency is:

$$\text{crystal freq} = \frac{432.0 - (\text{lowest freq. of tunable i.f.})}{13} \text{ MHz}$$

For example, another suitable i.f. would be 3.0 to 5.0 MHz, requiring a 33.0 MHz crystal, which is, in fact, available on the surplus market at the present time. The design for T2 would be essentially unaltered and that for T3 would be as follows:

Primary: 58 turns of 34 s.w.g. enamelled
Secondary: 5 turns of 28 s.w.g. enamelled (for 75 ohms output)

13 turns of 28 s.w.g. enamelled (for 600 ohms output) in the pot-core, as before.

The local oscillator chain consists of three stages, employing 2N3826s, the first as an overtone crystal oscillator in the 33 MHz region, the second tripling to 99 MHz, with double tuned interstage coupling circuits and the final stage quadrupling to about 396 MHz which is coupled into the first mixer by the mutual inductance between L3 and the secondary of T1.

Data for I.F. Transformer T3

Tunable i.f.	i.f. centre	Crystal freq.	T3 primary inductance	Turns	S.w.g.	Former	75 ohm coupling turns	600 ohm coupling turns
Four Metres								
1.8-2.4	2.08 MHz	34.15	205 μH	66	36	Pot-core	5	13
2.1-2.7	2.38	34.00	158	58	36	Pot-core	4	11
4.1-4.7	4.40	33.00	46.8	31	32	Pot-core	2	6
7.1-7.7	7.40	31.50	16.5	19	28	Pot-core	1	3
9.1-9.7	9.40	30.50	10.3	15	28	Pot-core	1	3
Two Metres								
1.8-3.8	2.62 MHz	35.55	369 μH	88	36	Pot-core	6	17
2.0-4.0	2.83	35.50	316	82	36	Pot-core	6	16
4.0-6.0	4.90	35.00	105	47	34	Pot-core	3	9
8.0-10.0	8.95	34.00	31.6	26	32	Pot-core	2	5
14.0-16.0	15.0	32.50	11.3	15	28	Pot-core	1	3
20.0-22.0	21.0	31.00	5.8	29 c/w	26	Aladdin former	2	5
24.0-26.0	25.0	30.00	4.1	27 c/w	26	Aladdin former	2	5
28.0-30.0	29.0	29.00	3.0	25 c/w	26	Aladdin former	2	5

VIEWED FROM LEAD END

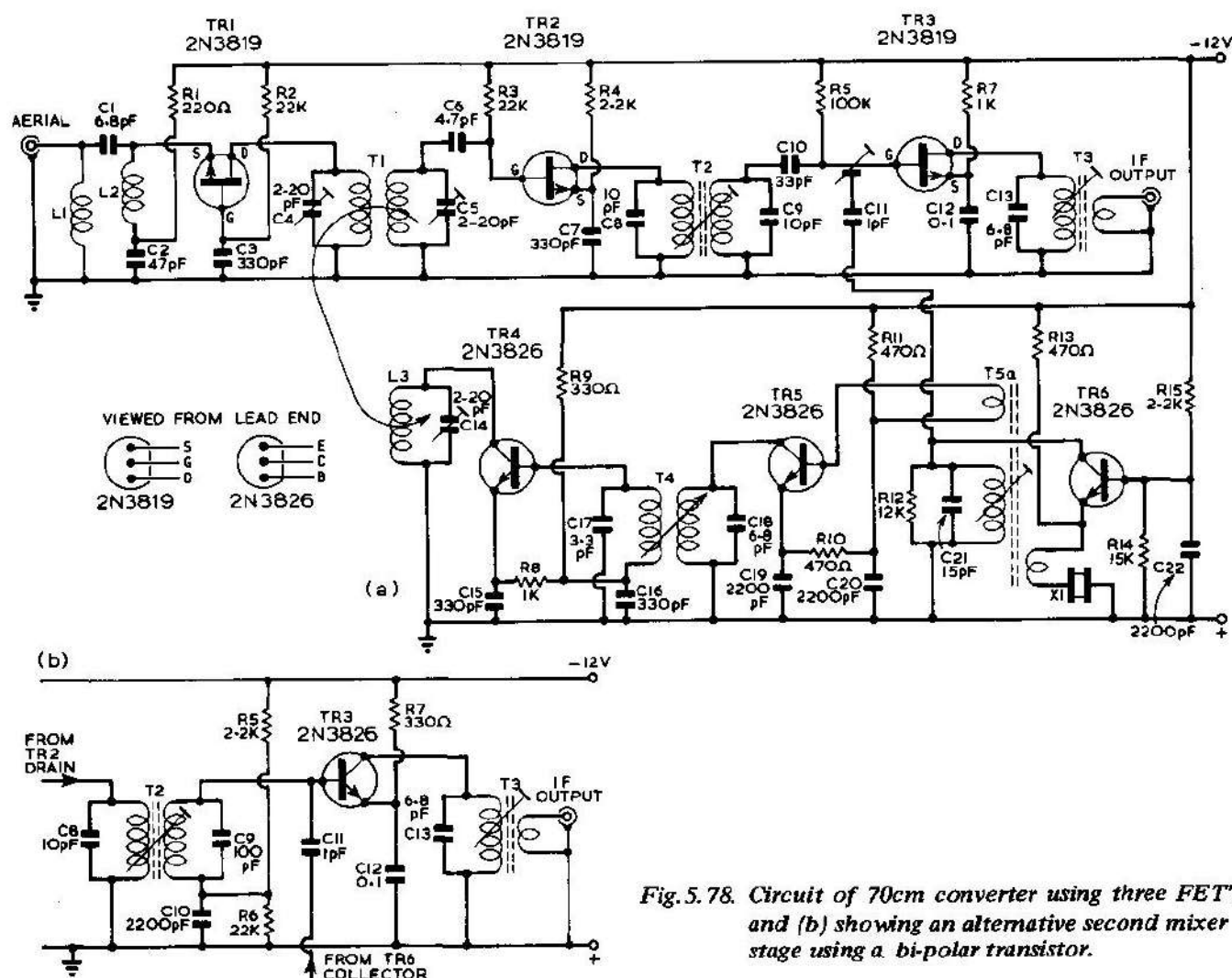2N3819 (S G D)   2N3826 (E C B)

(a)

(b)

Fig. 5.78. Circuit of 70cm converter using three FET's and (b) showing an alternative second mixer stage using a bi-polar transistor.

Coil Details For 70 cm Converter

L1	2 turns, 20 s.w.g. tinned copper, wound on ¼ in. mandrel, with ⁵/₁₆" long leads.
L2	As L1, but with ⅛" leads.
L3	As L1 but with one ⁵/₁₆" lead and one ¼ 'lead.
T1	Primary: as L1, but with one ⁵/₁₆" lead and one 1 in. lead. Secondary: as L1, but with one 3/16" lead and one ½ in. lead. Primary and secondary co-axial, earthy ends near together with approx. ⅛ " spacing.
T2	Primary: 13 turns, 28 s.w.g. enamelled c/w on 0.3 in. Aladdin former. Secondary 3 turns, 28 s.w.g. enamelled c/w (2N3819 second mixer); 4 turns 28 s.w.g. enam. c/w (2N3826 second mixer). Both coils wound initially upon ¼ in. drill and allowed to spring off. Earthy ends near together on former, separated by ⅛ " Secondary near to chassis end of former.

T3	Primary: 82 turns 36 s.w.g. enam., 41 turns wound in two adjacent sections of pot-core former (2.0 to 4.0 Mc/s i.f.), Secondary: 6 turns 28 s.w.g. (for 75 ohm i.f. output impedance) or 16 turns 28 s.w.g. (for 600 ohm i.f. output impedance), in remaining section.
T4	Primary: 5 turns 22 s.w.g. enam. c/w on 0.3 in. Aladdin former. Secondary: 5 turns 22 s.w.g. c/w on 0.3 in. Aladdin former. Coils mounted with earthy ends near together, space ³/₁₆". Primary near to chassis end of former. Both coils wound initially on ¼ in. drill and allowed to spring off.
T5	Main: 11 turns 28 s.w.g. enam., close wound on 0.3 in. Aladdin former. Emitter: 1 turn 28 s.w.g. enam. overwound at "cold" end of main winding. Output: 2 turns 28 s.w.g. enam., overwound at centre of main winding.

Components List For 70 cm Converter

C1,13,18	6.8 pF ceramicon
C2	47 pF ceramicon
C37,15,16	330 pF disc ceramic
4,5,14	2 to 20 pF tubular ceramic trimmers (Henry's of Edgware Road)
C6	4.7 ceramicon
C8	10 pF ceramicon
C9	10 pF ceramicon (FET mixer), 100 pF ceramicon (bipolar mixer)
C10	33 pF ceramicon (FET mixer), 2200 pF disc ceramic (bipolar mixer)
C11	1.0 pF ceramicon
C12	0.1 μF 30V disc ceramic
C17	3.3 pF ceramicon
C19,20,22	2200 pF disc ceramic
C21	15 pF ceramicon
R1	220 ohms
R2,3	22 K ohms
R4,15	2.2 K ohms
R5	100 K ohms (FET mixer), 2.2 K ohms (bipolar mixer)
R6	not required, (FET mixer), 22 K ohms (bipolar mixer)
R7	1 K ohms (FET mixer,) 330 ohms (bipolar mixer)
R8	1 K ohms
R9	330 ohms
R10,11,13	470 ohms
R12	12 K ohms
R14	15 K ohms
T2,4,5	0.3 in. diam nylon Aladdin former, with iron dust core
T3	Denco/Neosid pot-core (Henry's of Edgware Road)
TR1, 2	2N3819 FET (Texas Instrument Ltd.)
TR3	2N3819 FET or 2N3826 bipolar (see text)
TR4,5,6	2N3826 bipolar silicon planar transistor (Texas Instruments Ltd.)
X1	HC-6/U crystal (and holder) or HC-18/U crystal, as required. 6 Oxley Type SO1 (6BA) ceramic stand-off pillars. LEKTROKIT S.R.B.P. board and wiring pins, as required. Belling-Lee coaxial sockets, if required

Construction
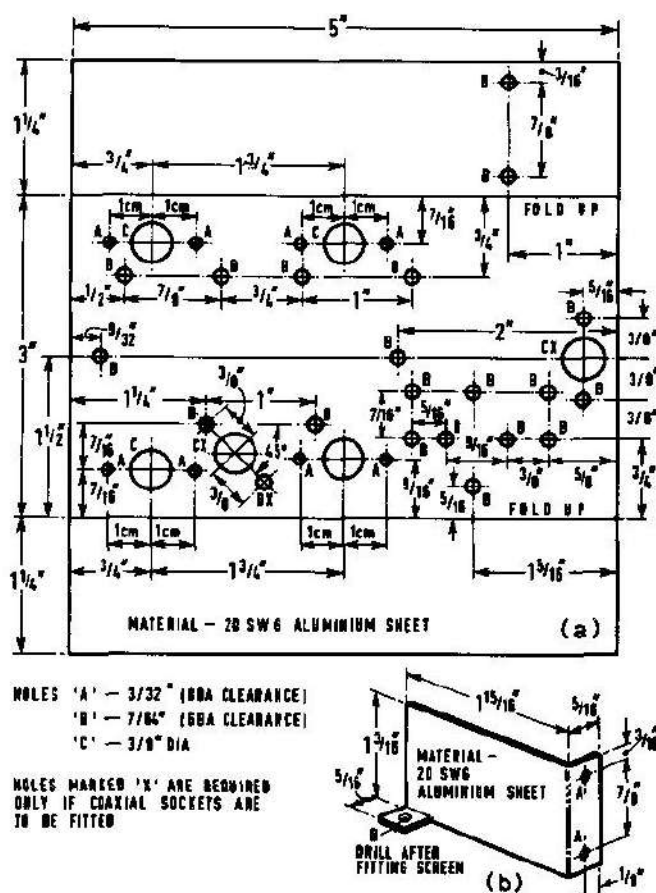
Fig. 5.79. *Dimensions of the chassis and screen for the input stage.*

Like the two v.h.f. converters, the 70cm unit uses a piece of Lektrokit board mounted inside a small, 20 s.w.g. aluminium U-channel chassis.

It is spaced about 1¼" away from the chassis, with one end fixed as previously, the other being secured by soldering a wiring pin to an earthed tag bolted to the chassis. Wiring pins in the board serve to support most components but those operating at signal or first oscillator frequency are mounted either upon ceramic pillars or on small tubular trimmers. A small screen between the input and output circuits of the r.f. stage ensures good stability. The single fixing hole in the screen tab should be marked and drilled only after the two holes have been drilled in the main flange, so allowing the screen to be fitted.

Apart from these details, most of the remarks concerning construction of the preceding converters apply also to this model. As before, the aerial input and i.f. output terminations may be either internal or external, using Belling-Lee coaxial sockets.

The coils associated with L3 and T1 are tuned with tubular trimmers. The type specified was chosen for its mechanical suitability but it has an unmodified capacitance range of from 2 to 20 pF. This minimum capacity is too high and should be reduced by removing about ¹/₁₆" of silvering completely from
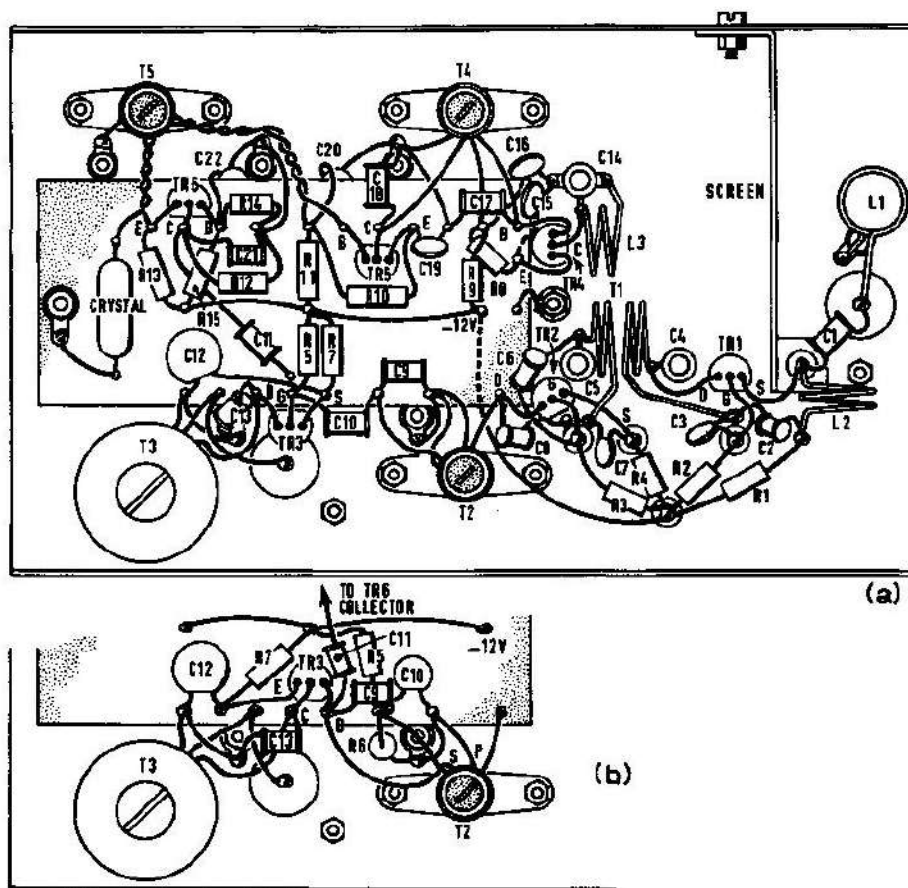
Fig. 5.80. Component layout for the standard converter, and (b) alternative mixer.

the normally clamped end of the trimmer using a file and a piece of medium grade emery cloth.

The coils for L3 and T1 are mounted directly upon their respective trimmers but their earthy ends are not brought to the base of the trimmer, instead they are returned to the stage common-earthing point (Fig.5.80). The gate lead of TR1 is shown, for clarity, in Fig.5.80 as being connected directly to the appropriate stand-off but, in fact, a shorter gate-earth lead results if C3 is grounded by a very short wire and then the gate of TR1 is connected as close as possible to C3 itself. The mixer, TR2, source lead should also be kept short. The by-passing of TR4 is improved if C15 and C16 are taken to their respective pins on the underside of the wiring board, before the latter is fitted. The negative supply wire to the TR1-TR2 stand-off may also be connected underneath the board in this way.

If an HC-6/U crystal is to be used, the crystal holder should have its lugs bent outwards and then soldered to the wiring pins provided. However, a miniature HC-18/U crystal should be fitted by passing the wires through a convenient pair of holes and then bringing the wires up through the board again for soldering to the pins. Very loose capacitative coupling is required between the crystal oscillator collector and the second mixer gate. This is accomplished by utilizing the capacitance between two adjacent pins on

the Lektrokit board in series with the 1 pF ceramicon C11. When the 2N3826 mixer is employed, C11 alone provides the correct oscillator injection level at the mixer base.

Measured Operating Conditions
All voltages are measured with respect to
the −12 volt supply line.

		Emitter or Source	Base or Gate	
TR1	R.F. Amplifier	0.5 to 1.5	—	
TR2	First Mixer	0.7 to 2.0	—	
TR3	Second Mixer	{ 0.7 to 2.0	—	(2N3819 FET)
		{ 0.8	1.1	(2N3826)
TR4	Second Mult.	1.2 to 1.8	—	
TR5	First Mult.	1.5 to 2.5	—	
TR6	Crystal Osc.	0.9 to 1.1	1.5	

Aligning the Converter

Again, most of the comments made previously still apply. The local oscillator chain should be aligned first. An r.f. detector and a set of absorption wavemeters are very helpful during this procedure. However, it is almost impossible to find the wrong harmonics using the coils and capacitors suggested.

When the oscillator chain is properly adjusted, tune the secondary of T2 until the drain current of TR3 increases slightly, as evidenced by a rise in d.c. voltage across R7. Then detune T2 secondary on the h.f. side until a small noise peak is heard. Next the slug in the primary of T2 may be adjusted to produce a further noise peak, due to the first mixer stage. A check with an absorption wavemeter will be worthwhile at this stage to ensure that T2 is tuned within the 35 to 37 MHz range. Incidentally, it will be noticed that the shot-noise output of the FET mixer stage is very low as compared with that produced by bi-polar mixers, until the local oscillator is coupled in and the input circuit tuned to resonance.

The next step is to disconnect the drain lead of TR1 and unscrew C4 fully. Then, resonate C5 and the secondary of T1 at the oscillator frequency by adjusting C5 either until the d.c. voltage measured across R4 kicks upwards slightly or else by coupling an r.f. indicator loosely into L3 and tuning C5 for a dip in indication. Detuning C5 slightly on the h.f. side should give rise to a small noise peak. Re-connect the TR1 drain lead and, with the aerial plugged into the input socket, increase the setting of C4 until a further noise is heard, or until local signals are peaked up. The primary of T1 will be found to be more sharply tuned than the secondary. In fact, the overall bandwidth of T1 may be controlled by adjusting the tightness of coupling between the two coils. The coil spacing should be changed as necessary, retuning C4 and C5, until the response of T1 is fairly flat over a band of frequencies 4 to 6 MHz wide, centred at about 433 MHz. Insufficient coupling can even result in self-oscillation of the r.f. stage when using some 2N3819s. It will probably not be found necessary to alter the component values in the input pi-coupler, but if a noise generator is available, C1 may be temporarily replaced by a 2 to 8 pF Mullard trimmer and L1 and L2 adjusted by bending them to optimize the input coupling circuits absolutely for the 2N3819 in use. Increasing the value of C1 will probably be found to reduce the loaded Q of the T1 primary to a certain extent.

A TUNABLE CONVERTER FOR 70cm AMATEUR TELEVISION
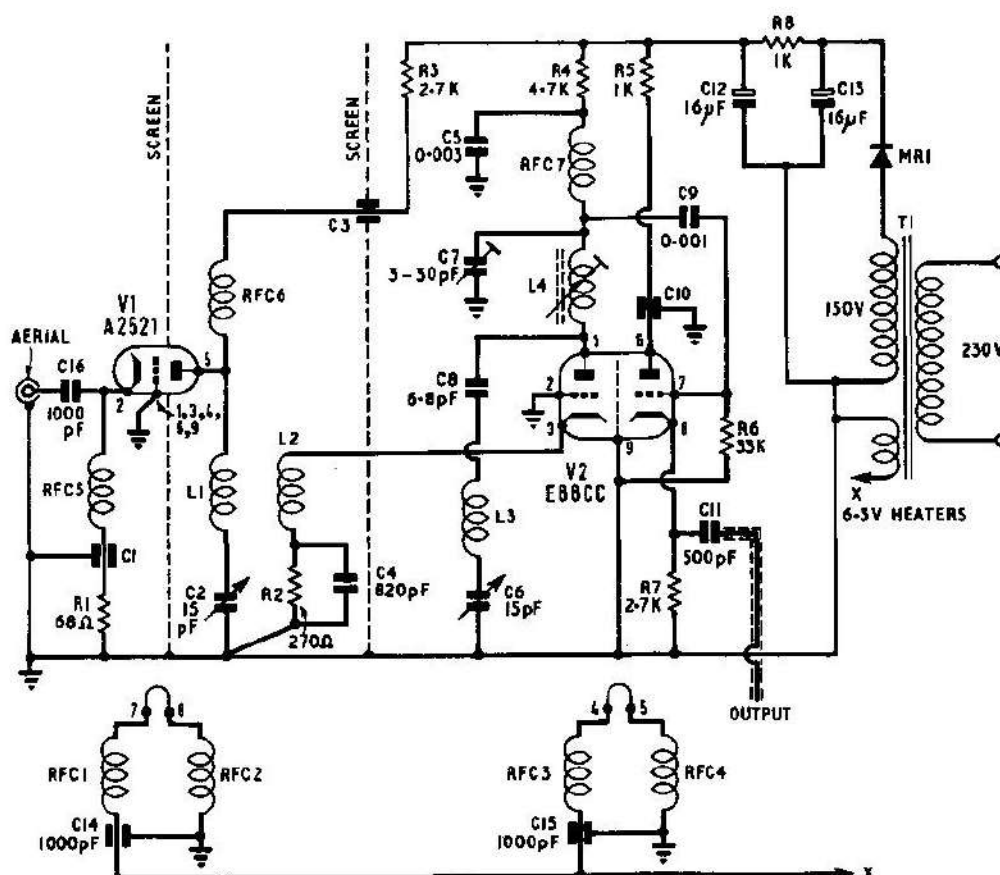
Fig.5.81. Circuit diagram of the converter including h.t. supply, and l.t. supply line.

Components Lists for 70cm TV Converter

Resistors

R1	68 ohms, ¼ watt.
R2	270 ohms, ¼ watt.
R3, 7	2.7K ohms, ¼ watt.
R4	4.7K ohms, ¼ watt.
R5	1K ohm, ¼ watt.
R6	33K ohms, ¼ watt.
R8	1K ohm, ½ watt.

Capacitors

C1, 3, 10, 14, 15	1000pF Radiospares feed through types.
C16, 9	1000pF disc ceramic.
C5	0.003μF disc ceramic.
C8	6.8pF ceramic.
C4	820pF ceramic.
C11	500pF ceramic.
C2, 6	15pF variable (2 moving—2 fixed plates).
C7	3-30pF Philips trimmer.
C12, 13	16μF 200V working.

Coils

L1	4½ in. of 14 s.w.g. tinned copper.
L3	2¾ in. of 14 s.w.g. tinned copper.
L2	3 in. of 20 s.w.g. p.v.c. covered.
L4	5 turns of 26 s.w.g. enamel ¼ in. diameter (Aladdin former).
R.f.c. 1, 2, 3, 4	10 in. of 22 s.w.g. enamel ³⁄₁₆ in. diameter self supporting.

T1 small mains transformer, 150V and 6.3 secondary.
V1 GEC A2521.
V2 Mullard E88CC.
2 p.t.f.e. valve holders B9A.

This converter will tune from about 395 MHz to 490 MHz, but the tuning rate is not excessive, even without a slow motion drive, when it is used as intended into a television receiver. Although it has proved quite satisfactory for reception of local 70cm phone stations it will not be sufficiently stable for general use if a selective communications receiver is used as the i.f. stage.

Circuit description

The r.f. stage is a conventional grounded grid amplifier using an A2521 (V1), with a half wave anode line (L1). An E88CC double triode (V2) is used as a self oscillating mixer and a cathode follower stage. The oscillator is a Colpitts circuit, feedback being through the inter-electrode capacitances of the valve.

The signal is coupled into the cathode of V2 by L2, RFC7, L4 and C7 filter the output from the mixer and select the i.f. signal which is fed into the cathode follower to give some isolation and allow low impedance output. The i.f. chosen was 56 MHz

(channel 3) to give a reasonable spacing from the London television transmitter on 45 MHz (Channel 1). If a different i.f. has to be used it should only be necessary to make a small adjustment to L4.

Construction

MATERIAL .... 22 SWG BRASS SHEET

(a) RF Stage

(b) Mixer
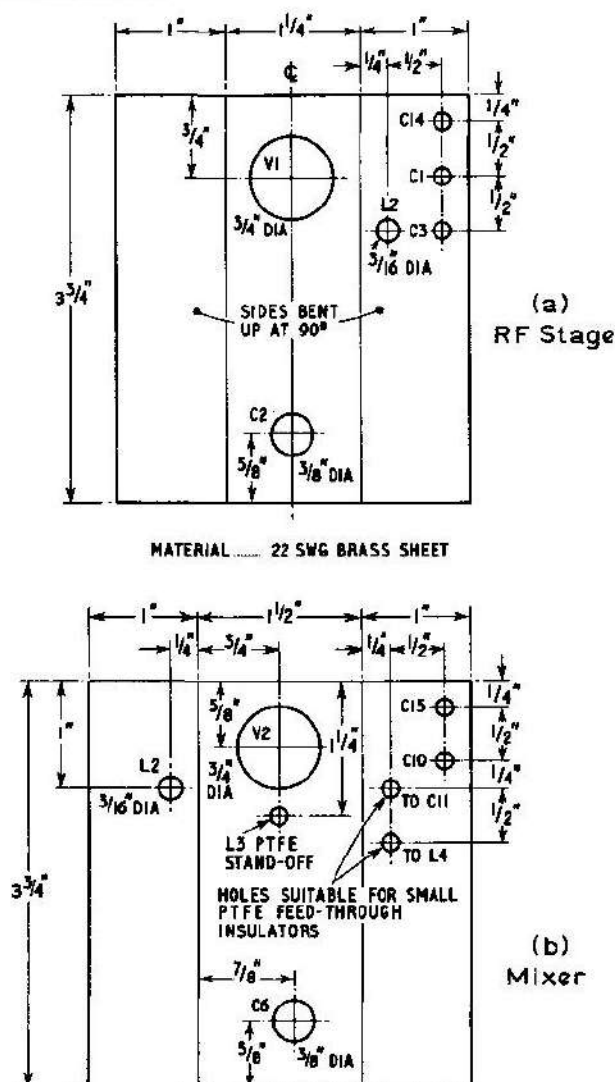
Fig. 5.82. Dimensions and drilling plan for the construction of the brass troughs.

The two brass troughs should be constructed first to the dimensions in Fig. 5.82. The valve holders may then be positioned so that the anode pin (5) of V1 is in line with L1 and the anode pin (1) of V2 is in line with L3. P.t.f.e. valve holders are strongly recommended. The feedthrough capacitors may now be soldered into position and the brass polished or silver plated. With the mixer trough spaced ½" from the r.f. stage trough and a small gap left at each end to allow for the wiring, both can be mounted in a 6" x 4" x 2¼" standard aluminium chassis. It is essential that a small screen is fitted to V1 holder across pins 3 and 6 as shown in Fig. 5.83. A scrap of tin plate is suitable and should be soldered to the sides of the r.f. trough. The r.f. stage will become unstable without this screen.
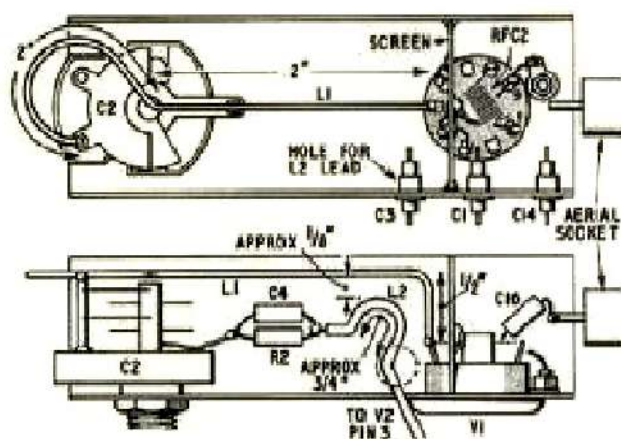
Fig. 5.83. Line drawing showing the position of components in the brass trough. The shaping of L1 and L2 is shown.

It is suggested that L2 is installed at this stage between pin 3 of V2 and R2/C4. R2/C4 should be earthed to the tag of C2; this part of the capacitor (C2) then forms a common path to the chassis and provides further coupling. The line L1 needs to be bent to accommodate its length of 4½" in the space available, the end nearest to C2 being shaped into a hook as shown in Fig. 5.83.

The resistors R1 and R3 are located outside the r.f. anode line trough. A small stand-off insulator with a low capacity to ground is used to support the 6.8 pF capacitor (C8) and the end of L3. The other components will fit into place if the brass troughs have been drilled as shown.

Testing

A 70cm signal is required, the third harmonic from a two meter transmitter will do for those without a signal generator. The mixer (C6) on the prototype tunes 70cm when about 75 per cent open and with the r.f. stage capacitor about 50 per cent open.

When a signal is located C7 should be tuned to peak up the i.f. circuit; a dust core will probably not be needed to tune L4 to 56 MHz. A 22K ohms damping resistor may be needed to tune L4 to increase the bandwidth though at the expense of the gain.

Performance

The noise factor has not been measured as the overall noise of the converter and the i.f. receiver will depend on the bandwidth of the receiver. Some re-alignment of the TV receiver to restrict the bandwidth to approximately 2 MHz is worthwhile, but this must not be overdone or the rise time of the sync pulses will be spoilt thus preventing the picture locking, to say nothing of the effect on quality.

Generally it is better to choose a TV receiver without flywheel sync, as amateur signals may not conform exactly with the British 405 line standard. A direct sync model is therefore strongly recommended. Some fringe area models have extra sync amplification and shaping which is useful.

GROUNDED GRID 70cm CONVERTER

In this converter, two grounded grid amplifiers are used at the signal frequency, followed by crystal diode mixer and i.f. input amplifier. The r.f. gain available is 25dB at a noise factor of 7dB with a bandwidth of 4 MHz.
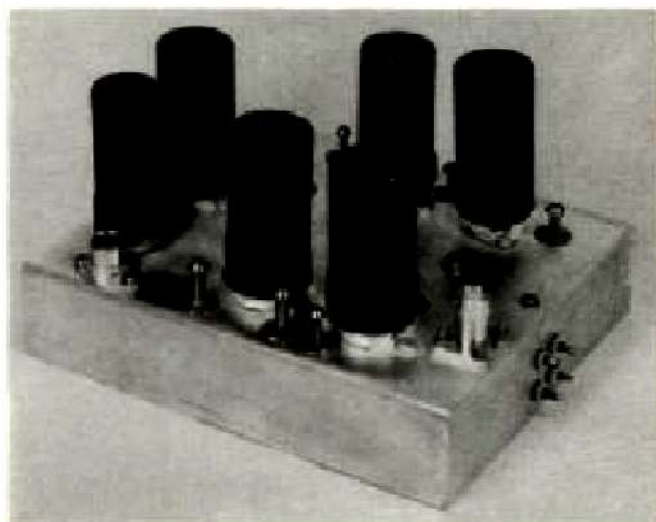
Fig. 5.84. A general view of the complete 70cm converter.

Circuit Description

The r.f. stages comprise two A2521 grounded grid amplifiers connected in cascade. The aerial is tapped into the input coil L9 and the tapping point and input circuit tuning adjusted for optimum noise factor. The coupling between the two r.f. stages consists of an r.f. transformer made from two identical coils L10 and L11 mounted side by side. An exactly similar transformer couples the output of the second r.f. stage into the crystal mixer. All the circuits are series tuned, except the coil feeding the mixer which is untuned. The layout of the transformers can be seen in the photograph in Fig.5.85.

Multiplier Chain

One half of an ECC85 double triode is used with a 20.3 MHz crystal as a Squier oscillator and the second triode is tuned to the fourth harmonic of the crystal at 81.2 MHz. The cathode of the following A2521 multiplier is tapped down the preceding anode
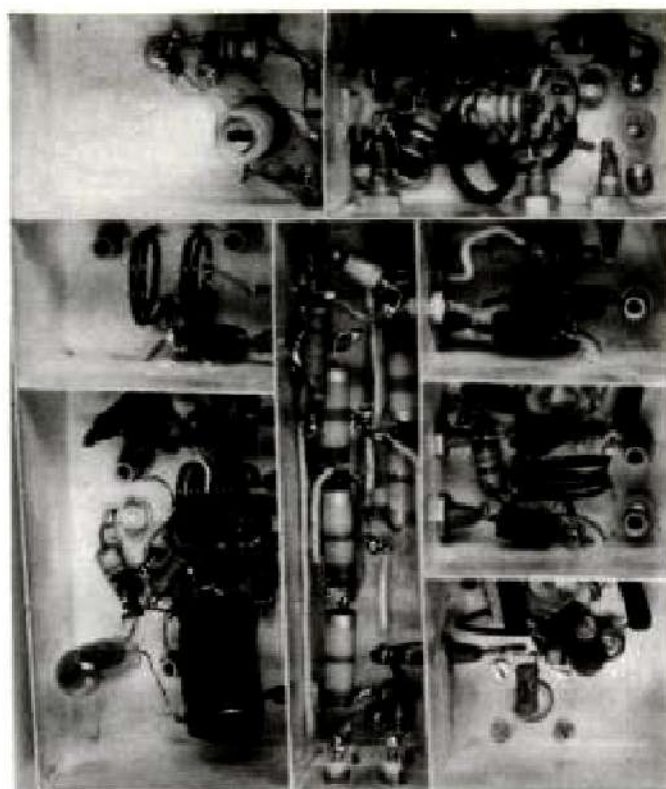
intermediate frequency and this is followed by a cathode follower in order to provide low impedance output which may conveniently be coupled to a tunable receiver at intermediate frequency.

Mechanical Construction

The chassis consists of a box 6" x 5" x 1¼" deep constructed of 20 s.w.g. brass sheet and silver plated. It is divided into compartments as shown in Fig.5.86 and the walls of the compartment are soldered to the main chassis. The transverse screens across the A2521 valve sockets must be made close fitting and the grid pins solidly earthed to the screen. P.t.f.e. feedthrough connectors are used to feed the u.h.f. signals into the mixer and feedthrough capacitors are used to filter the h.t. supplies into the various compartments.

Fig. 5.85. Underside view showing positions of the compartments.

coil to provide matching and its anode circuit is tuned to 406 MHz. The use of high orders of multiplication helps to reduce the number of unwanted harmonics of the crystal frequency, but some are nevertheless present, and these are attenuated by feeding the mixer diode from a tap on a tuned circuit which is loosely coupled to the A2521 anode circuit. The spacing between these coils should be as great as possible consistent with adequate diode current. The relative positions can be seen in Fig.5.86.

I.F. Stage

The wire-ended GEX66 mixer diode is followed by an ECC85 as a series cascode amplifier at

Fig. 5.86. Drawing showing layout details of underside of converter and compartments.

COIL DATA

L1–L6	23cm 24 s.w.g. enamelled copper wire wound into ¼ in. dia. air cored coil.
L7–L8	24 s.w.g. enamelled copper wire close wound on to 100kΩ0.25W resistor.
L9	¾ turn 18 s.w.g. enamelled copper wire wound into ¼ in. dia. air cored coil.
L10–L13	1¾ turns 18 s.w.g. enamelled copper wire wound into ¼ in. dia. air cored coil.
L14	44 s.w.g. silk insulated copper wire close wound on to 100kΩ0.25W resistor.
L15	25 turns 26 s.w.g. enamelled copper wire wound on to ¼ in. dia. former tapped 7 turns from earthy end.
L16	23 turns 24 s.w.g. enamelled copper wire wound on to ¼ in. dia. former and tuned with ferrite slug.
L17	25 turns 20 s.w.g. enamelled copper wire wound into ½ in. dia. air cored coil, tapped 5½ turns from crystal end.
L18	6½ turns 18 s.w.g. enamelled copper wire wound into ½ in. dia. air cored coil, tapped halfway.
L19	R.F. choke 3 μH.
L20–L21	24cm 24 s.w.g. enamelled copper wire wound into ¼ in. dia. air cored coil.
L22	26 s.w.g. enamelled copper wire close wound on to 100kΩ0.25W resistor.
L23	1¾ turns 18 s.w.g. enamelled copper wire wound into ¼ in. dia. air cored coil.
L24	1¾ turns 18 s.w.g. enamelled copper wire wound into ¼ in. dia. air cored coil, tapped halfway.

Fig. 5.87. Circuit diagram of the complete converter.

A 23cm CONVERTER

This is a grounded grid r.f. amplifier circuit and provides a gain of more than 13dB at signal frequency with a noise factor of 11.5dB at a bandwidth of 5 MHz.
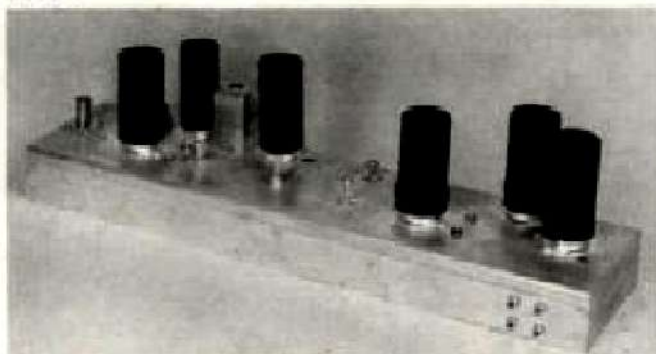

Fig. 5.88. General view of complete 23cm converter.

Circuit Description

The r.f. stage consists of a grounded grid A2521 trough line amplifier with tuned input and output lines. The input aerial impedance is matched by varying the position of the input tap along the cathode line, and by tuning the line.

The cathode line is a $3\lambda/4$ line short circuited at its extreme end by a capacitor of approximately 50 pF and tuned by an 0.5 to 3.0 pF trimmer mounted halfway along the line.

The anode line is a $3\lambda/4$ line short circuited and tuned by a 0.5 to 3.0 pF trimmer mounted halfway along the line. This is isolated from h.t. by a 47 pF capacitor. The output is tapped at the mechanical short circuit (the electrical r.f. short circuit being somewhere in the chassis) and fed directly into the crystal.

Multiplier Chain

This consists of one half of an ECC85 valve connected as a Squier oscillator controlled by a 55 MHz crystal. The second half of this stage is tuned to the third harmonic of the crystal frequency, thus giving 165 MHz output. This is fed into a grounded grid A2521 quadrupler operating in Class C, the output circuit being a $\lambda/2$ line tuned at its extreme end by a 0.5 to 3.0 pF trimmer. The 660 MHz output from this stage is tapped off about half-way along the
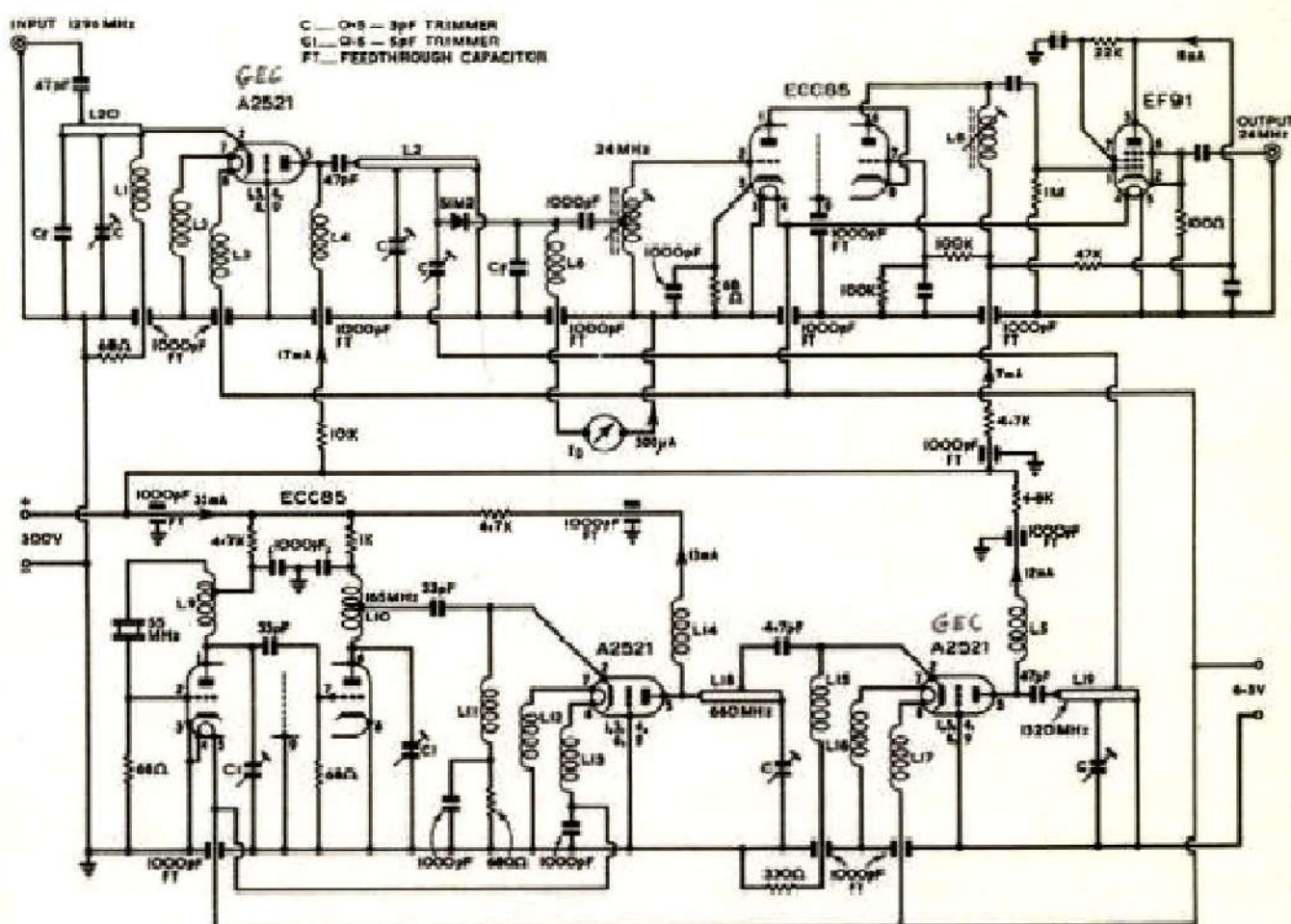
Fig. 5.89. Circuit of 23cm converter.

line (this tap should be adjusted for maximum crystal current) and fed into a grounded grid A2521 doubler also operating in Class C. The anode line of this stage is a $3\lambda/4$ line short circuited to the chassis at its extreme end and tuned by a 0.5 to 3.0 pF trimmer, h.t. isolation being achieved by a 47 pF capacitor at the anode pin.

The output is tapped off at the mechanical short circuit once again and fed into a fabricated variable capacitor, which couples the output into the crystal.

Mixer

This consists of an SIM2 low-noise diode mounted in a silver-plated brass holder, r.f. decoupling being achieved with a 50 pF capacitor formed at the base of the crystal holder. The output at 24 MHz is fed into a tapped input coil on which the tap should be adjusted for best noise performance; this occurs at about halfway along the coil.

I.F. Stage

This is an ECC85 connected in series cascode tuned to 24 MHz followed by an EF91/CV138

cathode follower buffer stage.

Construction

The chassis consists of a shallow box, 13″ long by 3¼″ wide and 1¼″ deep, constructed of 20 s.w.g. silver plated brass sheet. It is divided into compartments as shown in Fig.5.90, the walls of the compartments being solidly soldered to the main chassis. One wall of the chassis is made removable to facilitate the making of h.t. connections. The transverse screens across the valve sockets must be made very close fitting and the grid pins solidly earthed to the screen in all cases. A wiring jig should be inserted in the valve socket whilst this is done.

The valve sockets are all of p.t.f.e. and the input and output sockets are Transradio Type BNC and Belling Lee coaxial respectively.

The connections from the u.h.f. stages into the mixer are made by means of p.t.f.e. feedthrough capacitors. All power supply filter resistors and feedthrough capacitors are mounted in the power supply compartment.
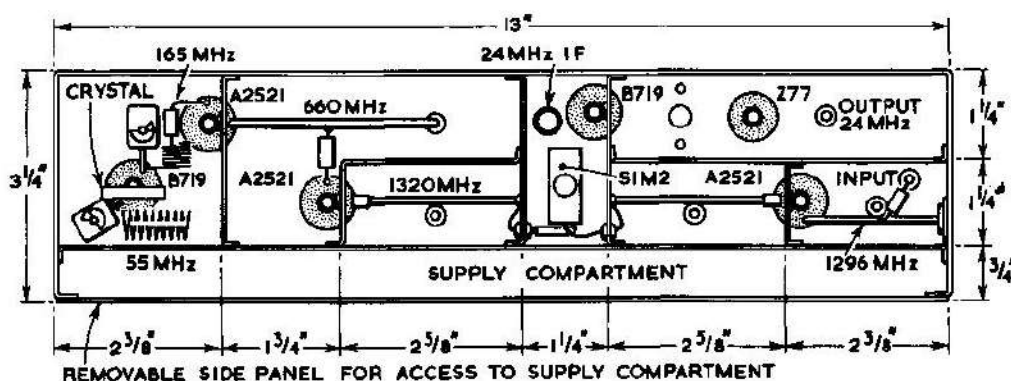
Fig. 5.90. Drawing showing layout of 23cm converter.

COIL DATA

L1–L5	8cm of 24 s.w.g. enamelled copper wire, wound into ⅛ in. dia. air-cored coil.
L6	24 MHz r.f. choke.
L7	30 turns 30 s.w.g. enamelled copper wire, wound on ¾ in. dia. former, tapped halfway and tuned with ferrite slug.
L8	17 turns 36 s.w.g. enamelled copper wire, wound on ¼ in. dia. former and tuned with ferrite slug.
L9	9 turns 18 s.w.g. tinned copper wire, wound into ½ in. dia. open-spaced, air-cored coil. Tapped 7 turns from anode end.
L10	3 turns 18 s.w.g. tinned copper wire, wound into ½ in. dia. open-spaced, air-cooled coil. Tapped 2 turns from anode end.
L11–L13	30 s.w.g. enamelled copper wire, close wound on to 100k Ω ½W resistor.
L14–L17	16cm of 24 s.w.g. enamelled copper wire, wound into ⅛ in. dia. air-cored coil.
L18	7½cm of ¼ in. dia. silver-plated brass rod mounted ¼ in. above chassis, tuned at extreme end, tapped 3½cm from open-circuit.
L19–L21	6cm of ¼ in. dia. silver-plated brass rod mounted ¼ in. above chassis, tuned at 3cm from short-circuit, tapped at short-circuit.
L20	5cm of ¼ in. dia. silver-plated brass rod mounted ¼ in. above chassis, tuned 2½cm from short-circuited, tapped 2cm from short-circuit.

A SIMPLE 23cm SINGLE CONVERSION CONVERTER FOR AN I.F. OF 24-26 MHz

Fig. 5.91. A 23cm converter.

Fig. 5.93. Signal and final oscillator chain. L6 and L7 are ⌇ lines soldered directly to the box at both ends and tuned by 2BA screws facing the centre of each line. The central screen has a cut-out ¼" long by ⅜" wide to allow signal diode and coupling loop to be fitted: this aperture couples the local oscillator into the diode. Aerial and signal diode CR2 coupling loops are ¼" wide copper strip ¾" long. The diode mounting plate is insulated from the box by polythene to form a by-pass capacitor. The plate is fixed by 6BA screws which must be insulated from the plate with bushes. Diode CR2 is IN21C or IN21, the latter gives slightly poorer performance.

The oscillator - multiplier chain, Fig.5.92, must provide sufficient injection, to allow light coupling to the mixer so as to avoid noise arising from it coupling into the signal circuit.

In order to make best use of the transistors, only frequency doubling is carried out at the higher frequency but tripling is used at the lowest frequency. For the final stage, however, a diode is the simplest arrangement.

It should be noted that since the collector capacitance varies with the collector voltage it is desirable to stabilize the operating voltage to avoid detuning effects.

The mixer unit consists of two half wave lines tuned to 1272 and 1296 MHz respectively. They are separated from one another by the intercircuit screen, the mixer diode being mounted so that it fits into a cut-out in the screen through which the local oscillator coupling is obtained. A low noise neutralised i.f. amplifier (24-26 MHz) using a GM290 transistor follows the mixer.

Fig.5.92. Oscillator chain. TR1, 2, 3 and 4 are 2N3663. CR1 is GEX66. RFC 7 turns ⅜" dia. 28 s.w.g. Crystal is 35 MHz

Circuit Description

TR1 is operated as a grounded base crystal oscillator with a 35.333 MHz crystal. The output from this is capacitively coupled to TR2 which triples to 106 MHz. TR3 and TR4 operate as doublers to reach 424 MHz. The final frequency of 1272 MHz is reached by the diode tripler CR1.

The emitter resistors connected to TR2, TR3 and TR4 are shown as 100 ohms. They may be 50-150 ohms since their primary purpose is to provide test points for setting up the circuit.

Construction

The whole converter can be built on a 6" x 4" plate of copper or brass or even copper clad fibre board which may form the lid of a box. Alternatively it may be fixed inside a standard diecast box.

The i.f amplifier may be built into the same box but should be screened by adding suitable partitions. The mixer unit should be attached to the box so that

the lead to CR1 is short and the tuning screws are accessible, see Fig.5.95.

Fig.5.94. The i.f. amplifier.

Fig.5.95. Underside view of 23cm converter showing component layout.

COIL DATA

L1 12 turns 28 s.w.g. tapped 1 turn from cold end, slug tuned ¼" former.

L2 6 turns 18 s.w.g. tapped ½ turn from hot end, air spaced.

L3 2½ turns 18 s.w.g. tapped ½ turn from hot end, air spaced.

L4 1½ turns 16 s.w.g. tapped ½ turn from hot end, air spaced.

L5 1½ turns 18 s.w.g., output taken ¾ turn from cold end

L6 4 inches of ¾" dia. rod or tube }
L7 4 inches of ¼" dia. rod or tube } Fig.5.94

L8 4 turns 28 s.w.g. wound over cold end of L9

L9 24 turns 28 s.w.g. slug tuned ¼" former

L10 24 turns 28 s.w.g. slug tuned ¼" former

L11 4 turns 28 s.w.g. wound over cold end of L10

All air spaced coils wound on ¼" dia. mandrel

Alignment

Adjust slug of L1, to give maximum current reading across TR2 emitter resistor, adjust trimmers

across L2 and L3 for the maximum reading across the preceding emitter resistor.

Typical readings across the emitter resistors of TR2, TR3 and TR4 are 10mA, 5mA and 5mA respectively.

L4 and L5 are adjusted by varying the turns spacing to obtain the maximum reading across the 100 ohm resistor in series with the RFC associated with CR1. A typical value is 6mA.

Following this the trimmer [L6] (2BA screw) is adjusted for maximum mixer diode CR2 current. L7 is now tuned to give maximum mixer diode current which will indicate that it is resonated at the oscillator frequency. The trimmer screw (2BA) of L7 should then be turned back about half a turn and finally adjusted for maximum signal with a 1296 MHz test signal fed into the aerial socket.

The operating value of mixer diode current is about 300 μA.

A 13cm (2300 MHz) CONVERTER

In this 13 cm crystal controlled converter, signals at 2300 MHz are mixed with the output of a crystal oscillator chain at 2280 MHz the i.f. produced being amplified by a wideband head amplifier covering 20-30 MHz. The circuit is shown in Fig.5.97. The upper half shows the i.f. head amplifier and also the trough line assembly which consists of the aerial and mixer circuits and the final stages of the multiplier chain.
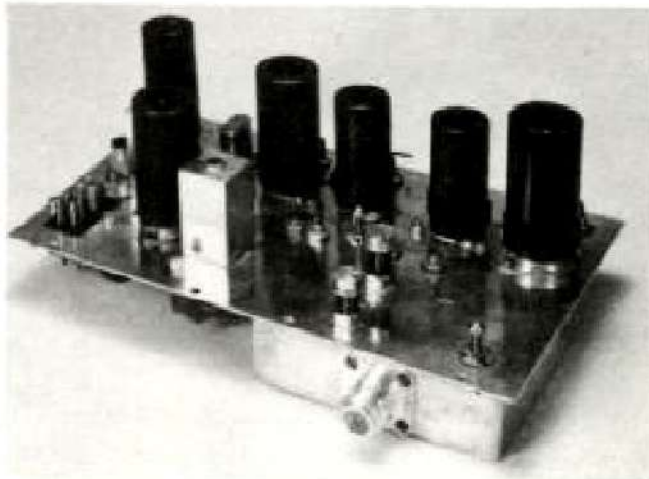
Fig.5.96. A 13cm converter.

Signals at 2300 MHz applied via the input loop L1 are tuned by L2, C12 and coupled to the mixer diode MR1 by L3. Also coupled to MR1 via L3 is the 2280 MHz output of the multiplier diode MR2. Intermediate frequency signals produced by MR1 are fed to the i.f. head amplifier V1. This is a cascode stage, the operating conditions of which are stabilised by components R3 to R6. Wideband transformers T2 and T1, which cover 20 to 30 MHz respectively, match the mixer diode MR1 to V1, and provide a 75 ohm output to the main receiver.
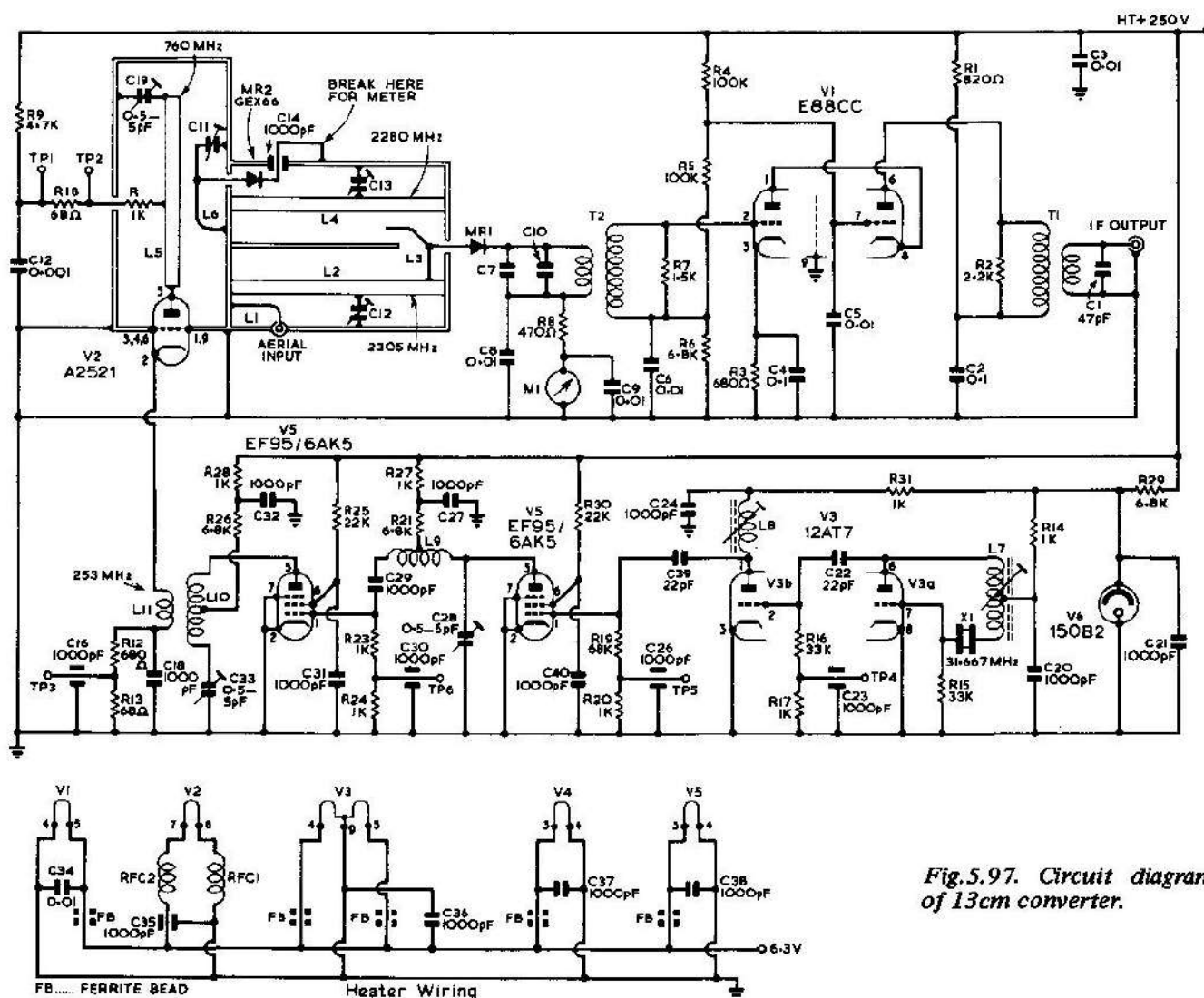
Fig.5.97. Circuit diagram of 13cm converter.

COIL DETAILS FOR 13cm CONVERTER

Coil	Tuned to MHz	
L1	–	½" 18 s.w.g. tinned copper wire spaced ⅛" from L2.
L2	2305	2.40" ¼" dia. brass rod, ends tapped 6BA, mounted ¾" clear of chassis.
L3	–	See Fig.5.99.
L4	2280	As L2.
L5	760	2.2" ¼" diameter brass rod, mounted ¼" clear from chassis, tapped approx.⅛" from anode end.
L6	–	Formed from wire end of MR2 and 18 s.w.g. wire. See Fig.5.101.
L7	31.7	25T 30 s.w.g. enam. copper wire closewound on 0.27" diam. former.
L8	63	8T 30 s.w.g. enam. copper wire closewound on 0.27" diam. former.
L9	126	10T 22 s.w.g. tinned copper wire wound on ¼" mandrel air-spaced to ¼", centre tapped.
L10	253	6T ditto¼" long, centre-tapped.
L11	–	1T 22 s.w.g. PVC covered wire wound over the centre of L10.
RFC1, RFC2		10" 24 s.w.g. enam. copper wire on ⅛" diam. mandrel, air spaced.
T1	20 to 30	primary: 7½T 34 s.w.g. enam. copper wire on 0.3" diam. former, 1.0 µH. secondary: 15T 34 s.w.g. pile-wound to occupy ⁵⁄₃₂" length wound adjacent to primary, 2.8 µH.
T2	20 to 30	primary: 18T 40 s.w.g. enam. copper wire on 0.3" diam. former, 3.9 µH. secondary: 4½T 40 s.w.g. immediately adjacent to primary, 0.43 µH.

Component details for 13cm Converter

C2, C3, C4, C5, C6, C8, C9, C21, C34, C39 0.01 µF disc ceramics

C7 value up to 10 pF to pad C10 to 51 pF (see text)

C10 capacitor formed between diode mount and trough wall. Dielectric polythene approximately 0.010″ thick selected to give capacity 40 to 51 pF.

C11 ⅜″ length ¼₆″ tube with 6BA screw inside p.t.f.e. insulation.

C14, C15, C17, C18, C20, C23, C24, C26, C30, C32, C35, 1000 pF feedthrough.

C12, C13 capacitor formed between OBA brass bolts and L2, L4.

C16, C27, C29, C31, C36, C37, C38 1000 pF disc ceramics.

C19, C28, C33 ceramic tubular trimmers 0.5 to 5 pF Erie 3116A.

R10, R13 metering resistors 10 to 100 ohms matched ± 2%.

MR1, SIM5 or CV 2155 (SIM2 and CV2154 similar but reversed polarity) MR2: GEX66.

The oscillator and earlier stages of the multiplier chain are shown in the lower part of Fig.5.97. V3A is used as a Squier oscillator, X1 being a 31.6667 MHz third overtone crystal. Stages V3B, V4 and V5 successively double to 253 MHz, and the output is fed via L10 and L11 to the troughline assembly. This assembly incorporates two further stages of multiplication, firstly, a grounded grid triode V2 tripling to 760 MHz, and secondly, a diode multiplier MR2, the 2280 MHz output of which is tuned by L4 and C13. The positive end of MR2 is decoupled through C14 and earthed with a link which may be removed for metering during aligning. Both V3A and V3B are fed from a 150v stabilised supply.
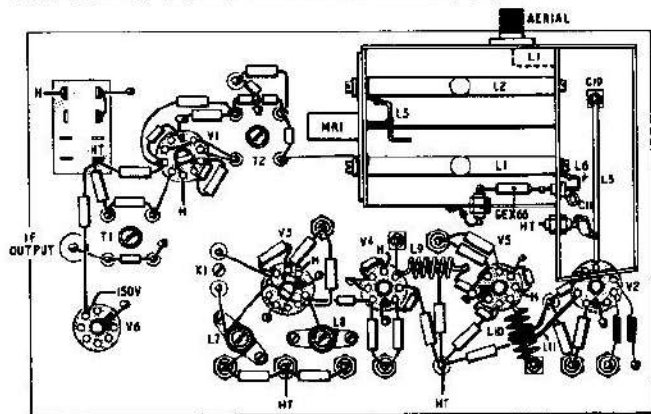
Fig.5.98. Layout of components for 13cm converter.

The general layout of the converter is illustrated by Fig.5.96 and Fig.5.98. The troughs are 1″ wide internally and 1″ deep, and are fabricated from 20 s.w.g. brass sheet. Details of the screen and diode mount are given in Fig.5.101. The trough components are temporarily jigged together using chromium plated instrument screws (which do not solder) screwed into the tapped ends of L2 and L4, and are located by a scrap valveholder in position V2. The assembly may then be soldered directly to the 18 s.w.g. brass chassis using a small gas flame. OBA brass nuts are soldered to the top of the chassis to act as threaded bearings for C12 and C13. 6BA solder tags are also soldered directly to the chassis to provide convenient earthing points, and in particular short earthing paths for the cathodes of V4 and V5 (pins 2 and 7). In connecting the valveholder for V2, pins, 3, 4 and 6 and the centre screen are soldered directly to the trough wall, and pins 1 and 9 to the solder-tag of the centre screen of the valve holder. P.t.f.e. valve holders are used, except for the stabiliser V6.
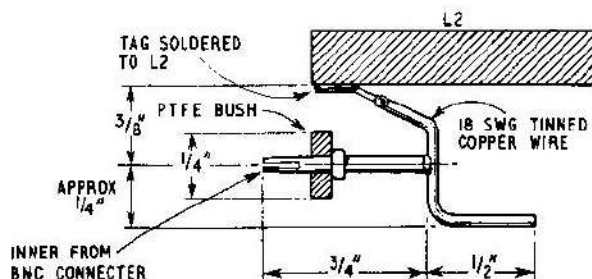
Fig. 5.99. Details of L3.
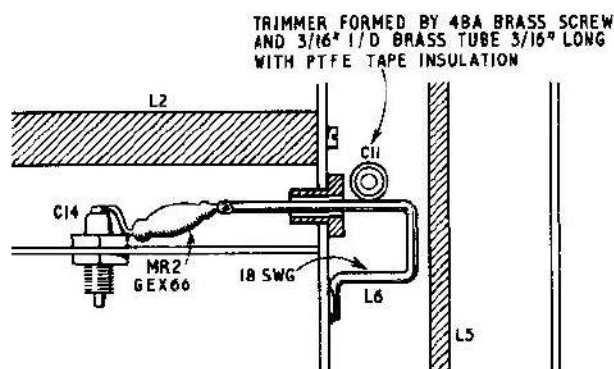
Fig.5.100. Details of L6.

The diode mount is held firmly against the trough wall by 6BA bolts screwed into the ends of L2 and L4, the bolts being insulated by p.t.f.e. or fibre bushes. Between the mount and the trough is fitted a piece of polythene sheet the thickness of which, approximately 0.010″, is chosen to produce a capacitor having a value of 40 to 51 pF. This capacitor, C10, which is both the r.f. bypass at 2300 MHz and also the tuning capacitor for the primary of T2, is padded to the value 51 pF by C7. The

connector to the mixer diode MR1 is the inner of a BNC socket, the p.t.f.e. insulation from which is also used in locating the inner within MR2. The adjustable tap on L5 consists of a connector removed from a paxolin octal valve holder.
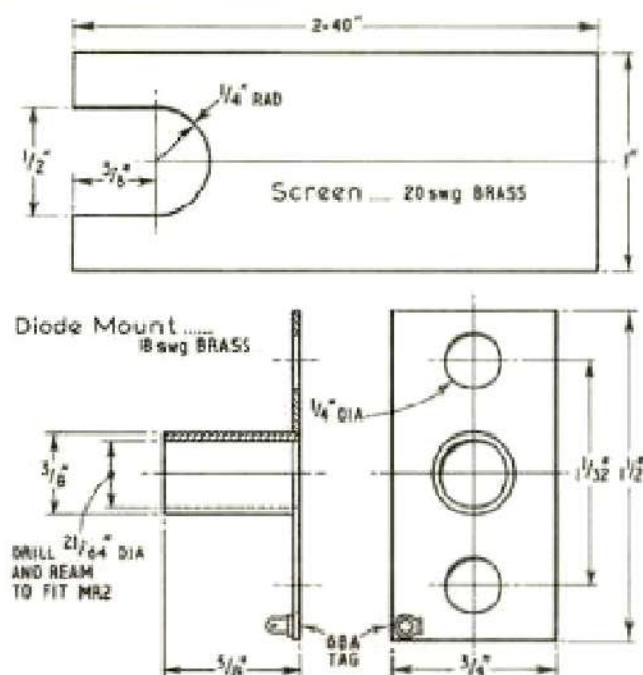
Fig.5.101. Details of screen and diode mount.

Alignment of the converter should present few problems provided that the critical dimensions of the trough circuitry i.e. the lengths of L2, L4 and L5 are within ± ¹⁄₁₆ ". With the valves plugged in, the anode circuits of V3A, V3B, V4 and V5 are tuned to the frequencies given in the table showing coil details using a g.d.o. H.t. may then be applied and the oscillator and multiplier circuits peaked by metering test points 4, 5, 6 and 3 in that order. The currents measured should be similar to those given in the table. The drive applied to the grid of V3, should be about 1mA, and is the difference between the cathode and anode currents measured from the voltage drop across the matched resistors R13 and R10 respectively. C19, C11, together with the tap on L5 and the coupling link to MR2 (L6) should then be adjusted to produce the maximum current in MR2 (up to a maximum of 40mA). The tuning point of C13 is mainly dependant on the position of MR2: that shown was found to be the optimum and resulted in the 0BA tuning bolt being unscrewed 2½ turns from touching L4. The small capacitive coupling between L3 and L4 is finally adjusted to give a mixer current in MR1 of 0.5 to 1mA. Final peaking should be done after L2 has been tuned to 2305 MHz. An approximate position for C12 is ¾ turn out from touching L2.

The current through V1 should be set to 10mA by altering R6, and T1 and T2 adjusted to give a uniform output over the range 20-30 MHz.
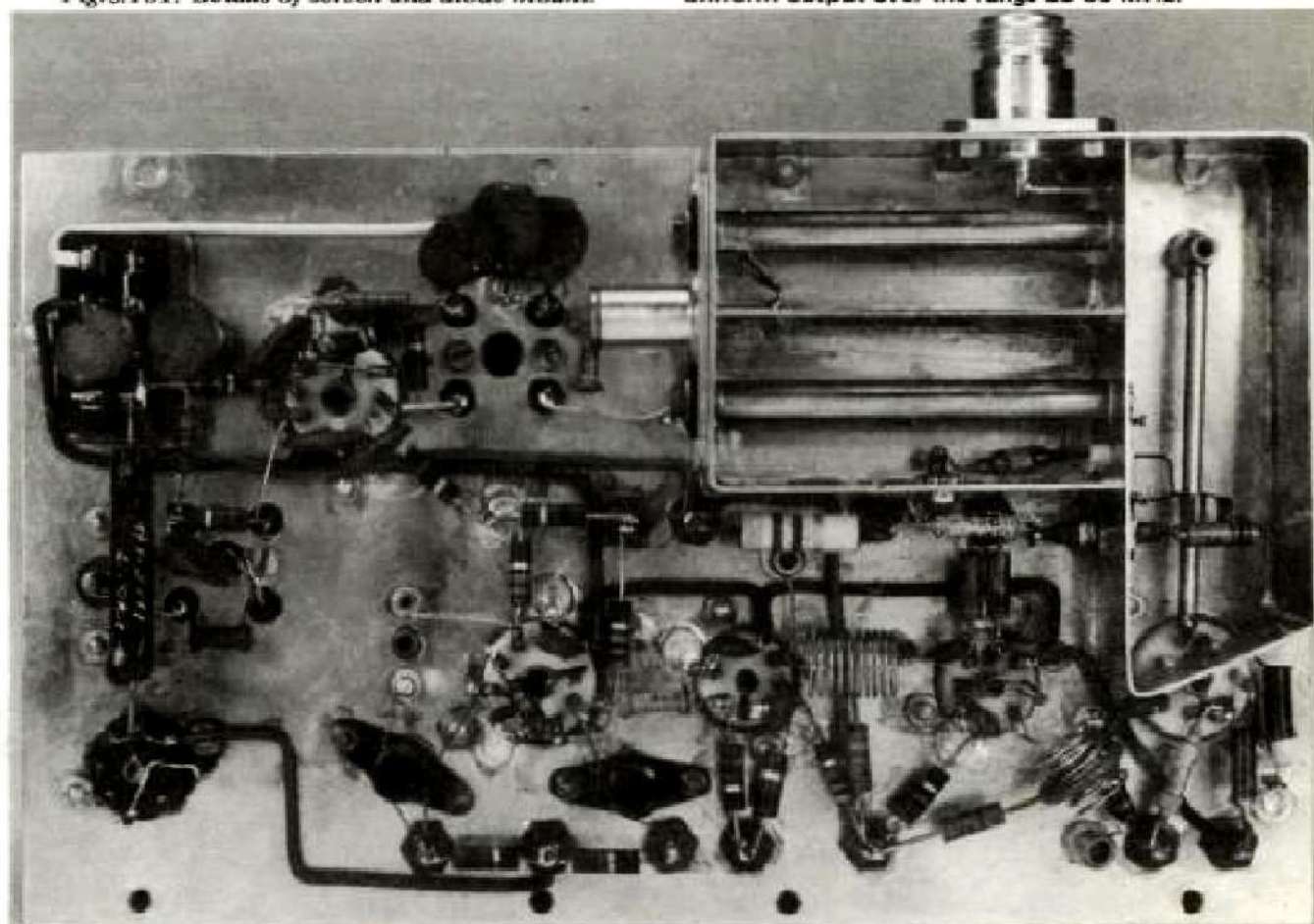
Fig.5.102. Underside view of 13cm converter.

Metering details for the 13cm converter

Meter between	Current mA
TP1/TP2	11.3
TP3/chassis	12.3
TP4 "	− 0.9
TP5 "	− 0.5
TP6 "	− 1.3
MR2 "	40
MR1	0.6

SEMICONDUCTOR DIODES AS SWITCHES

(a)

(b)
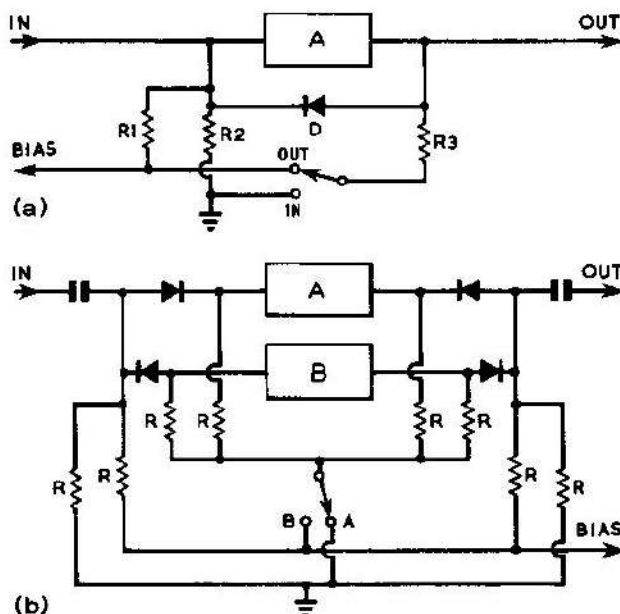
Fig.5.103. Two circuits illustrating the use of diodes as switches.

As mentioned elsewhere semiconductor diodes may be used for switching. They offer advantages over mechanically operated types because they can be fitted close to the circuit and the actuating switch or relay positioned where most convenient.

In Fig.5.103(a) the resistors R1 and R2 provide the bias voltage for the diode D, R3 connected to the other side of the diode provides r.f. isolation and allows the diode to be forward or back biassed, thereby causing the diode conductance to be either high or low, cutting out the circuit or leaving it in operation.

The actual value of the resistors is not important and values between 10K and 100K ohms are suitable. The fixed bias should be such that it is unlikely to be exceeded by the signals involved.

In Fig.5.103(b) is shown a double pole changeover, applied to two circuits. If only single pole changeover is required, then either the right or left hand half of this circuit may be used.

Diodes suitable for this type of application should have:-

(a) high forward conductance,

(b) low reverse conductance, and

(c) low capacitance.

Most types known as Gold Bonded Diodes will be suitable for use as switches.

CHAPTER 6

TRANSMITTERS

DESIGN CONSIDERATIONS

The early stages of a transmitter, often referred to as the exciter stages, are those circuits between the oscillator and the power amplifier. In these sections it is necessary to arrange for the control frequency to be multiplied to reach the final frequency and to build up the power level to a value suitable to drive the final amplifier.

This is normally done by frequency multiplying stages and including amplifiers if necessary, which are more likely to be needed at the higher frequencies due to lower efficiency as the frequency is increased.

The actual number of stages will of course, depend on the starting frequency (oscillator) and the final frequency required together with the multiplication factor used in each stage. Generally this is confined to doubling and trebling although sometimes quadrupling is used at lower frequencies.

In addition to the actual multiplier further amplifier stages may sometimes be required, for example as an isolation amplifier (buffer) between a variable frequency oscillator and the following stage to prevent frequency pulling, or as an amplifier after the final frequency multiplication to drive the power amplifier.

Decision on the oscillator and other exciter stages will be governed by the crystals or suitable variable frequency oscillator available. These two sources of frequency control may require a different starting frequency since the required stability of a variable frequency oscillator becomes more difficult to attain as the frequency is raised, whereas with crystals this is generally not a problem provided they are operated within their rating.

Some thought should also be given to the actual frequencies that are to be used, to ensure that if there is any radiation from these they do not cause interference. In this connection it is always advisable to operate exciter stages at low power level even if this

MULTIPLICATION FACTOR	4m BAND 70.025 - 70.7	2m BAND 144 - 146	70cm BAND 432 - 434*	23cm BAND 1296 - 1298*
x2	35.013 - 35.350	72 - 73		
x3	23.342 - 23.566	48 - 48.6		
x4 (x2x2)	17.507 - 17.675	36 - 36.5	108 - 108.5	
x5	14.005 - 14.140	28.8 - 29.2	86.4 - 86.8	
x6 (x2x3)	11.671 - 11.783	24 - 24.3	72 - 72.3	
x8 (x2x2x2)	8.753 - 8.837	18 - 18.25	54 - 54.25	
x9 (x3x3)	7.781 - 7.855	14.4 - 14.6	48 - 48.2	144 - 144.2
x10 (x5x2)	7.025 - 7.070	16 - 16.2	43.2 - 43.4	129.6 - 129.8
x12 (x3x2x2)	5.836 - 5.891	12 - 12.16	36 - 36.165	108 - 108.16
x16 (x2x2x2x2)		9 - 9.125	27 - 27.125	81 - 81.125
x18 (x3x3x2)		8 - 8.11	24 - 24.1	72 - 72.1
x20 (x5x2x2)		7.2 - 7.3	21.6 - 21.7	64.8 - 64.9
x24 (x3x2x2x2)		6 - 6.083	18 - 18.0825	54 - 54.08
x32 (x2x2x2x2x2)			13.5 - 13.653	40.5 - 40.563
x36 (x3x3x2x2)			12 - 12.05	36 - 36.05
x40 (x5x2x2x2)			10.8 - 10.85	32.4 - 32.45
x48 (x3x4x4)			9 - 9.04125	27 - 27.04
x64 (x4x2x2x2x2)			6.75 - 6.7813	20.25 - 20.2813

*Communication Section

makes necessary the addition of a driver/amplifier. Such an amplifier will not only provide an increase in power but also assist in attenuating in its tuned circuits unwanted harmonics generated earlier.

To assist in the selection of a starting frequency (crystal or v.f.o) the table shows the various frequencies and the multiplication needed to reach the individual bands.

From the table it will be seen that the use of a 72 MHz oscillator needs only a doubler to reach 144 MHz, a doubler and a trebler to reach 432 MHz and a doubler and two treblers to reach 1296 MHz. At this starting frequency separation of harmonics at the high frequencies is an easy matter.

5.836 - 5.891	x 3 x 2 x 2	12	70.025 - 70.7
	x 4 x 3	12	
7.781 - 7.855	x 3 x 3	9	
11.671 - 11.783	x 3 x 2	6	
	x 2 x 3	6	
6.00 - 6.6083	x 3 x 2 x 2 x 2	24	144 - 146
	x 4 x 3 x 2	24	
8.00 - 8.111	x 3 x 3 x 2	18	
9.00 - 9.22	x 2 x 2 x 2 x 2	16	
	x 4 x 2 x 2	16	
12.00 - 12.166	x 3 x 2 x 2	12	
18.00 - 18.44	x 2 x 2 x 2	8	
24.00 - 24.332	x 3 x 2	6	
36.00 - 36.48	x 2 x 2	4	
72.00 - 73	x 2	2	

Exciter and Low Power Stages

The valves most suitable for multipliers are generally those of high mutual conductance (high sensitivity). Since the grid and anode circuits are tuned to a different frequency, instability is not usually a problem provided there are no circuit conditions which could provide a tuned anode tuned grid oscillator. In some types there may be some chance of

parasitic oscillation which can usually be prevented by suitable isolation of input and output circuit components and adequate by-passing.

Pentodes are suitable for frequencies up to about 200 MHz but triodes are much more satisfactory at higher frequencies. The types generally are of the receiving valve class, such as pentodes CV138, EF80, 6AK5, E180F, 5763 or triodes 6J6, 6C4, EC8010, ECC85, 12AT7. In the performance tables typical results are given for 3 examples and these show that quite worthwhile power outputs are attainable under fairly low power input conditions.

Expensive valves, such as, E280F, E282F, E810F and D3A are pentodes specially designed for wide band applications and are not really suitable because their input and output capacitance is relatively high.

Of course, for the early stages triode pentodes such as ECF80 are suitable and can be a convenient means of reducing the equipment size which is important for portable and mobile equipment. The tuned circuits for the exciter stages, although handling relatively low power, should be designed for as high Q as possible so as to give the maximum attenuation of all unwanted frequencies, and similar methods to those proposed for power amplifiers may be used. They can however be relatively small for the lower frequency stages and up to about 100 MHz core tuned inductors are quite suitable. Above this frequency, self-supporting coils up to the frequency limit where these are permissible should be used. At the highest frequencies parallel lines, trough lines and cavities will have to be employed.

The use of link coupling is often a suitable means of coupling circuits together to make a compact layout which may otherwise be difficult mechanically.

In building circuits, the series resonance of such components as capacitors should be borne in mind and their connecting lead lengths should be as short as possible. Sometimes the self resonance is intentionally used, and the table shows some examples of self resonant frequencies of some commonly used capacitors.

SELF RESONANT FREQUENCIES OF TYPICAL CAPACITORS

Type	Maker	Lead Length	Frequency	
8 µF 450V. Electrolytic	Hunts	1 in.	0.26	MHz
0.1 µF 350V. paper foil (4702A)	Dubilier	½ in.	3.4	MHz
0.1 µF 200V. paper foil (CP45N)	T.C.C.	½ in.	3.46	MHz
0.01 µF (PZ) ceramic disc	Erie	¼ in.	17.8	MHz
0.1 µF (W99) metallized paper ...	Hunts	¼ in.	5.95	MHz
2200 pF polystyrene	G.E.C.	¼ in.	30.0	MHz
0.001 µF (W99) metallized paper	Hunts	¼ in.	43.6	MHz
0.001 µF (NY) ceramic disc... ...	Erie	¼ in.	53.0	MHz
100 pF (N750L) ceramic	Erie	½ in.	99.0	MHz
150 pF silvered mica	L.E.M.	¼ in.	114.0	MHz

EF91/6AM6 with anode voltage of 300V, screen
voltage of 250, and grid bias of − 20V.

Class of operation	f input (MHz)	f output (MHz)	Anode current (mA)	Screen current (mA)	Grid current (mA)	Power output (watts)
	60	60	11.5	3.2	1.5	1.9
Amplifier	120	120	11.5	3.2	1.7	1.7
	200	200	9.7	3.2	1.9	0.4
Frequency	30	60	10.0	3.2	1.5	1.5
Doubler	65	130	10.0	3.2	1.6	1.1
	90	180	9.0	3.2	1.8	0.25
Frequency	20	60	10.5	3.2	1.6	1.1
Tripler	30	90	10.0	3.2	1.7	0.9
	50	150	9.0	3.2	1.8	0.2

ECC91/6J6 with anode voltage of 200.

Class of operation	Frequency (MHz)	Cathode Current (mA)	Anode Current (Ia(tot) (mA)	Grid Current (Ig(tot) (mA)	Grid Resistor (Common) (K ohms)	Power Output (watts)
Amplifier (Grid bias	50	22	17.2	4.8	1.5	2.1
−14V	100	22	17.8	4.2	1.5	1.9
Cathode						
resistor	150	22	18.2	3.8	1.5	1.7
330 ohms)	200	22	18.8	3.2	2.2	1.4
	250	22	19.3	2.7	2.7	1.1
Frequency	50	22	16	6	15	0.95
Tripler	100	22	16.7	5.3	18	0.9
(Grid bias	150	22	17.2	4.8	22	0.8
−100V	200	22	17.7	4.3	22	0.7
Cathode						
resistor	250	22	18.2	3.8	27	0.6
120 ohms)						

EC 8010 with anode voltage of 200

Class of Operation	Frequency MHz	Cathode Current (mA)	Anode Current (mA)	Grid Current (mA)	Power Output (watts)
Amplifier (Grounded Grid)	432	39	29	5	3.7
Frequency Tripler from 144 MHz	432	34	30	4	2.8
Frequency Doubler from 432 MHz	864	31	30	1	1.4

R.F. Power Amplifiers

The type of power amplifier used will to a degree depend on the frequency of operation, and the efficiency will vary considerably over the range of 70 MHz to 1296 MHz.

At the low frequency end of this range the power amplifier stage may well be very similar to that used on the h.f. bands and substantially similar valves or transistors can usually be used, although recourse to series tuned circuits is often necessary because of the relatively high output capacitance.

As the frequency of operation is increased a change from the tetrode, to double tetrode (push pull), and on to grounded grid triodes is usually employed. Just where these changes occur depend on a number of factors and to some extent on personal choice.

In the table below a rough guide to the choice of tuned circuit is given and as can be seen there is not a clear cut change from one type of circuit for any particular frequency, the final decision usually depending on the actual valve or semiconductor used.

Probably the variety is greatest at 432 MHz not only in the circuit form, but also in the choice of amplifier valve. There are two popular amplifier arrangements which at the power levels usually encountered in amateur equipment, compete strongly for favour, these are the push pull double tetrode (QQVO3/20A) or the grounded grid triode (DET24). Similar power levels and efficiencies can usually be obtained with either arrangement.

In higher power stages the most popular is the forced air cooled single tetrode such as the 4X150A or 4CX250B. Two valves of this type may be used in a push-pull arrangement but the circuit becomes rather cumbersome and anyway such power levels are not generally required for amateur purposes.

Frequency	Valve/Semi-conductor	Tuned Circuit
70 MHz (4m)	Tetrode Transistor	Standard parallel or series tuned Lumped constants circuit
144 MHz (2m)	Tetrode	Series tuned
	Double Tetrode Transistor	Parallel tuned } Lumped constant circuits Linear Circuit
432 MHz (70cm)	Double Tetrode Grounded Grid Triode Varactor diode	Linear Circuit Slab Line Cavity or trough Lumped constants series tuned
1296 MHz (23cm)	Grounded Grid Triode Varactor diode	Cavity Linear circuit series tuned
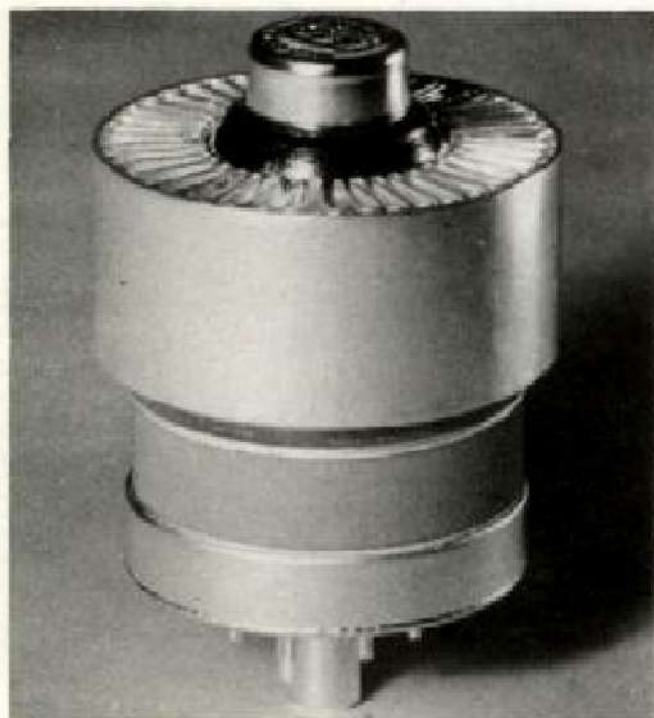
Power Frequency Doublers

Frequency multipliers, generally follow the practice normally used for the output frequency amplifier valve and circuit. Semiconductors such as varactor diodes are a simple and an entirely r.f. powered means of obtaining output at double or treble the input frequency at a high efficiency of 60% to 75%.

Although some triplers have been made in which spurious frequencies have been avoided, it should be remembered that since the operation of these devices is of the parametric type considerable care is needed not only to ensure that the input and idler frequencies are kept out of the output circuit but also that any lower unrelated frequencies and their harmonics are suitably reduced.

Suitable filters that are of narrow band width at the wanted frequency are essential. Of course, when a

Fig. 6.1. Photo of a 4CX250B.

trebler stage is driving an amplifier, much of the power of the unwanted frequencies is likely to be very much reduced by the amplifier's circuit selectivity.

Transistors

Transistors for substantial power levels at 144 MHz and above are relatively highly priced compared with valves of similar output power but they will undoubtedly become more competitive in the near future. They are attractive for portable and mobile use, but they are very sensitive to overloading and considerable care and precautions must be taken to remove or reduce the r.f. drive when the match to the output load is greater than about 2:1. This can probably be best arranged by building into the output circuit a directional coupler connected to an early

(a)

(b)

(c)

Fig. 6.2. Methods of neutralising.

drive stage in such a way that operation is in a similar manner to a.v c. or a.l.c. as used in a receiver or single sideband transmitter.

Circuit design for transistors should include some means by which parametric oscillation is avoided.

Tetrodes

The design of single tetrodes suitable for use at frequencies of 70 MHz and above, requires among other features, low anode to control grid capacitance in order to avoid instability. Most types will require neutralisation by some means or other, either by the capacity bridge method or screen tuning system.

Typical circuit arrangements for neutralisation are shown in Fig.6.2. In (a) the anode circuit is of a balanced type so that the neutralising capacitor NC is connected to the opposite side of the tuned circuit to the anode and thus forms with the valve capacitance a bridge. The NC may be either an adjustable capacitor or a suitable length of wire from the grid so placed to form a capacitor.

In (b) the out of phase voltage required is obtained from the grid circuit and is the method usually used when a pi-anode circuit is used. It is also applicable for series tuned anode circuits.

In (c) stabilisation is obtained by tuning the screen circuit in which there is a series inductance NL and tuning capacitor NC. In many cases NL is provided by the capacitor leads themselves and additional inductance is not needed.

In some specially designed types for operation in the frequency range 150 to 500 MHz, such as the 4X150A or 4CX250B, there is complete mechanical isolation between the grid and anode, the screen being brought out to an annular ring which when fitted into an appropriate socket with built-in bypass capacitor, can be thoroughly earthed with respect to r.f. The feedthrough capacitance, between grid and anode is then usually too small to cause self-oscillation in a properly designed circuit layout.

In this class of valve the screen should be supplied from a fixed voltage source interlocked with the anode supply, because the screen current varies fairly widely, due to secondary emission, and is often negative. In addition, as the frequency is increased it may be necessary to reduce the input power to the heater to prevent excessive internal heating.

Double Tetrodes

There are two distinct varieties of these valves, which may be classified as *American*, such as the 832A and 829B and *European*, such as the QQV03-20A and 6-40A.

In the American type the electrode system consists of two separate tetrodes in one envelope with or without a common screen bypass capacitor mounted inside, and neutralisation is required externally. These are now an old design of valve.
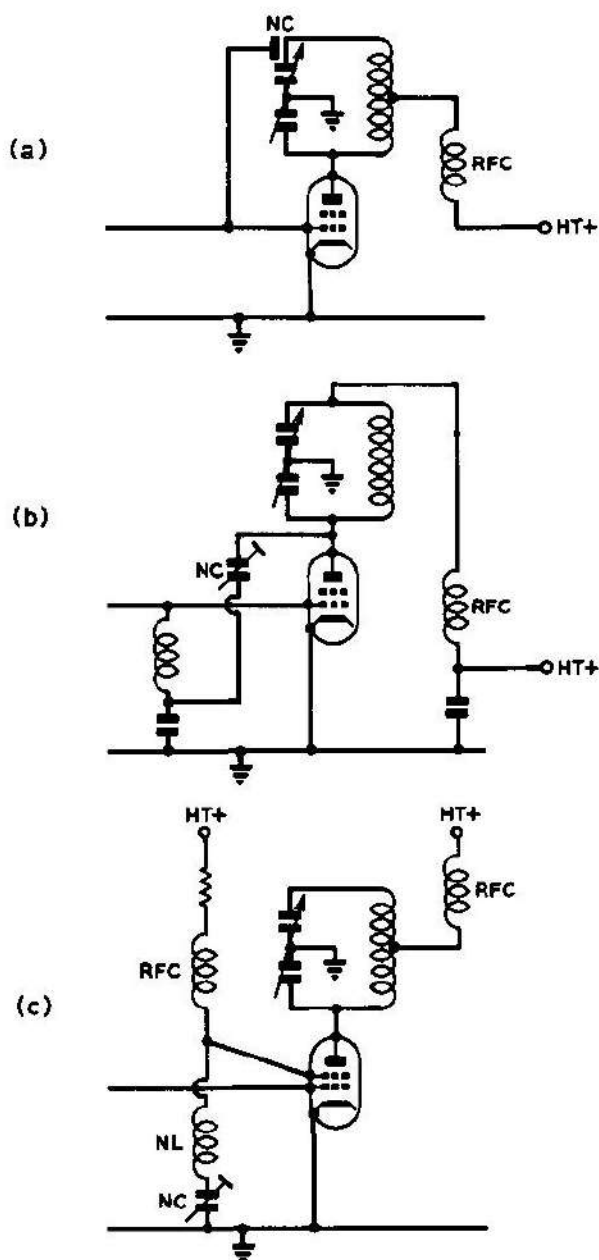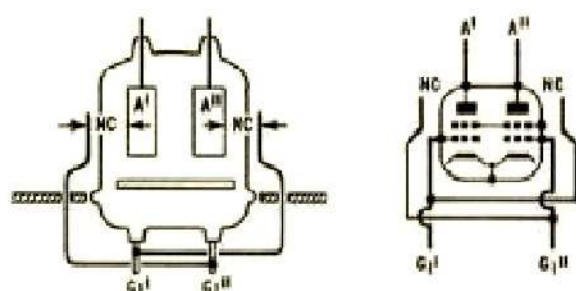
Fig. 6.3. A cross section of an 829B showing method of neutralising.

A typical method of the cross over neutralising arrangement is shown in the diagram in Fig.6.3. It usually consists of a wire connected to the grid circuit of one tetrode being placed near to the anode of the other tetrode, the capacity formed between the wire and the anode is adjusted by bending to give stability.
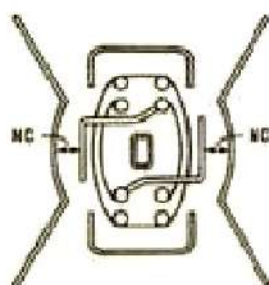
Fig. 6.4. A cross section of a European v.h.f. double tetrode such as a QQVO3-20A.

The European types such as the QQVO3-20A or QQVO6-40A are designed on the basis of a common cathode and a common screen grid. The separate systems are achieved by the use of single sided control grids and anodes. The design also includes the neutralising capacitors.

Fig. 6.5. Photo of a QQVO3-20A and QQVO6-40A.

The table below lists some of the more popular valves at present in use Their power output ranges from one watt upwards.

Make	Type No.	Base	Cathode			Limiting Values			Max. freq. full ratings (MHz)	Output at full ratings (W)	Max. freq. reduced ratings (MHz)	Output. at reduced ratings (W)
			Type	V	A	Pa	Va	Vg2				
Mullard U.S.A. CV	ECC91 6J6 858	B7G	IH	6.3	0.45	2 x 1.5	300	–	80	3.5	–	1.0 at 250 (MHz)
Mullard	EC56	Octal	IH	6.3	0.65	10	300	–	–	–	4000	0.5
Mullard	EC57	Octal	IH	6.3	0.65	10	300	–	–	–	4000	1.8
M-O Valve Mullard CV	DET22 TD03-10 273	-	IH	6.3	0.4	10	350	–	1000	2.8	3000	0.5
S.T.C. CV	33B/152M 1540	B9G	IH	6.3	0.92	15	375	–	300	28	–	–
M-O Valve Mullard CV	DET24 TD04-20 397	-	IH	6.3	1.0	20	400	–	600	23	2000	3.5
S.T.C. CV	3B/240M 2214	B8G	IH	6.3	1.1	24	375	–	200	24	–	–
M-O Valve CV	ACT22 257	-	IH	6.3	4.0	75	600	–	1000	90	–	–
Mullard CV	QQV02-6 2466	B9A	IH	6.3 12.6	0.8 0.4	2 x 3	275	200	200	6	500	5

Make	Type No.	Base	Cathode			Limiting Values			Max. freq. full ratings (MHz)	Output at full ratings (W)	Max. freq. reduced ratings (MHz)	Output at reduced ratings (W)
			Type	V	A	Pa	Va	Vg2				
Mullard CV	QV04-7 309	B9G	IH	6.3	0.6	7.5	400	250	20	7.9	150	6.3
Mullard U.S.A. CV	QQV03-10 6360 2798	B9A	IH	6.3 12.6	0.8 0.4	2 x 5	300	200	100	16	225	12.5
M-O Valve CV	TT15 415, 4046	B9G	IH	6.3	1.6	2 x 7.5	400	–	200	12	250	–
Mullard U.S.A. CV	QQV04-15 832A 788	B7A-	IH	6.3 12.6	1.6 0.9	2 x 7.5	750	250	100	26	250	18
M-O Valve Mullard U.S.A. CV	TT20 QQV03-20A 6252 2799	B7A	IH	6.3 12.6	1.3 0.65	2 x 10	600	250	200	48	600	20
Mullard U.S.A. CV	QV06-20 6146 3523	Octal	IH	6.3	1.25	20	600	250	60	52	175	25
Mullard U.S.A. CV	QQV06-40A 5894 2797	B7A	IH	6.3 12.6	1.8 0.9	2 x 10	750	250	200	90	475	60
Mullard U.S.A. CV	QQV07-40 829B 2666	B7A	IH	6.3 12.6	2.5 1.25	2 x 20	750	225	100	87	250	60
Mullard U.S.A. CV	QY3-65 4-65A 1905	B7A	DH	6.0	3.5	65	3000	400	50	280	220	110
Mullard U.S.A. CV	QY3-125 4-125A 2130	B5F	DH	5.0	6.5	125	3000	400	120	375	200	225
Mullard S.T.C. U.S.A. CV	QV1-150A 4X150A 4X150A 2519	B8F special	IH	6.0	2.6	150	1250	200	165	195	500	140
M-O Valve ST&C U.S.A. CV	4CX250B 4CX250B 4CX250B 2487	B8F special	IH	6.0	2.6	250	1500	300	175	235	500	225 (at 2kV)
ST&C U.S.A. CV	2C39A 2C39A 2516	- -	IH	6.3	1	100	1000	–	2400	20	–	–
M-O Valve CV	DET29 2397	-	IH	6.3	0.5	10	450 1000 (pk)	–	2300	3.5 (2300)	5000	1.5 (3800)

6.7

Driving Power

In Class C amplifiers, the driver should always be capable of providing sufficient power to allow for the range of operation required and to overcome the control grid circuit losses. This is particularly important when the amplifiers are operated in the v.h.f. and u.h.f. region where allowance must be made for circuit and radiation losses in addition to transit time losses within the valve itself.

With increased drive power requirements as the frequency is increased care should be exercised to ensure that excessive drive is not used. The r.f. drive voltage should not be greater than that required to give the maximum output and this can readily be seen if a power control is available in the driver stage and an r.f. output monitor connected into the output circuit. This may take the form of a simple r.f. voltmeter connected across the output coupling or a reflectometer set to read forward power.

Overloading

All oxide coated cathodes can fairly easily be damaged if the anode and grid are allowed to become excessively hot. The anode, for example, should never be allowed to become hot enough for it to be visibly red when viewed in normal room lighting.

In amplifiers operating within their rated conditions overheating is most likely to occur if there is a failure of r.f. drive when the grid bias is wholly obtained by the grid resistance method. It is wise to include a suitable cathode bias resistor to provide some part of the bias and its value should be chosen such that in the absence of r.f. drive the anode and screen currents are kept down to safe levels. Such cathode bias resistor must be by-passed effectively to maintain the cathode at substantially earth potential. Alternatively a clamp valve system may be used.

Cooling

Operation of any radiation cooled power valve at its maximum rating of input power, especially at or near its upper frequency limits, invariably requires an increase in air cooling to prevent overheating of the envelope. Simple fans which will provide the necessary additional air flow over the static convection air flow should be switched on by being interconnected with the power supply switch.

The extra cooling is particularly necessary in amplifiers which are fully screened, as they should be to avoid any appreciable losses introduced by free r.f. radiation. Adequate air inlet and outlet holes should be provided to ensure a free flow of air into and out from the enclosure.

Modulation of Double Tetrodes

Satisfactory modulation may be applied to the grid, screen or anode circuits, but for most purposes anode and screen modulation is normally preferred for the amplitude method. Frequency modulation may, of course, be used if desired and this form does offer operation at full rating, whereas a.m. requires suitable derating.
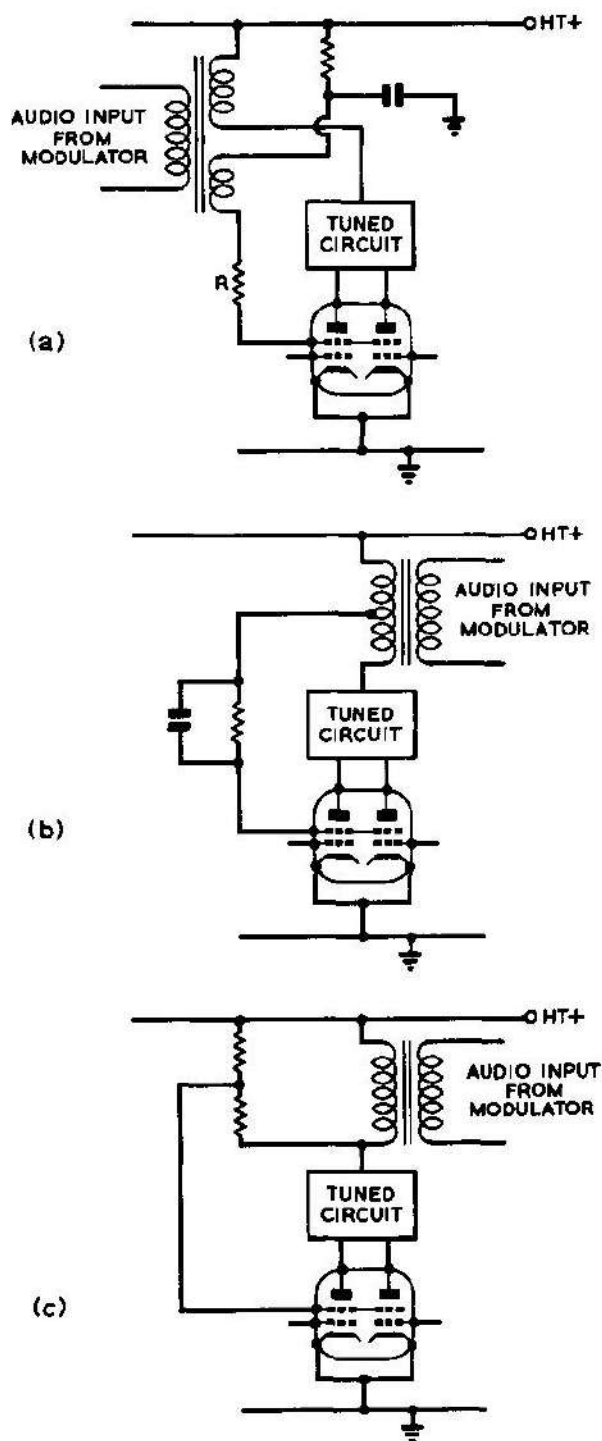
Fig. 6.6 Methods of applying modulation to the screen of a double tetrode.

Amplitude modulation can be applied to the anode by any of the usual methods and must also be applied to the screen by one of the following methods:

1. Series resistor from the modulated anode voltage.
2. Separate winding on the modulation transformer, with series resistor to the anode h.t. voltage as shown in diagram (a). The screen voltage dropping resistor must be by-passed to audio frequencies on the supply side. R is an r.f. stopper resistor.
3. A tapped modulation transformer may be used, in this case the screen dropper which must be connected in series with the tap must be by-passed to audio frequencies as shown in diagram (b).
4. In diagram (c) a resistive potential divider is placed across a standard modulation transformer, the values of the upper and lower resistor must be chosen to provide the proper screen voltage.
5. Fig.6.7 shows a further method of applying the correct voltage to the screen, together with the appropriate amount of modulation.

It should be remembered that for optimum operation the screen voltage is critical, and can never remain reasonably constant when supplied from a series resistor, since any change in anode current results in a change in screen current.
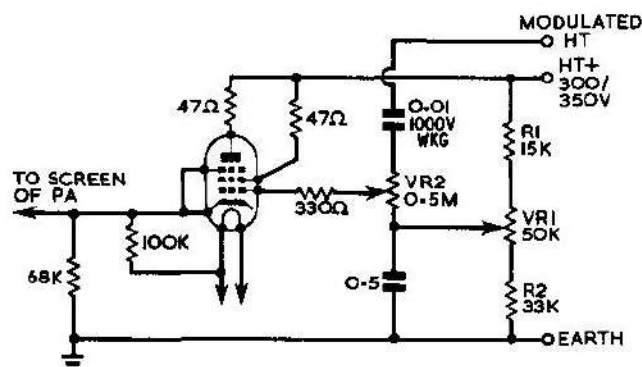
Fig. 6.7. Circuit showing a superior method of supplying both screen voltage and modulation. Valve may be a 6V6, 6CH6, 5763 or similar type.

The circuit of Fig.6.7 not only enables a definite value of screen voltage to be set, but also allows the correct amount of audio to be applied.

In operation VR1 controls the value of the screen voltage and the values of R1 and R2 must be chosen so that VR1 gives an adequate range. VR2 controls the amount of modulation applied to the grid of the cathode follower valve and hence to the screen grid. It should be adjusted so that the p.a. anode current meter does not move either up or down under modulation.

The only disadvantage of this method is that the heater supply to the valve must be from a separate source, since the heater - cathode insulation is usually insufficient for the voltages involved.

Whatever form is used it should be remembered that at peaks of modulation the anode and screen voltage is almost doubled and the driver must provide sufficient power under this condition. Modulation of the driver is sometimes useful, taking the audio from the screen grid of the amplifier.

Triodes

At frequencies above 432 MHz, triodes are more satisfactory as amplifiers and even at 432 MHz they offer a good alternative to the tetrodes especially in the power output range of 10 to 20 watts.
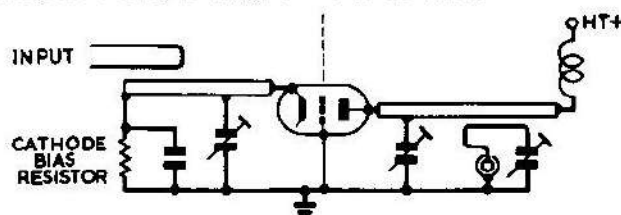
Fig. 6.8. A typical grounded grid amplifier.

The usual method of connection is grounded (earthed) grid and when valves of the planar type are used, the isolation between grid and anode is good enough to enable stable operation without trouble as long as the cathode (input) and anode (output) circuits are suitably separated. The drive power required is significantly higher than for a tetrode, but a considerable proportion of this is fed through into the output circuit and adds to the amplifier's total output power.

Since the drive power forms part of the output, it is necessary to modulate the driver stage as well as the amplifier, but because of this they cannot be driven from a frequency multiplier.

The best known valves for this type of use are the DET24 (20 watt dissipation) and 2C39A (100 watt dissipation). These two valves are very similar, except that the former relies on conduction cooling of the anode whereas the higher powered type makes use of forced air cooling. In addition, they are both suitable for use as triplers from 432 MHz to 1296 MHz.

In addition to the valve types mentioned above, there are a number of smaller types which are suitable either as drivers or lower power amplifiers. Such types are DET22, DET29, A2521, EC157, EC8010 and Nuvistor 8058 and of these both the EC157 and DET29 may be used at higher frequencies than 1296 MHz and with reasonable output up to about 4 or 5 GHz.

TRANSISTORS

General Considerations

The rapid developments in the semiconductor field have provided a number of relatively high pow-

ered transistors for use up to 432 MHz and although they are still expensive they should be considered carefully because their voltage requirements may well offer an overall economic benefit, especially for mobile or portable equipment.

There are, however, a number of important points which must be borne in mind when considering these:

(a) The output impedance is very low, and the tuning circuit must match this to the much higher aerial or output impedance.

(b) Overload protection must be provided to avoid destroying the device. Unlike valves which can accept considerable overloads without permanent damage transistor amplifiers should be protected by a fast acting protection circuit.

(c) Because of their sensitiveness to overload, when they are used as amplitude modulated amplifiers the de-rating is very considerable, so that frequency modulation is to be preferred, thus enabling a higher power level to be used.

(d) The output wave form must be substantially free from spikes which could cause over-loading.

(e) The circuit must be such that there is no parametric oscillation.

(f) The output circuit should not present a mismatch greater than a v.s.w.r. of 2:1 and the

protective circuit should be such as to reduce the r.f. drive as the mis-match increases, in a manner similar to a.g.c. or a.l.c.

(g) Adequate heat sink must be provided to keep down the temperature of the device and in high power stages some advantage may be obtained by using forced air cooling.

In considering a transistor power amplifier probably the most important feature which must be borne in mind is that of the impedance of the device compared with the established valve amplifier. The device impedance will be low and it is necessary to tap it well down the tuned circuit both for the input and output stages.

The use of pi-coupling is not always satisfactory owing to the large input capacitor that is needed due to the low impedance output of the transistor. This impedance may be between 1 and 5 ohms, and therefore to match this into the aerial impedance of 50 or 70 ohms the coupler will be very different from that used for a valve amplifier.

Usually it will be found more convenient to use a parallel tuned circuit and tap the transistor and the output connections on the inductor at suitable impedance points.
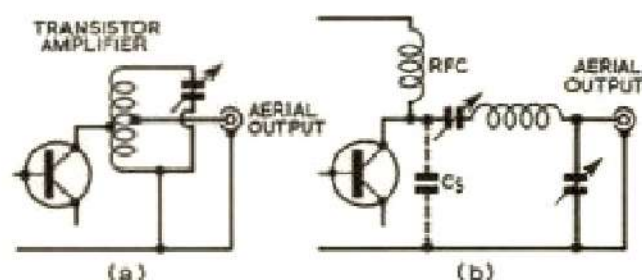
Fig.6.10. Two transistor amplifier circuits. (a) shows a parallel tuned circuit and (b) an L-circuit. Cs is a spike reducing capacitor.

Alternatively an L-circuit may be used, which in effect is a pi circuit, the input capacitance being that of the transistor amplifier itself. In this case however it is desirable to have a small amount of capacity at the input to keep down the development of spike voltages which may occur. It is always desirable that some check measurements are made to ensure that the maximum working voltage of the device is not exceeded.

Some typical transistor output circuits are illustrated in Fig.6.11.

For power amplifiers generally, high voltage types are likely to be the most useful, but as with most semiconductor devices, they are sensitive to excess voltages, and, unlike valves which can accept high voltage surges, often of considerable magnitude, without suffering permanent damage, must be protected if they are to be operated at near their maximum rating.

Because of the derating that is necessary when using amplitude modulation, it is worth considering
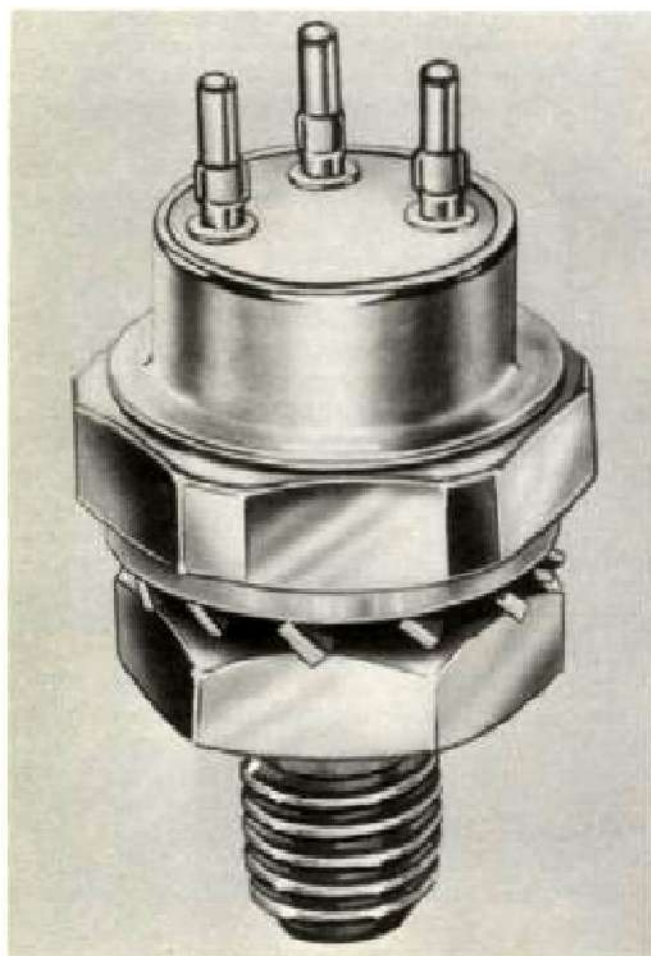
Fig. 6.9. A type BLY35 transistor.
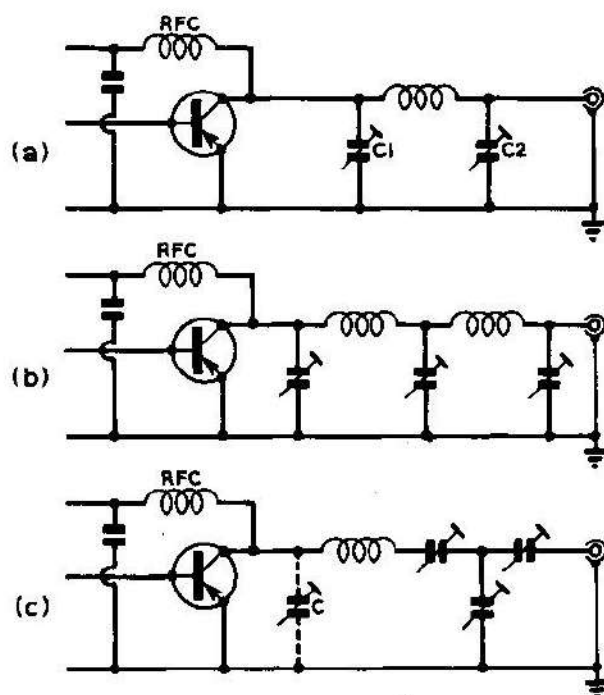
Fig. 6.11. (a) Simple Pi circuit. C1 should be quite small in order to prevent the development of spikes in the wave form as this prevents good impedance matching being achieved.

(b) In this circuit the impedance matching is done in two stages - say up to 5−15 ohms in the first section and 15−50 ohms in the second section. Better harmonic attenuation is also achieved.

(c) Preferred circuit for h.f. operation. A modified L circuit is generally more satisfactory than either of the previous two. Capacitance to prevent spikes is desirable.

the use of alternative modulation methods, such as n.b.f.m. or single sideband. In the former case full c.w. ratings can be used.

Apart from voltage surges and spikes, as with a valve the transistor amplifier should not be operated into a severe mismatch, the valve will usually survive but the transistor will not, and therefore some protective circuit should be incorporated which will reduce the r.f. drive under fault conditions. If the fault conditions are for any reason likely to produce a high v.s.w.r. in the output circuit the protective arrangement should be such as to be able to remove the drive or reduce it to zero.

Adequate heatsinks must be provided, especially in amplifiers of significant power rating and it may be necessary to provide forced cooling in order to keep the overall amplifier size down to reasonable overall dimensions.

The power gain of transistors, is at present, generally significantly lower than their valve equivalents and it may be necessary to have two or more stages at the operating frequency, but this in itself does not present any particular problem other than the additional circuits.

Since semiconductors are "transparent" devices and a transistor may operate on unwanted modes it is desirable to include a narrowband filter in the output circuit. This will at least reduce the chance of radiating spurious signals.

TRANSISTOR POWER AMPLIFIERS

The problem of obtaining the maximum transference of power from a transistor is essentially the same as with a valve, but special problems arise with the transistor because of its inherently lower impedance.

Whereas with a valve it is necessary to step down the impedance in order to match it into the feeder line, in the case of the transistor we have to step up to the feeder impedance and this involves components of vastly different values.

Unfortunately, component values computed by standard methods are often physically unattainable, especially at v.h.f. and u.h.f. and this makes necessary the use of unusual coupling networks in order to allow the use of more practical components for the impedance transformation.

Another problem that must be faced in transistor power amplifiers is the effects of off tune and excess loading which are most likely to occur during initial setting up but may occur during operation for one reason or another. In high impedance circuits (valve) the effects are usually adequately taken care of by the inherent overload capacity that most valves have, and generally only poor efficiency results.

In transistors this will not be the case, off tune and higher load conditions will present to the transistor a greater equivalent series impedance than does the correctly tuned and loaded circuit. The result of this is that the load line will cut the Ic Vc curves below the knee and in this region linear operation is impossible so that parasitics and other instabilities are most probable.

In most arrangements the circuits are a.c. coupled to the output and the collector is then fed with d.c. through an r.f. choke. Here again the valve practice of using a high Q component will not serve and the low Q type of suitable size must be used. The impedance required effectively to isolate the collector from the supply does not need to be greater than ten times the load presented to it by a correctly matched network.

Also it must be remembered when designing (and constructing) a transistor power amplifier that the emitter must be connected by a low impedance lead; 1" wide copper strip is useful for this purpose.

L-Pi Circuit

From much recent work the most satisfactory coupling network which lends itself to fairly straightforward calculation consists of a combination

of the series tuned "L" circuit and the conventional Pi circuit. In order to establish the component values it is necessary to calculate the component values in two parts. The first part being for the L section and then the Pi section components are added.
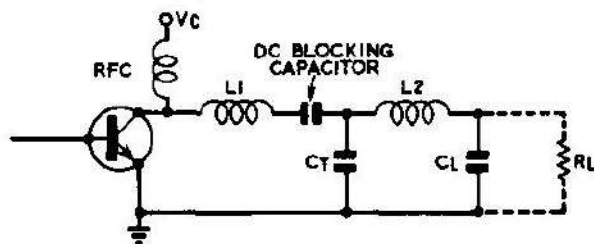
Fig. 6.12. The L-Pi circuit. CT is main tuning capacitor. CL loading capacitor. L1 is the network inductance and L2 the Pi-network inductance. RL is the load resistance and Vc the supply voltage.

The two parameters which must be determined initially are the large signal output capacitance of the transistor and its equivalent parallel resistance.

The output capacitance Cp can be obtained from the manufacturers data sheet and is usually quoted in the form of capacitance/frequency curves for various power levels.
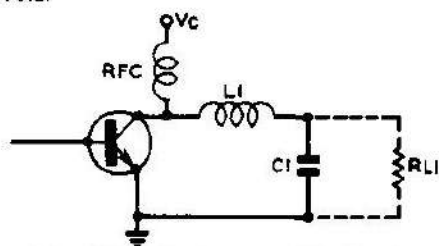
Fig. 6.13. Series tuned L circuit.

The output resistance Rp is not given but can be computed with sufficient accuracy by assuming a peak to peak r.f. voltage swing of twice the supply voltage Vc.

If P is the mean power output, then the equivalent parallel resistance of the transistor $Rp = \dfrac{Vc^2}{2P}$

The parallel capacitance Cp and resistance Rp are now converted into equivalent series circuits, see Fig. 6.15. These are:

Equivalent series resistance $Rs = \dfrac{Rp\,Xp}{Rp^2 + Xp^2}.Xp$

The equivalent series capacitive reactance Xs is:

$$Xs = \dfrac{Rp.Xp}{Rp^2 + Xp^2}.Rp$$

where $Xp = \dfrac{1}{\omega\,Cp}$

and $Xs = \dfrac{1}{\omega\,Cs}$
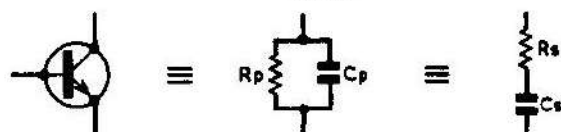
Fig. 6.14. Equivalent circuit.

The series impedance of the device, Zs, is therefore equal to Rs + jXs. For maximum power transfer to a load, the load impedance must be the conjugate of the source impedance, or Rs − jXs.
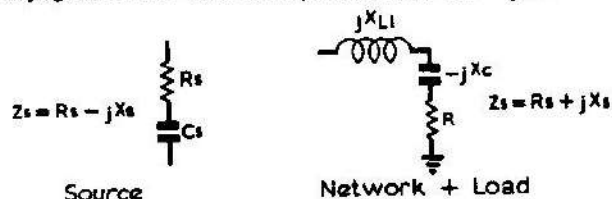
Fig. 6.15. Showing how the value of the load impedance is derived.

Since it is desirable that the network should provide good harmonic rejection and ease of tuning, a working Q of between 8 and 20 should be chosen as being suitable for use at v.h.f.

Since XL1 and Xc (Fig.6.15) may have a range of values, the desired loaded Q of the network may be obtained by a choice of the inductance L1.

The formula relating the inductive reactance, XL, the series resistance r, and Q is,

$$Q = \dfrac{XL}{R} \quad \text{and} \quad XL_1 = Q.Rs$$

the impedance of the source plus the inductive reactance XL1 is Rs + j (XL1 - Xs) thus, the impedance of R-jXc must be the conjugate to establish a match.

From this, by equating the real and imaginary parts

$$R = Rs$$

and $$Xc = (XL1 - Xs)$$

The values of C1 and RL1 in the L network may now be calculated by making the series to parallel conversion:

$$RL_1 = \dfrac{XC^2 + Rs^2}{Rs} \,.\, XC_1 = \dfrac{XC^2 + Rs^2}{Xc} \,,\, C1 = \dfrac{1}{\omega\,Xc1}$$

To complete the design of the whole network it only remains to calculate the pi circuit components from the formula given in the RSGB *Radio Communication Handbook*, these are:

$$XC_2 = \dfrac{R1}{Q}\left(1 + \sqrt{\dfrac{R2}{R1}}\right)$$

$$XC_3 = Xc_2\sqrt{\dfrac{R2}{R1}}$$

$$XL_2 = \dfrac{R1}{Q}\left(1 + \sqrt{\dfrac{R2}{R1}}\right)^2$$

$$C_2 = \dfrac{1}{\omega Xc_2} \,,\, C_3 = \dfrac{1}{\omega Xc_3} \,,\, L_2 = \dfrac{X_2}{\omega}$$
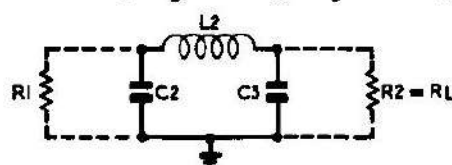
Fig. 6.16. Equivalent Pi-circuit.

Thus the values of the capacitive reactance, XC2 and XC3 and the inductive reactancce XL2, may be obtained by making R1 = RL and R2 = the load RL.

The two sections of the whole network (L + Pi) are connected together by lumping C1 of the 'L' circuit with C2 of the pi circuit in parallel to form the tuning capacitor CT, C3 is then the loading capacitor CL.

Note that the Q chosen for the pi section need not be the same as that used for the L section. Improved harmonic rejection will result with higher values of Q.

Example

'L-Pi' network design for a single 2N3632 transistor operating at 13.5 watts output c.w. at a frequency of 144 MHz into a 72 ohm resistive load. Supply voltage 28 volts.

From the manufacturer's data sheet ascertain the parallel equivalent output capacity Cp at the stated power and frequency. Cp = 22pF.

Parallel equivalent output resistance, Rp at stated power

$$Rp = \frac{Vc^2}{2.P} = \frac{28^2}{2.13.5} = 29 \text{ ohms}$$

Reactance of Cp $= \frac{1}{\omega Cp} = \frac{1}{2\pi f.Cp}$

where f = 144 MHz
and $\omega = 2\pi f = 9.1 \times 10^8$ thus,

$$Xp = \frac{10^{12}}{9.1 \times 10^8 \times 22} = 50 \text{ ohms}$$

By parallel to series conversion,

$$Rs = \frac{Rp.Xp}{Rp^2 + Xp^2} \times Xp = \frac{29 \times 50}{29^2 + 50^2} \times 50 = 22 \text{ ohms}$$

$$Xs = \frac{Rp.Xp}{Rp^2 + Xp^2} \times Rp = \frac{29 \times 50}{29^2 + 50^2} \times 29 = 12.7 \text{ ohms}$$

For a Q of 10, to find reactance of L_1
$$XL_1 = Q.Rs = 10 \times 22 = 220 \text{ ohms}$$
$$\therefore L_1 = 0.24 \mu H$$

From Xc = XL_1 - Xs = 220 - 12.7 = 207.3 ohms

To obtain the values of C_1 and RL_1 (Fig.6.13) the series combination of Xc and Rs must be converted to the parallel equivalent.

From the formula $Xc_1 = \frac{Xc^2 + Rs^2}{Xc}$

$$= \frac{2.07^2 \times 10^4 + 2.2^2 \times 10^2}{2.07 \times 10^2} = 208 \text{ ohms}$$

from which $C_1 = \frac{1}{\omega Xc_1} = \frac{10^{12}}{9.1 \times 10^8 \times 208} = 5.3 \text{ pF}$

and $\frac{RL_1 = Xc^2 + Rs^2}{Rs} = \frac{2.07^2 \times 10^4 + 2.2^2 \times 10^2}{2.2 \times 10}$

$$= 1.97 \text{k ohms}$$

This completes the L network calculations and the Pi network values must now be calculated. From the formula, given in the RSGB *Radio Communication Handbook* and using R1 = 1.97k ohms, R2 = 72 ohms with a selected Q of 15.

$$Xc_2 = \frac{R_1}{Q}\left(1 + \sqrt{\frac{R2}{R1}}\right) = \frac{1.97 \times 10^3}{15}$$

$$\left(1 + \sqrt{\frac{72}{1.97 \times 10^3}}\right) = 156 \text{ ohms}$$

from which C2 $= \frac{1}{\omega Xc_2} = \frac{10^{12}}{9.1 \times 10^8 \times 156} = 7.1 \text{ pF}$

$$Xc_3 = Xc_2 \sqrt{\frac{R_2}{R_1}} = 156 \sqrt{\frac{72}{1.97 \times 10^3}} = 29.7 \text{ ohms}$$

from which $C_3 = \frac{1}{\omega Xc_3} = \frac{10^{12}}{9.1 \times 10^8 \times 29.7} = 37 \text{ pF}$

$$XL_2 = \frac{R_1}{Q}\left(1 + \left(\sqrt{\frac{R_2}{R_1}}\right)\right)^2$$

$$= \frac{1.97 \times 10^3}{15}\left(1 + \sqrt{\frac{72}{1.97 \times 10^3}}\right)^2 = 186 \text{ ohms}$$

from which L2 = 0.204 μH

Combining C_1 and C_2
CT = $C_1 + C_2$ = 5.3 + 7.1 = 12.4 pF
The final L-Pi network becomes
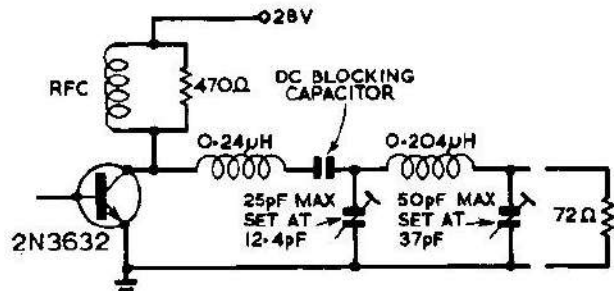
Fig. 6.17. The final L-Pi-network.

Similarly for two 2N3632 transistors in parallel operating at 20 watts on 144 MHz the following values can be calculated:
Cp (for pair) 44 pF   Rp = 19.6 ohms
Xs = 9.5 ohms and Rs = 12.1 ohms
XL_1 = 182 ohms and L1 = 0.2 μH for Q = 15
Xc = 172.5 ohms
Xc_1 = 174 ohms and C1 = 6.35 pF. RL_1 = 2.5K ohms (for Q = 20 in pi-circuit).
C_2 = 7.5 pF
C_3 = 43 pF
L_2 = 0.19 μH

VARACTOR MULTIPLIERS

The theory of Varactor operation being unfamiliar to many, it is worthwhile having a brief

look at how they work. Fig.6.18 shows a p-n junction formed by butting a piece of type n semiconductor material, that is, material doped so that it has spare electrons available, and a piece of p type material, which has spare "holes" (missing electrons). This is the familiar semiconductor diode. Spare electrons near the junction in the n region are attracted into the p region where, after wandering about for a while, they combine with holes and are neutralized; conversely some of the holes near the junction on the p side find their way to the n region and combine with electrons there. The volume of material near the junction is thus swept clear of holes and electrons, and forms a sort of no-man's land - the depletion layer where to all intents and purposes there are no charge carriers. No net current flows to the outside world, and the whole thing behaves like a capacitor with the depletion layer as the di-electric between the conducting p and n region "plates". It is a slightly charged capacitor, because the charge separation process produces a space charge or "contact potential" across the junction. This is about 0.5 volt for silicon.
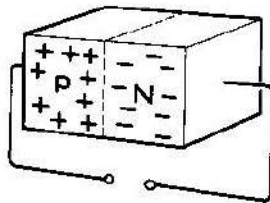
Fig. 6.18. A junction of p type and n type semiconductor materials.

If we connect an external d.c. supply in such a sense as to aid this space charge, i.e. we reverse-bias the junction, then the increased electric field sweeps more carriers out of the bulk material and increases the width of the depletion layer. Thus the junction capacitance is reduced, just as when the plates of a parallel plate capacitor are pulled apart. Conversely, reducing the space charge by forward bias decreases the width of the depletion layer and increases its capacitance, until finally the depletion layer width becomes zero, the plates in effect, touching and the device conducts in the forward direction. This, of course, happens at about 0.5 volt for a silicon diode.
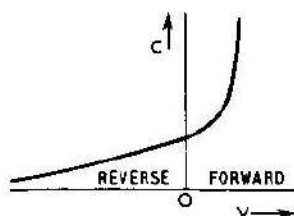
Fig. 6.19. A typical plot of capacitance against applied voltage for a semi-conductor junction.

Thus the junction diode behaves, when reverse biased, like a voltage variable capacitor: the limits to the capacitance swing are set by reverse breakdown of

the junction at one end, and forward conduction at the other. Actually, at v.h.f. Varactor diodes may be driven quite hard into the forward region without appreciable current flowing, because the applied voltage reverses again and sweeps most of the injected carriers back to the side from which they started, before they have time to recombine on the other side. This results in a large increase in capacitance swing and hence power handling capacity.
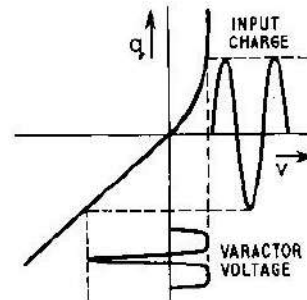
Fig. 6.20. The curve of junction stored charge plotted against voltage being non-linear, severely distorts a sinewave input charge, thus producing harmonics.

If we now plot capacitance against voltage for a semiconductor junction we obtain a curve similar to that shown in Fig.6.19, although the exact shape depends on the way that the doping levels vary near the junction.

The charge on a capacitor is given by the expression $q = CV$, and Fig.6.20 shows a plot of charge, q, against voltage. If we superimpose an input sinewave of charge on the curve, we see that the resultant Varactor voltage waveform is spiky - in other words, it contains harmonics. This is the way in which the "classical" Varactor produces frequency multiplication, but there is another important mechanism which deserves a brief mention. It was said above that if a diode is driven for a short time into the forward region, when the applied voltage reverses against those carriers which were injected from one side of the junction into the other will not have had time to combine with opposite charges there, and will return to their side of origin appearing as a brief pulse of reverse current (see Fig. 6.21). This recovered
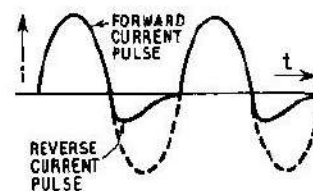
Fig. 6.21. The reverse current over-shoot of a normal p-n junction.

charge which is a nuisance in ordinary diode circuit applications, is useful in Varactors as we have already seen, as it represents a welcome increase in effective capacitance swing. However, if the impurity doping

levels near the junction are suitably profiled, it can be arranged that the stored charge all comes back in one lump, giving a very fast cut-off of reverse current, as in Fig.6.22. This phenomenon, known as "step recovery" results in the production of a great deal of extra useful harmonic energy, and is exploited in some modern Varactor diodes to increase the harmonic output power available.
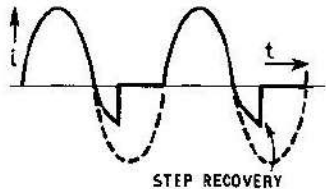
Fig. 6.22. If the semiconductor layers are suitably doped, the reverse current pulse can be tailored to produce an extremely rapid recovery, thus improving the harmonic output.

In order to make use of these effects (harmonic generation by non-linear capacitance and by step recovery) in a practical circuit, it is necessary to arrange to feed current into the Varactor at the input frequency, and to extract the harmonic energy generated in some way. Varactors may be used in either series or shunt circuit configurations, but the shunt arrangement is more convenient for high power use as it allows one side of the device to be solidly connected to a good heat sink - usually the chassis. Filters are needed to isolate the input and output, and these may simply be series tuned circuits, suitably tapped to provide impedance matching between the source, Varactor and load.
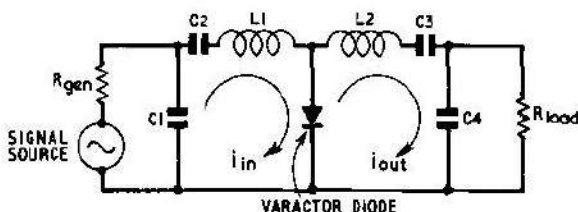
Fig. 6.23. A basic varactor doubler circuit.

Thus, we arrive at the basic circuit of Fig.6.23 in which C1, C2 and L1 and the diode capacitance form a tuned circuit series resonant at the input frequency, and C3, C4, L2 and the diode capacitance (usually quoted at two-thirds of the breakdown voltage value for this purpose) in series resonance at the output frequency. With the addition of a bias resistor across the Varactor to induce a little forward current to flow on peaks of forward drive, Fig.6.23 would form a perfectly good frequency doubler. For higher order multiplication, greatest efficiency is obtained by short-circuiting the Varactor at the second harmonic frequency with an "idler" circuit, and thus allowing second harmonic currents to flow in the diode, where mixing occurs with the fundamental to produce sum and difference frequencies. In this way, tripling,

quadrupling, and so on can be achieved, although at progressively lower efficiencies owing to idler circuit losses.
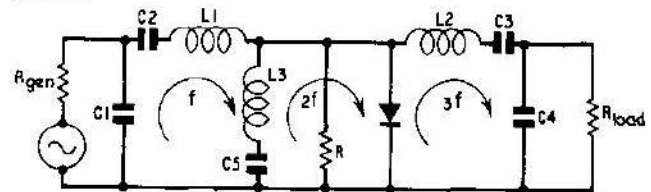
Fig.6.24. The circuit of Fig. 6.23. with two refinements: a bias resistor (R) to improve the doubling performance, and an idler circuit tuned to the second harmonic, to facilitate higher order multiplication.

Because of the harmonic relationship between the amateur v.h.f./u.h.f. bands, we are mostly interested in tripling, and Fig.6.24 shows a typical tripler circuit. This is developed from Fig.6.23 simply by the addition of an idler circuit comprising L3 and C5, series resonant with the diode capacitance at twice the input frequency. The bias resistor R is not critical and is usually chosen between 68-200 K ohms. Efficiencies from 60 to 80 per cent are attainable when tripling from 144 to 432 MHz with this circuit, using devices such as the Mullard BAY 66 (12 watts maximum input), BAY 96 (40 watts) or the Motorola IN4387 (40 watts), but it is not quite suitable, as it stands, for connection direct to an aerial, because the single tuned output circuit does not give sufficient attenuation of the fundamental and unwanted harmonic signals appearing across the Varactor. It is, therefore, necessary to provide at least a double tuned output circuit, and at 70cm this can conveniently consist of a stub in a coaxial cavity, with a series tuned input loop taking the place of L2 and C3 in Fig.6.24. With proper adjustment, spurious signals will be reduced to about 40dB below carrier, but if the odd milliwatt of spurious emission causes interference to local 2m stations, an additional high-Q break may be used.

Practical Varactor Power Multipliers for 70cm
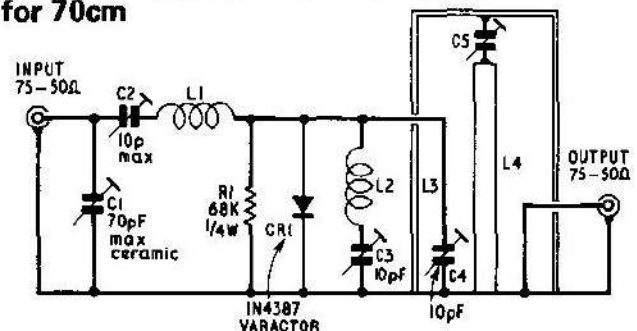
Fig. 6.25. The circuit of the tripler. C1, 70pF max. ceramic; C2, 10pF max.; C3, C4, 10pF beehive; C5, two 1" diam. discs, approx. ⅛" apart; D1, IN4387 Varactor diode; L1, 3½ turns 16 s.w.g. ½" diam. (0.2 uH); L2, 2 turns, 16 s.w.g. ¼" diam. (0.06 uH); L3, 2", 16 s.w.g. L4, 3¼", ¾" o.d. brass tube; R1, 68 K ohms ¼ watt.

Construction and Alignment

The complete circuit of the multiplier is shown in Fig.6.25 and by now should need no further explanation. The unit is built in a copper box 6" long by 1½" square, brazed or soldered up from 16 to 18 s.w.g. sheet. A partition 2½" from one end forms two compartments, the Varactor with its input and idler tuned circuits being mounted in the smaller one, while the longer compartment forms the output cavity. Fig. 6.26 shows the layout and leading dimensions, but these are not critical provided that the output cavity dimensions are not varied too much, consequently a detailed metalwork drawing is not given. A 4½" x 3½" x 2" Eddystone diecast box with the cavity formed by a copper partition would undoubtedly work just as well.

Setting up the multiplier requires, in addition to a source of r.f. at 2m, an absorption wavemeter to cover 288 and 432MHz, and a good dummy load or r.f. power meter. With the load connected to the Varactor output, apply about 10 watts of r.f. to the input and peak up the input circuit (C2, L1 and C1) for maximum absorbed power, using the 2m transmitter p.a. anode current meter as an indicator. Next set the wavemeter to 288 MHz, and bring a probe, loosely coupled to it, near to the junction of L2, C3 and adjust C3 for maximum idler voltage at this point (i.e. L2, C3 in series resonance). These first two steps can be carried out more rapidly if a g.d.o. is used initially to set the input and idler to the correct frequencies.

set the idler to series resonance by the method suggested above, it is best left alone, and subsequent alignment confined to the input and output circuits. If a reflectometer is available, it may be used to adjust C1 for optimum match to the feeder cable, again at the proper power. In practice, adjustment of C1 so that the 2m transmitter p.a. does not detune as the coupling loop is brought in, seems to be satisfactory.

After a short period of operation at full power, check the Varactor stud temperature. At the maximum allowable r.f. input level of 40 watts, the 1N4387 will produce about 25 watts of r.f. output and will, therefore be dissipating 15 watts as heat; at this dissipation the maximum permissible stud temperature is 100°C. The box shown in Fig.6.26, if made of 16 s.w.g. copper, will have a thermal resistance of about 5°C/watt in free air, and at 15 watts will therefore hold the Varactor stud at 100°C in an ambient of 25°C. This, of course, is on the limit of the device rating, and is uncomfortably hot in any case, so it is recommended that the tripler unit be mounted to a thick copper or brass bracket bolted directly on the Varactor stud.

The power levels quoted above are c.w., and in fact the Varactor may limit, due to reverse voltage breakdown, at a power level slightly above the manufacturer's rated maximum. This means that although the 1N4387 or BAY96 will produce 25 watts of c.w. power, the peak power level attainable will not be much more than this. With the 1N4387 it is necessary to reduce the input power from 40 to 20
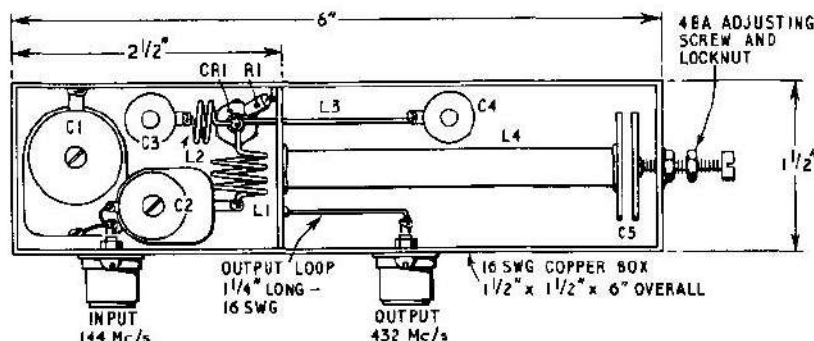
Fig.6.26. Layout and principal dimensions of the 70cm Varactor tripler.

Next, peak up C4 and C5 for maximum r.f. out, using the wavemeter to confirm that this is on 432 MHz and not 576 MHz. With the cavity dimensions shown, the plates of C5 will be about ¼" apart. Now increase the power level in steps, readjusting the trimmers each time because of the changing Varactor capacitance which forms part of the input, idler and first output tuned circuits. It will be found that correct adjustment of the idler circuit produces a peak in the output power, but this is not a reliable way of adjusting the idler since a number of spurious peaks will occur if C3 is varied over its range, owing to interaction between the various tuned circuits. Having

watts in order to produce reasonably symmetrical 100 per cent a.m. This diode, which is a graded-junction step recovery device, will, in fact, multiply an a.m. signal very well with good linearity, but the abrupt junction types, such as the BAY66 detune more rapidly with changes of drive level, and would probably be slightly less satisfactory for a.m. use.

Phase or frequency modulation is cheaper to achieve and more suitable (because of voltage breakdown and hence peak power limitations) for Varactor and transistor transmitters. Furthermore, when it is fully appreciated by amateurs that future transistor field day transmitters provide a choice of 8

watts f.m. or 2 watts a.m. at 70cm for example, f.m. may become a little more popular. If it is necessary to use a.m. with a Varactor tripler, it is advisable to tune it up with a 100 per cent sinewave modulated signal at the normal power level while watching the demodulated envelope on an oscilloscope. Look for peak clipping, and for discontinuities in the scope pattern due to dynamic detuning as the applied power varies, over the modulation cycle. These can be eliminated by adjusting the input power level and finding a compromise setting for the input tuned circuit. These complications are unnecessary, of course, with n.b.f.m.

An Alternative Design
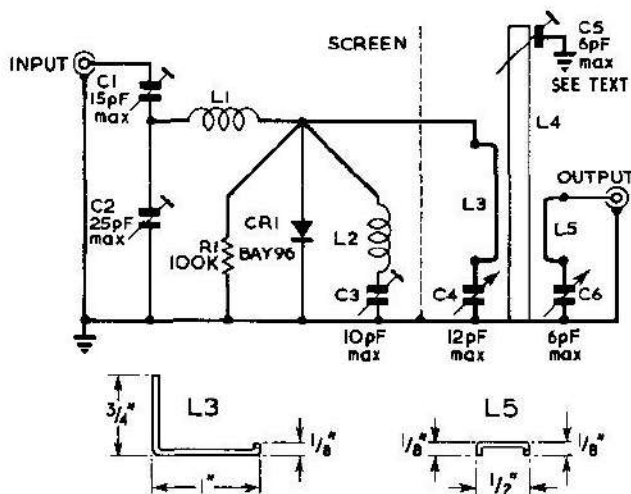
Fig.6.27. The 70cm tripler circuit. L1, 6 turns, 18 s.w.g.¼" diam., ½" long; L2, 3 turns, 14 s.w.g. ¾" diam. ¼" long; L3, 18 s.w.g., shaped as shown, and spaced ¾₂" from L4, ¼" o.d. ¾₆" i.d. copper tube 4½" long; L5, 18 s.w.g., as drawing, spaced ¾₂" from L4.

Another type to that just described is based on a standard diecast box. The circuit is very similar, except that the high Q filter consists of a tuned quarter wave line, shown as L4 in the diagram (Fig.6.27). The Varactor used is a BAY96 and with 15
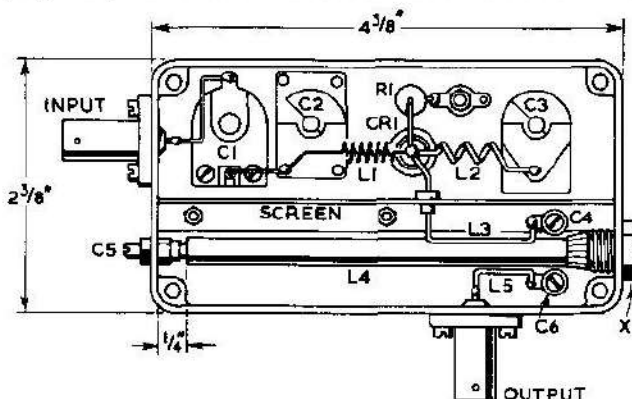
Fig.6.28. Layout of the tripler in the diecast box. The diode is bolted directly to the base. CR1 shows position of the diode.

watts input, 9 watts output can be expected when properly adjusted.

Good quality amplitude speech can be passed through this multiplier provided the peak percentage modulation of the drive is held to about 80% maximum. The procedure for setting up and alignment is the same as for the previous unit.

MODULATION

There are many methods of impressing speech or other audio frequencies on an r.f. carrier. Those normally used are:-

Amplitude modulation - in which the audio frequency varies the amplitude of the r.f. output.

Frequency modulation - in this case the actual frequency of the carrier is varied at the audio frequency.

Single sideband - this is a special type of amplitude modulation where the r.f. carrier is not radiated, nor is the unwanted sideband but the carrier has to be re-inserted at the receiver.

Amplitude modulation is the most widely used method in amateur communication on the v.h.f. and u.h.f. bands. It is probably the easiest system to operate but it does have a strong disadvantage that it is relatively expensive both in terms of power consumed, as well as size and weight of the equipment.

In addition, valves and semiconductors have to be derated to allow for the peak voltage swings that must occur, and this is particularly important in semiconductors because of their sensitivity to over voltage which can permanently damage them. The valve can sustain substantial over voltage temporarily without being permanently damaged.

No matter what system is employed, the process of modulation produces additional frequencies above and below that of the carrier wave. Thus, the modulated carrier wave consists of a band of frequencies as distinct from the single frequency of the carrier. The bandwidth depends upon the modulation system and the frequency of the modulating signals.

The bands of frequencies produced above and below the carrier wave frequency by complex modulating signals (i.e. composed of many different frequencies as in speech or music) are known as the upper and lower sidebands respectively. In the case of amplitude modulation the highest sideband frequency is equal to the sum of the carrier frequency and the highest modulation frequency, similarly the lowest sideband frequency is the difference between the carrier frequency and the highest modulation frequency. Thus the total bandwidth occupied is equal to twice the highest frequency in the modulating signal. For example, if the highest frequency in the modulating signal is 15 kHz and the carrier frequency is 1000 kHz. The sideband will extend from 1015 kHz

to 985 kHz. The total bandwidth occupied is therefore 30 kHz.

Modulation Depth

The amplitude-modulated wave is shown graphically in Fig.6.29. Here (a) represents the unmodulated carrier wave of constant amplitude and frequency which when modulated by the audio-frequency wave (b) acquires a varying amplitude as shown at (c). This is the modulated carrier wave, and the two curved lines touching the crests of the modulated carrier wave constitute the modulation envelope. The modulation amplitude is represented by either x or y (which in most cases can be assumed to be equal) and the ratio of this to the amplitude of the unmodulated carrier wave is known as the modulation depth or modulation factor. This ratio may also be expressed as a percentage. When the amplitude of the modulating signal is increased as at (d), the condition (e) is reached where the negative peak of the modulating signal has reduced the amplitude of the carrier to zero, while the positive peak increased the carrier amplitude to twice the unmodulated value. This represents 100 per cent modulation, or a modulation factor of 1.
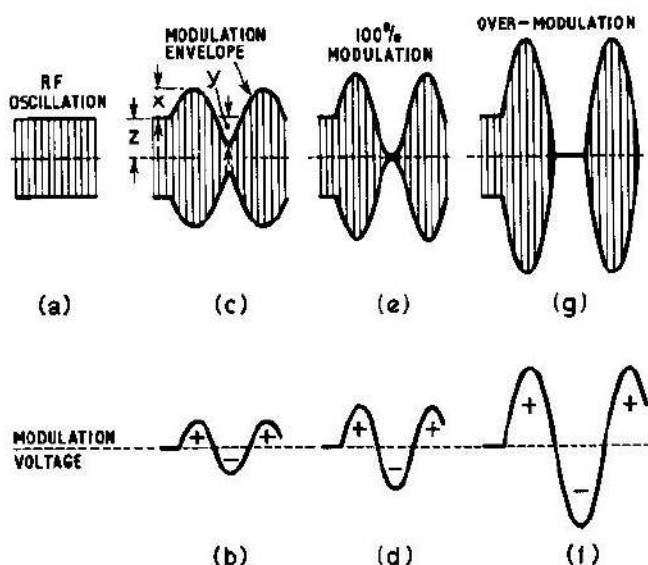
Fig. 6.29. Graphical representation of an amplitude modulated wave.

Further increase of the modulating signal amplitude as indicated by (f) produces the condition (g) where the carrier wave is reduced to zero for an appreciable period by the negative peaks of the modulating signal. This condition is known as over-modulation. The breaking up of the carrier in this way causes distortion and the introduction of harmonics of the modulating frequencies which will be radiated as spurious sidebands; this causes the transmission to occupy a much greater bandwidth than necessary, and considerable interference is likely

to be experienced in nearby receivers. The radiation of such spurious sidebands by over-modulation (sometimes known as splatter or spitch) must be avoided at all costs.

Modulation Power

In the special case of a sinusoidal modulating signal corresponding to a single pure tone, it can be proved mathematically that the effective power in such a wave at 100 per cent modulation is 1.5 times the unmodulated carrier power. Thus, in order to modulate the carrier fully with a sinusoidal wave, the average power in it must be increased by 50 per cent. This extra power must be supplied from the modulator. For example, to modulate fully a radio frequency stage operating with a d.c. power input of 150 watts, the amount of audio frequency power required would be 75 watts.

It must not be assumed, however, that the aerial current of a fully modulated transmission will increase by 50 per cent. The relationship between the modulated and unmodulated aerial current for sine wave modulation is given by -

$$Im = Io \sqrt{1 + \frac{m^2}{2}}$$

where   Im = r.m.s. value of modulated aerial current

           Io = r.m.s. value of unmodulated aerial current

           m = modulation factor

Thus, for 100 per cent modulation by a sinusoidal signal -

$$Im = Io \sqrt{1 + \tfrac{1}{2}}$$
$$= 1.226 \, Io$$

In other words, the aerial current will increase by 22.6 per cent.

The position is somewhat different when the modulating signal consists of the "peaky" waveform of speech. Assuming that the peaks drive the transmitter into full modulation, the 22.6 per cent increase will occur at these peaks, but for most of the remainder of the time the modulation depth is much lower; the increase in aerial current will also be much lower.

Effect of Modulation on Aerial Current

Depth of modulation (per cent)	Ratio: a.f. power: d.c. power	Increase in aerial current (per cent)
100	0.5	22.6
90	0.405	18.5
80	0.32	15.1
70	0.245	11.5
60	0.18	8.6
50	0.125	6.0

The average modulation depth when the peaks fully modulate will be of the order of 30 per cent, and the average increase in aerial current as seen on a typical ammeter will then be only a few per cent.

Modulation depth may also be expressed in terms of the a.f. power actually supplied and the unmodulated d.c. power supply to the modulated stage; thus -

$$m = \sqrt{2A/W}$$

where
A = a.f. power supplied,
W = d.c. input power.

The table gives the values, calculated from the expression, for the amount of a.f. power required for various depths of modulation. It will be seen that to produce 70 percent modulation requires only one half of the a.f. power required for 100 per cent modulation. The corresponding increases in aerial current are shown in the same table.

Linearity of Modulation

Ideally, for all modulation depths up to 100 per cent the difference in the amplitudes of the r.f. output between the crests and the troughs of the modulation involved should be proportional to the amplitude of the modulating signal, i.e. the modulation characteristic should be linear.
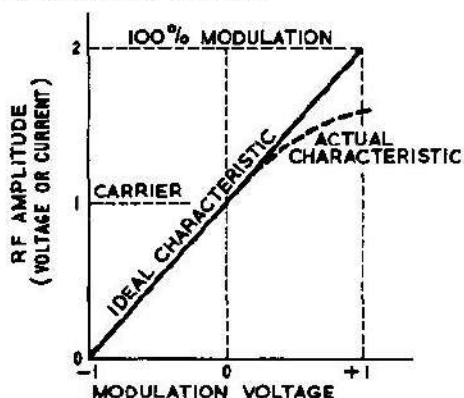
Fig.6.30. The ideal modulation characteristic is a straight line. Often the actual characteristic is found to be non-linear, and the form of non-linearity shown here is due to the failure of the r.f. voltage (or current) amplitude to rise to twice its normal value at the positive peaks of the modulating signal.

Non-linearity is most often manifest as a flattening of the crests of the modulation waveform, and this causes considerable distortion. It may be minimized by careful design and correct adjustment of the modulated stage, particularly with respect to the amount of r.f. grid drive and the aerial loading. Linear and non-linear modulation characteristics are shown in Fig.6.30.

Bandwidth of a Modulated Wave

For the faithful reproduction of speech and music it is necessary to transmit frequencies in the whole range of the audio spectrum (i.e. approximately 40-15,000 hertz per second). The total bandwidth for this purpose would therefore be 30 kHz. For a communication system, however, it is the intelligibility and not the fidelity which is of prime importance, and experience has shown that for the intelligible transmission of speech it is sufficient to transmit frequencies up to about 2.5 or 3 kHz. Thus, the transmitted bandwidth need not exceed approximately 5 kHz. In the overcrowded conditions of the present day amateur bands it is obviously important to ensure that no transmission occupies a greater bandwidth than is necessary for intelligible communication.

Modulating Impedance

The impedance that an r.f. stage which is being modulated presents to the source of the modulating signal, i.e. the modulator, is called *modulating impedance*. It is the ratio of the anode voltage and anode current of the r.f. stage or

$$Zm = \frac{Va}{Ia} \times 1000 \text{ ohms}$$

where
Zm = modulating impedance
Va = anode voltage of r.f. stage
Ia = anode current of r.f. stage (in mA)

Mathematical Representation of Sidebands

The mathematical equation for a carrier wave of constant frequency which is amplitude-modulated by a signal of constant frequency is -

$$e = E_0 (1 + m.\sin 2\pi f_m.t) \sin 2\pi fc.t$$

where
m = modulation factor
fm = frequency of modulating signal
fc = frequency of carrier wave
E_0 = amplitude of unmodulated carrier

This equation may be expanded, giving -

$$e = E_0 \sin 2\pi fc.t + m\frac{E_0}{2} \cos 2\pi (fc - fm)$$
$$t - m \frac{E_0}{2} \cos 2\pi (fc + fm)t$$

Inspection of this expanded form shows that it is made up of three separate terms. The first, $E_0 \sin 2\pi fc.t$, represents original carrier, while $m \frac{E_0}{2} \cos 2\pi (fc-fm)t$ and $m \frac{E_0}{2} \cos 2\pi (fc + fm)t$ correspond to the lower and upper sideband frequencies respectively which are the result of applying a modulating signal of frequency fm. The total bandwidth of this amplitude modulated wave is $(fc + fm) - (fc - fm)$ or $2fm$; i.e. the bandwidth is equal to twice the modulating frequency.

It should be noted from the last equation that the carrier wave is not fundamentally essential to communication since all the intelligence is contained

in the sidebands. The carrier wave can therefore be suppressed and need not be transmitted; indeed it is sufficient to transmit only one of the sidebands. This method is often used and is known as single sideband operation (s.s.b.). Although it requires more complicated equipment than ordinary amplitude modulation, it has the obvious advantage of transmitting the same intelligence within a smaller bandwidth and without a carrier.

Typical Circuit Arrangements

For simplicity the diagrams show a single tetrode but the arrangements apply equally to the double tetrode or triode r.f. amplifiers. Anode and screen modulation of a tetrode is shown diagramatically in Fig.6.31 (a) and (b). In (a) the modulator is coupled to the r.f. power amplifier by a modulation transformer which should either be of the

correct ratio to match the relative impedances or a multi ratio type so that the correct impedance match can be chosen.

The screen supply with its dropper resistor R may be connected to anode at point A or to an appropriate tap on the modulation transformer or choke at point B.

In (b) the use of a simple choke is shown, as a means of impressing the modulation on the r.f. amplifier. In this arrangement to obtain 100% modulation it is necessary to reduce the voltage applied to the modulated amplifier with respect to the modulator. This is done by inserting Rx in the feed line and Cx is provided to by-pass the audio round the resistor.

The screen of the amplifier may be connected to the point A or a tapping point B on the choke. This method is generally only used for low power transmitters.

The modulation transformer has to match the impedance of the amplifier to the specified modulator load impedance, and that of the amplifier may be found by:

$$\text{Amplifier Impedance } Za = \frac{Va}{Ia} \times 1000 \text{ ohms}$$

where   Va = d.c. anode voltage
        Ia = d.c. anode current in mA.

Screen Modulation

The efficiency of screen modulation is lower than anode or anode/screen methods, the main drawback being that the steady voltage must be reduced so that under maximum excursions it does not exceed the maximum rating for the valve. The modulation depth cannot reach 100% and generally 75-80% is the maximum that can be achieved.

The modulation power however is very low and only a few watts are required for maximum modulation.
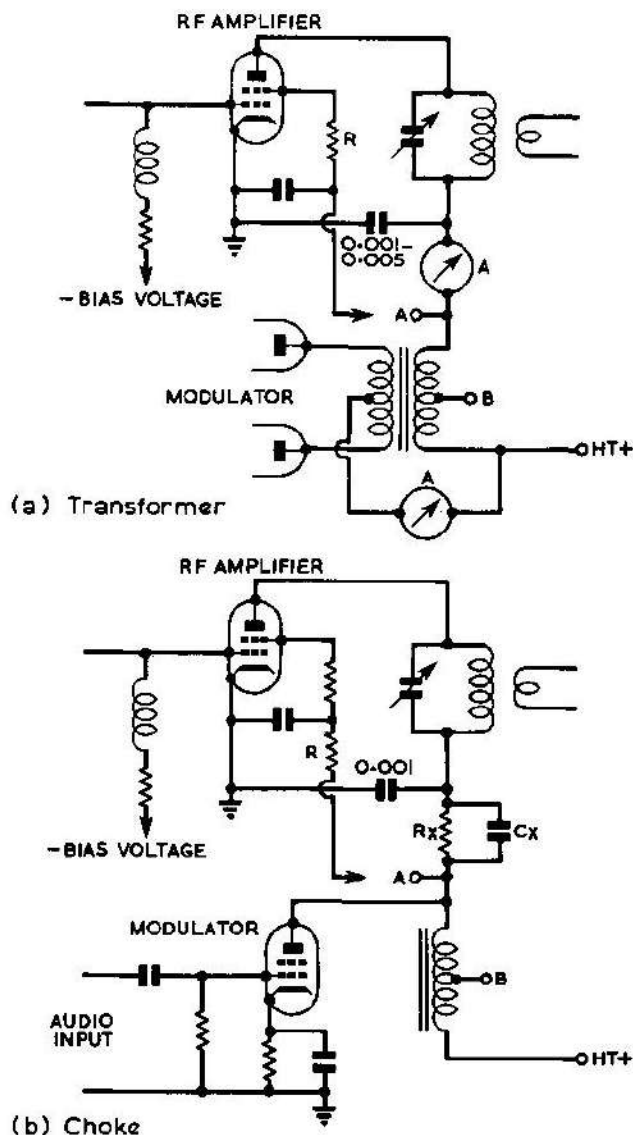
Fig.6.31. Anode and screen modulation, (a) Transformer method (b) choke method.
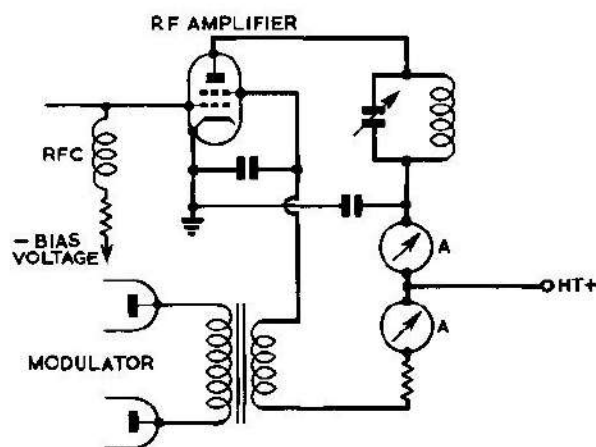
Fig.6.32. Screen modulation.

Screen (Clamp Valve) Modulation

When a clamper valve is used to limit the input to an amplifier in the absence of drive, such as may occur under keying conditions for c.w. this can be used for modulation, by switching the control grid of the clamper valve from the grid circuit of the amplifier to the source of audio input from a speech amplifier
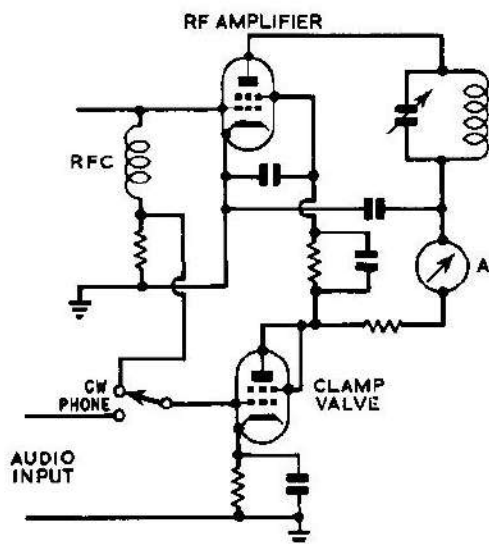
Fig. 6.33. Screen modulation using a clamp valve.

Screen (Series Gate) Modulation

Probably the most satisfactory method of screen modulation - that known as series gate modulation - is both simple and effective. The advantages of this method are:-

(a) The standing power amplifier anode current can be set to any desired level by the variable resistor VR1 in the cathode of the first section of the double triode series gate valve.

(b) Over-modulation on positive peaks cannot occur because the audio input voltage applied to the first section of the double triode will cause limiting and thus the voltage applied to the amplifier screen cannot rise too high.

(c) Splatter caused by break up of the carrier due to over modulation on negative peaks cannot occur because the screen voltage, which is set by the position of VR1, cannot fall below this value. A simple and effective means of speech clipping is available by suitable adjustment of the audio input voltage by the speech amplifier gain control and the standing amplifier screen voltage as set by VR1.

(d) It can be applied to existing c.w. or n.b.f.m. transmitters and to s.s.b. linear amplifiers, with carrier inserted, when an a.m. signal is required.
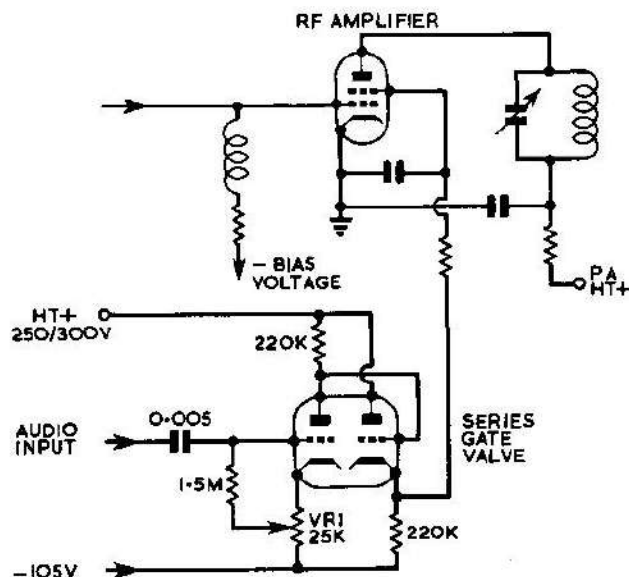
Fig. 6.34. Series gate modulation.

With the circuit shown, the connections of the cathodes of the series gate valve are returned to a negative voltage. This is necessary to enable the full screen voltage of the amplifier to be taken from the cathode of the second triode of the series gate valve, because of the voltage drop by the valve itself. The most satisfactory double triode for this use is the 12BH7.

An alternative arrangement would be to use a higher h.t. voltage and to use a lower impedance valve such as a triode pentode in which the pentode section is connected as a triode. A valve such as the ECL82 or ECL86 would enable a much higher voltage output available from the cathode to feed the amplifier screen, without the need for the negative voltage.
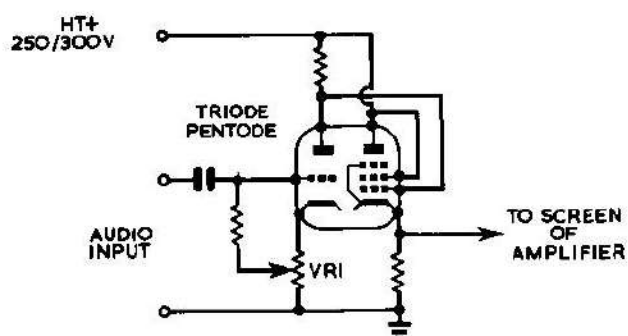
Fig. 6.35. Circuit of a series gate modulator using a triode pentode valve with separate cathodes.

The chief disadvantage of the series gate circuit is that due to the relatively high voltage output from the cathode of the control valve, it is desirable to supply the heater from an isolated supply in order to avoid voltage breakdown between heater and cathode.

A Practical Series Gate Modulator for a QQVO6-40A Power Amplifier

The Series Gate Modulator for a QQVO6-40A double tetrode power amplifier is comprised of two valves only and is shown in Fig.6.36. V1a is a straightforward a.f. voltage amplifier feeding V1b, this stage operating as a d.c. control valve and a.f. voltage amplifier directly coupled to V2 which acts as a cathode follower feeding the screen of the power amplifier.
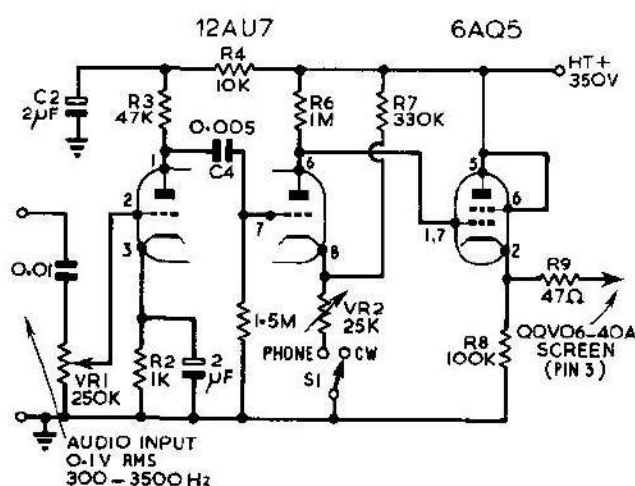
Fig.6.36. Circuit of a series gate modulator for a QQVO6-40A valve.

To appreciate the way the circuit works it is best to consider it in two ways, the d.c. control function and the a.f. modulating function. The d.c. control is effected by V1b as follows. With no a.f. input to the modulator and with VR2 set to minimum resistance, V1b has no bias and therefore a large voltage drop occurs across R6, thus the anode voltage of V1b is reduced to a very low value, approximately +45V. As the grid of V2 is directly coupled to the anode of V1b the cathode of V2 is set at approximately +40V and so the screen of the power amplifier is at the same potential, thus the r.f. output is low. If now VR2 is set to maximum resistance then V1b is biased to approximately cut off and therefore the anode voltage of this valve rises to a much higher value than previously, consequently the power amplifier screen reaches approximately +250V and a large r.f. output is available. By varying VR2 over the full range a very convenient and effective power control is available.

The a.f. modulation operates in the following manner. Assume that the grid of V1b is set to −1V by VR2, then the d.c. voltage on the power amplifier screen is low and the r.f. output is also low. With an a.f. signal applied to C1 and amplified by V1a so that it is of a value of 1V peak at the grid of V1b, this signal will then be further amplified by V1b and fed to the screen of the power amplifier through V2, the cathode follower. This a.f. signal will now modulate the low r.f. output to approximately 95%. If the a.f.

signal is now increased grid current will flow in grid of V1b and a negative charge will develop on C4 proportional to the peak value of the a.f. signal at this point. This additional bias will cause the anode voltage of V1b to rise and so the mean screen voltage of the power amplifier will rise and the r.f. output will increase. The increased a.f. signal that has caused the increase in the r.f. output will also appear at the screen of the power amplifier superimposed on the higher mean potential and the larger carrier is now modulated to approximately 95%. The final result is a fully modulated carrier controlled v.h.f. power amplifier.

The only critical components in the modulator are C4 and R5 (1.5M) as these determine the period that the increased screen voltage is retained once the a.f. signal has disappeared and this is important with speech signals. The optimum values are those given in Fig.6.36. R9 must be located immediately adjacent to pin 3 of the holder for the QQVO6-40A.

The switch in the cathode return of V1b permits the screen voltage of the power amplifier to rise to the maximum and is used for setting up or for c.w. operation. The method of operation is to set S1 to c.w. and tune the amplifier in the normal way. Set S1 to phone and adjust VR2 to give approximately a fifth of the maximum r.f. output obtained in the c.w. position. Apply the a.f. signal and set the gain control VR1 so that the amplifier anode current indicates 0.7 of the full c.w. anode current and this will ensure that the amplifier is fully modulated.

A feature of this type of modulator is the very wide audio response available. The particular circuit shown here has been tested from 20 hertz to 12 kHz and is subtantially flat, therefore any pre-amplifier should restrict the input to frequencies of 300 to 3500 hertz. If no pre-amplifier is used then a low pass filter should be inserted between the cathode of V2 and the screen of the power amplifier.

FREQUENCY AND PHASE MODULATION

When the frequency of the carrier is varied in accordance with variations in a modulating signal, the result is frequency modulation (f.m.). Similarly, varying the phase of the carrier current is called phase modulation (p.m.).

Frequency and phase modulation are not independent, since the frequency cannot be varied without also varying the phase, and vice versa. The difference is largely a matter of definition.

The effectiveness of f.m. and p.m. for communication purposes depends almost entirely on the receiving methods. If the receiver will respond to frequency and phase changes but is insensitive to amplitude changes, it will discriminate against most forms of noise, particularly impulse noise such as is set up by ignition systems and other sparking devices.

Special methods of detection are required to accomplish this result.

Modulation methods for f.m. and p.m. are simple and require practically no audio power. There is also the advantage that, since there is no amplitude variation in the signal, interference to broadcast reception resulting from rectification of the transmitted signal in the audio circuits of the broadcast receiver is substantially eliminated. These two points represent the principal reasons for the use of f.m. and p.m. in amateur work.

Frequency Modulation

Fig.6.37 is a representation of frequency modulation. When a modulating signal is applied, the carrier frequency is increased during one half-cycle of the modulating signal and decreased during the half-cycle of opposite polarity. This is indicated in the drawing by the fact that the r.f. cycles occupy less time (higher frequency) when the modulating signal is positive, and more time (lower frequency) when the modulating signal is negative. The change in the carrier frequency (frequency deviation) is proportional to the instantaneous amplitude of the modulating signal, so the deviation is small when the instantaneous amplitude of the modulating signal is small, and is greatest when the modulating signal reaches its peak, either positive or negative.

As shown by the drawing the amplitude of the signal does not change during modulation.
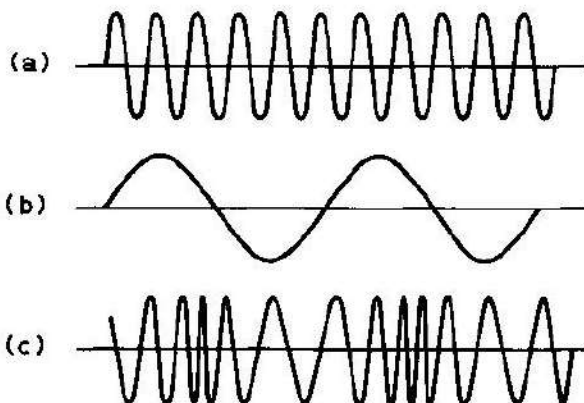
Fig.6.37. Graphical representation of frequency modulation. (a) Unmodulated carrier wave. (b) Modulating signal. (c) Frequency modulated carrier wave (i.e. its frequency is varied in sympathy with (b)).

Phase Modulation

If the phase of the current in a circuit is changed there is an instantaneous frequency change during the time that the phase is being shifted. The amount of frequency change, or deviation, depends on how rapidly the phase shift is accomplished. It is also dependent upon the total amount of the phase shift. In a properly operating p.m. system the amount of phase shift is proportional to the instantaneous amplitude of the modulating signal. The rapidity of the phase shift is directly proportional to the frequency of the modulating signal. Consequently, the frequency deviation in p.m. is proportional to both the amplitude and frequency of the modulating signal. The latter represents the outstanding difference between f.m. and p.m. since in f.m. the frequency deviation is proportional only to the amplitude of the modulating signal.
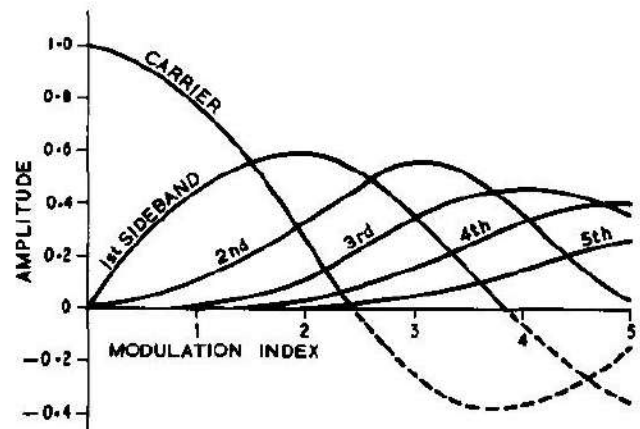
Fig.6.38. Variation in relative amplitude of the carrier and the low order sidebands in a frequency modulation system.

The amplitudes of the sidebands produced are proportional to the modulation index; i.e. the amplitudes of the higher order side frequencies (4th and above) increase from a value which is negligible when the modulation index is 1.0 to appreciable values when the modulation index is greater than 5. The manner in which the sidebands vary is shown in Fig.6.38 where their variation is plotted for values of the modulation index up to 5, it will also be seen that when the modulation index is 2.4, the carrier itself disappears, i.e. its phase reverses compared with the phase when modulation is absent. Further disappearances occur at higher values of modulation index.

It follows therefore that the question of the bandwidth required in the transmitter and receiver of an f.m. system has not such a straightforward answer as in the case of amplitude modulation.

In frequency modulation there is no condition equivalent to over-modulation. An increase in the amplitude of the modulating signal will merely cause an increase in the deviation produced. In the case of large deviations this effect will of course introduce difficult problems into the design of the r.f. circuits of the transmitter and in the receiver, but these conditions do not occur in amateur communication equipment where the deviation is kept relatively small.

6.23

At present the recommendations on the use of frequency modulation by amateurs are as follows:

(a)   The maximum deviation shall not exeed 2.5 kHz.

(b)   The maximum effective modulating frequency shall be limited to 4 kHz and the a.f. input to the frequency modulator at frequencies higher than 4 kHz shall be not less than 26dB below the maximum input at lower frequencies.

(c)   The carrier frequency shall be at least 10 kHz within the limits of the frequency band in use.

The deviation ratio permitted is therefore 2.5 kHz divided by 4 kHz or 0.625. At this value, the carrier and the first-order sidebands are appreciable, and the second-order sidebands are almost small enough to be neglected while the third and higher orders may be ignored. Therefore, to a first approximation the bandwidth occupied may be considered as equal to twice the deviation or 5 kHz. In other words, it is roughly equivalent to the bandwidth required for a "communication-quality" amplitude modulated transmission. In contrast to the large deviations used in high-fidelity broadcasting for which an international standard of 75 kHz has been adopted.

Comparison of F.M. and P.M.

Frequency modulation cannot be applied to an amplifier stage, but phase modulation can; p.m. is therefore readily adaptable to transmitters employing oscillators of high stability such as the crystal-controlled type. The amount of phase shift that can be obtained with good linearity is such that the maximum practicable modulation index is about 0.5. Because the phase shift is proportional to the modulating frequency, this index can be used only at the highest frequency present in the modulating signal, assuming that all frequencies will at one time or another have equal amplitude. Taking 2500 hertz as a suitable upper limit for voice work, and setting the modulation index at 0.5 for 2500 hertz, the frequency response of the speech amplifier system above 2500 hertz must be sharply attenuated to prevent sideband splatter. Also, if the "tinny" quality of p.m. as received on an f.m. receiver is to be avoided, the p.m. must be changed to f.m., in which the modulation index decreases in inverse proportion to the modulating frequency. This requires shaping the speech-amplifier frequency-response curve in such a way that the output voltage is inversely proportional to frequency over most of the voice range. When this is done the maximum modulation index can only be used at some relatively low audio frequency, perhaps 300 to 400 hertz in voice transmission, and must decrease in proportion to the increase in frequency.

The result is that the maximum linear frequency deviation is only one or two hundred hertz, when p.m. is changed to f.m. To increase the deviation for n.b.f.m. requires a frequency multiplication of 8 times or more.

It is relatively easy to secure a fairly large frequency deviation when a self-controlled oscillator is frequency-modulated directly. (True frequency modulation of a crystal-controlled oscillator results in only very small deviations and so requires a great deal of frequency multiplication). The chief problem is to maintain a satisfactory degree of carrier stability, since the greater the inherent stability of the oscillator the more difficult it is to secure a wide frequency swing with linearity.

Modulation Depth

Percentage of modulation in f.m. and p.m. has to be defined differently than for a.m. Practically, "100 per cent modulation" is reached when the transmitted signal occupies a channel just equal to the bandwidth for which the receiver is designed. If the frequency deviation is greater than the receiver can accept, the receiver distorts the signal. However, on another receiver designed for a different bandwidth the same signal might be equivalent to only 25 per cent modulation.

In amateur work *narrow-band* f.m. or p.m. (frequently abbreviated n.b.f.m.) is defined as having the same channel width as a properly modulated a.m. signal. That is, the effective channel width does not exceed twice the highest audio frequency in the modulating signal. N.b.f.m. transmissions based on an upper audio limit of 2500 hertz therefore should occupy a channel not significantly wider than 5 kHz.

F.M. and P.M. Sidebands

The sidebands set up by f.m. and p.m. differ from those resulting from a.m. in that they occur at integral multiples of the modulating frequency on either side of the carrier rather than, as in a.m. consisting of a single set of side frequencies for each modulating frequency. An f.m. or p.m. signal therefore inherently occupies a wider channel than a.m.

The number of *extra* sidebands that occur in f.m. and p.m. depends on the relationship between the modulating frequency and the frequency deviation. The ratio between the frequency deviation, in hertz, and the modulating frequency, also in hertz, is called the modulation index. That is:

$$\text{Modulation index} = \frac{\text{Carrier frequency deviation}}{\text{Modulating frequency}}$$

Example: The maximum frequency deviation in an f.m. transmitter is 2500 hertz either side of the carrier frequency. The modulation index when the

modulating frequency is 1000 hertz is

$$\text{Modulation index} = \frac{2500}{1000} = 2.5$$

At the same deviation with 2500 hertz modulation the index would be 1, at 100 hertz it would be 25 and so on.

In p.m. the modulation index is constant regardless of the modulating frequency; in f.m. it varies with the modulating frequency, as shown in the above example. In an f.m. system the ratio of the maximum carrier-frequency deviation to the highest modulating frequency used is called the deviation ratio.

TYPES OF F.M. TRANSMITTERS

Frequency modulated transmitters may be divided into two broad classes:

(i)   Those which use a quartz crystal to determine the centre frequency.

(ii)   Those which operate by the direct frequency variation of the master oscillator.

Transmitters in the first category are, in general, complicated in design and may involve the production of a phase-modulated signal which is then converted to a frequency-modulated one, the use of specially cut crystals or complicated chains of phase shifters, balanced modulators, frequency multipliers and mixers. However, they are capable of providing large deviations at extremely high degrees of stability of the centre frequency. For details, the reader is referred to the standard works on frequency modulation.

In the second category, the frequency of the master oscillator or v.f.o. is varied directly by placing a variable reactance across the tuned circuit of the oscillator. This of course implies that the frequency stability of the oscillator itself must be as high as possible. The stability of the present-day variable-frequency oscillator as used in amateur transmitters is usually adequate, and this method of producing frequency modulation is therefore the one best suited to amateur use.

The Variable-Reactance Valve Modulator

The basic circuit of the variable reactance or reactor type of modulator is shown in Fig.6.39. Essentially it consists of a valve connected across the tuned circuit of an oscillator in such a way that it behaves like a variable capacitance or inductance. In the circuit shown, a potentiometer consisting of a capacitor, C2 and a resistor R in series is connected from anode to earth and also, of course, across the tuned circuit. The centre point is taken to the grid of the valve. Provided that the resistance R is large compared with the reactance of C2 at the resonant

frequency of the tuned circuit, the grid will be fed with a voltage which is very nearly 90° out of phase with the voltage at the anode. A similar phase-shift but of the opposite sign may also be obtained by interchanging C2 and R, in which case the reactance of C2 would need to be large compared with the resistance R.

Since the anode current of a valve is in phase with its grid voltage, the current flowing through the reactance valve is 90° out of phase with the voltage across the tuned circuit due to the 90° phase-shift introduced by the combination of C2 and R. Since the valve current also flows through the tuned circuit, the current through the tuned circuit is 90° out of phase with the voltage across it, which is equivalent to the result produced by connecting a reactance across the tuned circuit. This reactance may be either capacitive or inductive according to the configuration of the RC2 potentiometer. In the arrangement shown in Fig.6.39 the reactance will be inductive, while if R and C2 are interchanged, the reactance will be capacitive. In either case, the effect of this reactance will be to change the resonant frequency of the tuned circuit.
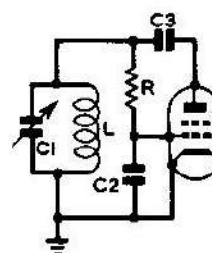
Fig.6.39. Basic circuit of a variable-reactance modulator. LC1 is the tuned circuit of the master oscillator.

The value of the reactance thrown across the tuned circuit and hence the change in resonant frequency will obviously depend on the value of the anode current of the reactance valve, in other words it will depend on the voltage applied to the grid of the valve. Thus the variable reactance valve presents a simple means of varying the frequency of an oscillator in sympathy with an a.f voltage applied to the grid of the reactance valve. The amplitude of the grid voltage will govern the magnitude of the change in frequency, i.e. the deviation produced will be determined by the amplitude of the modulating signal and the rate at which the oscillator frequency is varied will be equal to the frequency of the voltage applied.

A reactance valve modulator may be used with any type of oscillator. It is normally connected directly across the tuned circuit of the oscillator as shown in Fig.6.39 although in the case of a series-tuned circuit, such as the Clapp oscillator, it may be connected across the tuning capacitor.

Almost any tetrode or pentode valve will function as a reactor in this manner. Since the reactor is a voltage-operated device, a voltage-amplifying type is required rather than a power type. The long grid
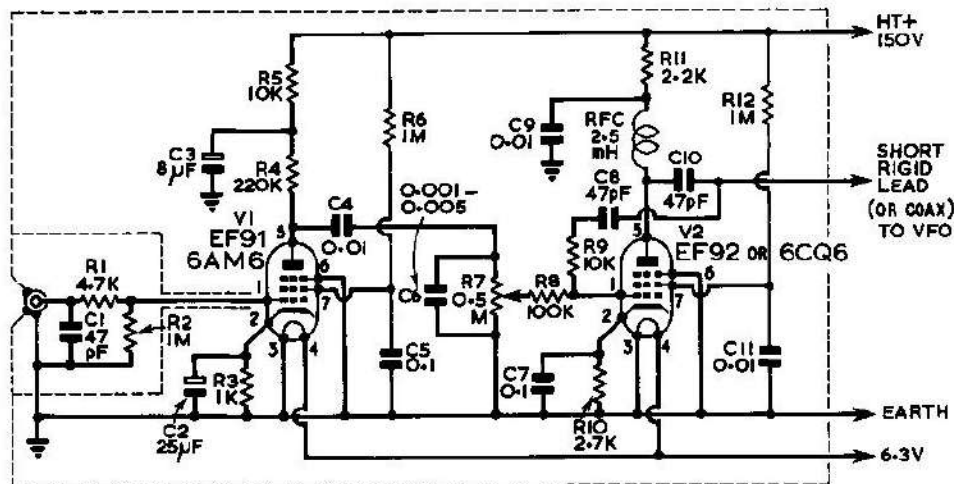
Fig.6.40. A typical reactance valve modulator. C1 - 47 pF (Ceramicon) C2 - 25 µF, 25V (electrolytic) C3 - 8µF, 350V (electrolytic) C4, C9, C11 - 0.01 µF C5, C7 - 0.1 µF C6 - 0.001-0.005 µF (mica) C8, C10 - 47 pF (mica) RFC - 2.5 mH r.f. choke V1 - EF91, or equivalent, V2 - EF92, or equivalent, R1 - 4.7 K ohms, ¼W, R2 - 1 M ohms, ¼W, R3 - 1 K ohms, ¼W, R4 - 220 K ohms, ¼W, R5 - R9 - 10 K ohms, ¼W, R6 - R12 - 1 M ohms, ¼W, R7 - 0.5 M ohms potentiometer (linear), R8 - 100 K ohms, ¼W, R10 - 2.7 K ohms, ¼W, R11 - 2.2 K ohms, ¼W.

base of the variable-µ valve is required for optimum performance. Alternatively, a hexode or similar multi-grid valve may be used, and the r.f. voltage from the reactance potentiometer and the a.f. voltage from the speech amplifier can then be applied to different grids. In this way, there is less risk of instability which could be caused by feedback between the r.f. and a.f. circuits. It will be realised from the foregoing account of the operation of reactance modulators that the sensitivity of the reactor stage will be dependent on the mutual conductance of the reactor valve, i.e. the higher the mutual conductance the greater will be the change in anode current and hence in deviation for a given input voltage to the grid. The reactor must therefore be chosen with regard to the degree of a.f. amplification available or intended. In the interest of overall stability and ease of adjustment a valve having low or medium mutual conductance is generally to be preferred.

In the potentiometer arrangement shown in Fig.6.39 the capacitor C2 may conveniently consist of the input capacitance of the reactor valve plus the stray capacitance at this point, although of course no adjustment of C2 is then possible and the control must be effected by varying the resistance R. C3 is merely a d.c. blocking capacitor.

The h.t. supply to the reactor and associated circuits, i.e. the speech amplifier and the v.f.o. must be very well smoothed to prevent the production of spurious amplitude and frequency modulation at the h.t. ripple frequency and also any unwanted change of mutual conductance of the reactor valve. Likewise great care must be taken with the layout to prevent hum pick-up.

A Practical Reactance Modulator

The circuit of a typical reactance valve modulator is shown in Fig.6.40. This consists of two

stages, a speech amplifier, V1 and a reactor valve V2. The speech amplifier is intended for use with a crystal microphone or other low-output type and is therefore designed to have a high gain. The microphone input is shielded right up to the grid of V1. It would be advisable to reduce the gain of the first stage if a high-output microphone is used. The reactor stage V2 is a pentode having a variable-µ characteristic. The phase-shift network consists of the resistor R2 and the stray capacitance between the grid of V2 and earth, i.e. the input capacitance of V2 plus the wiring and valve-holder capacitances. The a.f. input to the reactor is controlled by the potentiometer R2. This control is, of course, the deviation control. A certain amount of top cut is introduced by the capacitor C6 shunting R7.
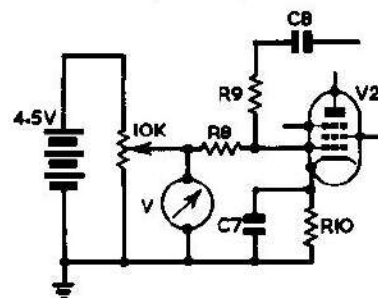
Fig.6.41. Test circuit for determining the static characteristic of a variable-reactance modulator. V2 must be isolated from the speech amplifier. The circuit components are identical with those shown in Fig.6.40.

C8 and C10 are blocking capacitors. The output lead which connects to the tuned circuit of the v.f.o. should be rigid and as short as possible to prevent any unwanted variation of the frequency. Alternatively coaxial cable may be used if the length of the lead exceeds an inch or two. With care and by the use of miniature components, the whole unit can be made on

a very small chassis so that it may be located conveniently close to the v.f.o. The screening afforded by the chassis should be as effective as possible and the h.t. supply (150v) preferably should be stabilized by a voltage regulator tube such as the VR150/30 or OA2.

In setting up the reactor stage the first step is to determine its characteristic, i.e. the relationship between the voltage applied to the grid of the reactor and the resulting deviation of the frequency of the v.f.o. This may be done quite simply by isolating the reactor and applying a variable d.c. voltage to its grid as shown in Fig.6.41. The frequency of the v.f.o. must first be measured by an accurate frequency meter, zero voltage being applied to the reactor grid. When the grid voltage is increased, the frequency of the v.f.o. will be found to change. The change in frequency for a known change in grid voltage should be measured by the frequency meter, this should be continued up to a maximum frequency deviation of about 4 kHz or so. Next, the polarity of the battery from which the grid voltage is obtained should be reversed and the test repeated. Frequency changes in the opposite direction should now be obtained. The results should be plotted on a graph and the curve produced should be found to resemble that shown in Fig.6.42. The characteristic should be reasonably linear over a total frequency swing of 5-6 kHz.
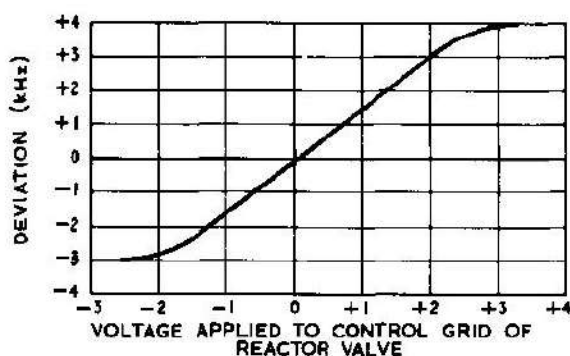
Fig.6.42. Static characteristic of the reactor valve of the n.b.f.m. unit shown in Fig.6.40.

Simple Frequency Modulators

A very simple method of achieving frequency modulation of an oscillator is shown in Fig.6.43. This circuit was originally used for making small frequency deviations (100-200 hertz) of commercial c.w. transmitters and it is quite similar in operation to the reactance modulator. A small variable condenser C2 is connected across the tuned circuit of an oscillator through a diode V1. Variation of the current through the diode will vary the shunting or detuning effect of C2, and thereby the resonant frequency of the tuned circuit will be varied. The current through the diode is modulated by the application of an a.f. voltage to the anode of the diode.
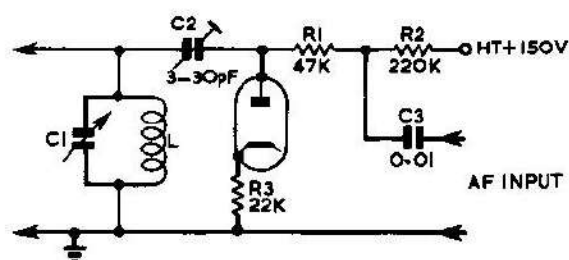
Fig.6.43. Simple frequency modulator having a reasonably linear characteristic over small deviations.

C1, L · tuned circuit of v.f.o.
C2 - 3-30 pF trimmer
C3 - 0.01 µF
V1 - EA50 or D1 (alternatively, double diode with sections paralleled)
R1 - 47 K ohms,½ W
R2 - 220 K ohms, ½ W
R3 - 22 K ohms, ½ W

The value of C2, can be adjusted to give the required deviation on each band. Preferably, the characteristics should be determined as described earlier. The linearity given by this arrangement, although not good, is adequate over the small frequency deviation required for amateur communication purposes.
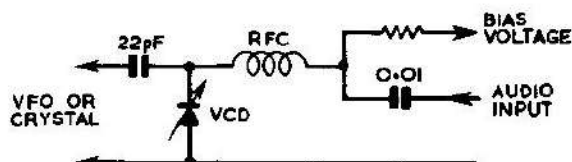
Fig.6.44. A basic circuit which shows the minimum requirements of a variable capacitance diode suitable for applying to a v.f.o. or crystal oscillator.

The valve diode in this circuit may be replaced by a semiconductor diode or a variable capacitance diode. The latter will be more satisfactory but most other types display variable capacitance characteristics. The operating point should be chosen to suit the capacitance variation required, this can be set by applying an appropriate bias voltage to the diode.

Typical Frequency Modulators

The following typical circuit arrangements have been in use for a number of years and very satisfactory results have been achieved. Reference is given to the amateurs who have developed and used the individual circuits for use on the 144 MHz band.

In the following circuits the operation may be summarised as follows:-

Circuit	Modulation Method
6.45	Variable capacity diode in series with capacitor across the crystal.
6.46	Variable capacity diode in series with capacitor across the crystal.
6.47	Reactance valve across an inductance in series with the crystal.
6.48	Reactance valve applied to the crystal oscillator anode circuit.
6.49	Reactance valve applied to the crystal oscillator anode circuit.
6.50	Audio modulation of anode voltage of the variable oscillator and frequency variation produced by this voltage change on the oscillator.
6.51	Variable capacity diode applied to crystal oscillator output.

Fig.6.47. Reactance Valve Modulator by G6TA. In this circuit the inductance in series with the crystal is varied by the reactance valve so causing frequency variations.

Fig.6.45. A variable capacitance diode modulator by G6AG. In this arrangement the variable capacitance diode is in series with a fixed capacitor and the operating bias voltage is provided from a potential divider.
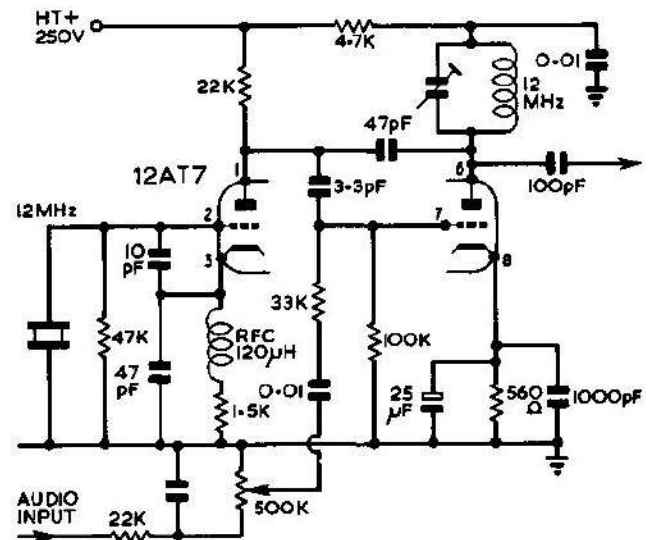
Fig.6.48. Reactance valve modulator by G3EDD using a double triode. In this figure and the next there is no direct action on the crystal itself.

Fig.6.46. A variable capacitance diode modulator by G6RH. Bias voltage is provided from the potential divider.

Fig.6.49. Reactance Valve Modulator by G3AWS/G3SLF using double triode as combined crystal oscillator/modulator.

Fig.6.50. Direct frequency modulation of a Franklin variable frequency oscillator by G2HCG. In this arrangement the audio modulation is applied to the anode voltage of the oscillator valves.

The general form taken by the first two circuits are probably the most useful for the amateur. Some care is, however, necessary in setting the bias required for the best linearity and this will vary with the diode used. The bias for these diodes will generally be in the region of 4 to 5 volts which can usually be obtained by a potential divider across the oscillator anode supply.

Any of the other circuits will give satisfactory results but they are obviously rather more complex, but nevertheless can be relied upon to give very adequate modulation and, when properly adjusted, only slight detuning from the carrier frequency is needed when using an a.m. detector on the receiver.

A LOW POWER TRANSMITTER for 70 and 144 MHz

This is a description of a simple two band transmitter, the band changing being accomplished merely by changing crystals and resetting the tuning controls. The block diagram in Fig.6.52 illustrates the principle involved.

Circuit Details

The circuit diagram is shown in Fig.6.53. The crystal oscillator stage uses a Z77 pentode. The anode coil is wound on a 7/16" diameter Aladdin former without a dust iron core (see coil table) and is tuned by a 100 pF tuning capacitor C1 of 10 pF minimum capacity, a Jackson Bros type C804.

The crystal oscillator circuit used is very stable and keys well, as shown, in the cathode lead. A test point is provided at the earthy end of the grid leak so that oscillation of the crystal may be checked by connecting a 0.1 mA meter between the test point and chassis across the 2.2K ohms resistor. A current of about 100 μA will be indicated when the crystal is oscillating.

The first tripler stage uses a 6CH6 valve which is provided with some cathode bias to limit the anode current to a safe value under key-up conditions. A 0-2 mA meter is included on the transmitter panel for measuring grid currents in the tripler and p.a. stages. A five-way rotary switch connects the meter with the appropriate grid circuits; the first tripler grid current being indicated when the switch is in position 1.

A push-pull output is obtained from the anode circuit of the 6CH6 tripler and circuit balance is preserved by the 4.7 pF capacitor connected between earth and the end of the tuned circuit remote from that to which the anode is connected. The centre tap of the anode coil is by-passed to earth and h.t. is supplied through a 220 ohm decoupling resistor. The tuning capacitor C2 (Jackson Bros type C801) is very

Fig.6.51. In this circuit a transistor oscillator is phase modulated by a variable capacity diode, a deviation of about 5 kHz is usually possible at an output frequency of 145 MHz.

L1 = 31 turns close wound 26 s.w.g. 7 mm former slug tuned.

This coil will cover 12 MHz crystal frequency with the normal slug, but if an 8 MHz crystal is to be used a coil with suitably increased turns will be needed.

In the first three circuits and the last but one the modulation is by direct frequency variation of the oscillator and may be regarded as pure frequency modulation. In the fourth and fifth circuits the modulation is by change of frequency of the anode circuit by phase change and is strictly a form of phase modulation and it is necessary in this type of circuit to make some frequency correction of the audio output.

Fig.6.52. Block diagram of the 2N4 transmitter. Band changing is achieved simply by inserting an appropriate crystal and resetting four tuning capacitors.
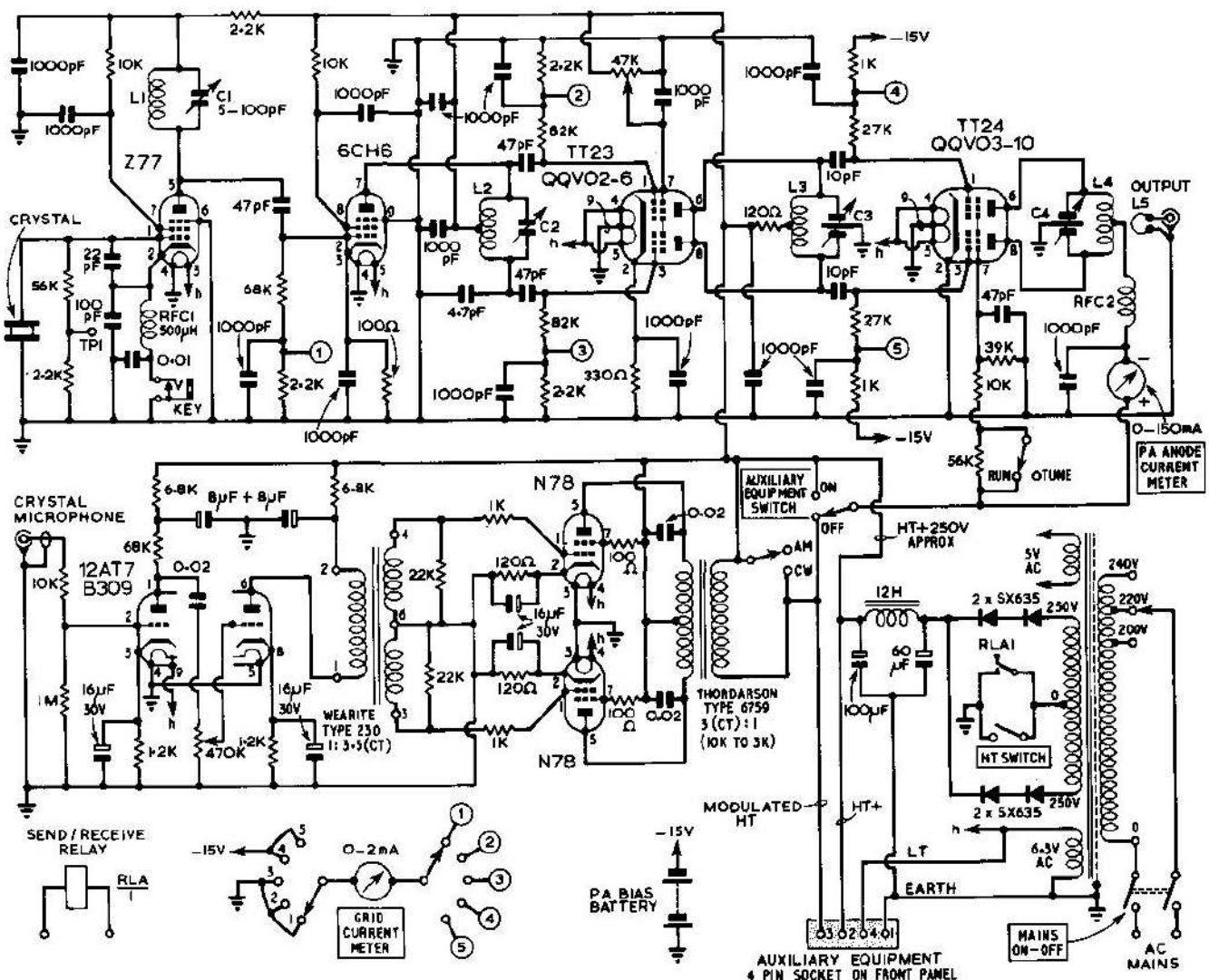
Fig.6.53. The complete transmitter circuit. C1, 10-100pF Jackson Bros C804. C2, 10-100 pF Jackson Bros C801, with vanes removed to leave 7 rotor and 7 stator vanes. C3, 5-70 pF Jackson Bros C808, with vanes removed to leave 8 rotor and 7 stator vanes per section. C4, 5-70 pF Jackson Bros C808, with vanes removed to leave 5 rotor and 6 stator vanes per section. The modulation transformer used is an American Thordarson type 6759 matching 10 K ohms to 3 K ohms. The key jack socket in the cathode lead of the Z77 must short circuit when the plug is withdrawn.

similar to that used in the oscillator anode circuit, but is of the pre-set type to which an insulated shaft has been added. Some vanes have been removed so that seven fixed and seven moving plates remain.

The second tripler stage is of the push-pull type and uses a QQVO2-6 valve which has very low input and output capacities allowing a reasonable L to C ratio to be maintained on both 2m and 4m. Cathode bias is provided to limit the anode current to a safe value under key-up conditions. The grid current of each half of the valve may be checked by setting the grid current meter switch to positions 2 and 3. The screen grid feed to the valve is taken from a potentiometer so that the drive to the p.a. may be controlled. Because the dynamic resistance of the tuned circuits is much lower on 4m than 2m, the drive, which is sufficient on 4m, is excessive on 2m and may be reduced to the correct value by adjusting the screen potentiometer accordingly.

The push-pull anode circuit of the second tripler is tuned by a split stator capacitor C3, a Jackson Bros type C808, 5-70 pF, stripped down so that it has eight moving and seven fixed plates per section. With the capacitor thus modified the circuit just covers the range necessary to cover both 4m and 2m.

The push-pull p.a. stage uses a QQVO3-10 twin tetrode valve. Battery bias of − 15 volts is used although a small negative bias supply derived from the mains could be used instead. Grid current in the two halves of the valve may be measured by setting the meter switch to positions 4 and 5. A *Tune-Run* switch arranges for the screen voltage of the p.a. valve to be reduced while tuning up the transmitter. Anode and screen modulation is used and the anode current of the p.a. is measured on a 0 to 150 milliammeter.

The tuning capacitor in the anode circuit of the p.a. is of the same split stator type as in the previous stage but with only five moving and six fixed plates per section. Tank coil details are given in the coil table. A single turn link coil is coupled to the aerial socket.

The modulator stage uses a pair of N78s in push-pull with transformer input. A 12AT7 twin triode with both halves in cascade forms the speech amplifier and provides ample gain to give full modulation from a crystal microphone.

The power supply is quite straightforward and provides 275 volts h.t. on load.

Coil Details

L1	22 turns, 30 s.w.g. enam., wound on 7/16" diam. Aladdin former.
L2	10 turns, centre tapped, 22 s.w.g. enam., wound on 7/16" diam. Aladdin former with dust core.
L3	3 turns, centre tapped, 18 s.w.g. tinned, 1/2" i.d., 1/2" long, self-supporting.
L4	6 turns, centre tapped, 14 s.w.g. tinned, 3/4" i.d., 3/4" long, self-supporting.
L5	1 turn in centre of L4.
RFC1	500 μH r.f. choke.
RFC2	50", 36 s.w.g. wound on ½ watt Erie resistor.

VALVE CURRENTS

Stage	Current 4m and 2m
Oscillator grid	100 μA
First tripler grid	1mA
Second tripler grid	1mA *
P.a. grid	1.5mA *
P.a. anode	60mA

* each half

Adjustment

Set the *Tune-Run* switch to Tune and plug in a suitable crystal for 2m. Check that the crystal is oscillating by connecting a milliammeter between the test point and earth in the oscillator grid leak lead. As with other stages, typical grid currents will be found below.

Tune the anode circuit of the oscillator until maximum grid current is obtained in the 6CH6 stage with the meter switch set to 1. The tuning position should be with the tuning capacitor nearly at minimum capacity. Check that the frequency is twice the crystal frequency by means of an absorption wavemeter.

Next tune the anode circuit of the 6CH6 stage for maximum grid current in the QQVO2-6 first tripler stage with the meter set to 2 or 3. Again, the tuning point should be near minimum capacity of the variable capacitor and a check should be made that the circuit is tuned to six times the crystal frequency using the absorption wavemeter. The grid currents in the two halves of the QQVO2-6 should be nearly equal. If they are widely different try altering the value of the 4.7 pF balancing capacitor.

Next tune the anode circuit of the QQVO2-6 (second tripler) stage for maximum grid current in the p.a. valve with the meter set to 4 or 5. Again the capacitor should be near minimum. Check the correct multiplication with the wavemeter. Set the grid current to about 1.5mA in each half of the valve by means of the screen potentiometer in the second tripler screen grid feed circuit.

Finally tune the p.a. to resonance with no load by setting the p.a. anode tuning capacitor for minimum anode current. The tuning point should be with the capacitor very near minimum capacity; some adjustment of the coil inductance may be necessary, and this is easily accomplished by squashing the turns together or pulling them apart. Switch from Tune to Run and connect the load. The output link may then be adjusted so that at resonance the p.a. valve draws the desired anode current.

The same procedure should then be followed using a crystal suitable for 4m. The correct tuning points will be found with the tuning capacitors set near their maximum capacity. Having once determined the tuning points for 2m and 4m marks can be made on the front panel and band changing simply consists of plugging in the appropriate crystal and setting the four capacitors to the correct marks.

A 70 MHz Power Amplifier

This amplifier is intended for operation up to the maximum permitted power and to be driven by a self-contained transmitter or transceiver. It contains its own power supply and the only transmit - receive switching provided is for the r.f. circuit.

The amplifier is a single 6146 or QVO6-20 with a transformer input coupling and a standard pi anode circuit, although this circuit is of necessity somewhat restricted in the range of output impedances to which it will match, nevertheless, it is more than sufficient to cover the usual aerial feeder lines of 50 or 75 ohms.

Provision has been made on the front panel for the control of the screen volts and this, together with the variable bias, allows a wide range of d.c. inputs to suit any particular need.

The amplifier gain is such that, with only 100mW drive, 20 watts output can be obtained with 40 watts input, thus enabling it to be operated as an amplifier for c.w. or n.b.f.m. from a low power semiconductor exciter. Satisfactory a.m. phone can be obtained even at this low drive level providing a suitable modulator is available.

The pi-output circuit is conventional and includes a point for use as modulation monitor or r.f. output indicator, the latter being essential if proper tuning of the circuit is to be achieved.
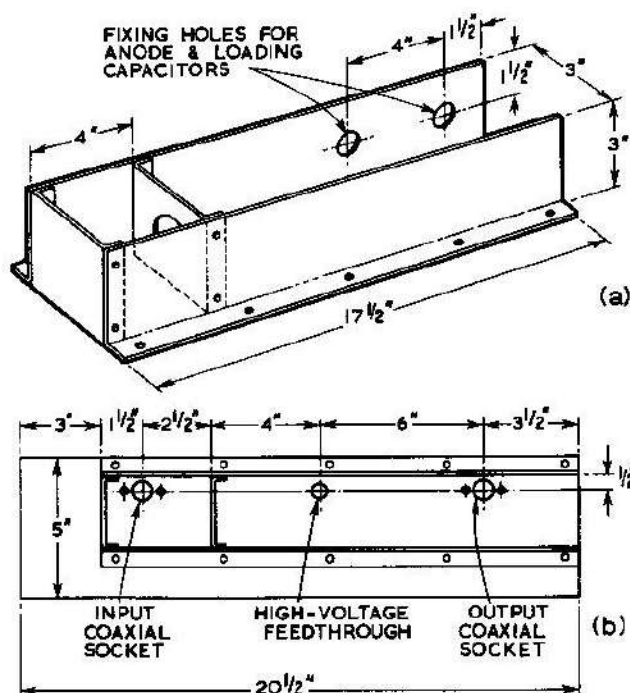
Fig.6.55. General arrangement of screen trough for the 70 MHz amplifier.

Circuit Description

The power supply section of the amplifier is self evident, a separate transformer being used for the h.t. thus allowing primary switching. The full circuit diagram is shown in Fig.6.54.

The screen supply to V1 (6146/QVO6-20) is controlled by V2 (N78, EL84 operating as a cathode follower. VR2 sets the d.c. voltage, and VR3 sets the audio level.

The use of a cathode follower to supply the screen of the power amplifier is superior to the clamp type control commonly used because, as explained, it allows both the d.c. and a.c. (modulation) levels to be set accurately.

The circuit enclosed by the dotted line is a simple clipper which is sufficient to prevent splatter due to accidental occasional over modulation. If it is not intended to use a.m. then the input to the grid of V2 can be taken directly to VR2 and the unwanted components eliminated.

The meter with the six position switch enables the voltage and current to be read for the grid, screen and anode of V1.

The input circuit to V1 is by centre tapped transformer and this provides the out-of-phase voltage for neutralising. The neutralising capacitor consists of a ½" square copper flag placed near the bulb in the region of the anode.

Construction

The r.f. section of the amplifier is built into a trough with a cross screen for mounting the valve socket, which is arranged so that valve is mounted horizontally within the screen as shown in Fig.6.55.

The input end of the screen is closed but the output end is left open. If necessary both the output end and the top may be enclosed in order to reduce radiation, and if this is done, suitable ventilation must be provided.

COIL DETAILS

L1 2 turns insulated wire over centre of L2
L2 4½ turns 18 s.w.g. ¾" i.d. ½" long
L3 2 turns 22 s.w.g. wound on 10 ohm resistor
L4 40-50 turns 26 s.w.g. ¼" i.d. close wound
L5 See text
L6 5 turns 18 s.w.g. ¾" i.d. ¾" long
L7 As L4
L8 10 H choke 120mA
L9 40 turns scramble wound on ¼" former 1" long
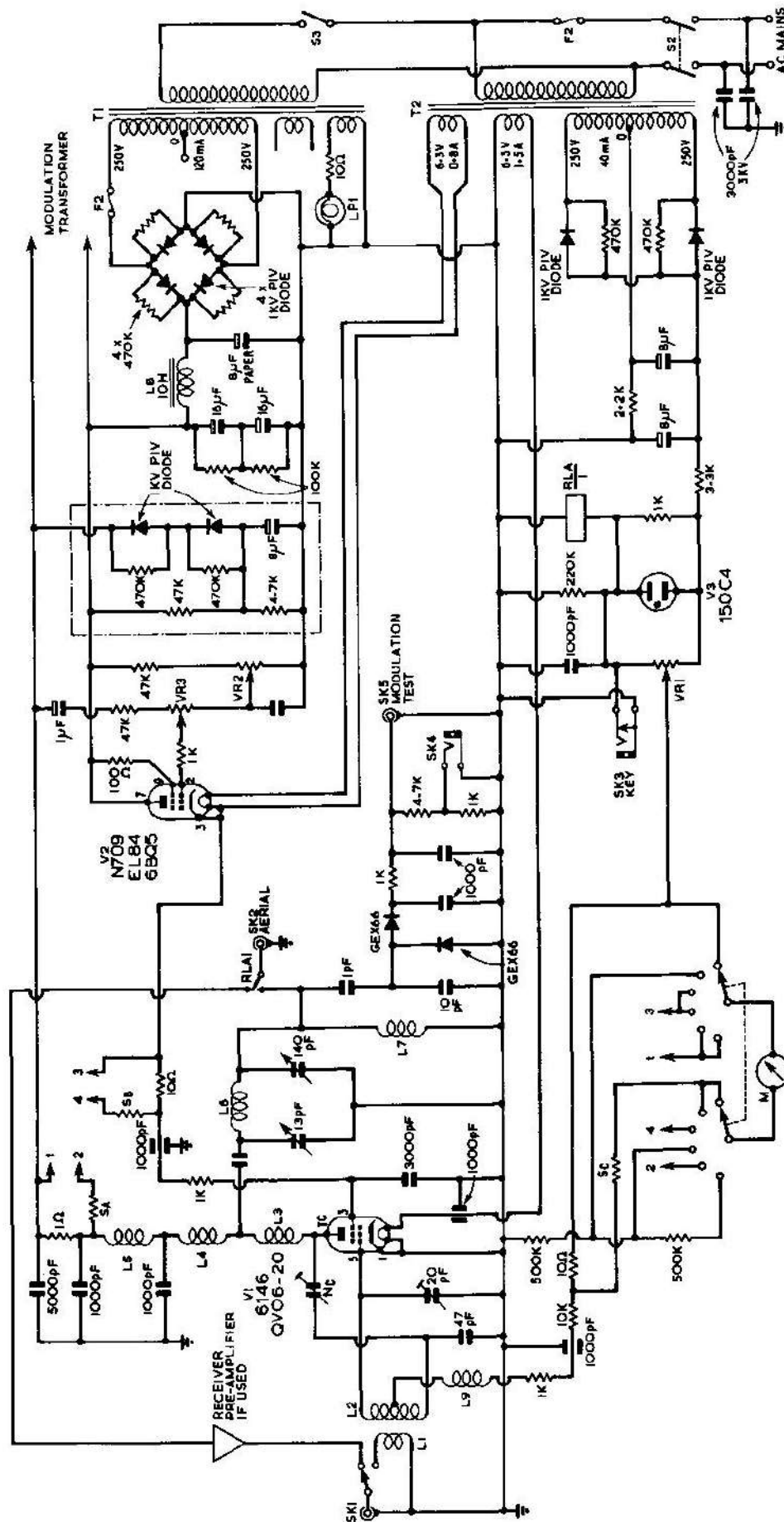
Fig. 6.54. Circuit diagram of a 70 MHz p.a. A series valve screen supply is used here and may be recommended for use in any similar circuit.

The h.t. feedthrough is in fact a filter consisting of the two 1000 pF ceramic capacitors connected together by L5 which is a short length of copper wire with six ferrite beads cemented to it. This is enclosed in an insulation tube and fixed into a panel fixing bush.

Care should be taken to ensure that the leads to the screen bypass capacitor of V1 are as short as possible. The resistor in series with L9 is to prevent t.p.t.g and other oscillation. The remainder of the assembly can follow any convenient form.

Setting Up

After checking the grid circuit and tuning it to about the centre of the band, switch on valve heaters and adjust VR1 for maximum voltage and VR2 to minimum.

Adjust VR2 for 150 volts on to V1 screen and with a dummy load in the output socket, set VR1 to give anode current of, say 40mA.

Neutralising should be carried out carefully in the following manner. First disconnect all power to the p.a. stage except heater voltage, and apply drive to the input. Couple a sensitive detecting device to the p.a. tank circuit and tune this circuit for maximum reading. Then adjust the neutralising capacitor for minimum indication, with an occasional check on the grid circuit tuning.

The detecting device could be a loop and diode rectifier with a suitable meter as the indicator. Since the amplifier is operated with the h.t. on at all times and only r.f. switching used for send-receive, it is most important that the neutralising is properly achieved.
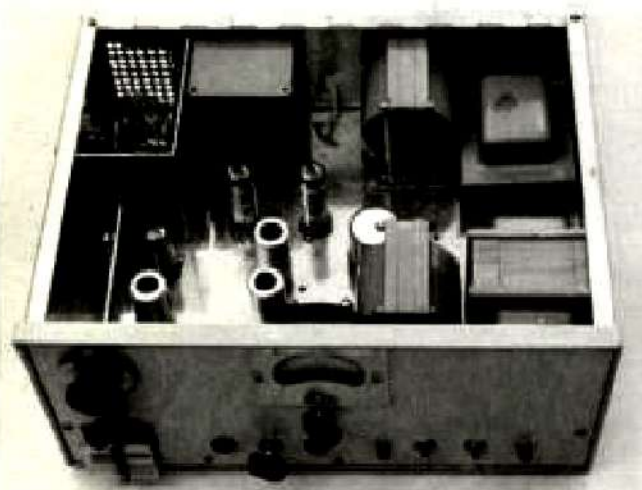
Fig.6.56. Top View of the 144 MHz transmitter. The power supply components are located in the right half of the cabinet. The modulation transformer is seen next to the p.a. screen which is located in the top left corner of the picture.

A GENERAL PURPOSE TRANSMITTER FOR THE 144 MHz BAND

This transmitter is suitable for most general purpose communications work and will be found eminently suitable for the home station. The design has some features not often included which will be found useful for operating convenience.

The circuit consists of a multi-crystal oscillator, using one half of a 12AT7 (V1). The second half of which is not used and is available for other use such as a phase modulator for the anode circuit of V1. This is followed by two frequency multipliers V2 and V3 using EF184 valves, tuned to 72 and 144 MHz respectively. The 144 MHz circuit is series tuned. Link coupling is used from the last frequency multiplier circuit to the input of the power amplifier.

The input circuit of the power amplifier is self resonant, and the anode circuit is a conventional lumped circuit with a series tuned output. The general arrangement of the anode and output circuits are shown in Fig.6.58.

A simple r.f. voltmeter circuit is connected across the output socket to enable easy tuning for maximum r.f. output. It should be noted that since maximum r.f. rarely coincides with minimum dip of the anode tuning when using double tetrodes, every effort should be made to ensure complete stability by preserving as far as possible the circuit symmetry.

The speech amplifier-modulator has been arranged for both crystal and moving coil microphones. As the impedance and voltage output of moving coil microphones are generally lower than the crystal types, a transistor amplifier provides the additional gain required together with the matching for medium impedance microphones, typically 200-600 ohms. Should a low impedance microphone such as one of 2-15 ohms impedance be used, it will be necessary to use a transformer to step up to about 600 ohms (microphone to line transformer). The negative voltage for this transistor pre-amplifier is obtained from the partial bias supply for the power amplifier.

The speech amplifier proper consists of V5 and V6, both being EF86 pentodes strapped as triodes. The first is a simple voltage amplifier while the second is used as a phase splitter to give the necessary push-pull input to the triodes of the triode pentodes used as the modulator.

The modulation transformer is a UMO or other equivalent type, adequate audio output is given by the amplifier to modulate fully the power amplifier running at an input of 18 watts. A modulation indicator is included.
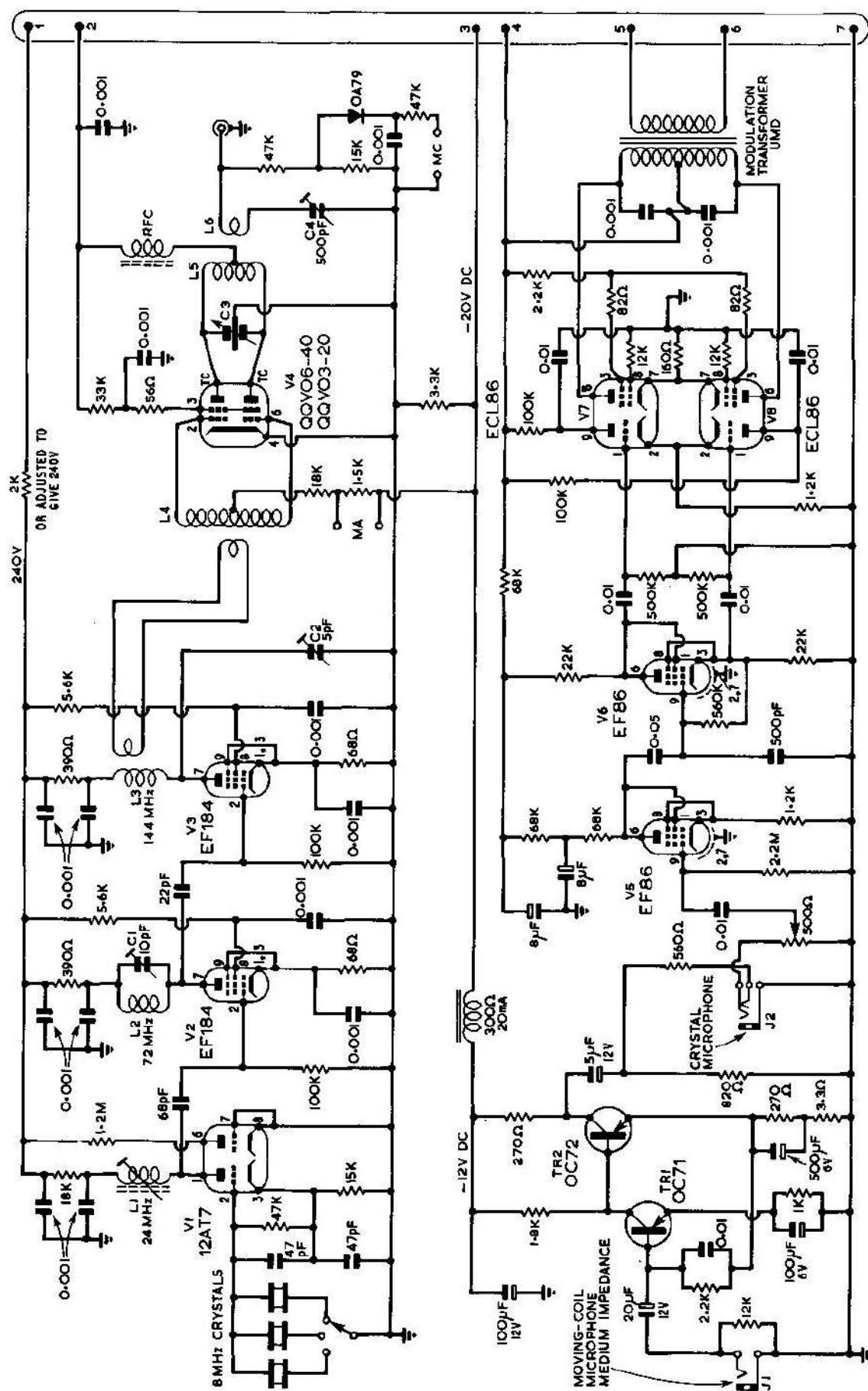
Fig. 6.57. A general purpose 2 metre transmitter using switched crystals in the 8 MHz range. An ECL86 is used as the modulator valve. R = 2K or should be adjusted to give 24v. If a low impedance microphone is used a transformer will be needed to step up to about 600 ohms, J2 is the input for a crystal microphone.
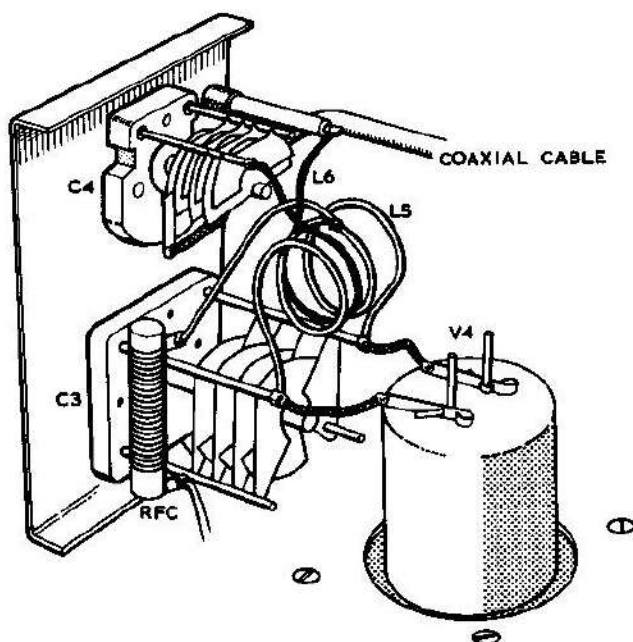
Fig. 6.58. The p.a. circuit arrangement. To assist in the proper adjustment of the power amplifier anode tuning capacitor a 2:1 reduction gear was used in the original but this may, of course, be omitted or a simple epicyclic reduction drive substituted. The p.a. anode circuit is enclosed in a screen box above the chassis as shown in top view photograph, Fig.6.56.

Details of coils and capacitors

Coil	Turns	Wire	Diam.
L1	25	28	¼" dust iron core
L2	4½	18	½" ½" long
L3	4	18	½" ¾" long
L4	4	18	½" 1¼" long
L5	3	14	1"(od) ¾" long
L6	1	16	1 ⅛" (o.d)

Link coils between L3 and L4 1 turn ¼" i.d. 26 s.w.g.

C1	10pF
C2	5 pF
C3	15 + 15 pF
C4	50pF
RFC	60 turns 30 s.w.g on 3/16" ferrite former

Power Supply

A complete power supply unit is detailed in Fig.6.60 in which all supplies are provided. Little description is needed since individual requirements will vary with available components. The modulation monitor circuit is shown in this diagram also.

It will be noted that the relay supply and the supply for the transistor preamplifier is obtained by series connecting a number of heater supplies. In the prototype T2 had 4 volt heater windings but a standard 6.3v and 5v type is obviously suitable.

The on/off switching is by relay for both h.t. supplies and the aerial change-over, and a contact for

relay switching the receiver is provided. If the whole unit is required for mobile or portable operation, the relays should have suitable voltage coils.

Typical valve currents

R.F. Section			
	V1	12 AT7	1-2mA
	V2	EF184	18-19mA cathode current
	V3	EF184	18-19mA cathode current
	V4	QQV03-20	60mA
Audio Section			
	V5	EF86	1.5mA
	V6	EF86	2mA
	V7/8	ECL86	triode 1.5mA pentode 35mA

Construction

The overall dimensions of the cabinet are 14½" x 11" x 7" high. The component layout and construction are clearly shown in the photographs (Figs.6.56 and 6.59). For convenience the crystals are plugged into the six-way socket, through the front panel, seen on the bottom left of Fig.6.56. Ventilation of the power amplifier valve is provided by drilling holes in the back of the cabinet. The aerial changeover relay is mounted on the outside of the back for convenient cable access.

Setting up the meter for its various functions:

1. for position MA p.a. grid current, shunt for 10mA f.s.d.
2. MB p.a. anode current, shunt for 100mA f.s.d.
3. MC r.f. output meter used on basic 1mA range
4. MD modulation percentage meter used on its basic range, as a voltmeter with VR series adjusted to read f.s.d. when points X1 and X2 with h.t. at 300 ± 5V f.s.d. is equal to 100% modulation

A MEDIUM POWER AMPLIFIER FOR 144 MHz

The amplifier described is typical of the design from which fairly high power at good efficiency may be obtained from the popular double tetrode QQV06-40A (5894).

The anode circuit consists of quarter wave lines which are enclosed to prevent loss by radiation. In cases where space is important a lumped constant circuit may be used in place of the linear circuit with only small loss of efficiency.
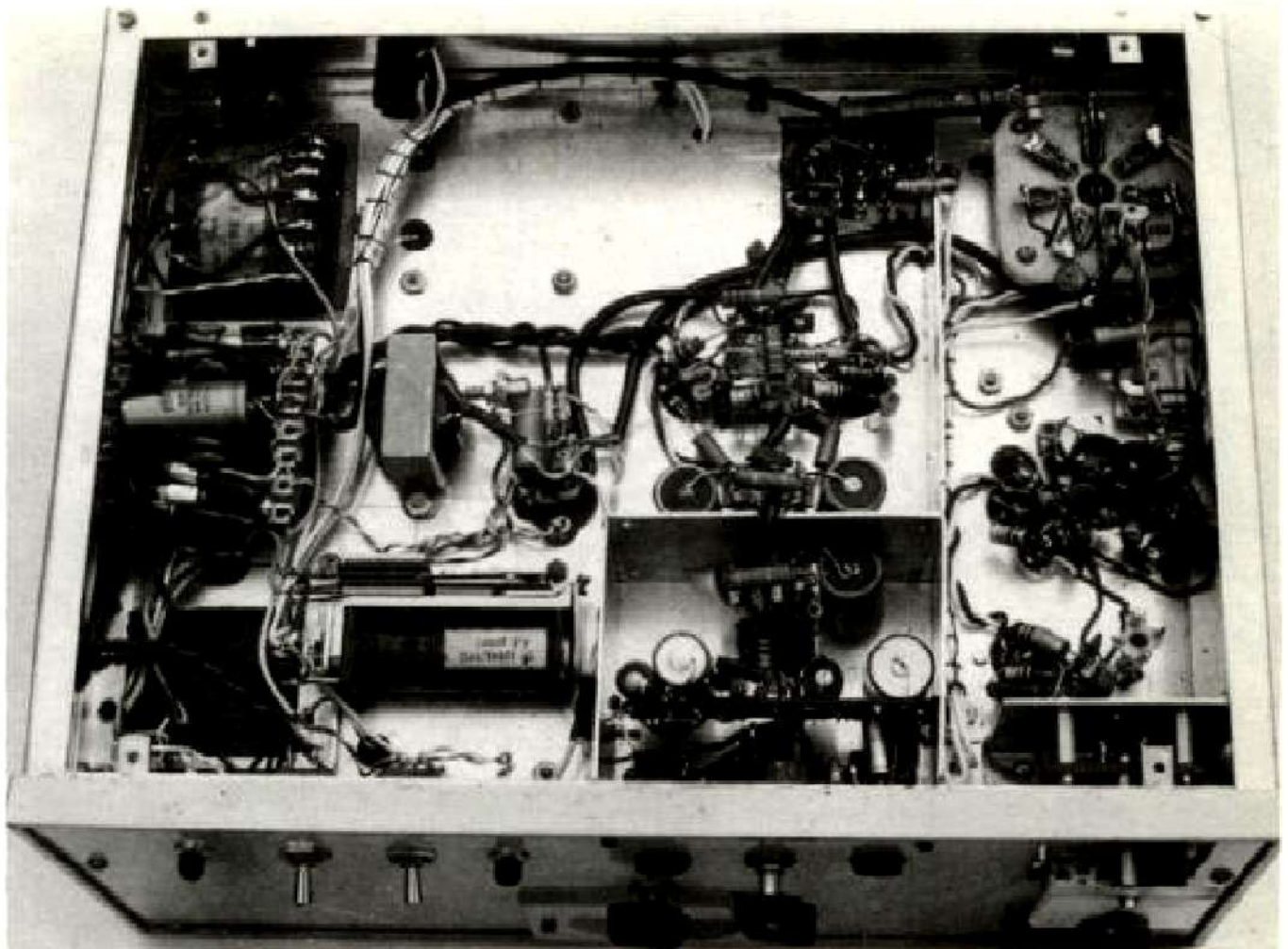
Fig. 6.59. Underside view of the general purpose transmitter showing the under chassis screening separating the r.j. section from the speech amplifier stages. The transistor pre-amplifier and the first audio amplifier valve circuits are enclosed by the box seen to the left of the relay.

Before detailing the amplifier it is useful to note the power capability of this valve when used as a Class C amplifier on this band.

From these figures rt may be seen that the efficiency is higher for higher anode voltages and that there is a considerable down rating needed when amplitude modulation is used compared to c.w. or n.b.f.m.

	c.w or n.b.f.m.		Anode and Screen Modulated (a.m.)		
Anode voltage	400	600	400	600	V
Screen voltage	250	250	250	250	V
Grid voltage	-60	-80	-70	-80	V
Anode current	2 x 100	2 x 100	2 x 75	2 x 75	mA
Screen current	2 x 8	2 x 9	2 x 9.0	2 x 9.0	mA
Grid current	2 x 3	2 x 3.5	2 x 9.0	2 x 2.0	mA
Power output	56	90	41	67	W
Power in load *	47	79	35	57	W
Efficiency	70	75	69	75	%

* this assumes a typical transfer efficiency
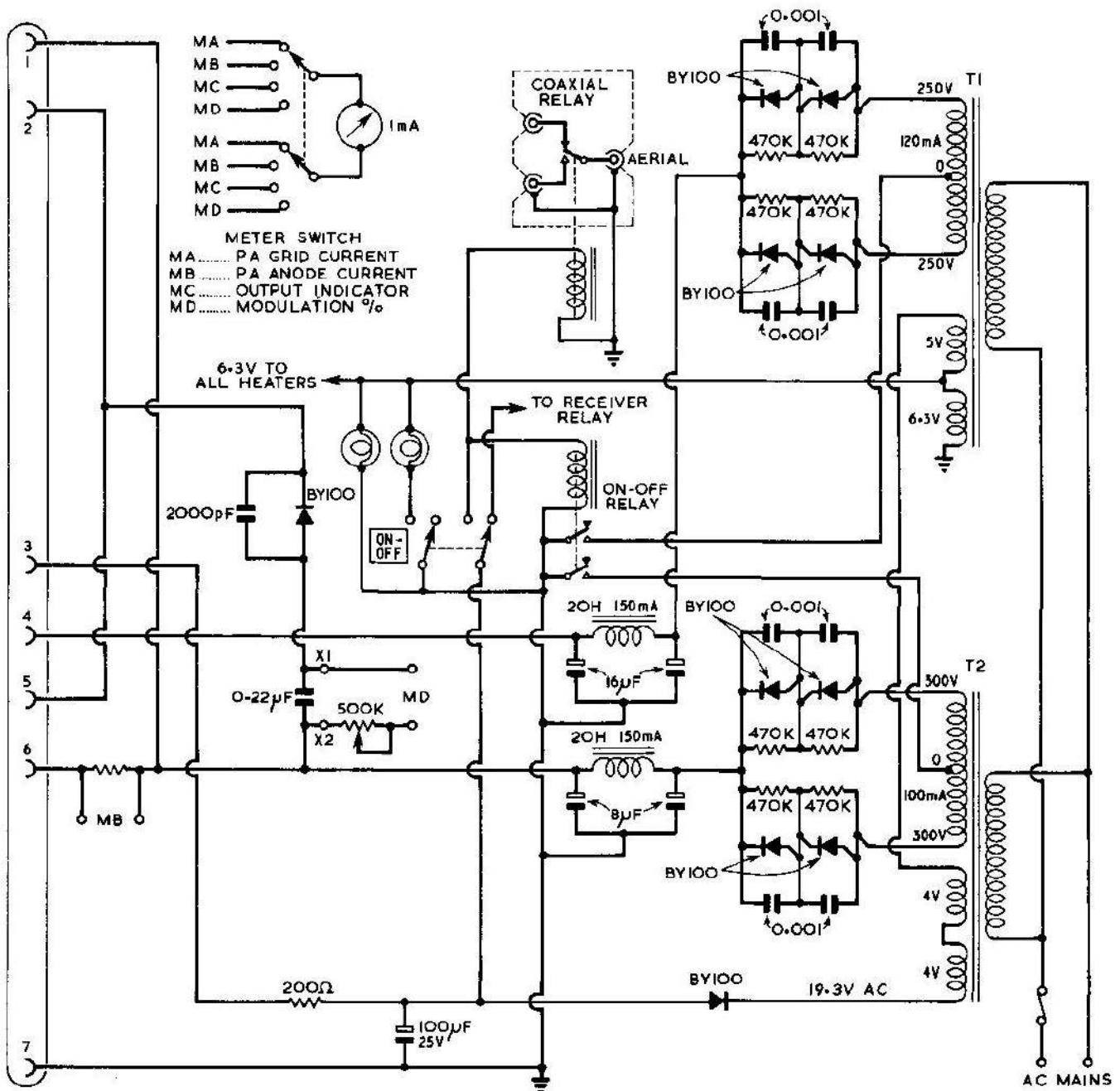
Fig. 6.60. Power supplies for the transmitter shown in Fig. 6.57. Meter switch at MA shows p.a. grid current; at MB p.a. anode current; at MC output indicator, and at MD modulation percentage.

Circuit Description

The circuit is shown in Fig.6.61. R.f. drive is applied to the push-pull grid circuit L2, C1. Grid bias is obtained from current flow through R2 when excitation is applied. For this reason a tetrode clamp valve V2 (6V6) is connected between the amplifier screen and earth, so that when the p.a. is driven the negative bias developed across R2, is also applied to the grid of the clamp valve, biasing it beyond cut-off. Removal of the drive, and hence removal of the bias from V2 grid, causes this valve to conduct heavily, resulting in a greatly increased voltage drop across R4. The screen voltage of V2 drops to a value such that the anode current of V1 is well within the rated dissipation.
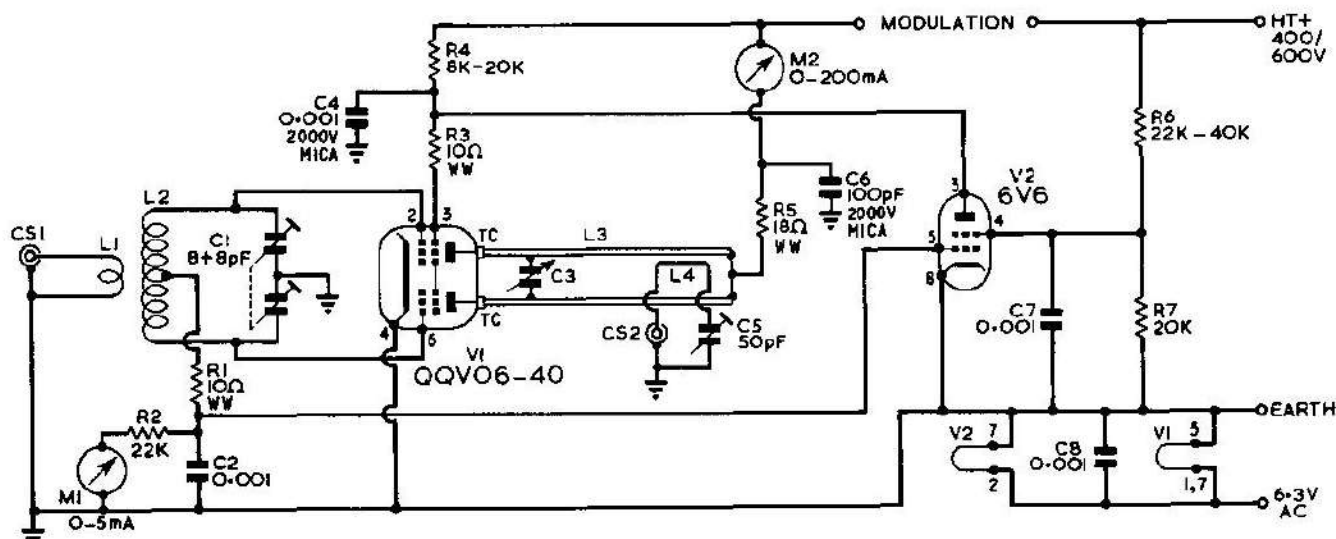
Fig.6.61. Circuit of the 2m p.a. showing the linear anode configuration.

L1   1 turn 20 s.w.g. ¾" diam.
L2   4 turns 16 s.w.g. ⅝" diam. 1¼" long
L3   2 tubes 8" long ¼" diam. spaced 1" apart
L4   Output coupling loop, 2¾" long 1" wide 16 s.w.g.
R1 R3 10 ohms wire wound

The anode circuit L3 is a quarter wave line, tuned by C3, which is tapped down the line a short distance from the anode end of the line. R.f. output is taken from a loop L4 which is coupled to L3 adjacent to the shortcircuited end.

The value of the screen dropping resistors for V1 and V2 (R4 and R6 respectively) is chosen according to the h.t. voltage so that the screen dissipation of the valves is not exceeded. The range of values shown is suitable for voltages between 400 and 600. Resistors R1 and R3 in the grid and screen circuits respectively are wire wound to reduce any tendency to parasitic oscillation in these circuits.

Fig.6.61A. Layout of the tank circuit for the 144 MHz. p.a.

Construction

The amplifier is built on an aluminium chassis 14" long x 5" wide x 3" deep; the output circuit compartment is 3½" wide and 4½" high. Dimensions are not critical and may be chosen to suit any particular installation. Complete isolation between grid and anode circuits is achieved by mounting the valveholder below a 2" diameter hole in the chassis on ⅜" pillars, so that the horizontal screening disc inside the valve is level with the top of the chassis.

The anode lines can be easily fabricated and consist of two lengths of ¼" diameter brass tubing 8" long, spaced 1" apart. At the short circuited end the two tubes are joined by a strip of heavy gauge (e.g. 12 s.w.g.) brass, in which are drilled two ¼" diameter holes centred 1" apart, into which the tubes are pushed after filing to fit and then soldered. H.t. (via R5) is connected to the midpoint of the short circuiting strip. At a distance of 1½" from the open circuited end of the lines the anode tuning capacitor C3 is mounted; each plate of this capacitor consists of a brass disc 1¾" diameter attached to a threaded stud which engages in a tapped hole in a brass block soldered to the lines. A fine thread (e.g. 40 t.p.i) should be cut on the studs to ensure smooth movement of the capacitor plates. A slot should be cut at the end of one stud to engage with a small metal blade fixed to the end of a ¼"diameter polystyrene rod which forms the tuning control; this rod may be spring-loaded to avoid any tendency of the blade to jump out of the slot as the control is rotated.

It will be found that resonance occurs when the two plates are approximately ¼" apart and they should be adjusted so that this distance is obtained when the front plate is at the mean position of its

6.39

travel. The threaded studs should be a reasonably tight fit in the tapped holes to ensure that there is no electrical discontinuity between the capacitor and the lines. It is advisable for the rear plate to be locked in position after resonance has been determined.

Anode connectors must be used and can conveniently be made of ⁵⁄₁₆″ brass rod ¾″ long similar to those employed for the 70cm transmitter. Copper strip ¼″ wide and 2¼″ long is used to join the end of the lines to each anode connector. The anode circuit lines are supported at the end by a ¼″ thick paxolin plate to which the short-circuiting strip is screwed, and by two polystyrene blocks mounted on an aluminium bracket located 4½″ from the end of the lines.

Operation

To set up the unit apply heater voltage only and connect a load to CS2. Excitation may then be applied to CS1 and C1 tuned to resonance. The output of the exciter and coupling between L1 and L2 should be adjusted to give a grid current of 4mA. H.t. may then be applied to the amplifier and C3 tuned; the dip in anode current at resonance should correspond to maximum r.f. output. Coupling between L3 and L4, and the capacitance of C5, should then be adjusted to load the stage to the required rating. Finally, the value of R2 and R4 should be adjusted to obtain maximum r.f. output for a given d.c. input to the stage, without exceeding the rated anode and screen dissipation of the valve. After the anode circuit has been tuned, h.t. voltage may be applied without grid drive, as the clamp valve effectively reduces the input to anode and screen of the valve to a very low value.

A 10 WATT TRANSMITTER FOR 70cm USING A GROUNDED GRID AMPLIFIER

It is common practice to use double tetrode valves such as the QQVO3-20A/TT20 for transmitters operating in the frequency range 200-500 MHz. Towards the upper frequencies, the driving power requirements for such valves rise rapidly, complete stability is not easy to achieve and the balanced anode circuit is often an inconvenience. An alternative approach is to use grounded grid triodes which are inherently more stable and which give an equally good power gain above 400 MHz. The pin-based A2521 is readily usable in this arrangement.

Fig. 6.62. Circuit diagram for the 10 watt 70cm transmitter. An alternative valve for V1 is the 12AT7, and for V2 the EF91 or 6AM6.
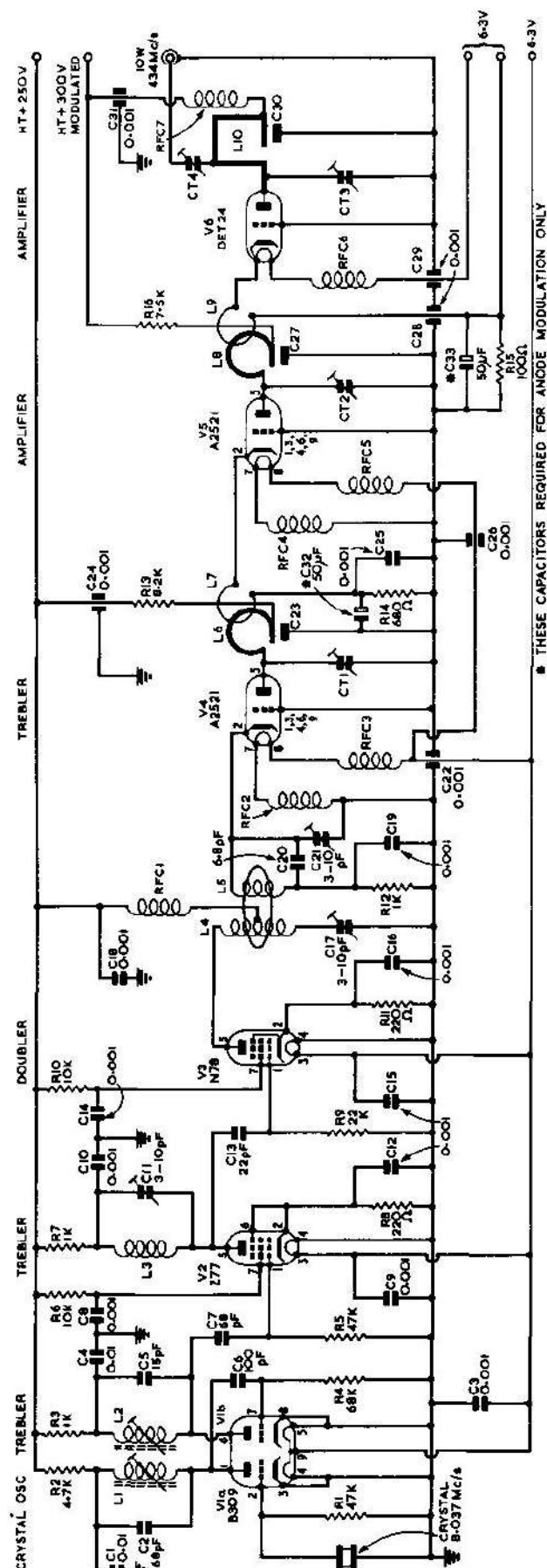
The transmitter now described uses a pair of A2521 followed by a DET24 disc seal valve to raise the output power to 10 watts when anode modulated or 15 watts on c.w. Disc seal valves are normally used with coaxial circuits, but at frequencies up to 500 MHz extremely simple lumped circuits may be used.

Circuit Description

The crystal controlled multiplier uses three valves to produce 750mW at a frequency of 144 MHz from an 8 MHz crystal oscillator. Series tuning of the N78 (V3) anode allows more latitude in the construction of the coil than would be possible with parallel tuning. The anode coil of the N78 and the cathode coil of the A2521 trebler (V4) are coupled by a single turn of p.v.c. covered wire pushed into each coil. The anode coil of the A2521 trebler and amplifier stages are identical single turn loops and each is coupled to the following cathode by an untuned loop adjacent to the anode coil. The DET24 (V6) anode circuit consists of a square loop from which the output is taken through a coupling capacitor close to the anode. All three grounded grid stages employ cathode bias; as one side of the heater and cathode of the DET24 are common, this necessitates a separate 6.3V heater supply for this valve.

Since an appreciable part of the power output reaching the aerial is fed through from the A2521 amplifier, it is necessary to anode modulate both the DET24 and the A2521 amplifier (V5) and to bypass the cathode resistors of both stages to audio as well as radio frequencies. The anode supply to the A2521 from the modulated h.t. line should include a resistor to reduce the voltage to 160V at 18mA. If anode modulation is not applied, then the A2521 may be fed directly from the 250V h.t. line and the DET24 anode supply may be increased to 400V. The A2521 anode current should never exceed 18mA, nor the grid current 6mA.

Typical performance figures for the DET24 output stage are:[1]

Class C Telegraphy

Va	350	300	250	V
Ia	92	81	73	mA
Ig	38	39	40	mA
Pload	15.0	12.8	10.3	W

VALVE ANODE MOUNT, INTEGRAL WITH LOOP AND MEASURING 1.1/2" x 1.1/2" SQUARE WITH A 1.1/8" DIA CENTRAL HOLE. CLAMP PIECE IS 3/4" LENGTH OF 1.1/8" I/D x 16 swg COPPER TUBE, WITH SQUARE COPPER FLANGE 1.3/8"x 1.3/8" SOLDERED TO TUBE. ANODE MOUNT IS TAPPED 6BA AT FOUR CORNERS COINCIDENT WITH 6BA CLEARANCE HOLES IN CLAMP FLANGE.

THIS SIDE OF CHASSIS SHOULD BE DETACHABLE TO FACILITATE CHANGING THE DET24 VALVE

VENTILATION HOLES 1/4" DIA DRILLED IN TOP AND SIDE OF CHASSIS

SEALECTRO CONTACT FINGERING TYPE CF/01/45

LOOP CLAMPED TO CHASSIS WITH 6BA BRASS SCREWS AND NUTS WITH P.T.F.E. BUSHES FOR INSULATION
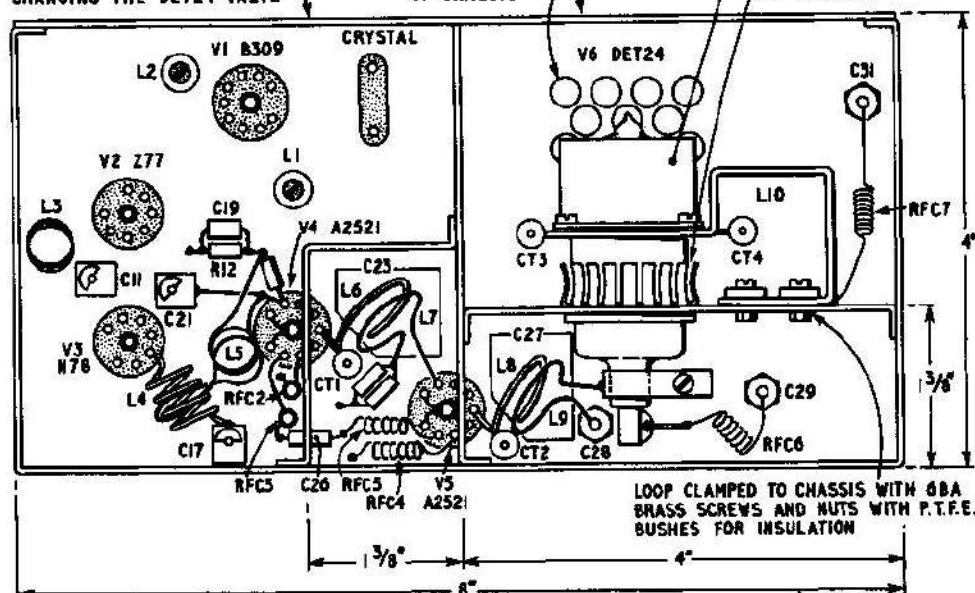
Fig.6.63. Layout of the underside of the complete transmitter, showing the position for coils, associated capacitors and resistors.

Class C Telephony anode modulated (carrier conditions)

Va	300	250	V
Ia	77	68	mA
Ig	30	34	mA
Pload	10.9	8.8	W
Pmod	11.6	8.5	W
Zmod	3.9	3.7	K ohms

Construction

The crystal controlled multiplier is conventional and simple to construct, following established principles.

Copper screens are soldered in the positions shown in Fig.6.63 and pins 1, 3, 4, 6, and 9 of the two A2521 Valveholders should be soldered directly to the appropriate screen. The anode circuits of the A2521 tripler and amplifier are identical. A length of 10 s.w.g. (¼" diameter) copper wire is bent around a ¾" mandrel; one end is straightened over a length of approximately ¼", and this is soldered to the top surface of a small copper plate measuring ¾" x 1"; the other end of the loop is cranked to allow connection to the anode pin of the valve base and at the same time allow a small copper tube (CT1 and CT2) to be soft soldered to the loop as near as possible to the anode but projecting over the chassis, clear of the valveholder. A hole is then drilled and tapped 6BA through the top of the chassis in line with the axis of the copper tube; a 6BA brass nut is soldered on to the

Fig.6.64. (a) Cooling clamp for the cathode of V6. (b) Method of bending metal to form L10, the anode loop for V6. The diameter in (a) is 0.39".

top surface of the chassis, over the tapped hole, to give increased stability to the 6BA screw which forms the tuning control.

The copper plate which forms the bypass capacity to earth (C23 and C27) should be "tailored" to provide clearance of the valveholder and tuning control. The plate is insulated by a thin mica sheet and fixed to the chassis by two 6BA nylon screws. It is necessary to countersink the 6BA clearance holes on the underside of the chassis to prevent voltage flashover around the edges of the screw holes. In the case of the A2521 amplifier, one nylon screw is replaced by a brass screw and p.t.f.e. bush in the chassis to provide an external connection for the anode supply to that stage.

The DET24 anode circuit is made from a strip of 16 s.w.g. copper (Fig.6.63). The anode mount is 1½" wide and the strip is trimmed over the length of the loop to 1¼" wide. The anode flange of the valve seats onto the anode mount and the anode clamp, which acts as an additional heat sink, is fixed to the mount by 6BA screws at each corner. It is essential to keep the temperature of the valve anode seal below 140°C and it is necessary to provide good heat conduction to the chassis, the integral mount and loop of copper together with the use of thin mica for insulation, meet this requirement. Nylon screws are not suitable for clamping the loop to the chassis screens due to the temperature of the chassis, so that it is necessary to use p.t.f.e. bushed holes and 6BA brass screws and nuts. The grid contact is formed by soldering contact fingering directly to the chassis screen and arranged to give firm contact with the grid ring of the valve. A clamp on the cathode tube of the valve assists in keeping it cool and provides a convenient anchorage for the coupling loop.

It is essential to provide a well-fitting base plate to the chassis in order to prevent direct radiation from the output circuit. One side of the chassis should be made removable so that the anode clamp of the DET24 can be unscrewed if the valve has to be changed.
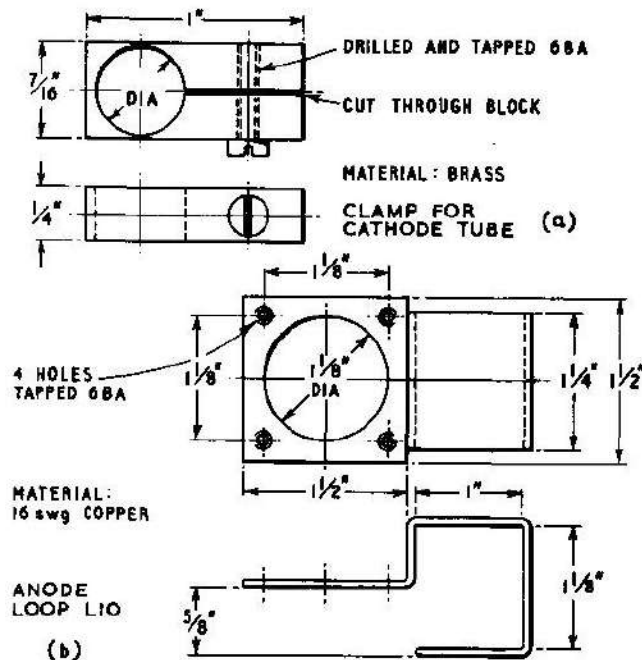
70cm HIGH POWER AMPLIFIER WITH BOX CAVITY ANODE CIRCUIT

Cavity Design

The interest in using a cavity for one valve lies mainly in the simplicity of construction, since the cavity has no advantage in either efficiency or physical size over the alternative coaxial line circuit.

It can be shown that an unloaded cavity, square in shape, has a resonant frequency equal to $\frac{C}{a\sqrt{2}}$ where C = 300 x 10^6 and a = length of the cavity in metres. Using this formula as a starting point, the actual cavity size when loaded with a valve is
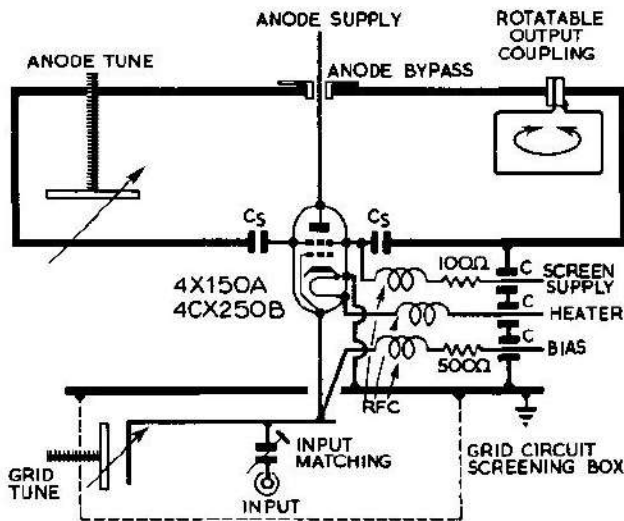
Fig.6.66A. Circuit of the 70cm box cavity power amplifier. C_S is built-in; $C = 1000$ pF feed-through.

determined by "cut-and-try" methods. In the cavity to be described, the size is fixed to resonate at 440 MHz, and a trimmer capacitance added which allows the cavity to be tuned from 420 MHz to 440 MHz. A d.c.

blocking capacitor is fitted to the top surface of the cavity, and includes a spring finger assembly to make contact with the valve anode. An Eimac type SK-610 air system socket is clamped by a ring to the inside lower surface of the cavity. With the valve in place the cavity is then effectively between anode and screen. A loop attached to a type "N" socket is arranged to rotate within the cavity to couple the output feeder.

The input circuit is a simple trough line of half-wave electrical length, tuned by a trimmer capacitor at the end remote from the valve. Input matching is achieved with a fabricated trimmer capacitor in series with the input feeder. The whole grid circuit is conveniently housed in an Eddystone die-cast aluminium box, the bottom of which is fixed to the skirt of the air system socket. The space between the diecast box and the lower surface of the cavity is enclosed, and becomes an r.f. "dead space" Heater and d.c. screen connections to the valveholder are made within this space.

Construction

Fig.6.65 shows the basic construction of the anode cavity resonator, which measures 13½" square
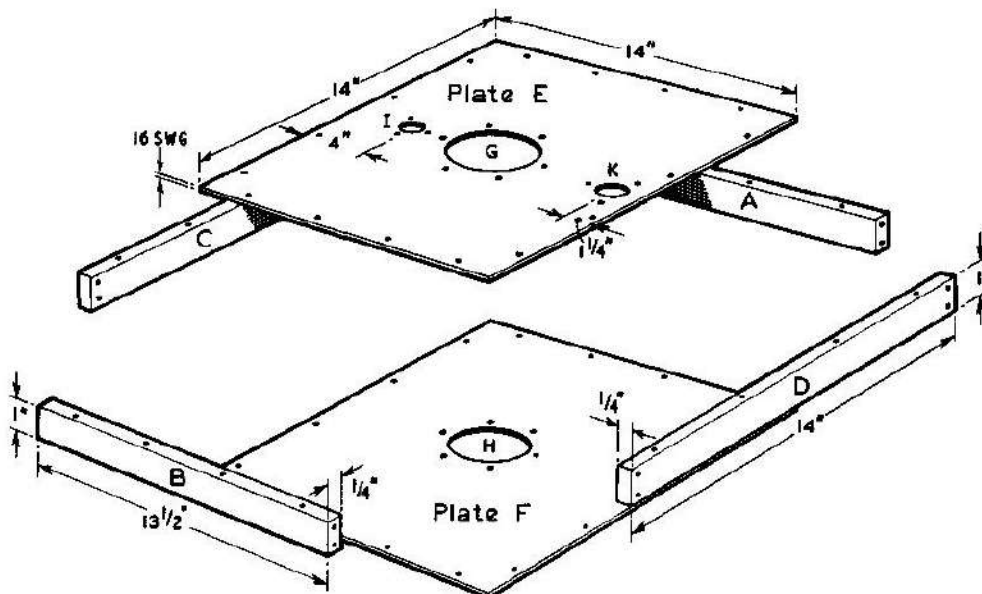
Fig. 6.65. Construction of the main cavity for the high power 70cm amplifier.

A, B, C, D,1" x ¼" brass bars making the side walls of the cavity

A, B, 13½" long

C, D, 14" long

E, F, Top and bottom plates of cavity made of ⅛" aluminium, 14" x 14"

G on E, 3" diameter hole centrally placed with six 4 BA clearance holes equally spaced on 4" P.C.D. to enable the d.c. blocking capacitor to be clamped on the outside of the plate.

I on E, ¼" diameter hole spaced 4" from edge of plate with three 6 BA clearance holes equally spaced on ¾" P.C.D. to permit fixing of the trimming capacitor support.

K on E, Output coupling hole 1" diameter space at 1¼" from edge of plate with three 4 BA clearance holes equally spaced on 1½" P.C.D.

H on F, 2½" diameter hole centrally placed with six 6BA clearance holes equally spaced on 3" P.C.D. to permit the valveholder to be clamped on the inside of the plate.
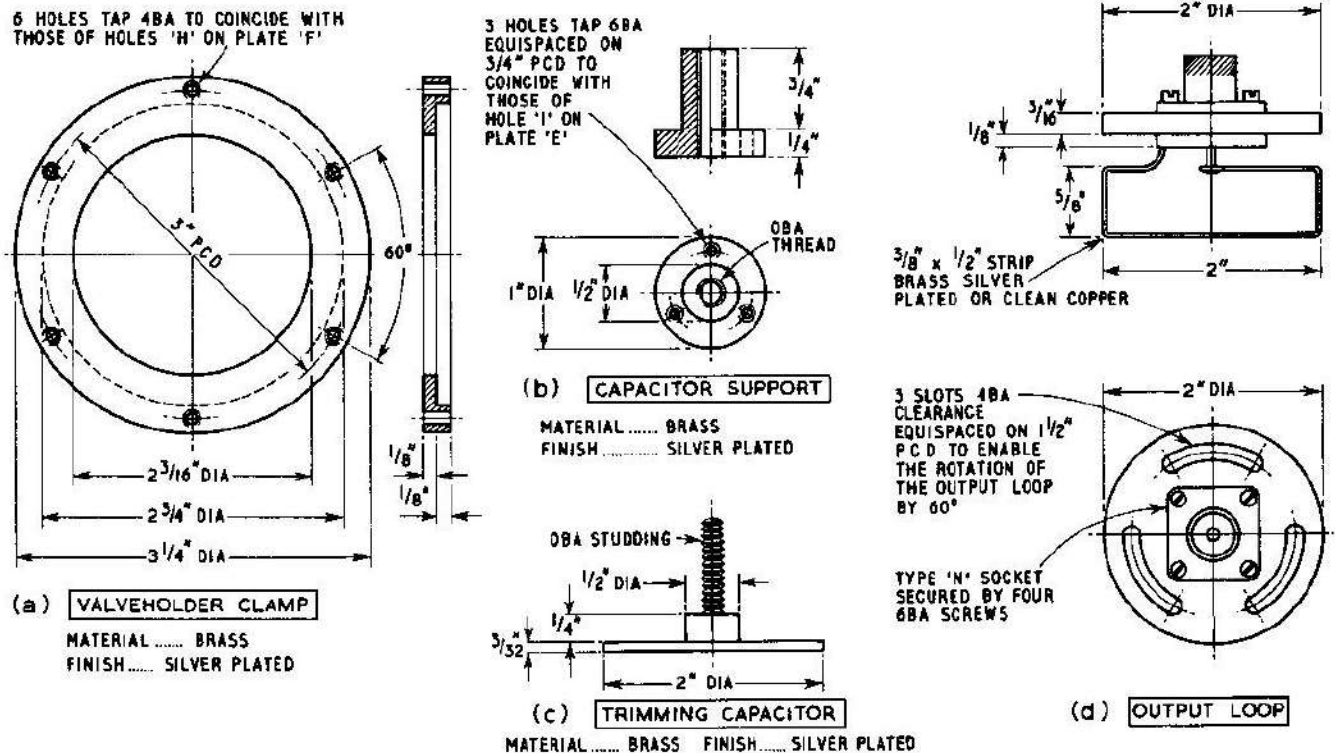
Fig.6.66. (a) Valveholder clamp (b) Capacitor support (c) Trimming capacitor (d) Output loop.

by 1″ deep internally. The internal dimensions control the resonant frequency of the cavity, and this should be borne in mind if it is desired to replace the 1″ × ¼″ brass bars specified by aluminium "U" section which may be either fabricated from sheet or extruded. The trimmer capacitance mounting is placed as close as possible to the valve for maximum effect, and the output loop assembly is placed as near the outer edge of the cavity as possible. The reason is that the voltage and current distribution within the cavity is such that the voltage is a maximum at the centre of the cavity, and where capacitative loading has maximum effect, and the current is a maximum at the outer edge of the cavity where inductive loop coupling will be a maximum. Figs.6.66 (a), (b) and (c) show the construction of the valveholder clamp and trimmer capacitor. The valveholder clamp is shown as a complete ring which ensures intimate contact between the socket and the cavity. In the event of difficulty in fabricating this ring, an alternative would be to use the three small clamp pieces supplied with the valveholder. It may be necessary, however, to fabricate three additional clamp pieces to ensure good contact around the periphery of the valveholder.

The main requirement as far as the trimmer capacitor is concerned is that the disc face should rotate parallel with the surface of the cavity.

Fig.6.66 (d) shows the output loop assembly which should be fitted to the cavity with the loop in line with the valve for maximum coupling.

The anode capacitor plate with ¾″ diameter fixing holes to facilitate fitting of the insulating bushes is illustrated in Fig.6.67. This assembly may be
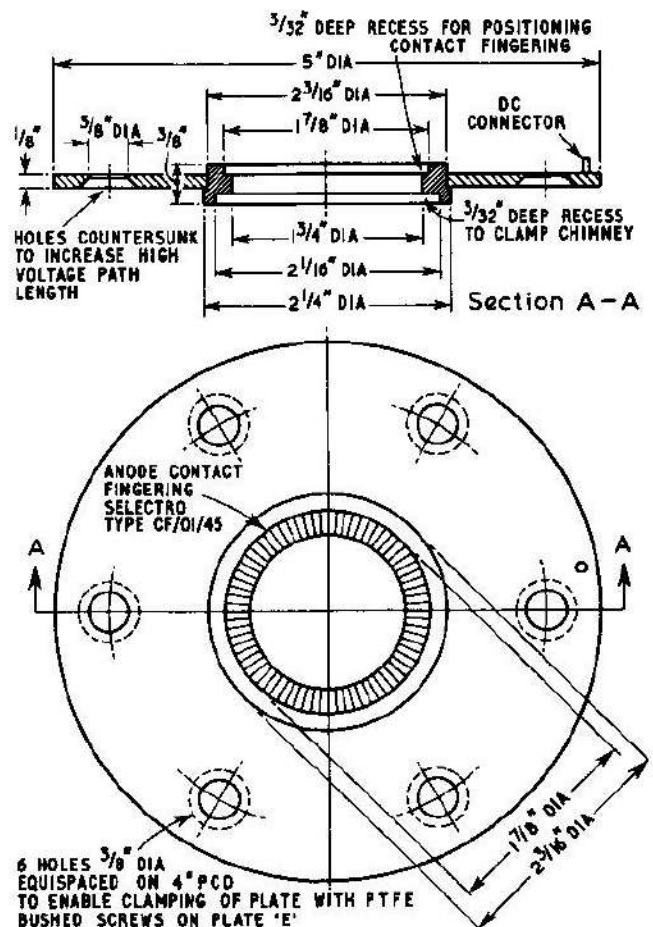
Fig.6.67. Anode capacitor plate.
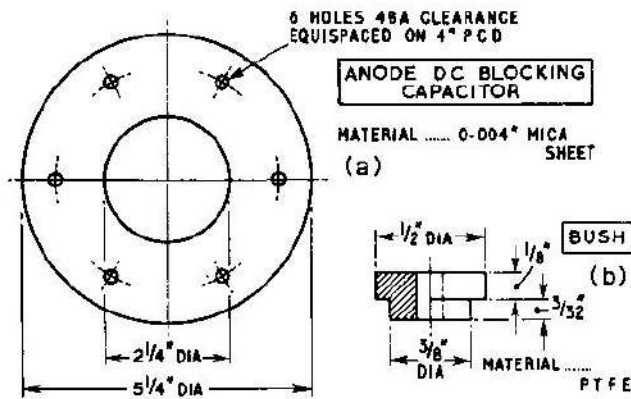
Fig. 6.68. (a) Anode d.c. blocking capacitor (b) P.t.f.e. bush.
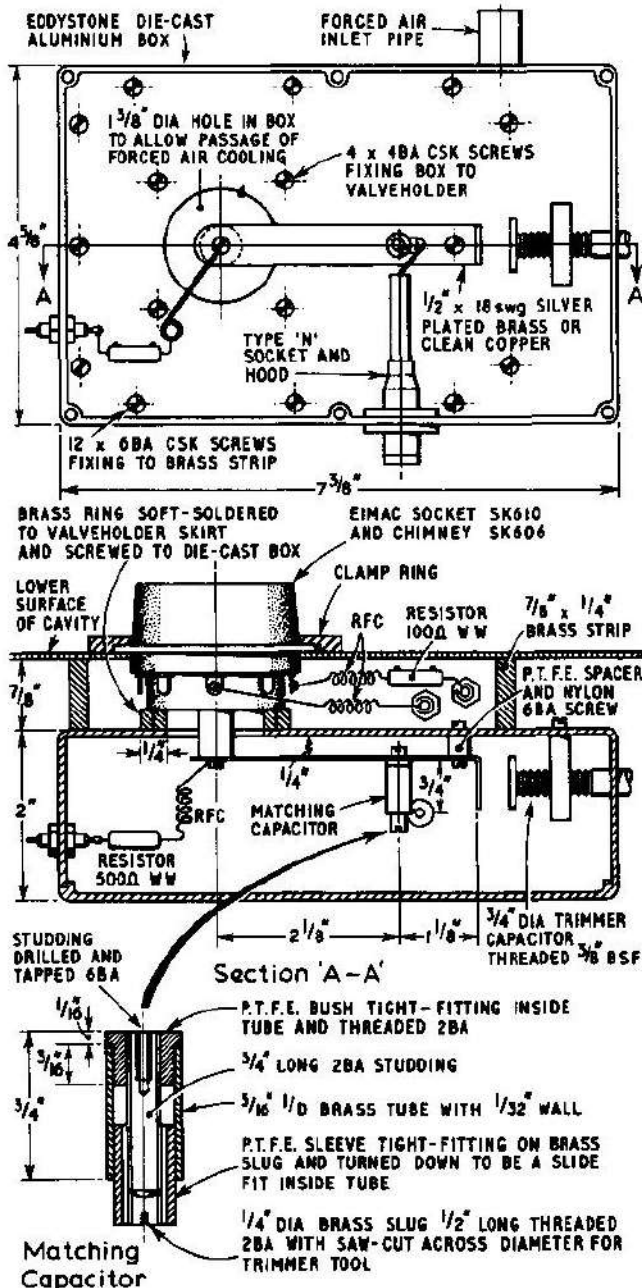
Fig. 6.69. Construction and layout of the grid circuit for the 430 MHz amplifier.

simplified by using a square form rather than circular, but should be of the same effective area, e.g. 4½" x 4½". A recess is provided for fitting spring finger contacts which make connection to the anode. A recess on the underside is provided to locate the ceramic chimney, part of the air system socket.

The insulator for the d.c. blocking capacitor and the insulator bushes are shown in Fig.6.68. Mica was used on the prototype cavity, but p.t.f.e. or polythene sheet would be equally satisfactory. Similarly, the insulator bushes were p.t.f.e. on the prototype but since no r.f. voltage appears across these bushes other materials may be used, e.g. Tufnol, Perspex or polystyrene.

Fig.6.69 shows the layout of the grid circuit within the diecast aluminium box. The grid line is fabricated from ½" wide by 18" s.w.g. strip and one end is bent through a right-angle to form one plate of the trimmer capacitor. The line is screwed to the grid connector of the air system socket with a short tube to space the line ¼" from the bottom of the box, and a ¼" p.t.f.e. spacer serves to support the other end of the line. The type "N" input socket is fitted with a type "N" hood on the inside of the box which, in conjunction with a short length of 50 ohm coaxial cable, preserves the continuity of the input feeder to the matching capacitor. The d.c. grid connection consists of an r.f. choke and wire wound resistor in series between the grid of the valve and a 1000 pF feed-through type capacitor.

The complete amplifier is illustrated in Fig.6.70. The r.f. dead space between the cavity and the grid circuit is enclosed by ⅛" x ¼" brass strip and fitted with two 1000 pF feedthrough capacitors for the screen and heater leads. An r.f. choke is fitted in the heater lead, and another r.f. choke with 100 ohms wire wound resistor in series connected in the screen lead.

Operation

It is necessary to monitor the anode, screen and grid currents to the valve, and meters should be connected in circuit. A forced air supply should be connected to give a minimum flow of 7.5 c.f.m. through the anode cooler even with only the heater supply switched on. After allowing at least 30 seconds for the heater to warm up, the grid bias, anode and screen voltages should be applied in that order.

The grid circuit may then be tuned and matched to the drive source, using the anode current meter as an indicator. Due to secondary emission effects, grid current may not be indicated even when full drive is applied, and may even be negative. In this respect it is necessary to fit a bleed resistor across the negative bias supply drawing approximately 20mA to stabilize the supply under negative grid current conditions. These remarks also apply to the screen supply. The bias voltage should be −80 volts and the screen voltage 250 volts. The drive requirement is 15 to 20 watts.

The grid circuit may be similar to that previously described and built into a cast box. Above this a plate is mounted and spaced to provide the necessary room for the supply leads.

A flange (detail L) is made to fit into the lower rim of the valve socket skirt, this is fixed to the cast box and completely isolates the input and output circuits, as well as providing a suitable air duct for the cooling air which must be blown up from the underside of the valve socket and through the anode cooling fins. The valve socket is attached to the extra top plate by the normal clips.

The anode circuit outer, Detail A, is attached to the top plate by four eyebolts at the bottom and then

Fig.6.70. Showing assembly of the complete 70cm amplifier.

With an anode voltage of 1000 volts and current of 200mA, an r.f. output of approximately 100 watts may be obtained.

A COAXIAL CAVITY FOR 70cm

As an alternative to the box type cavity a tubular or coaxial type offers a considerable space saving, but of course a significant amount of machine work will be needed.

Fig.6.71 shows the general arrangement of the anode circuit and Fig.6.72 gives the mechanical details of the various component parts. Detail (A) gives dimensions for either a 4X150A or 4CX250B valve.

Fig.6.71. General arrangement of anode circuit of the coaxial cavity.
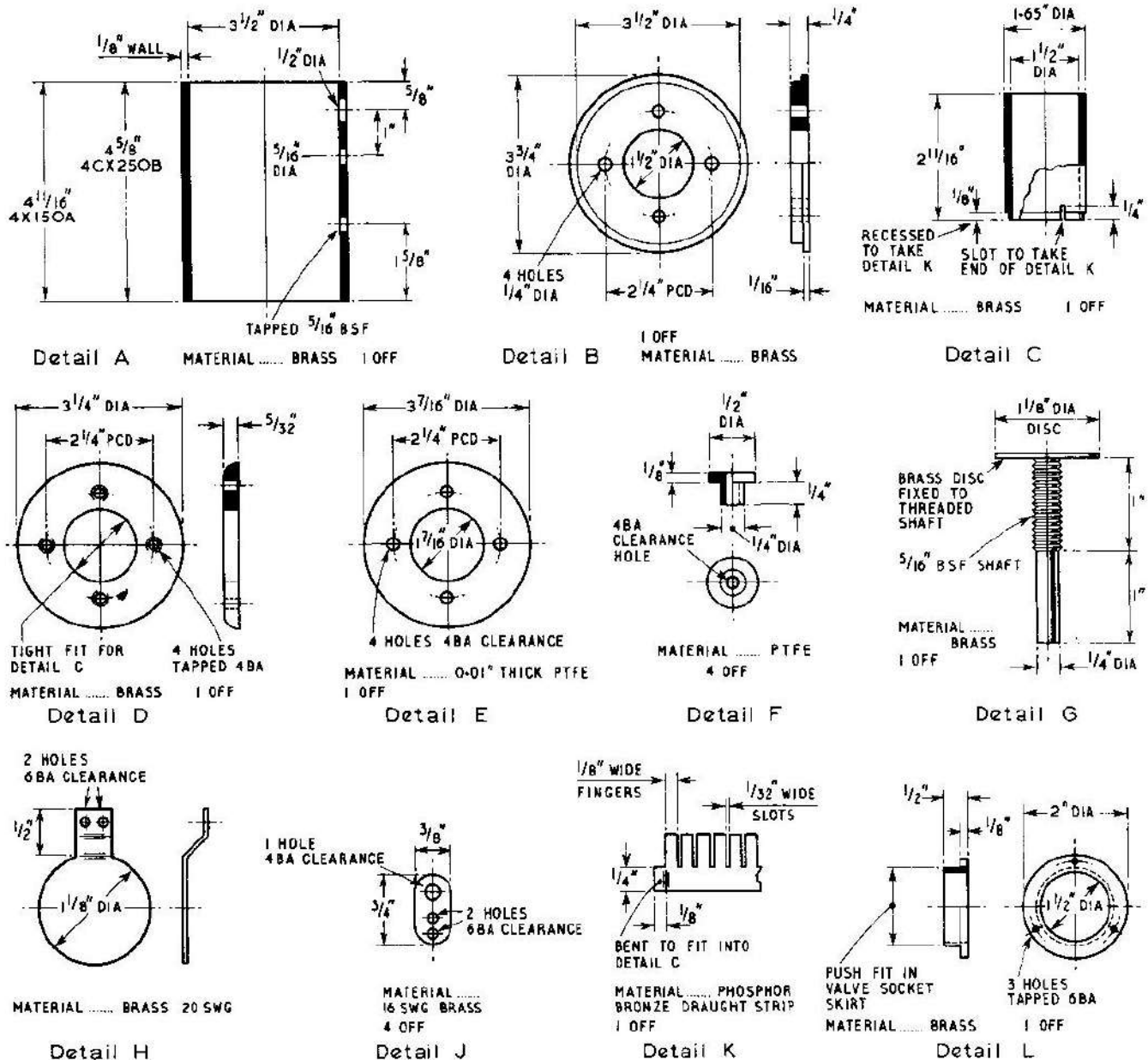
Fig. 6.72. Mechanical details of construction of the coaxial cavity.

the inner tube assembly consisting of the anode line (C) is soldered or brazed to the inner top disc (D) which is bolted to the outer top disc (B) being isolated from it by the p.t.f.e. washer (E). The fixing bolts of the two top discs are insulated from the outer top disc by p.t.f.e. bushes (F). The whole assembly is then attached to the outer tube by four lugs into which eyebolts at the top end of the tube locate.

Tuning of the cavity is provided by a disc type capacitor as detail H and G. The adjustable element (G) should be provided with some tensioning device such as a spring locating into the screw thread. Since there is the full h.t. between these two plates one of them should be covered with suitable insulating material to prevent a possible short circuit.

Output from the cavity is by the conventional series tuned loop. The loop spacing from the anode line should be adjusted to give the maximum output; this does not need close coupling to the centre tube and it should be kept as far away as possible consistent with good efficiency.

23cm TRIPLERS

At this frequency the choice of valves is for all practical purposes limited to two basic types, the DET24 for low power and the 2C39A for high power. The DET24 is a conduction cooled type with an anode dissipation of 20 watts while the 2C39A has an air cooled anode of 100 watt rating.

Two designs of circuit are described, a cavity type and a strip (or slab) line type, both of which are based on the 2C39A valve.
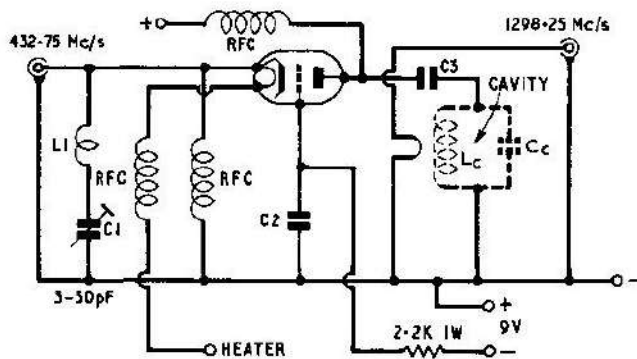
Fig.6.73. Circuit of the tripler cavity. C2 and C3 are formed by the anode and grid plates in the cavity; Cc and Lc represent the cavity; L1 is formed from thin copper strip ¾" wide and 2¾" long bent to a 'U' shape; r.f. choke, 2½" of 20 s.w.g. enam. copper, wound on ¾" diam. former.

Cavity Type

From the circuit diagram of Fig.6.73 it will be seen that the valve operates in the grounded grid mode, the drive being applied directly to the cathode. The 23cm output is taken from a loop Fig.6.77 which couples into the cavity.

Construction

The cavity is constructed from 20 s.w.g. sheet brass, the general details of which, together with part of the assembly order, are shown in Fig.6.74. Full details of the main body of the cavity are shown in Fig.6.75. This is constructed by forming up the four sides of a flat sheet to produce an open box. It should be noted that the corners are not soldered. This permits easy movement of the grid tray which is a sliding fit inside this open box. It also allows a good contact to be made between the grid tray and the box when the side fixing screws are tightened.

The position of the grid tray in the box is the coarse adjustment for the anode circuit. Fine tuning is achieved by means of a tuning paddle which is also illustrated in Fig.6.75. This consists of a ⅜" x ¾" metal plate which is soldered into a slot cut into the end of a ¼" operating spindle.

Output from the cavity is taken via the ½" inside diameter tube mounted directly opposite the fine tuning control. Into this tube slides the aerial coupling probe illustrated in Fig.6.77. It is important to ensure that the probe unit is a reasonably tight fit into the tube on the side of the cavity and to this end, if possible, the two parts should be constructed from telescopic tubing. Since the diameter of the outer tube is not critical to within ½₂" it is permissible to cut a lengthwise slot in it to allow it to be closed up slightly and thus ensure a tight fit round the probe.

The anode and grid plates are shown in Figs.6.78 and 6.79 and respectively, and as will be seen they are quite straightforward. Careful attention must be paid to the contact fingers for the valve, and it is important to ensure that the final size is suitable for the particular sample of the valve being employed. The fingering is constructed by making suitably spaced cuts with a very fine saw. This strip is then formed into a circle, fitted to the inside of the hole provided, and soldered into position. Small adjustments to the
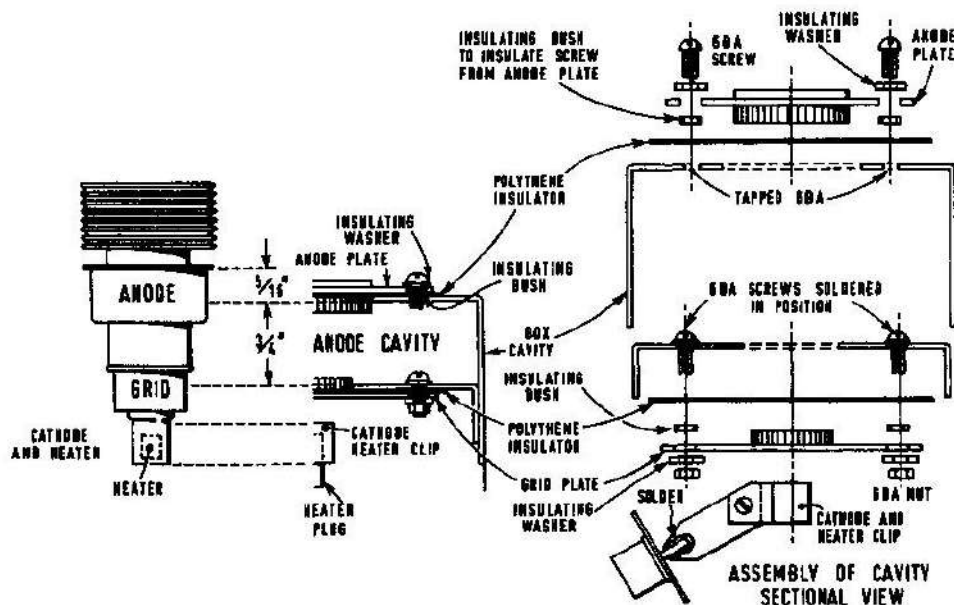
Fig.6.74. Assembly of cavity. The position of the valve is very important; it should be placed for maximum output, but this will affect the tuning. Should the tuning paddle not correct the tuning, the grid tray will need to be moved slightly.
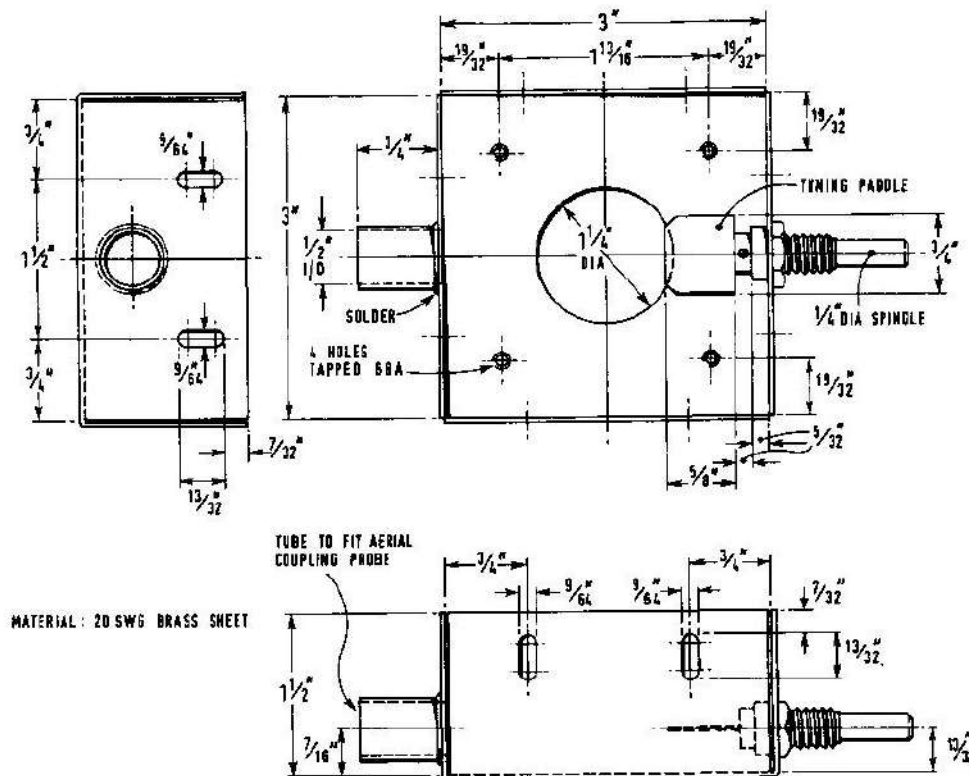
Fig. 6.75. Box cavity body.

final size may be achieved by slightly bending the contact fingers.

Polythene sheet of a thickness of 0.008" suitably cut to shape, is used as insulation between the anode plate and the cavity body. This is satisfactory for anode voltages up to 600 volts. Between the grid plate and the grid tray, a similar sheet of polythene is used.
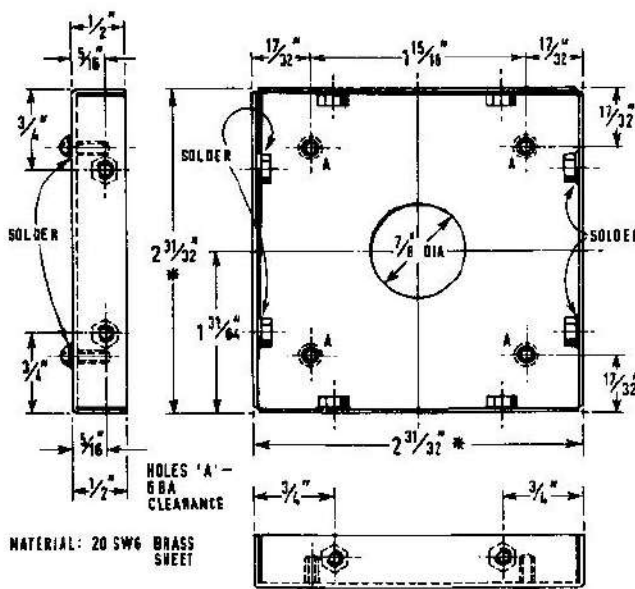
Fig. 6.76. Grid tray.

The coaxial input socket is mounted on the bracket shown in Fig.6.80 and this is held on to one side wall of the cavity by means of two of the screws which hold the grid tray in position. The lead from the coaxial socket to the cathode of the valve is made from thin copper strip ¾" wide. If copper strip is not available brass may be used so long as it is thin. It is important that the lead be not too rigid as otherwise there may be a danger of damaging the valve.
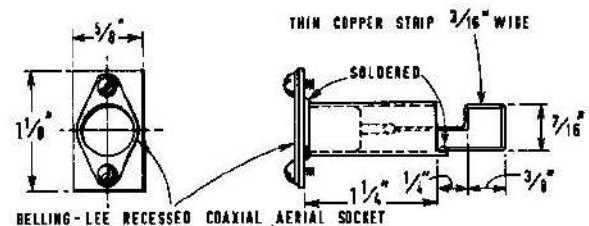
Fig.6.77. Aerial coupling probe.

The input inductance (see L1 in Fig.6.73) consists of a length of brass or copper strip, 2⅛" long by ¼" wide formed into a U with arms of equal length. With the base of the cavity uppermost, this U-shaped inductance is inverted and soldered at one end to the coaxial socket, at the other end to the 3–30 pF Philips trimmer. The earthy side of this trimmer is soldered to a tag located under one of the two 6BA screws in the corner of the grid tray farthest away from the coaxial input socket.

The 2.2 K ohm grid resistor is soldered to a tag secured by a short 6BA screw to the centre of one side

of the grid plate. It is particularly important that this screw, and its counterpart on the anode plate, are filed off flush with the underside of the plate, otherwise the polythene insulation will be damaged, and a short circuit may occur.
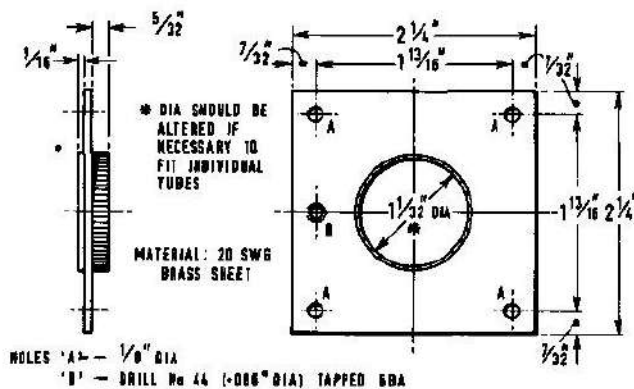
Fig. 6.78. Anode plate.

Adjustments and Operating Conditions

Connect the 9V bias supply to the valve with a 0—50mA meter in series with the negative lead. Apply a 6V source, capable of supplying 1A, to the heater of the valve. Do not attempt to apply h.t. at this stage, but connect the anode of the valve from the solder tag on the anode plate via a 100mA meter to the cavity body.

If 70cm drive is applied to the cathode, grid current will flow, and some anode current will show on the 100mA meter. The trimmer C1 and the inductance L1 should be adjusted to give maximum grid current. This should be in the region of 25mA. The earthing point of the trimmer may need changing to a different point on the grid tray if insufficient grid current is noted. There are a number of combinations of the capacity of C1 and the inductance of L1 which will give resonance at 70cm, the correct ratio of these two is quite critical in order to achieve maximum drive to the tripler. Once the grid current is of the correct order, then the meter in the anode circuit should read between 30 and 60mA.
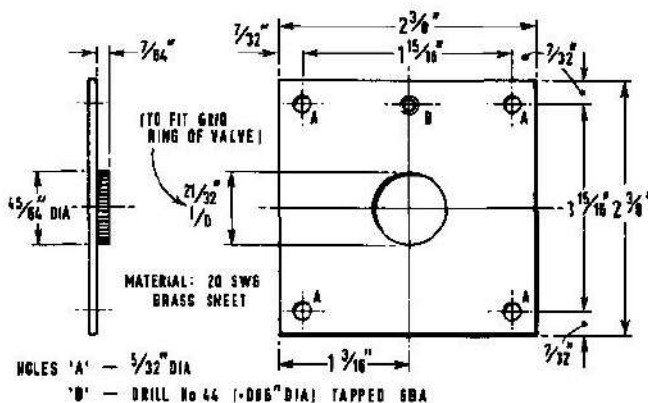
Fig. 6.79. Grid plate.
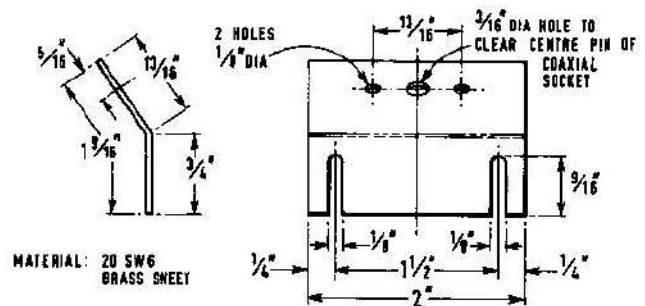
Fig. 6.80. Bracket for the Belling-Lee coaxial input socket.

When a suitable level of grid current has been achieved, attention can be turned to resonating the anode cavity. For this it is necessary to have some means of indicating 23cm output. One possibility is to use a 6V 0.04A flash lamp bulb soldered directly to a coaxial plug. Alternatively, and in many respects better, a diode probe on a coaxial line, as illustrated in Fig.6.81 can be used.

First, set the tuning paddle to an angle of 45.° to the plane of the anode plate. Slightly loosen the four grid tray fixing screws and then very carefully move the position of the grid tray until resonance is found. Resonance will be denoted by a small dip in anode current. If the diode probe on the coaxial line is in circuit a reading will be obtained on the diode meter. Lock the grid tray into position, and check that the tuning paddle will resonate the cavity more critically. The anode may now be disconnected from the chassis and taken to a supply of about 350V positive via an r.f. choke. The anode current may lie anywhere between 50 and 100mA depending on the valve.
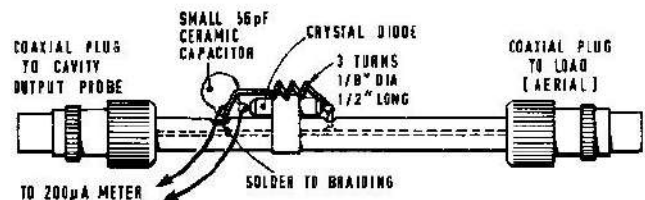
Fig.6.81. Diode probe on a coaxial line.

Acceptable levels of modulation may be achieved by modulating only the 70cm drive to the 23cm tripler. For c.w. operation, the 70cm stage can be keyed, and under key-up conditions, the anode current in the tripler should fall to zero.

Strip Line Type

This alternative design is particularly attractive because the amount of engineering is considerably less. As will be seen from the circuit diagram, the input circuit is slightly different from the cavity type. In this case the cathode of the valve is tapped down the inductance of a parallel tuned circuit which

correctly set up offers some advantage over the arrangement of the previous circuit.
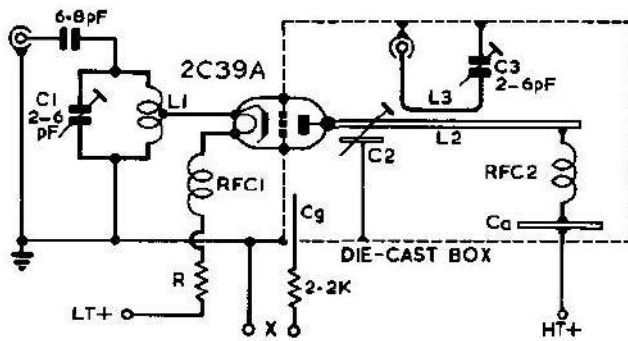
Fig.6.82. Strip line 23cm tripler.

L1 = 2 turns ⅛" wide strip ¼" dia. ¾" long cathode tap at ¾ turn from earth end
L2 & L3 = Anode strip line and output coupling. See Fig.6.84

Cg grid to chassis by pass capacitor
Ca HT to chassis by pass capacitor
}insulation 0.004/6" thick mica or other suitable material
R.f. choke 1 8 turns ¾" diam. 18 s.w.g.
R.f. choke 2 7 turns ⅛" diam. 22 s.w.g.
R value to suit so that valve heater voltage is 5.5v
C1 and C3 2-6 pF concentric trimmer (Mullard CO04/EA)
C2 Adjustable plate trimmer see Fig.6.84.
X position for grid current meter

Construction

The anode circuit is enclosed in a standard diecast box 4 ⅝" x 3 ⅜" x 2 ¼"The lid is replaced by a copper or brass plate, ¹⁄₁₆" thick with a U
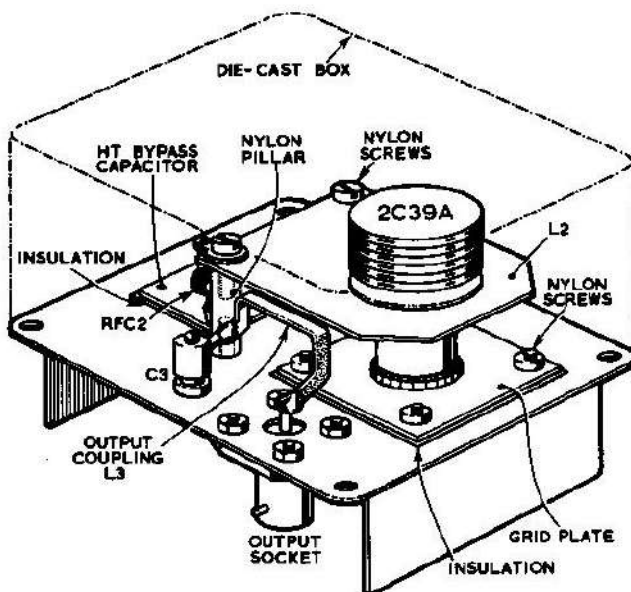
Fig.6.83. General arrangement of 2C39A tripler.

shaped screen of the same material soldered to the new plate, as shown in the illustrations Fig.6.84.

The plate forming the anode strip line is mounted ³⁄₁₆" above the chassis plate by two insulated pillars, and the h.t. supply to this is connected to a tag under one of the fixings, through the r.f. choke (RFC2) to the h.t. bypass capacitor centre fixing screw insulated under the plate, through the plate to the supply socket.

The grid which is at r.f. earth potential is connected to a 2" square plate insulated from chassis and fixed by three nylon and one metal screws. Note that the insulation is on the underside.

The screw is connected to the 2.2k resistor, which is in turn connected to the chassis plate. When it is required to read grid current this connection is lifted off earth and a meter connected between it and the chassis plate. The insulation of both the h.t. and grid bypass capacitors is 0.004/6" thick mica or other similar material.
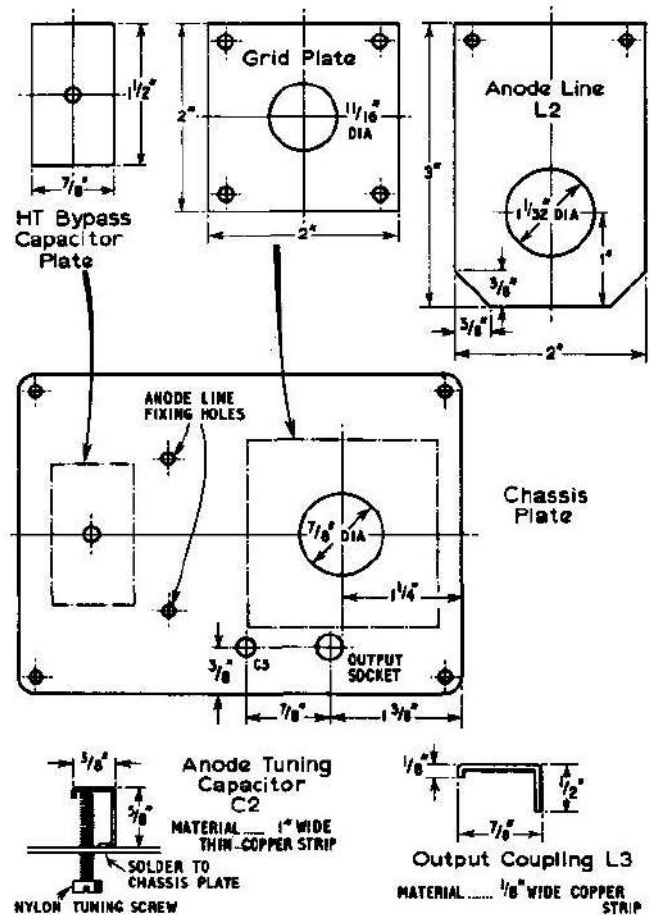
Fig.6.84. Construction details of anode and grid circuits.

As can be seen from the top view photograph Fig.6.85 the input circuit L1 C1 is fitted into the top left-hand corner close to the valve and the input socket. Contact to the valve anode and grid is made by use of a phosphor bronze fingering ¹⁄₃₂" long

soldered to the anode and grid plates, the fingering should face the same direction.

The completed unit is operated with valve anode downwards, i.e. standing on the box in the normal manner.

Adjustment

For full power operation, a drive power of 10 watts is needed. The input circuit is tuned for maximum grid current which when correctly adjusted should be between 30 and 40mA. When the input has been properly adjusted apply h.t. of 250-450 volts to the anode and adjust the anode tuning capacitor and output coupling circuit for maximum output. The indicator described for the previous unit should be used to assist in tuning up the anode circuit.
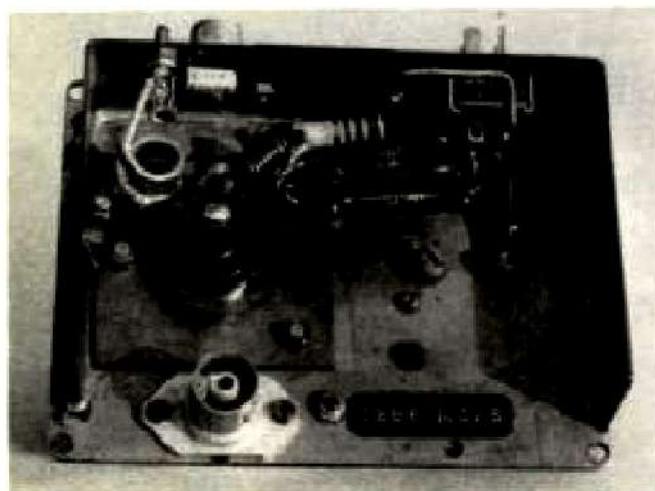
Fig. 6.85. Photo of complete 1296 MHz tripler.
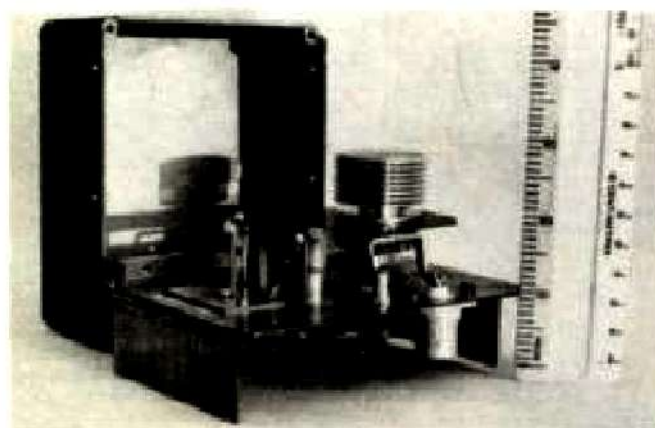
Fig. 6.86. Underside view of tripler.

With inputs up to 50 watts no forced cooling is needed, but if the valve is to be operated at its full input rating provision for forced air cooling must be made. Since the diecast box forms a cavity at the operating frequency, any holes which are cut in the sides of the box to allow cooling to be blown across the anode radiator, should be covered with mesh to prevent disturbance of the r.f. field.

For local (short range) working sufficient output can be obtained with the h.t. lead connected to chassis (no anode voltage).

SIMPLE LOW POWER TRANSISTOR TRANSMITTERS FOR 70 and 144 MHz.

Transmitter for 70 MHz

The crystal oscillator operates at 7.788-7.855 MHz with the collector circuit tuned to the crystal frequency. The second and third transistors function as triplers to reach the output frequency in the range 70.025-70.7 MHz. The output from TR3 is coupled to TR4 operating as a grounded base class B amplifier. Forward bias of about 0.1 volt is provided by the potentiometer across the supply, and VR is adjusted for maximum r.f. output. The output circuit is a simple pi-coupler to match the output impedance to the feeder, usually 50 or 75 ohms.

At a supply voltage of 15 volts TR1 should take 4mA, TR2 and TR3 between 8 and 10mA each, with this power sufficient r.f. drive should be available for a p.a. current of 25mA.

The setting up of the transmitter, should preferably be carried out stage by stage using a simple r.f. voltmeter to indicate proper tuning. The use of a sensitive absorption wavemeter is essential to ensure that each stage is operating on the correct frequency and to make checks to ensure that no additional spurious frequencies are present in the output of each stage.

An output power of 100mW should be available but higher output could be obtained by use of an alternative transistor. In significantly higher power amplifiers the use of the L-pi circuit is recommended to achieve satisfactorily the impedance step-up required.

Transmitter for 144 MHz

In this transmitter the crystal oscillator operates at 24 MHz, the second stage trebles to 72 MHz, and the third transistor TR3 operates as a frequency doubler to the final frequency in the 144 MHz band. The output is taken directly from this stage to the aerial feeder. It is most important as mentioned in the description of the previous transmitter to ensure that no spurious frequencies are present in the output. The use of an amplifier at the final frequency would be desirable and one similar to that forming part of the 500mW transmitter that follows this description would be suitable.
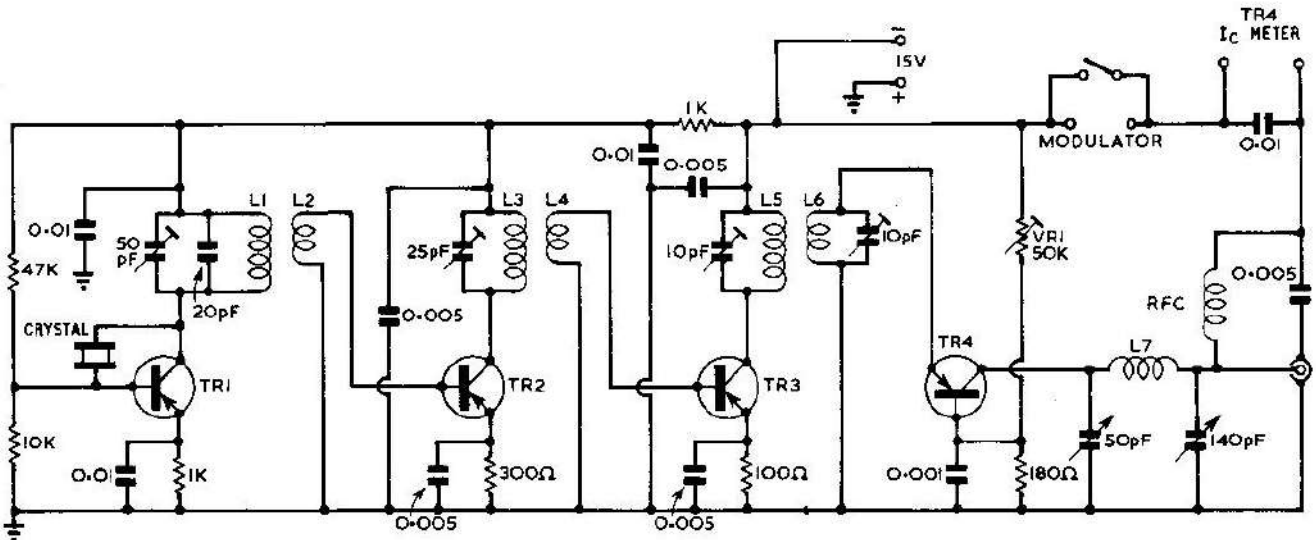
Fig. 6.87. A 70 MHz transmitter.

L1   42 turns 28 s.w.g. enam. ⅛" diam.
L2   8 turns, 28 s.w.g. enam. overwound at decoupled end of L1.
L3   18 turns, 28 s.w.g. enam.
L4   5 turns, 28 s.w.g. enam. overwound at decoupled end of L3.
L5   12½ turns, 22 s.w.g. enam., 0.3" diam.
L6   5 turns, 26 s.w.g. silk covered, interwound at low potential
      end of L5.
L7   8 turns, 16 s.w.g., silver plated 0.3 μH ⁷⁄₁₆" diam., ⅞"
      long.
RFC   42" 28 s.w.g. enam., on ⅜" diam. former
TR1   2G401, Texas; OC170, Mullard
TR2   2G402, Texas; OC171, Mullard
TR4   2G110 Texas
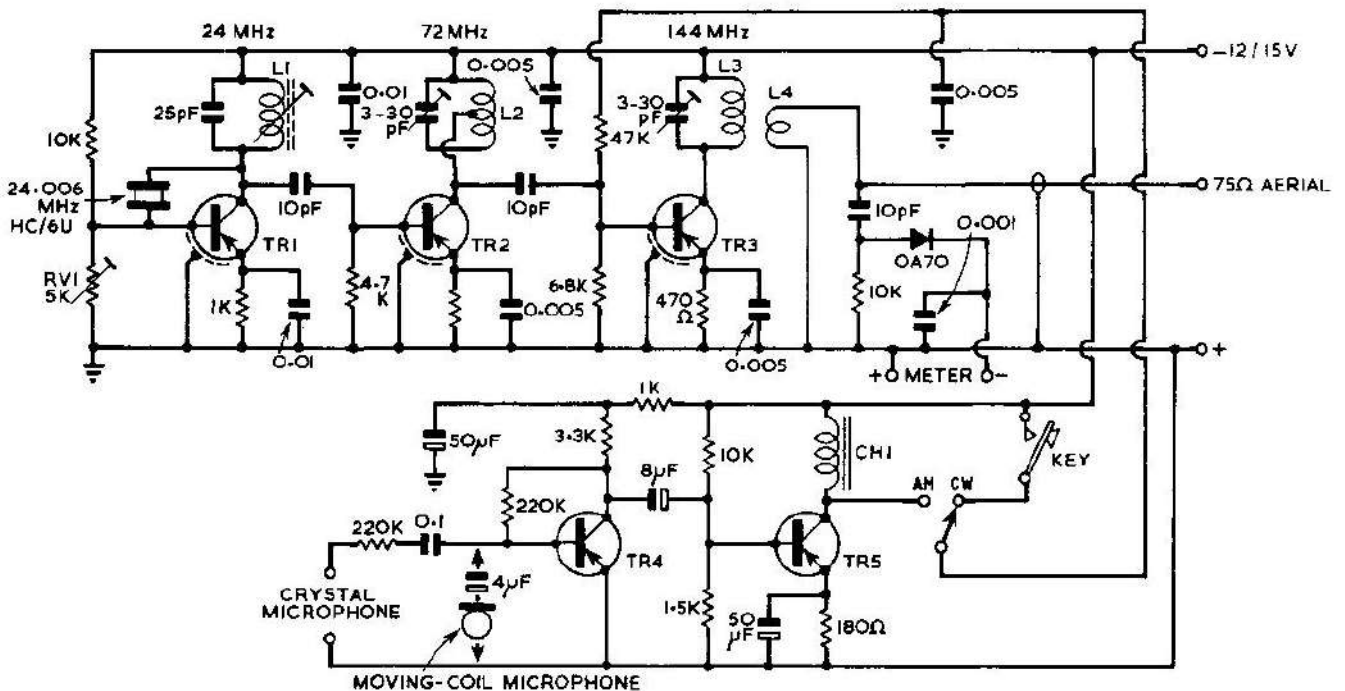
Fig. 6.88. A five transistor transmitter for 2 metres.

Coil Details (144 MHz)

L1	25T	30 s.w.g. close wound on ¼" diam. former with dust core.
L2	8T	26 s.w.g. centre tapped ¼" diam.
L3	5T	26 s.w.g. spaced one wire diam. ¼" diam.
L4	2T	26 s.w.g. insulated wound over L3.
CH1		small a.f. choke with d.c. resistance less than 250 ohms.

A 500mW Transmitter for 144 MHz.

In this more elaborate design the crystal oscillator operates at 48 MHz and therefore requires less multiplication to reach the final frequency. The second stage operates as an amplifier at the crystal frequency and this is followed by a tripler stage, with a pi-coupler to the 144 MHz driver stage. The output stage in this case is a push-pull pair of transistors with modulation applied to both the emitters and collectors. In this amplifier a parallel tuned circuit is used but some advantage may be gained by the use of a large transistor coupled to the output circuit by means of an L-pi circuit.

Coil Details

L1 8 turns 6mm diam. tapped at 4th turn, dust core

L2 4½ turns 6mm diam., dust core

L3 3 turns insulated wound over L2. 6mm diam.

L4 6 turns 6mm diam, dust core.

L5 4 turns insulated coupled to centre of L6. 6mm diam.

L6 2 + 2 turns each side of L5 centre tapped and at centre of each half.

L7 2 + 2 turns each side of L8 10mm diam.

L8 2 turns in centre of L7 10mm diam.

All coils wound with 18 s.w.g. wire.

RFC1 60" 36 s.w.g., wound on small insulated resistor as former.

RFC2 19" 36 s.w.g., wound on small insulated resistor as former.
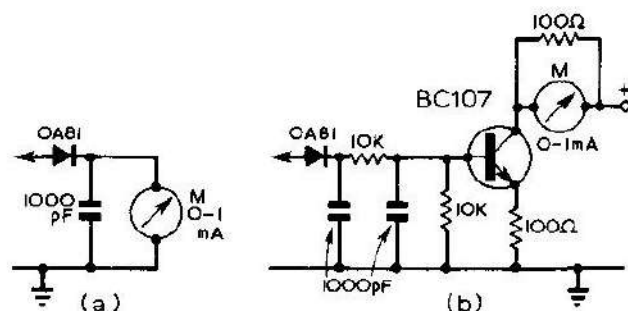
Fig.6.89. A simple r.f. voltmeter suitable as an indicator for tuning the various stages of these low power stages. For higher sensitivity a transistor d.c. amplifier may be used ahead of the meter.

OVERLAY TRANSISTORS

Although data is given in this Manual for other types of transistors in some of the equipment described there is doubt that the overlay (planar) silicon transistors are the most useful for transmitter applications. The devices now available here and from the USA offer quite large power output up to and above 430 MHz.

Overlay transistors can provide - if care is taken - a very much worthwhile increase in output power and efficiency at u.h.f. due to their low series resistance compared with other transistors. All transistors have very limited heat dissipation and they can pass a great deal more current than is good for them.

With an overlay type of structure, as with most other types of transistors, the collector-base capacity varies with the voltage on the collector, and that means that such a device can use the non-linear capacitance of the collector-base junction to generate harmonics of the input frequency, like a varactor. An overlay transistor, because of its low series resistance can generate harmonics at an efficiency comparable to a good varactor with the advantage that the output power is more than the drive.
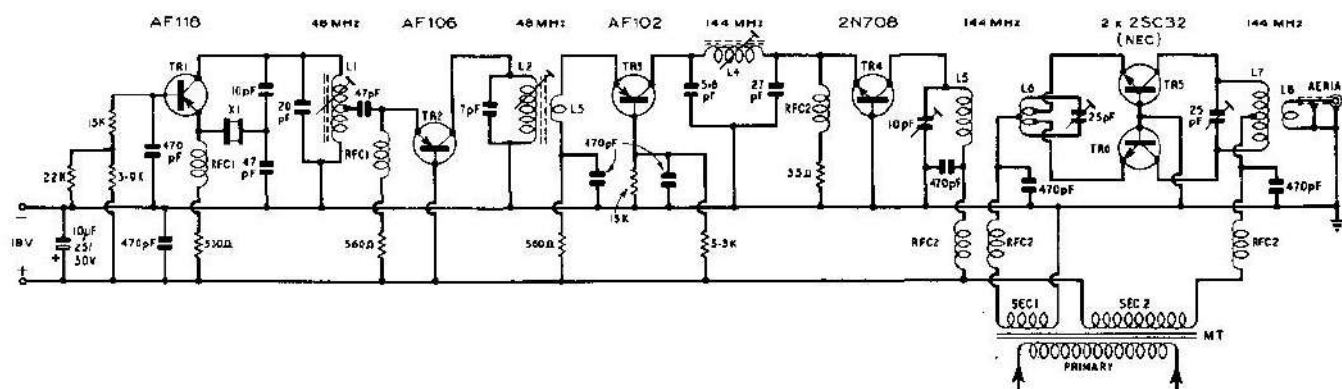
Fig.6.90. A 500 milliwatt 2 metre transmitter. The crystal is 48 MHz and the modulation transformer is a rewound miniature output transformer. Secondary 1 is 15 turns. Secondary 2 is 800 turns.

An overlay transistor operates in two completely distinct ways as a frequency multiplier. It must give gain at the input fundamental frequency and functions as a varactor diode to make harmonics.
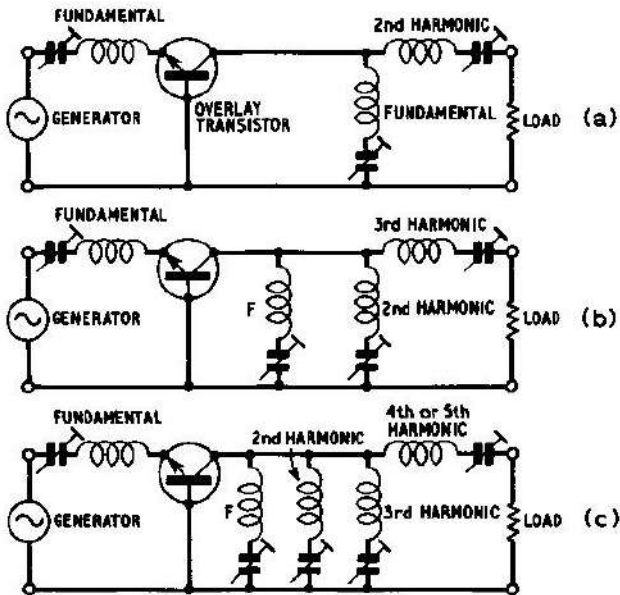
Fig. 6.91. Overlay transistor multipliers. The collector-base junction is a shunt connected varactor diode. (a) 2nd harmonic, (b) 3rd harmonic, (c) 4th and 5th harmonics.

The collector-base junction in an overlay transistor and many other types of transistor is very much like a varactor, and so the same circuit conditions apply to it with one exception, if one can call it that. A transistor is a three terminal device: the input is applied to a separate electrode from the one in which the output is recovered. Because of this, it is necessary to have a fundamental frequency idler in the output circuit to get gain at this frequency or, the same thing, to get fundamental current flowing through the collector-base junction. This idler develops the drive for a non-linear capacitance to make harmonics.
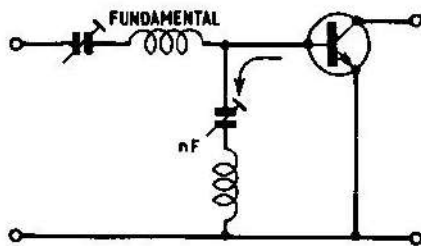
Fig. 6.92. Common emitter input circuit.

All the circuits in Fig.6.91 use transistors in the common-base mode, which is very like the shunt configuration of a varactor. If the common-emitter is decided upon, a series-resonant circuit becomes essential between the base and ground for whatever final frequency is used in the output. The common-emitter input circuit is shown in Fig.6.92. This is more akin to the series configuration of a varactor.

In the common-base multiplier there is one less tuned circuit to bother about than in the common-emitter mode. If the input drive level is low, the common-base circuit is considerably more efficient and provides more gain. If the input power is pushed up, the efficiency goes through a maximum and then starts to decrease and is overhauled by the common-emitter circuit for efficiency and output power at large input drives (around ¾W in the 2N4012). So the common-base arrangement saturates at lower power levels than the common-emitter. Therefore it seems wise to use the common-base circuit for low power stages and common-emitter for the high power or p.a. stage.

All non-linear circuit elements are especially prone to instability and it is difficult to know where to begin so as not to discourage constructors.

High frequency oscillations may be rather a nuisance. These oscillations can be very close to the multiplier output frequency and they become very evident when the drive power is removed. Sometimes they have no centre frequency and one gets a band full of noise. This type of instability can usually be cured by use of as short lead as possible between the base or emitter (as the case may be) and earth.
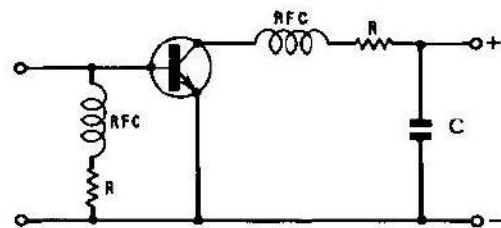
Fig. 6.93. Small resistance in power supply leads to stop low frequency oscillations. C must have a low impedance at both signal frequency and below.

Low frequency oscillation may occur due to the higher gain at lower frequencies, these may be frequencies at which the r.f. chokes could be resonant and it is best to add a low value resistor in series with the chokes as shown in Fig.6.93 to overcome the problem.

Parametric oscillations can occur if these l.f. oscillations are allowed to interfere with the harmonic output (if C in Fig.6.93 has not got a low impedance to the l.f. oscillations as well as to the signal frequency).

The use of pi or L coupling will usually be satisfactory for matching into and out from the transistor. It is however important, as mentioned elsewhere in this Manual, to include a suitable filter in the output to attenuate the unwanted frequencies to a level which will not cause out of band interference.

Fig.6.94. High Q or Bandpass filter.

The minimum should be some form of high Q circuit as shown diagramatically in Fig.6.94 but it is preferable to use a complete band pass filter possibly of the form given later in the Manual.

A circuit for a practical 144 MHz exciter using overlay transistors is shown in Fig.6.95 and component details are given in the caption.

The circuit for a suitable amplifier, and its component details is illustrated in Fig.6.96.

Fig.6.95. A 144 MHz exciter with approximately 0.5 watt output. L1, 6 turns, 18 s.w.g. spaced wire diam. ¾" diam.; L2, 2 turns, 22 s.w.g. insulated, interleaved with L1; L3, 7 turns, 18 s.w.g. spaced wire diam., ¼" diam., tuned with dust core. L4, 5 turns, 18 s.w.g., spaced wire diam., ¼" diam., self supporting; RFC1 + R5, 30 turns, 28 s.w.g. enam. nichrome wire, on ⅛" x ½" ferrite core; TR1, 2N703; TR2, 2N3553; X, 72 MHz overtone crystal; ZD1, 14-volt Zener diode, ½ watt.

Fig.6.96. A 144 MHz amplifier. L1, 1 turn, 22 s.w.g. insulated, interleaved with L2, ¼" diam.; L2, 2 turns 18 s.w.g. spaced wire diam., ¼" diam., dust core tuned, L3, 5 turns, 18 s.w.g. spaced wire diam., ¾" diam., air-spaced; L4, 5 turns 18 s.w.g., tapped 1 turn, spaced wire diam., ¼" diam., dust core tuned; L5, 1 turn, 22 s.w.g. interleaved with L4, ¼" diam.; RFC1, 2, as Fig. 6.95. TR1, 2N3553. A heat sink will be necessary, but as the collector of this transistor is connected to its case, some insulating material must be used.

Fig.6.97. A 432 MHz 150mW exciter using a 2N3553 by a GM290 mounted into a standard diecast box. The circuit for this unit is shown in Fig.6.98.
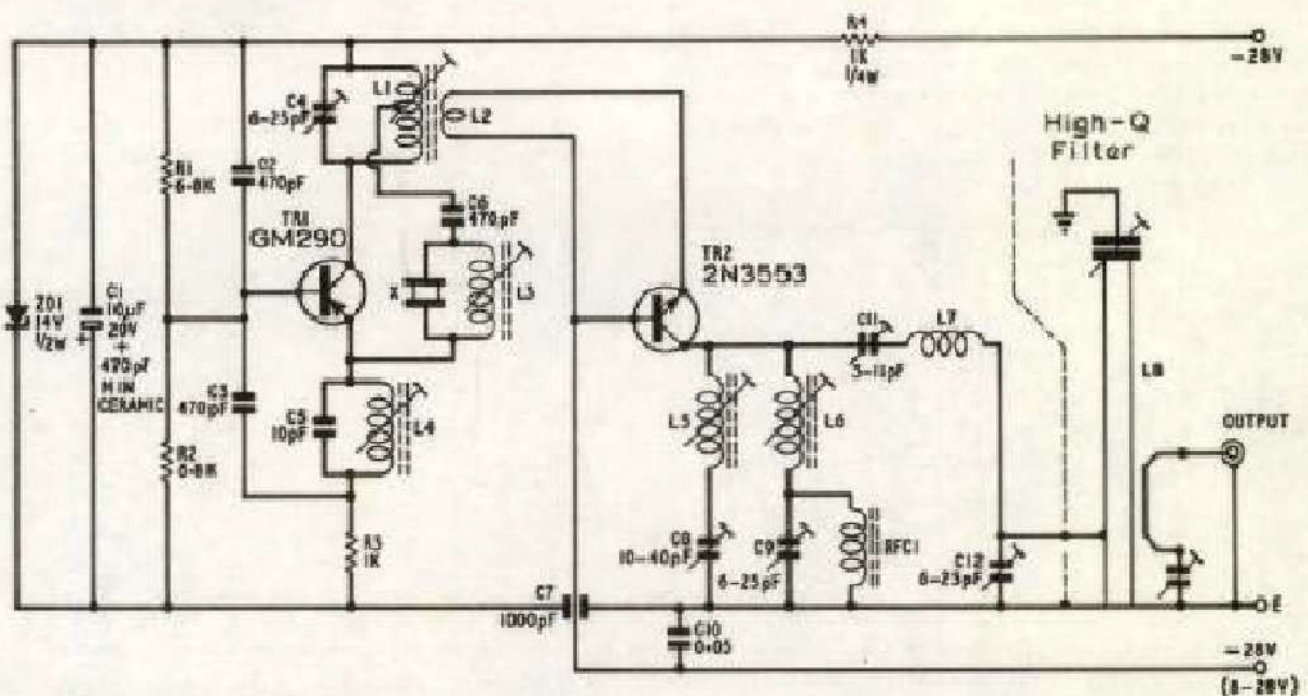
Fig.6.98. A 432 MHz exciter with an output of 150mW. L1, 3 turns, 18 s.w.g. spaced wire diam., ¼" diam., tapped 1 turn from h.t. end, dust core tuned; L2, 1 turn, 18 s.w.g. insulated, interleaved with L1, twisted length not greater than ¾" to TR2; L3, 6 turns, 18 s.w.g. spaced wire diam., ¼" diam., dust core tuned; L4, 5 turns, 18 s.w.g., spaced wire diam., ¼" diam., dust core tuned; L5, 4 turns, 18 s.w.g. spaced wire diam., ¼" diam., dust core tuned; L6, 2 turns, 18 s.w.g. spaced wire diam., ¼" diam., dust core tuned; L7, 1¼ turns, 14 s.w.g. ¼" long, ¼" diam., L8, Any high Q filter (see RSGB Radio Communication Handbook, page 5.30); RFC1, as Fig.6.95. TR1, GM290; TR2, 2N3553; X, 144 MHz overtone crystal; ZD1, 14-volt Zener diode, ½ watt.
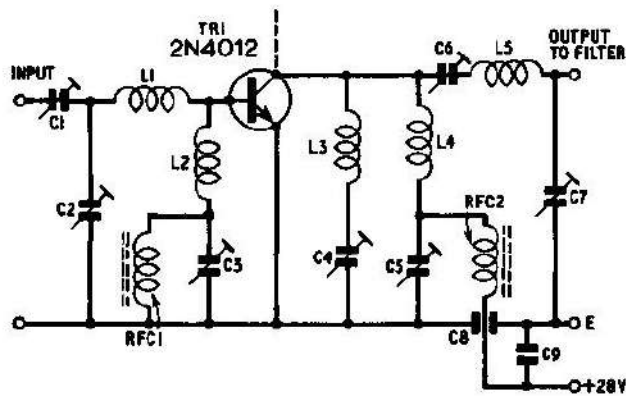
Fig.6.99. A doubler or tripler to 432 MHz. Doubler (216 MHz input): C1, 5 30pF; C2, 5-30pF; C3, 2.5-5pF air spaced; C4, not used; C5, 3-11pF air spaced, C6, 3-11 air spaced; C7, 6-25pF; C8, 1000pF; C9, 0.05µF ceramic; L1 2 turns, 18 s.w.g. spaced wire diam., ¼" diam.; L2, 2½ turns, 18 s.w.g., ⅛" spacing ¼" diam.: L3, not used; L4 4 turns, ¼₆" spacing ¼" i.d.; L5, 2 turns, 14 s.w.g.,³⁄₁₆" spacing ¼" diam.

Tripler (144 MHz input); C1, 8-50pF; C2, 8-50pF; C3, as doubler; C4, 10-40 pF; C5, 6, 7, 8, 9, as doubler; L1, 3 turns 18 s.w.g., spaced wire diam., ¼" diam.; L2, as doubler; L3, 4 turns 18 s.w.g. ¼₆" spacing, ¼" diam.; L4, 3 turns, 18 s.w.g., ¼₆" spacing, ¼" diam.; L5, as doubler. RFC1, as Fig.6.95, RFC2, ¼" x ½" ferrite core, 10 turns, 24 s.w.g. nichrome. Transistor, 2N4012; make certain that this transistor has enough heat sinkage and that the case is at r.f. earth potential.

For use at 70cm, a suitable overlay transistor circuit is given in Fig.6.98, and again the component details and coil sizes are to be found in the caption. An alternative doubler or tripler circuit is shown in Fig.6.99 with the coil details below.

A HIGHER POWER EXCITER TRANSMITTER FOR 70cm USING OVERLAY TRANSISTORS

In this example four stages are used to reach the final frequency. The crystal oscillator TR1 operates at a quarter of the output frequency, a variable capacity diode VD pulls the frequency and this could be used either for frequency shift keying or n.b.f.m. by applying a suitable d.c. or a.f. voltage to it. The voltage applied to the oscillator stage is stabilised by two zener diodes connected in series.

The second stage TR2 operates as an amplifier at the crystal frequency, this is followed by TR3 and TR4 both operating as doublers tuned to twice and four times the crystal frequency respectively. An output power of about five watts is obtainable, this is sufficient to drive a DET24 grounded grid amplifier to 15 watts output which, if required, will drive a 4CX250B up to around 190 watts using a good cavity tuned circuit.

Coil Details

L1 8 turns 22 s.w.g.	L6 7 turns 22 s.w.g.
L2 7 turns 22 s.w.g.	L7 4½ turns 22 s.w.g.
L3 7 turns 22 s.w.g.	L8 4½ turns 22 s.w.g.
L4 4 turns 20 s.w.g.	L9, L10 3 turns 22 s.w.g.
L5 2½ turns 18 s.w.g.	L11 2 turns 22 s.w.g.

L1 to L8 are wound on ceramic formers of ¼" diameter. L9, L10 and L11 are r.f. chokes.
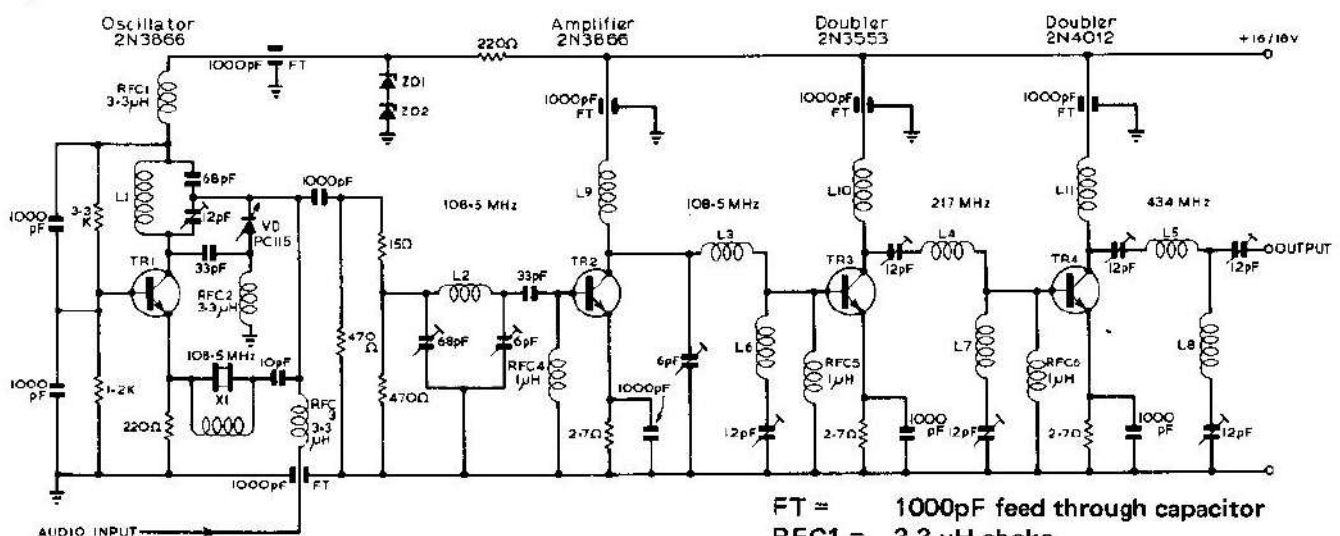
Fig.6.100. Circuit of higher power 432 MHz exciter.

FT =	1000pF feed through capacitor
RFC1 =	3.3 µH choke
RFC2 =	1.0 µH choke
X =	108.5 MHz
TR1 =	2N3866 or similar
TR2 =	2N3866 or similar
TR3 =	2N3553 or similar
TR4 =	2N4012 or similar
VD =	PC115 varicap diode

COAXIAL LINE AMPLIFIER FOR 432 MHz

This amplifier uses a 4CX250B, the anode and grid circuits of which are both coaxial lines tuned by disc capacitors with inductive input and output coupling.

The construction follows standard practice and consists of a box made of a main "U" shaped body with the ends closed and a screen across the box for mounting the valve socket and for isolation of the anode and grid circuits. The end plates and the screen should preferably bend around all four edges to facilitate fixing to the main body. The end plate to which the anode line is attached must be arranged so that it is easily removable for valve changing.

The box is 8¼" long and 3½" square (inside) across section, using 18 swg copper or brass, the lid should be bent on the long sides so that there is a good contact between the lid and the body.

The anode line, Fig. 6.103A, is made of 1¾" diameter (outside) tube with 20 swg wall thickness, at one end a 2¾" diameter flange is fitted and at the other, eight equidistant slots are cut for a distance of 1" from the end. After these slots have been cut, the open end should close down to a diameter of 1.6" to make good contact to the valve cooler. The inner edge of the "fingers" should be chamfered to assist fitting to the valve. The fixed plate of the anode tuning capacitor is attached centrally to one of the end fingers by soldering, the position of this is arranged to be opposite to its moving plate.
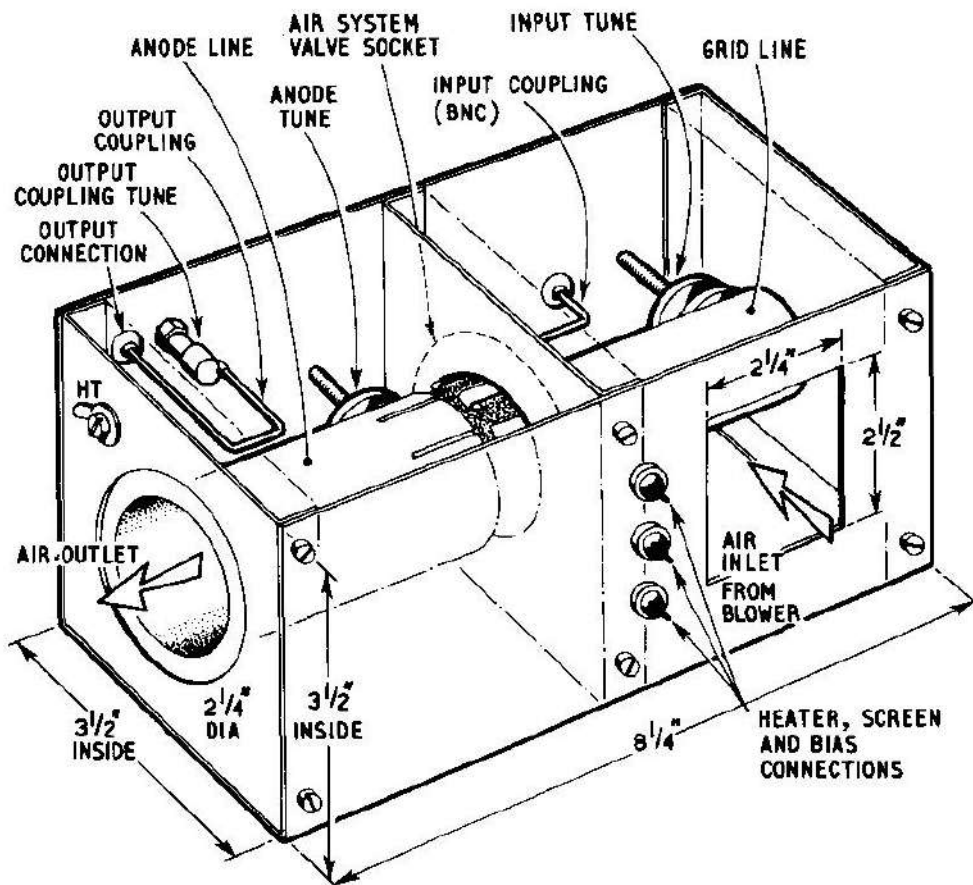
Fig. 6.101. General arrangement of the coaxial line amplifier.

The end plate to which the mounting flange is attached has a hole to match the inside diameter of the anode line for the air outlet. Insulation between the line and the end plate should be a mica ring, which, with its high dielectric constant, will provide a substantial by-pass capacitor. For safety it is desirable to cover the hole in the end plate with an open mesh to prevent accidental contact with the live anode line.

The grid line is made of ⅛" diameter tube of 20 swg wall. One end is closed by a disc soldered into the tube and a clearance in centre of this is needed so that the line can be directly attached to the grid connection of the valve socket. As with the anode line, the fixed plate of the grid tuning capacitor is fixed directly on to the tube at 2 ¼" from the valve end of the line, see Fig.6.103B.

The valve socket is fitted to the central screen, with the screen by-pass capacitor flange of the socket on the anode side of the screen and fixed by the three clamps provided with socket. The position and size of the input and output couplings are shown in Fig.6.102.

Connections to the heater (live side), screen and bias supplies are made through insulated terminals and rf chokes. The bias connection is made on to the grid line at a point of minimum rf voltage, approximately 0.8" from the valve end of the grid line. The ht connection to the anode is taken by one of the insulated screws used to fix the anode flange to the end plate, an rf choke should be fitted externally.
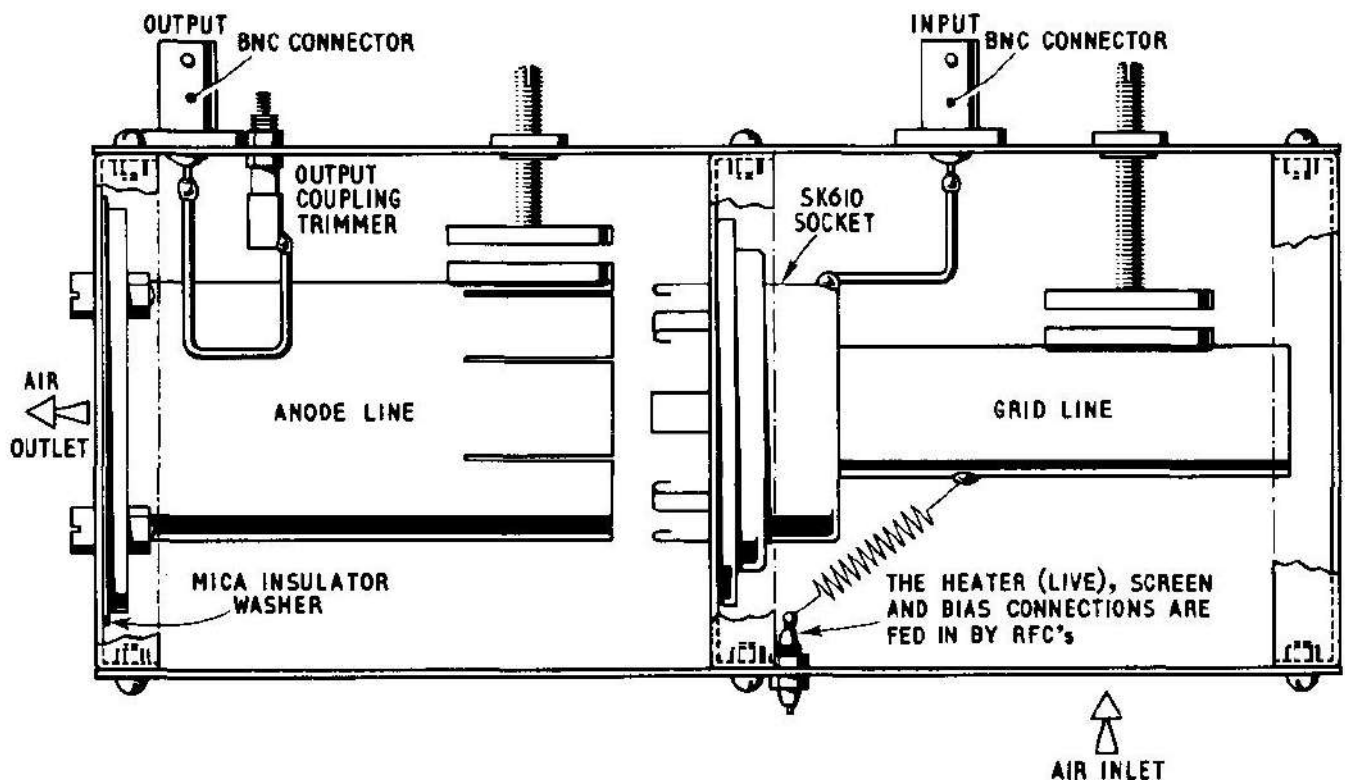
Fig. 6.102. Layout of the amplifier.

SOLDER

1⅛" DIA

⅛"

2¾" DIA

1¾" O/D

8 SLOTS
1/32" WIDE

1"

3 13/32"

Anode Line..... A

MATERIAL ..... COPPER OR BRASS — TUBE 20 SWG WALL THICKNESS

1⅛" DIA

7/8" O/D

2 1/16"

3⅛"

CLOSED END DRILLED FOR FIXING TO
GRID TERMINAL OF VALVE SOCKET

1⅛" DIA

SOLDER

⅛"

Grid Line.... B

MATERIAL ..... COPPER OR BRASS
TUBE 20 SWG WALL THICKNESS

⅛"

2BA STUDDING

1⅛" DIA

1 13/32" ANODE
1⅞" GRID

Tuning Plate
MATERIAL .....
COPPER OR BRASS
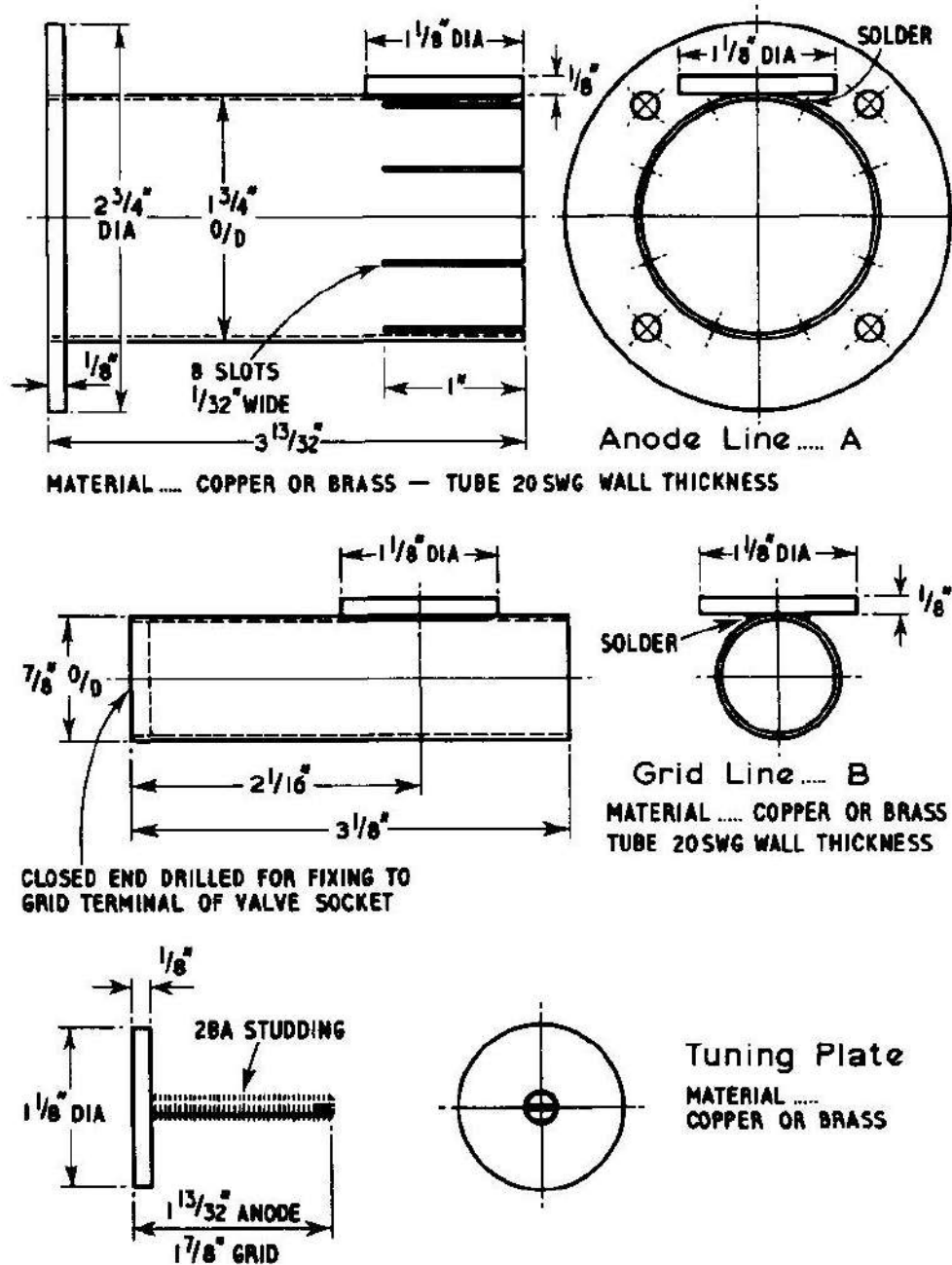
Fig. 6.103. Mechanical details of the anode line (A), grid line (B) and tuning plate.

MOBILE

TRANSMITTER — RECEIVER FOR 144 MHz.

Transmitter r.f. Section

The r.f. stages of the transmitter, Fig. 7.1, are built on a separate 16 s.w.g. aluminium chassis measuring 4" x 10¼" x 1½" and are connected to the other sections by a miniature 8-way plug and socket.

One section of a QQV03-10 double tetrode (VI) is used as a Squier oscillator operating on the third overtone of an 8 MHz crystal. The other section of the valve operates as a frequency tripler to 72 MHz. A stabilized 150 volt supply is used for the oscillator and it was found necessary to operate the tripler from this supply in order to reduce the grid drive to the following stage to a reasonable level. The grid resistors

are connected to the 12 volt negative supply to provide protection in the event of oscillator failure. Adjustment of the oscillator feedback can be made by altering the spacing between L1 and L2. The QQV03-10 is internally neutralized for push-pull operation but this appears to have no significant effect on the operation of the circuit.

The anode circuit of the tripler is inductively coupled to the grid of an EL85(V3) operating as a frequency doubler to 144 MHz. The grid current of this stage should be approximately 1.2mA. For c.w. operation the cathode circuit of V3 is keyed.

The final stage is a QQV03-20A(V4) with a self resonant grid circuit. The p.a. tank circuit is above the chassis adjacent to the valve anodes with the aerial change-over relay (RY1) alongside.

The EL85 can provide a grid drive current of approximately 1.5mA, which is slightly less than the
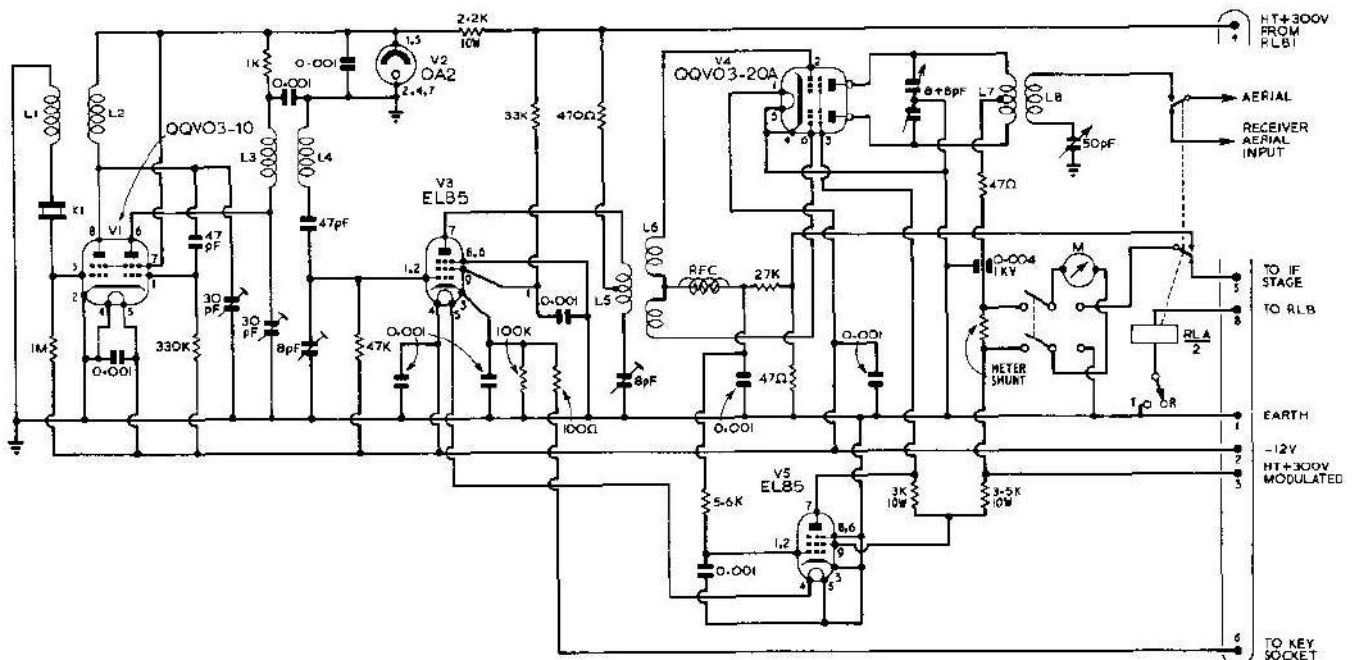
Fig.7.1. Circuit diagram of the r.f. section of the transmitter, L1, 2¾ turns 20 s.w.g. enam, close wound on ⅜" dia, former; L2, 15 turns 20 s.w.g. enam. close wound on ⅜" dia. former and spaced ⅛" from L1; L3, 3 turns 18 s.w.g. enam. ½" i.d. ³⁄₁₆" long; L4, 5 turns 18 s.w.g. enam. ½" i.d. ¾" long, close coupled to L3. L5, 4 turns, 16 s.w.g. enam. ½" i.d. ¾" long, tapped at 2 turns; L6, 2 + 2 turns 14 s.w.g., silver plated ½" i.d. 1 ¼" long with ⅛" gap at centre for L5; L7, 2 + 2 turns ¾" i.d. 12 s.w.g. silver plated 1 ¼ " long with ¼" gap for L8 at centre; L8, 3 turns 16 s.w.g. ½" i.d. ¼" long, RFC, 40 turns 36 s.w.g. enam. close wound on 1 watt resistor; RY1, relay with one low loss change over contact (Magnetic Devices Ltd), X, FT243 type crystal in 8 MHz range.
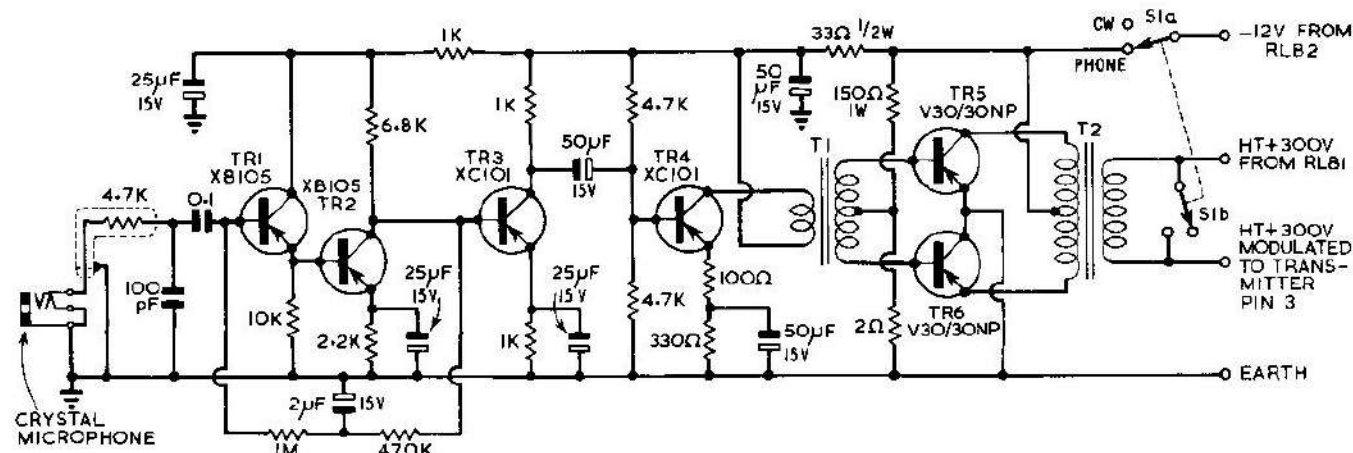
Fig. 7.2. Circuit of the transistor modulator. T1, driver transformer matching 120 ohms to 30 ohms centre tapped (Gardners Radio Type GR.33445); T2, 15W modulation transformer matching 20 ohms centre tapped to 4000 ohms ratio 1 + 1: 28 (Gardners type 33446); Suggested alternatives for the transistor types shown in the diagram are: OC71 (XB105), AC128 (XC101) and 2-AD149 (V30/30NP) mounted on 3" square heat sink made of 16 s.w.g. aluminium.

recommended value, but has proved to be quite adequate. An increase in drive power could be obtained by replacing the EL85 with a QV03-12(5763), but there would be a considerable increase in power consumption. The EL85 has the advantage that the heater consumption is only 6.3 volts 0.2A.

A second EL85(V5) is used as a clamp valve for the p.a. screen grid. To obtain a satisfactory clamping action the screen grid of the EL85 is connected to a tap on the p.a. screen dropping resistor.

A double pole change-over relay (RY1) with one low loss contact set performs the aerial change-over and also switches a 0-1mA meter from the receiver to the transmitter. Metering facilities are provided for the p.a. anode and grid currents during transmission. The meter operates as a signal strength indicator during reception.

The transmitter power consumption is 12.6 volts at 1.25A and 300 volts at 170mA.

Modulator

The modulator consists of a three-stage transistor amplifier, class A driver stage and class B push-pull output stage (Fig. 7.2). The quiescent input current is approximately 0.3A., rising to 2A., for an audio output of 12 watts.

A high input impedance is provided by the common collector stage (TR1). The first three stages are directly coupled and to provide d.c. stabilization the bias for the first stage is derived from the collector of TR2. Negative current feedback is applied to the driver stage (TR4) by the unbypassed emitter resistor. The value of the resistor was selected to give a suitable overall gain for the particular microphone in use (an Acos Type MIC 39-1). The driver and modulation

transformers were made to order and full details can be obtained from the manufacturer.

The output transistors are mounted on insulated 3" x 3" 16 s.w.g. aluminium heat sinks on either side of the driver transformer, while the other stages are mounted on a paxolin panel 4" x 2½" at the rear of the main chassis.

In order to avoid oscillation due to r.f. feedback to the modulator input the first four stages are completely screened. Particular care is necessary with the screening of the microphone input lead.

Alternative transistors which would be suitable, with slight modification of some resistor values are OC75, GET113 (TR1,2) OC72, GET114 (TR3) OC35, GET573 (TR4,5,6).

Receiver R.F. Section

This section, Fig. 7.3, of the receiver is built on a 16 s.w.g. aluminium sheet measuring 6" x 4". Although a crystal controlled converter followed by a tunable first i.f. would be ideal it was not considered to be practical in this case. In order to avoid harmonics of the local oscillator appearing in the range 144-146 MHz a high intermediate frequency would be necessary and several tunable i.f. stages would then be required to give reasonable image rejection. To avoid this complication a simple self-excited oscillator and fixed i.f. amplifier have been used. A first i.f. of 5.2 MHz was chosen as commercial transformers are available for this frequency.

An E88CC cascode stage (V1) gives high r.f. gain combined with a reasonable noise factor. The aerial tapping point, grid tuning and adjustment of the matching coil (L2) are made to obtain optimum noise factor.
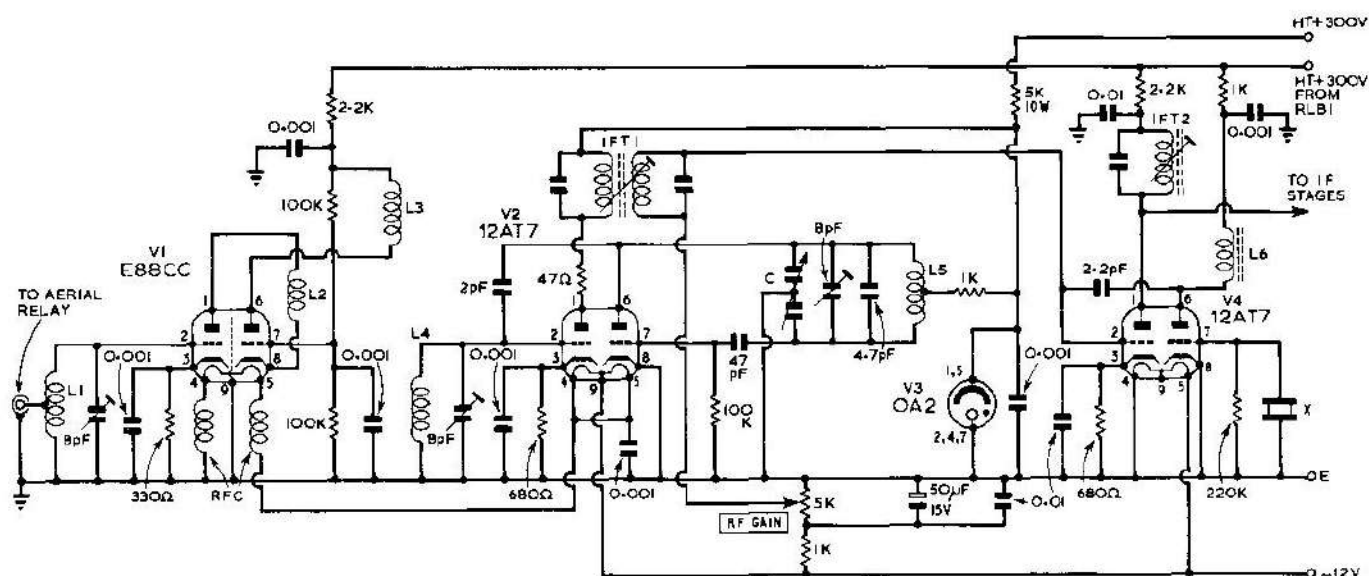
Fig. 7.3. Receiver r.f. section. C, 2pF per section split stator capacitor (1 rotor plate per section); L1, 3 turns 18 s.w.g. 5/16" i.d. 3/16" long tapped 1 turn from earth; L2, 4 turns 22 s.w.g. enam. 1/4" diam. close wound; L3, 6½ turns 20 s.w.g. enam. 3/16" i.d. close wound and coupled to earthy end of L4; L4, 4 turns 18 s.w.g. 5/16" i.d. 3/8" long; L5, 2 turns 16 s.w.g. 5/16" i.d. approx. 3/8" long, tapped at 1 turn; L6, 40 turns 36 s.w.g. enam. close wound on 1/4" diam. dust cored former; RFC, 25 + 25 turns 28 s.w.g. enam. bifilar wound on 1/4" diam. former; IFT1, 5.2 MHz i.f. transformer (Eddystone); IFT2, one winding of standard 470 kHz i.f. transformer; X, 5.675 MHz crystal type FT243; in V1 pin 9 is connected to the shield but not to the heater and in V2 pins 4 and 5 should be connected.

A 12AT7 (V2) is used as the first mixer-oscillator and control of r.f. gain is obtained by varying the bias on the mixer grid. Oscillator injection is provided by a capacitance of about 2pF connected between the oscillator anode and mixer grid. This consists of two short lengths of insulated flex twisted together and adjusted to give maximum sensitivity. Tuning of the mixer grid circuit has a small effect on oscillator frequency. The oscillator tuning capacitor (C) is mounted on top of the chassis with L5 above it. It is a 10 + 10pF receiver type split-stator capacitor with most of the rotor plates removed leaving only one plate per section. The 8pF preset capacitor (mounted below the chassis) and the dimensions of L5 can be adjusted to give suitable bandspread. The slow motion dial from a RF26 unit was found to be suitable for the oscillator tuning, although it was necessary to mount a 1/4" bush on the front panel to support the spindle of the capacitor in order to eliminate backlash.

To reduce oscillator drift the first mixer-oscillator is permanently connected to a 150 volt stabilized h.t. supply. There is considerable drift immediately after switching on but after a few minutes it is acceptable.

The second mixer-oscillator (V4) is another 12AT7 with which the 5.675 MHz crystal gives a second i.f. of 475 kHz. The mixer output is capacity coupled to the first 475 kHz i.f. stage.

Receiver I.F. and Audio Stages

Transistors are used in the remaining stages of the receiver, which are assembled on an 18 s.w.g. brass strip 3" high x 8¼" long. The transistors and i.f. transformers are mounted along the centre of the strip and tag strips along the top and bottom. The component layout is very similar to the circuit diagram. A printed circuit would, of course, be ideal for this section of the receiver.

In order to obtain sufficient sensitivity three neutralized i.f. stages are used (Fig. 7.4). A.g.c. is obtained by returning the base potentiometer of TR1 to the diode detector output. In addition, a reverse biased diode is connected to the collector of TR1. The effect of the normal a.g.c. is to reduce the collector current of the first stage, thus reducing the diode bias voltage. For large input signals the diode conducts, reducing the gain of the first stage and increasing the bandwidth. The 1mA meter connected in the emitter circuit of this stage acts as a sensitive S meter while for c.w. reception a crystal controlled b.f.o. is included. An XA103 transistor was used for the b.f.o. as it was available at the time of construction but an XA101 could also be employed.

The push-pull transformerless output stage matches directly to a 15 ohm loudspeaker and provides approximately 750mW audio output, sufficient to produce ample volume under all conditions. Alternatively several pairs of high impedance headphones can be connected in parallel across the output.

Most manufacturers produce transistors for these stages, e.g. OC45, GET873 (TR1-4) OC71, GET114 (TR5) GET115 (TR6-7), although some changes in component values would be required if transistors other than the Ediswan types specified were used.
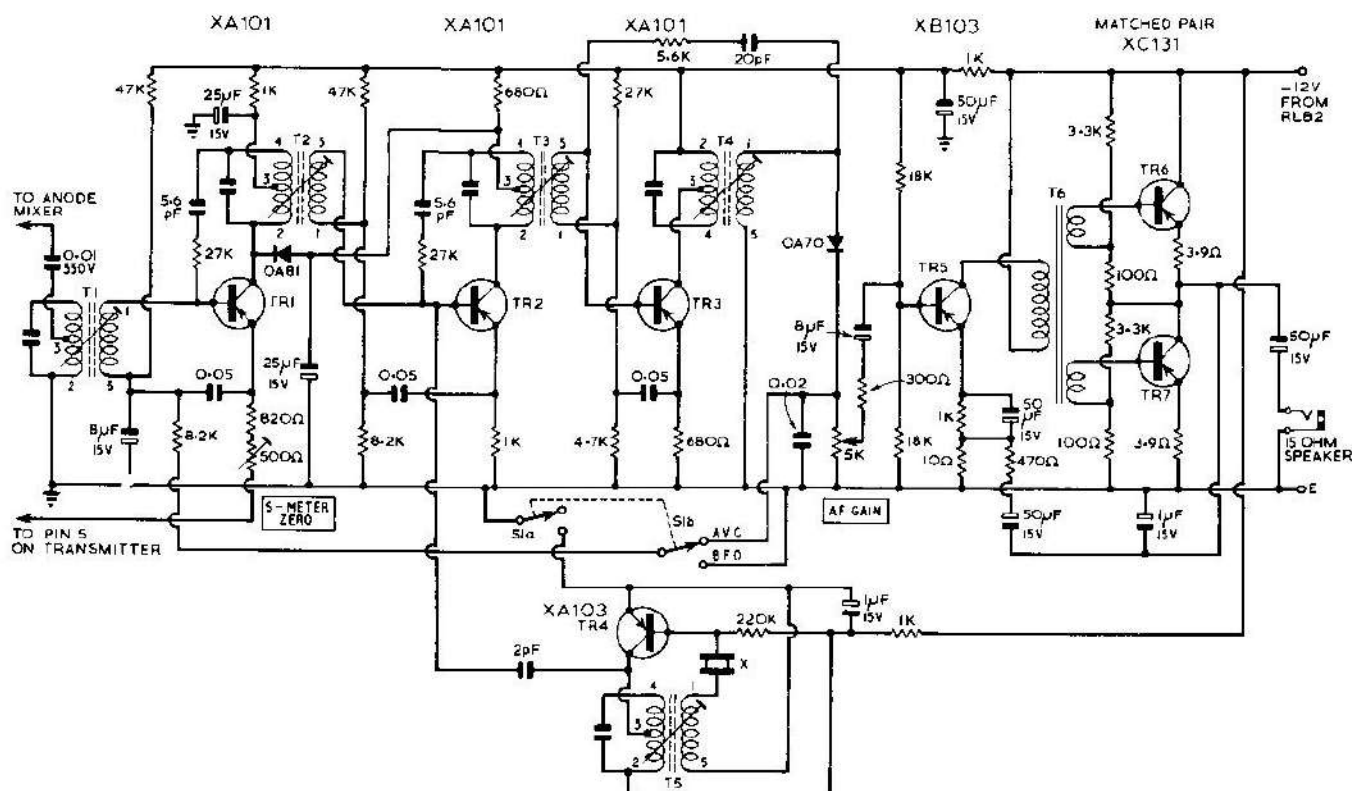
Fig. 7.4. Circuit of the i.f. and audio stages of the receiver, T1, 2, 3, Weymouth Radio type P50/2CC; T4, 5, Weymouth Radio Type P50/3CC; T6, Weymouth Radio type LFDT1; Suggested alternatives for the transistor types shown in the diagram are: OC45 (XA101), AC126 (XA103), OC71 (XB103) and 2-AC128 (XB131); X, crystal type FT241A marked 25.5 MHz.

General Details

The receiver and modulator are mounted on a 16 s.w.g. aluminium chassis 8" x 10¼" x 1½" and the complete unit is enclosed in a case of perforated 18 s.w.g. sheet steel. The overall dimensions are 12¼" x 10½" x 5½' high.

Details of additional power supply switching arrangements are given in Fig. 7.5. One set of contacts of the d.p.c.o. relay (RY2) switches the h.t. between receiver and transmitter, while the other switches the 12 volt negative supply from the i.f. stages to the modulator. The receiver oscillator is permanently connected to the h.t. supply.
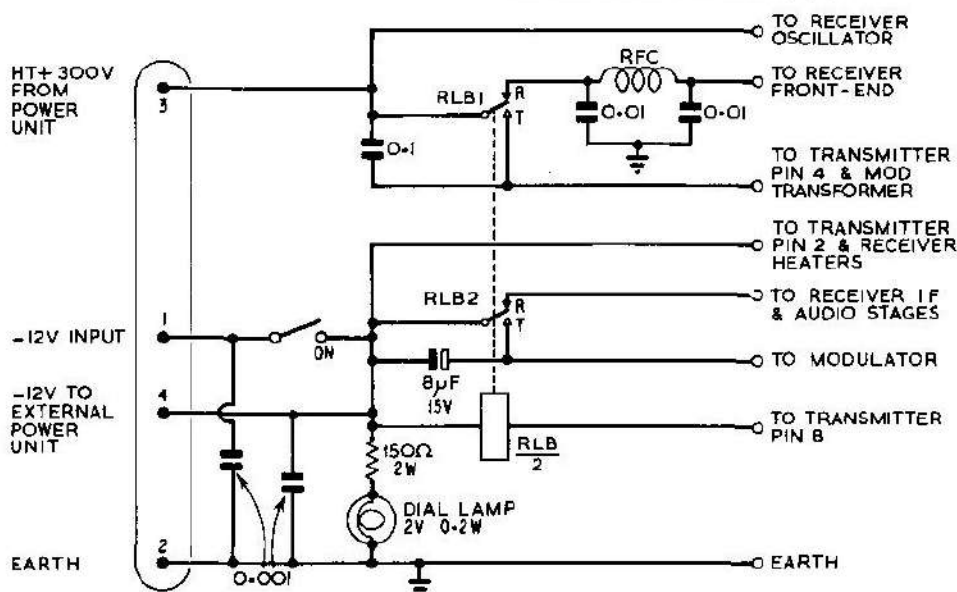
Fig. 7.5. Control circuits for the 144 MHz transmitter receiver. The r.f. choke RFC is an Eddystone type 1010 and the relay RY2 a model 596 EB. d.p.c.o. type manufactured by Magnetic Devices Ltd.

The 12 volt supply for an external h.t. power unit is taken from the on-off switch. A rotary generator would be quite satisfactory but the improved efficiency and regulation of a transistor d.c. converter are a considerable advantage.

Any h.t. supply voltage in the range 200-300 volts may be used, but when voltages below 250 volts are used, the series resistor for the stabilisers will have to be reduced.

AN ALTERNATIVE DESIGN FOR A TRANSMITTER—RECEIVER FOR 144 MHz.

This transceiver is also a hybrid design in which valves and semiconductors are used, but the individual applications are different from the previous unit.

From these two designs there is sufficient data provided for an all semiconductor piece of equipment, with the exception of the r.f. section of the transmitter. Where the maximum use of semiconductors is required, the early stages of this r.f. section can be re-designed from data provided from the transmitter chapter, but it is unlikely that at present the power amplifier itself would be changed from either the QQV03-20A or 6146 used in these designs.

The circuit diagrams are self explanatory and any details are to be found in the captions. The coil data is listed in the table.

Transmitter (Fig.7.6)

Coil	No. of turns	Diam	Wire	Length	Trimmer	
L1	16	¾"	26	close wound	25pF	
L2	5	½"	16	½"	15pF	
L3	3	½"	14	½"	15pF	} close coupling
L4	4	½"	14	¾"	15pF	}
L5	4¼	⅝"	14	¾ centre tapped	20pF	
L6	1	1 "	16	-	50pF	
	48	½"	34	close wound	-	

Receiver (Fig.7.9)

Coil	No of turns	Diam	Wire s.w.g.	Spacing	Tap
L1	2½	⅜"	18 s.w.g.	1 dia	½ from earth
L2	3	¾"	18	½" long	-
L3	2½	⅜"	18	½" long	½ from earth
L4	15	¼"	34	close wound	-
L5	3	¼"	34	close wound	-
L6	16	¼"	34	close wound	2 and 4 turns from base end
L7	2	⅝"	14	½" long	centre
L8 L9	Electroniques Stabcoil No. 27LZ				
L10 L11	Electroniques Stabcoil No. 27MX				
L12	20	¼"	34	close wound	5 from crystal end
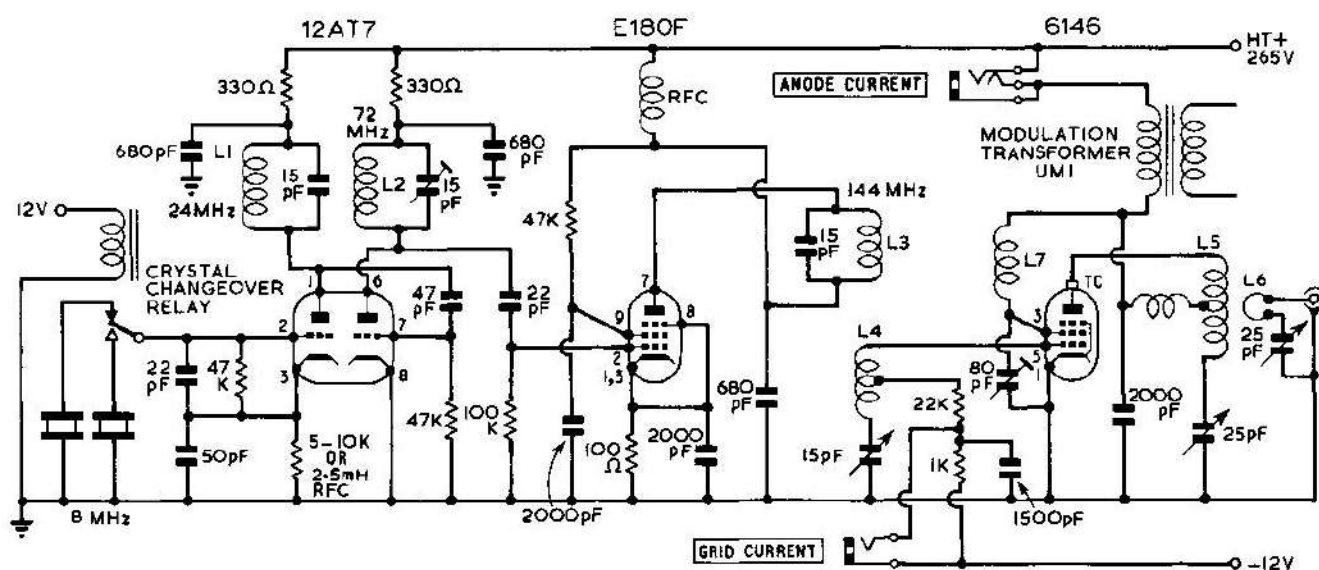
Fig. 7.6. Circuit of transmitter section of an alternative design for a transceiver.
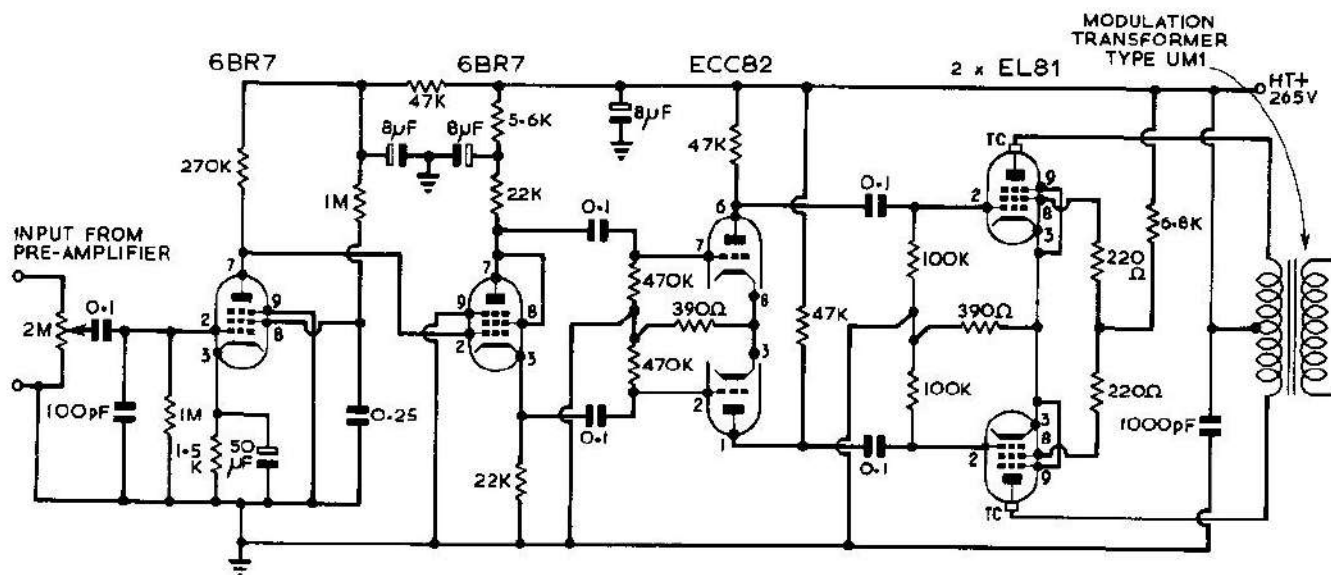
Fig. 7.7. Modulator circuit for use with r.f. section shown in Fig. 7.6.
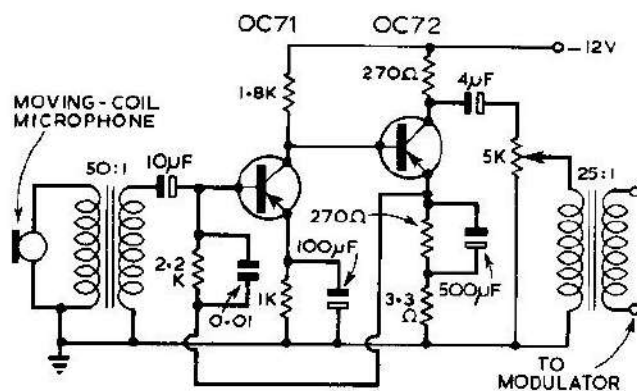
Fig. 7.8. Microphone pre-amplifier for use with modulator shown in Fig. 7.7.

Fig. 7.9. The receiver section of the transceiver showing the front end, i.f. amplifier and audio sub-section. Suggested alternatives for TR2 and TR6 are types AF178 and AC128 respectively.
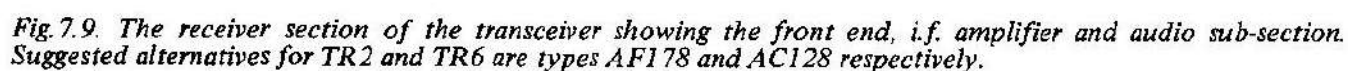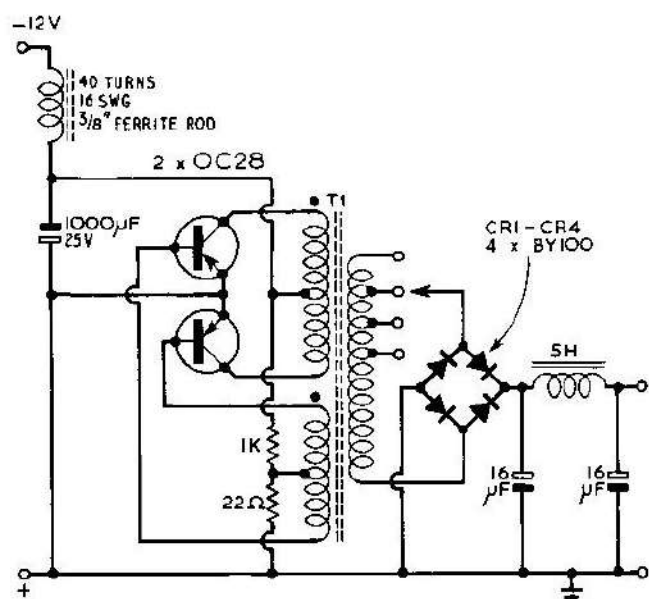
Fig. 7.10. Power supply for the transceiver. Transformer wound on Ferroxcube core FX1079 primary-inner, feed back-outer.

Primary	32 + 32T	16 s.w.g.	Bifilar wound
Feedback	14 + 14T	16 s.w.g.	Bifilar wound
Secondary	860T	26 s.w.g.	tapped at 680, 740, 800T

Consumption	Standby	2.3A
	Transmit	7.8A
D1,D2,D3,D4,	BY100 silicon diodes	
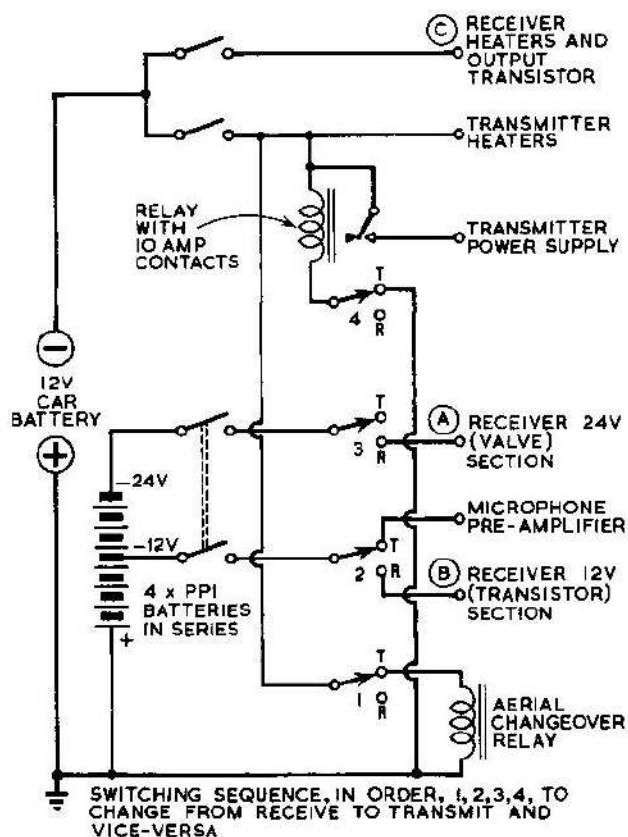
Fig. 7.12. The wiring for a suitable control unit for the transceiver.
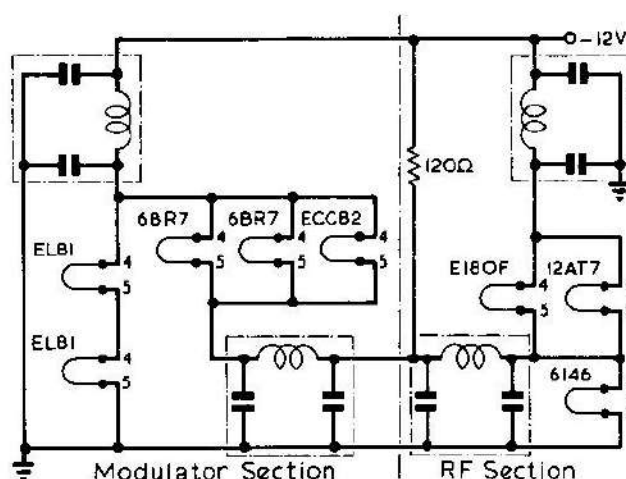
Fig. 7.11. Showing wiring of the heater circuit for the transmitter section shown in Figs. 7.6 and 7.7.

From the two equipment designs given in this Chapter, there is sufficient material available for an all valve or all transistor design with the exception of the r.f. section of the transmitter which in both cases is valved.

If an all transistor design is required then the data given under *Overlay Transistors* in Chapter 6 provides much of the necessary information. However, for power levels comparable with the designs given here it is more economical to use a valve power amplifier driven by a transistor exciter.

The provision of the h.t. voltage for such an amplifier does not present any particular problem and for equal power outputs the loading on the vehicle battery is likely to be somewhat lower when using a valve amplifier than with transistors.

In this connection it should be remembered that the gain of transistors are generally significantly lower than power amplifier valves and this means the additional transistor amplifier stage would almost certainly be needed to drive the final amplifier.

CHAPTER 8

SINGLE SIDEBAND

A 144 MHz S.S.B. or C.W. TRANSMITTER

This transmitter is suitable for operation with an s.s.b. or c.w. exciter. Although details are given for particular frequencies, there is of course no reason why other suitable combinations should not be used, but care should be taken to ensure that the unused sum or difference frequency does not occur in other service bands, such as Air Traffic Control or television.

In the unit described, which is the complete r.f. section requiring an s.s.b. generator operating at 7 MHz with a tuning range of ± 1MHz if it is to cover the whole 2 metre band. The normal sideband channel used however is 145.41 MHz.

As can be seen from the various illustrations the components are mounted above and below the chassis which should preferably be of the type allowing a bottom cover to be fitted. Various interstage and other screens are fitted in appropriate positions. The overall dimensions are 17" x 10" x 3" chassis with an 8¾" high panel, the length may be either 17" for a table top cabinet or 19" for standard rack mounting.

The circuits shown in Fig. 8.1 and 8.2 respectively give details of the two main sections.

The crystal oscillator V1 is an ECF80, the triode section of which operates as a Squier oscillator, with its anode voltage stabilised by an OA2 stabiliser V3. The pentode section of V1 is operated as a tripler to the final frequency of 137.7 MHz.
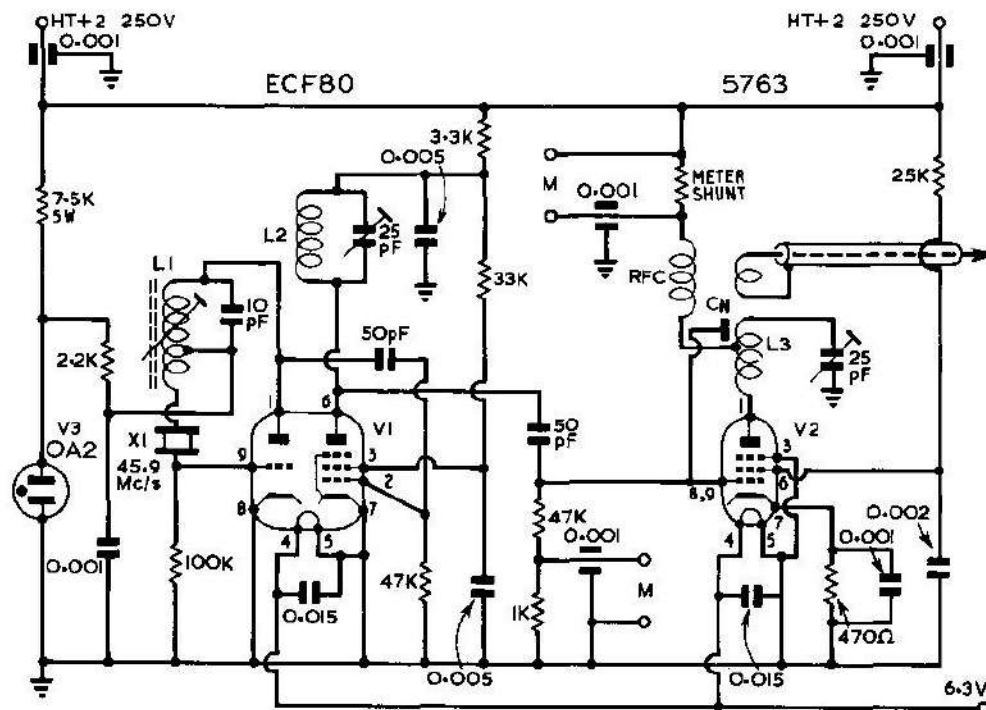
Fig.8.1. Circuit of the oscillator section of the s.s.b. transmitter. A fixed frequency oscillator multiplier chain with an output at 137.7 MHz. L1, 10 turns, tapped at 3 turns, 24 s.w.g., ⅛" i.d.; L2, 1 turn, 14 s.w.g. ¼" i.d. L3, 5 turns, tapped at 1½ turns, 14 s.w.g. ½" i.d. All decoupling capacitors are disc ceramics, and feedthrough capacitors are 1000pF, other capacitors are ceramicons.
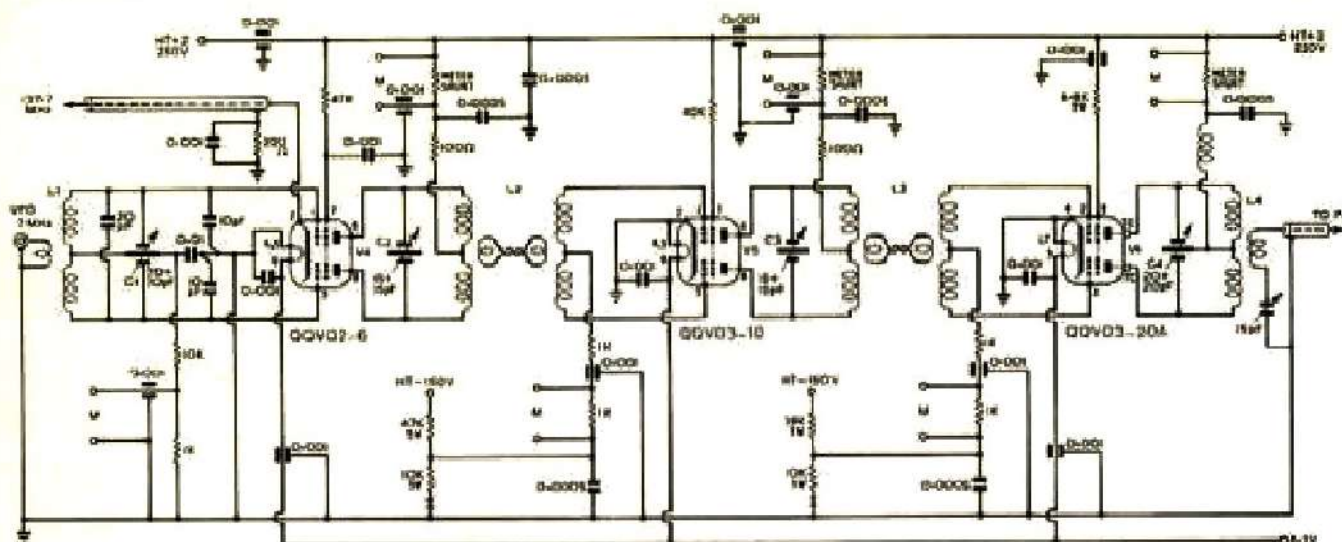
Fig. 8.2. Circuit of the mixer and driver section. L1, 23 + 23 turns, 22 s.w.g. with a 2 turn link in centre, 18 s.w.g., ½" diam. All other coils are 4 turns each section, 14 s.w.g. ½" diam., the length being adjusted to tune. Decoupling capacitors are disc ceramics, 0.0005 µF types are stand-offs, feedthroughs are 0.001 µF, and all others are ceramicons.

V2 (5763) is a buffer amplifier with a series tuned balanced anode circuit so as to provide the out of phase voltage needed for neutralisation. H.t. is fed in at the centre tap of the coil.

The grid and anode circuits are suitably screened from one another, and with careful neutralising the stage is perfectly stable. V4 is a QQV02-6 which is a small double tetrode and is used as a balanced mixer. The s.s.b. signal at 7 MHz is fed into this valve at the two control grids in push-pull and the 137.7 MHz signal is fed into the common cathode.
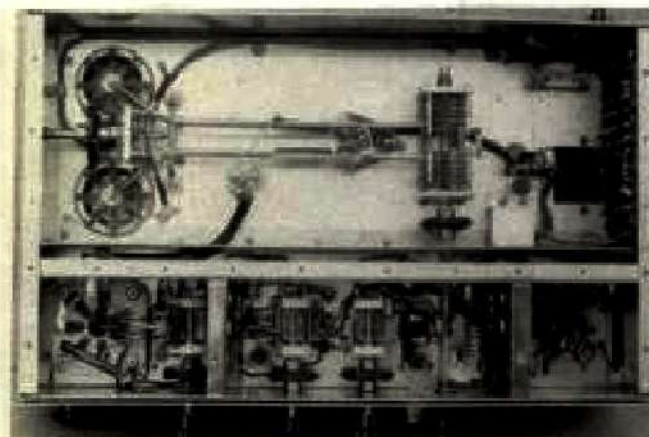
Fig. 8.3. Photograph showing the layout of components and grid lines beneath the chassis of the transmitter.

The anode circuit is tuned to 144 MHz and being push-pull cancels much of 137.7 MHz signal. V5, a QQV03-10 and V6, a QQV03-20, are both operated as push-pull linear amplifiers under AB1 conditions.

The coupling between stages is by 2 turn links. The anode circuit of V6 is above the chassis, with the output coupling series tuned and connected to the input coupling loop of the power amplifier by means of a short length of coaxial cable. Fixed bias is applied to the grids of V5 and V6. 25V for V5 and 30V for V6 is provided from the two potentiometers across the 150v negative supply.

The power amplifier valves V7 and V8 may be either 4X150A or 4CX250B and have tuned linear grid and anode circuits. The grid circuit is a half wave arrangement tuned at the remote end. At the valve end a differential capacitor is used to balance the input and the valves have 50v negative fixed bias applied to the grid in a similar manner to that used for the driver stages.

As shown in Fig. 8.4 the power amplifier valves are neutralised by small capacitors consisting of bolts passing through suitable feed through insulators fitted close to the respective valve anodes. In addition two small disc capacitors are included to balance the screen currents. Heater supply for the power amplifier valves is set to 6 volts from a 6.3 volt supply. Bias for all stages is taken from a suitable supply using appropriate potential dividers, those shown are for a 150 volt supply. The points X in these dividers can be opened to provide a suitable means of stand-by switching without removing the h.t. to the driver amplifier stages. A suitable relay would do this and could be operated by the VOX or TR switch.

Test points are provided for all the various stages, and with a two way 11 position switch these can be read directly. An additional position for an r.f. output indicator would be an advantage.
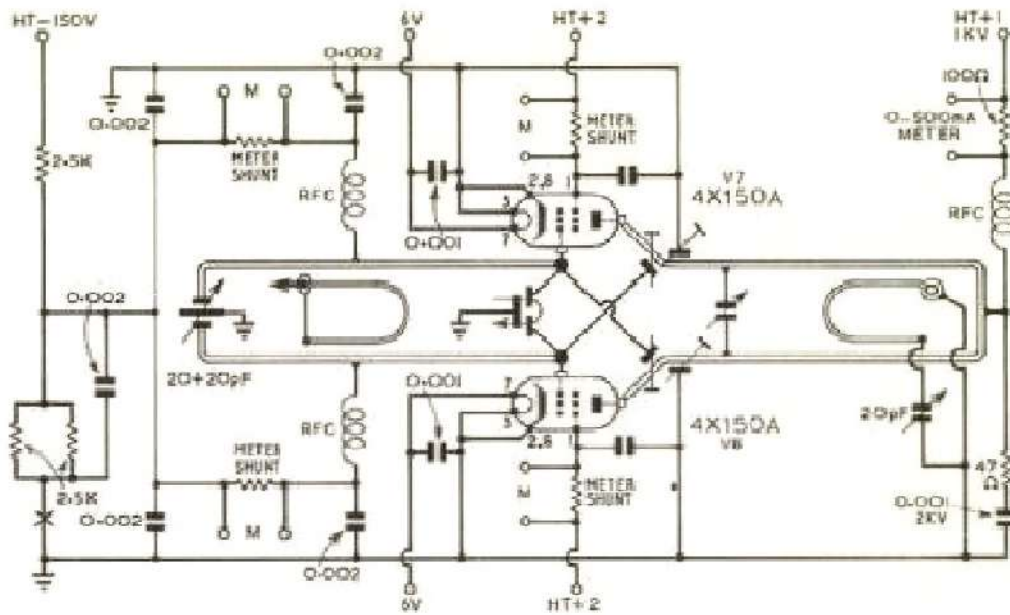
Fig. 8.4. The p.a. section. Grid lines, 5/16" diam. copper tube, 9" long, 1" spacing (centre to centre); link 3" long, 1" spacing, 14 s.w.g.; anode lines, 3/8" plated tube; 10" long, 1" spacing, centre to centre; tuning discs 1 1/2" diam. 0.002 μF capacitors are stand-off button types, and the screen decoupling is built into the bases.
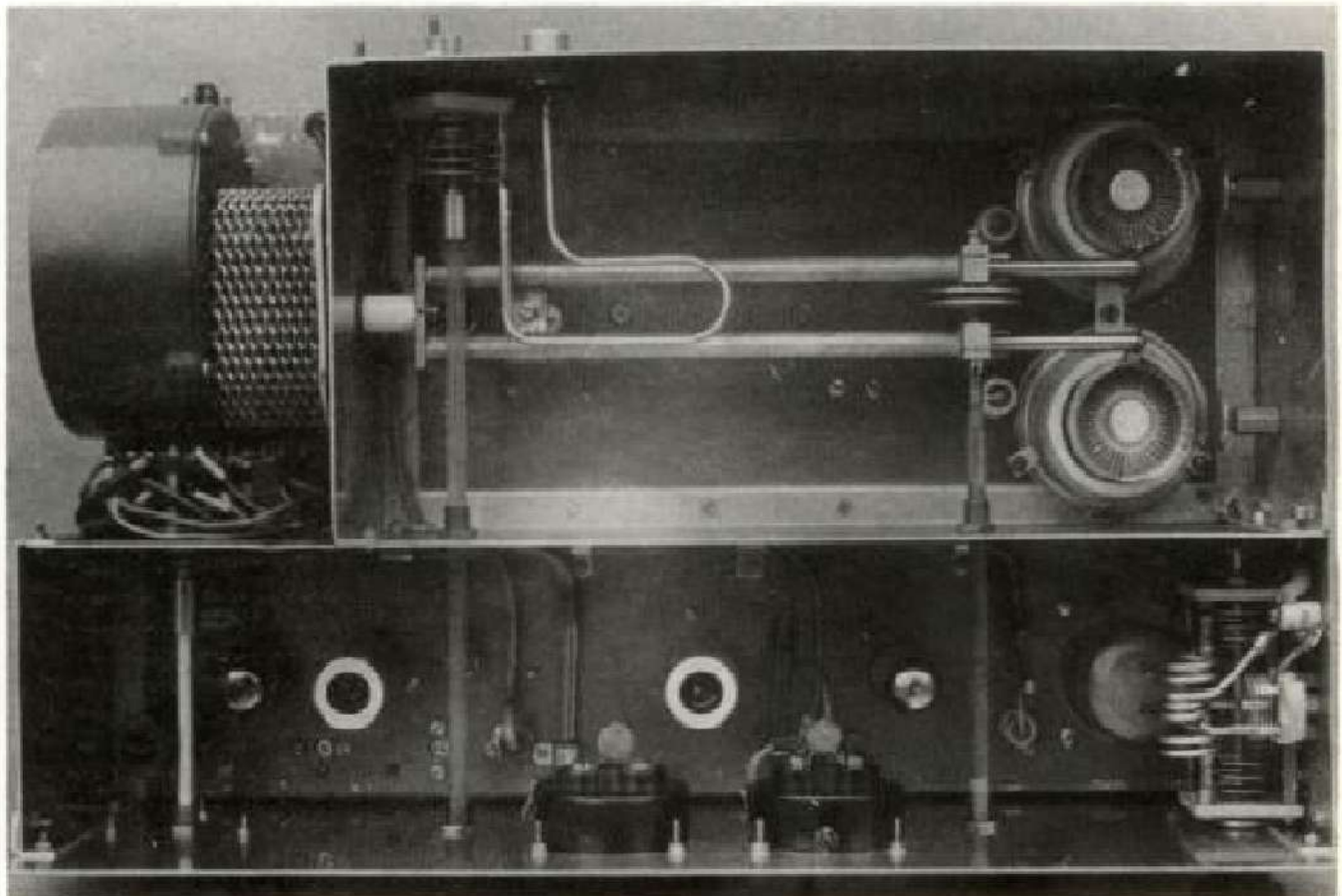
Fig. 8.5. Top view of the transmitter clearly showing the relative positions of the valve anode lines, controls and coupling loop.
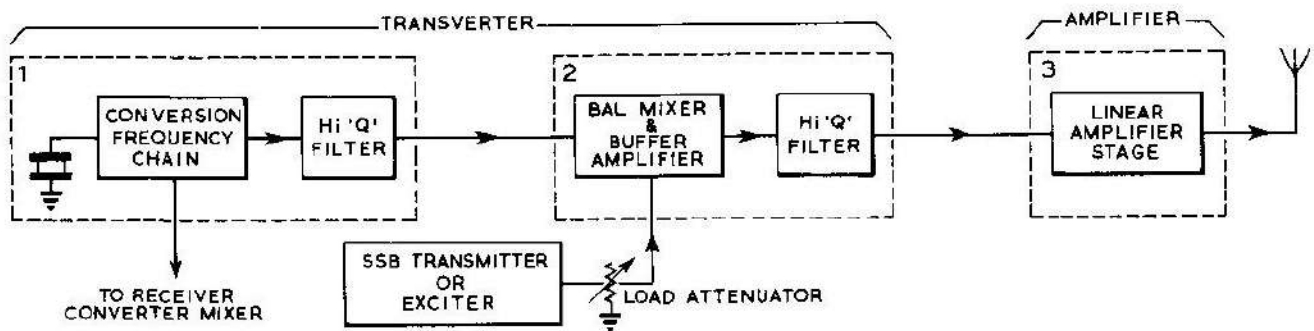
Fig. 8.6. A block diagram of the transverter stage arrangement for the transmitter section.

THE TRANSVERTER

V.h.f. operators are familiar with receiver converters which transpose a v.h.f. or u.h.f. band signal down to a much lower frequency which can then be received on a communications receiver or specially designed i.f. tuning unit. Receiver converters are usually crystal controlled so that the stability of the communications receiver and tunable i.f. amplifier is virtually the limiting factor as to performance on the v.h.f. band chosen. Such converters are in common use on v.h.f. and u.h.f. frequencies up to about 1296 MHz and basically they all consist of three sections whether they are solid state or valve units. The three sections are (a) an r.f. amplifier stage, or stages; (b) a mixer and (c) a conversion frequency oscillator chain.

For reception work, a popular frequency range for the communications receiver to tune over is the 10 metre band, i.e. from 28 to 29.7 MHz. This fits in with 2 metres particularly well as that band also is 2 MHz wide under the European frequency allocation. Taking a few simple figures for a typical receiver converter for 2 metres, we have first of all an incoming signal somewhere between 144 and 146 MHz, this is mixed with conversion frequency of 116 MHz and the difference product is extracted at the mixer output to give an i.f. of 28 to 30 MHz. Now it does not matter which way round we take these three sets of

frequencies as far as the mixing process is concerned, and it is quite legitimate to take a variable frequency of 28-30 MHz, mix it with 116 MHz and then take the sum product which will give us 144—146 MHz at the mixer output. This set of conditions then is the basic principle behind all transverter circuits and the process is only an extension of the converter circuits which are found in any multiband s.s.b. transmitter covering the h.f. and l.f. bands.

Transverter Design

The basic concept of a transmitter converter is simple enough, Fig. 8.6, and in practice too there is no complicated circuitry. Most v.h.f. operators like to make their own apparatus to fit in with some general scheme of cabinet design or matching layout. This being the case, the following paragraphs will be set out into a series of sections which can be used in any physical combination as long as the individual sections are not altered to any great extent.

Conversion Oscillator Chain

There are numerous combinations of valve types and coupling conditions between stages which could be employed, but there are also several circuits which should be avoided for one reason or another. Taking
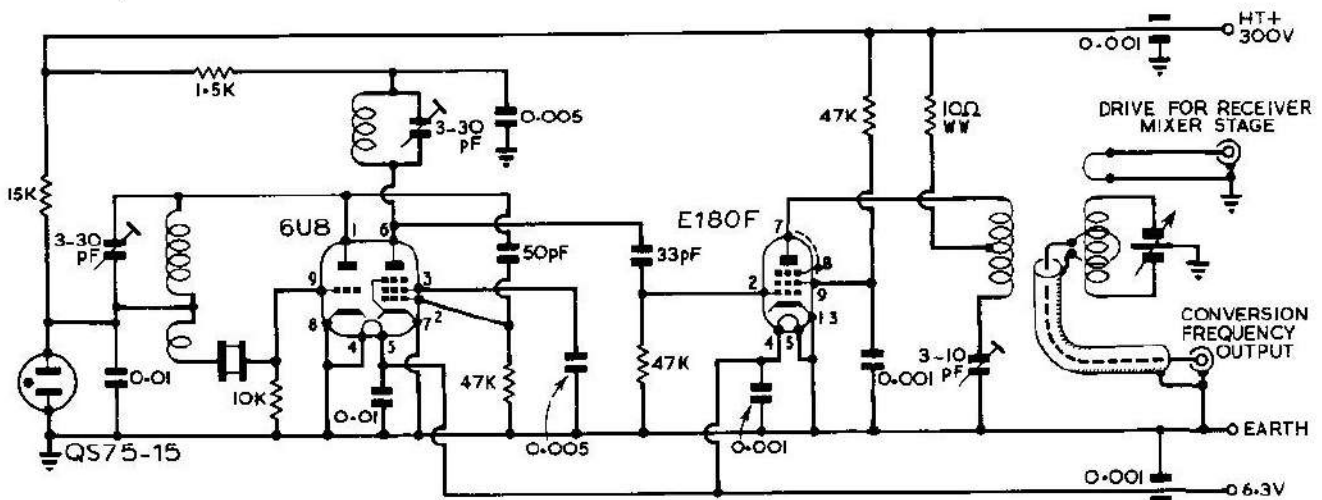
Fig. 8.7. Circuit of the conversion frequency chain and filter.

the crystal stage first of all, it is advisable not to use a low frequency to start from, such as the range 6–8 MHz, as this is likely to produce a whole host of unwanted signals around the crystal chain sub-chassis which could get into the mixer with consequent poor results. In a valve oscillator the crystal should preferably be operated at an extremely low level and with a regulated h.t. supply. The evacuated B7G crystals or the FT243 type are very good as oscillators at around 20 MHz when used in the right circuit. Care should be taken to ensure that the crystal is well away from any source of heat, and it should, preferably, be mounted out of the back of the chassis so that it operates at near ambient temperature.

The oscillator multiplier chain should be simple and the last multiplier have a tuned circuit of fairly high Q which virtually means using the series tuning arrangement of L and C. The sub chassis will have quite high level multiples of the oscillator and all its tuned harmonics present, and as it is imperative that we have only one frequency at the output of this unit before we inject it into the mixer stage, the use of a high Q tuned filter circuit should be incorporated somewhere in the feedout between the oscillator unit and the mixer grid input tuned circuitry. All supply leads to the crystal unit must be properly by-passed against r.f. radiation. The circuit diagram to achieve these results is given in Fig. 8.7. This is a well tried arrangement which is easy to construct and provided a grid dip oscillator or frequency meter is available during the lining up process the circuit requires no particular explanation.

Mixer Circuit

The next stage, which is shown in Fig. 8.8, has only a single valve, but since the whole working of the transverter depends on it being made and set up

correctly it will be described in somewhat greater detail than the simpler stages. A balanced mixer must be used and there are several combinations of circuit which could be employed with varying degrees of intermodulation cleanliness. The circuit which has finally been chosen has given good results over several years now and represents a satisfactory compromise between simplicity, ease of adjustment and good unwanted frequency suppression.

From the circuit diagram Fig. 8.8 it will be seen that the conversion frequency is fed into the sub-chassis via its high Q tuned filter, and inductively loosely coupled to the grids of a double tetrode (QQV03-10). The recommended frequency for this is about 117 MHz so that a normally available 28 MHz frequency from the s.s.b. exciter can combine with this to produce 145.41 MHz, the s.s.b. calling channel on 2 metres. It is also possible to use 14 MHz with 131.4 MHz conversion frequency, but a little more care has to be used in this case as will be explained later in this section.

The s.s.b. signal is injected into the cathode circuit of the balanced mixer valve and this will in effect grid modulate the stage. No tuned circuits are used in the cathode, only resistors, and this assists the flexibility of frequency change when necessary. The important point about using this method is that the s.s.b. drive and any harmonics or other unwanted signals are attenuated by the balancing action of the two tetrode halves with respect to the anode configuration which is push-pull connected, as opposed to the s.s.b. signal which is in push push. As with all mixers, the levels of operation are very important to the correct working of the device in order to prevent intermodulation products. A 10dB ratio between conversion frequency and s.s.b. drive is considered to be adequate for most amateur purposes, but this ratio must not be exceeded in an attempt to
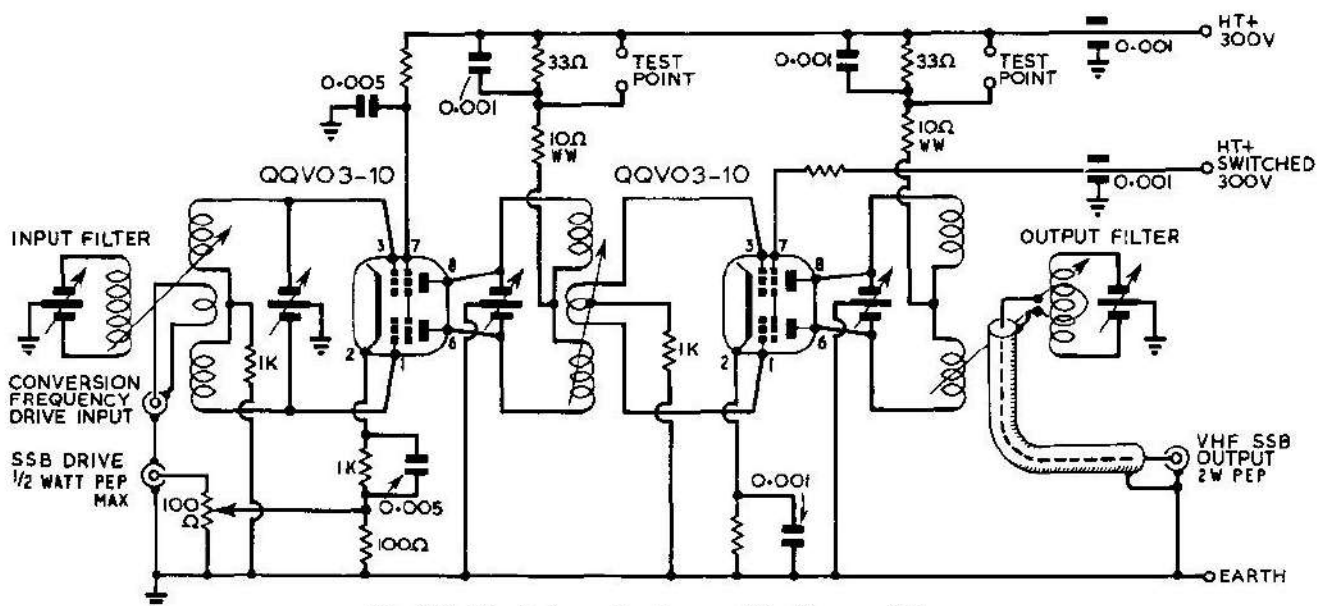
Fig.8.8. The balanced mixer and buffer amplifier.

get more drive to the linear stages. The correct approach to this is to make sure that there is sufficient gain after the mixer to give the required drive to the linear amplifiers which follow the transverter and hence the incorporation of the third stage which is a r.f. buffer amplifier. About ½ watt of s.s.b. is required to be fed to the mixer cathode circuit terminating resistor and this must be regarded as the top limit and the lowest value possible should be used. As an alternative it is quite acceptable to use two v.h.f. tetrodes separately as the balanced mixer but this is more complicated than with the single double tetrode.

Buffer Amplifier

The third stage which has to be built is the buffer amplifier which follows the mixer and it is usually convenient to construct it as part of the mixer sub-chassis which will then match up in size to that of the oscillator multiplier and so aid the final layout. Again a QQVO3-10 is used for convenience and its grid circuit is not deliberately tuned but simply consists of a 2 turn link which is coupled loosely to the anode of the balanced mixer to give just enough drive volts for Class A operation. Screening is incorporated to ensure that the amplifier is stable. The power output from this stage should be about 2 watts p.e.p. This is more than adequate for most linear amplifiers up to 200 watts rating.

The buffer amplifier not only amplifies the s.s.b. signal linearly but provides quite a substantial amount of attenuation to the conversion frequency which is inevitably present at the mixer output circuit. Subsequent tuned circuits in the linear amplifiers will satisfactorily eliminate the remaining conversion frequency passed through. To satisfy the very stringent needs for reducing all out of channel frequencies to the lowest practical level, a high Q tuned circuit is employed as a simple filter after the buffer anode either at the input to the linear amplifier or alternatively adjacent to the buffer anode circuit itself. This also helps to give a band pass characteristic to the set up which is advantageous if c.w. and s.s.b. working are required at widely different frequencies within the band. This filter is not of so much importance when 28 MHz drive is used, but when it is intended to use 14 MHz from a simple exciter, it becomes essential. The selectivity of the tuned circuits and the filter is the most expedient way of achieving good intermodulation suppression with adequate out-of-band rejection sufficient to meet all GPO requirements for amateur working. The foregoing details give the background to the design and circuit of the transverter.

The P.A. Stage

What kind of stage follows will depend on what the individual operator requires from his installation,

but probably the most useful medium power linear would be the QQVO6-40A operated in Class AB1 with about 750 volts on its anode. A circuit for this valve is given which will fit into quite a small chassis and cabinet if it is desired to make a complete unit rather than have several different units. Although the power supply for the transverter can be incorporated within the same unit, this is perhaps not too advisable because of the heat factor and in general, it is better to have a separate p.s.u. in the same manner as many commercial s.s.b. transmitters and transceivers are arranged.
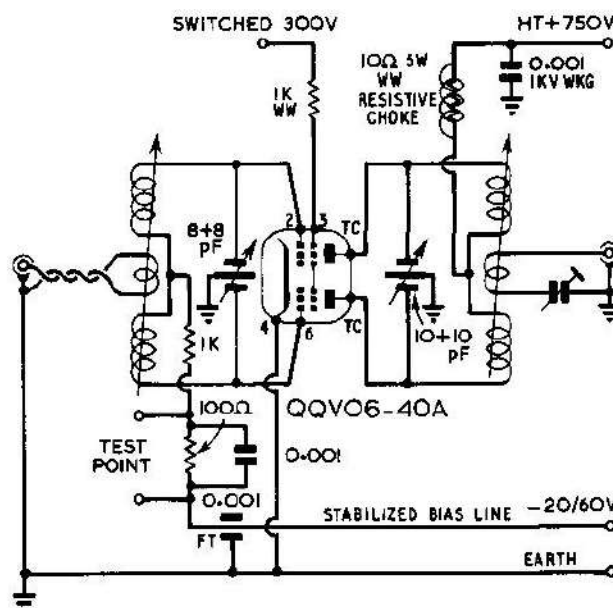
Fig.8.9. The linear amplifier.

The receiving side of s.s.b. is just as important as the transmitter. Whether transceive operation or split receiver working is going to be used, will have to be for the operator himself to choose, but there is little doubt that transceive arrangements are by far the most effective operationally. No matter which method is adopted, the receiver converter set-up will be the same. Any existing v.h.f. converter can be used, of course, provided that the i.f. is correct for the s.s.b. receiver. It is better to take a little r.f. from the last multiplier in the transverter oscillator chain and use this for receiver converter oscillator frequency injection.

A single turn loop very loosely coupled to the E180F anode coil with coaxial cable to the receiver converter chassis is all that is needed. The original last multiplier anode coil in the receiver converter can be link coupled to this new r.f. feed source and this will give the minimum of modification to the original circuit and layout. The oscillator chain on the receiver converter is disabled of course. The receiver converter i.f. out is fed to the s.s.b. receiver or the transceiver input and the rest is normal operating technique as on the h.f. bands.

Control Switching

There are many ways of achieving this but the main point to consider is that all control should be done from and by the s.s.b. prime mover. This is because if you are accustomed to a particular facility you will naturally want to have the same facility on v.h.f., such as VOX or Press to Talk.

The switching for the transverter and linear is simple and all that is required is to disable the screen circuit of the linear and the h.t. supply to the QQV03-10 buffer amplifier when on *receive*. Aerial changeover to the transverter must be operated from auxiliary contacts on the s.s.b. transmitter control circuit relay or switches.

Lining Up The Transverter

All tuned circuits should be checked with the grid dip oscillator and the output of the buffer should be terminated in a 2 watt flashlamp bulb. The oscillator chain must be tested for stability and output power by means of a good receiver and flashlamp loop or a v.t.v.m. With no oscillator or s.s.b. drive the mixer anode current should be about 15 to 20mA and the buffer current will be about 40mA. Check for any sign of self oscillation by tuning all capacitors and ensuring that the anode currents do not change.

With the conversion frequency strip only operating drive the anode current of the mixer up to 30mA with all multiplier circuits of the oscillator chain resonated, including the drive filter and the input coil of the balanced mixer. It is likely that anode current will be driven up to 60mA or more on first doing this, in which case reduce the coupling from the filter to the mixer input, still ensuring correct 'on the nose' tune, until the anode current is 30mA. Switch on s.s.b. with carrier to about ½ watt output or so and observe the mixer anode current, which will rise slightly. Set the s.s.b. exciter to normal working with maximum output into load, or from drive stage, into the transverter variable drive attenuator control which is now set to give a peak rise of mixer anode current between 'carrier on' and 'no carrier' of no more than 2mA. With this amount of s.s.b. carrier (or tone) drive, peak up the mixer anode and the buffer anode to give one to two watts in the bulb load.

Adjust couplings to ensure that no grid current is present on the buffer stage and adjust maximum r.f. transfer from buffer anode to its temporary load. Reduce s.s.b. drive to suppressed carrier conditions and see that lamp load goes out immediately. Connect the linear amplifier to the transverter output and drive with carrier inserted as in the previous test. Tune the 145.41 MHz input filter to resonance as well as the linear grid circuit and adjust the coupling between the two for the start of grid current to the linear. Overcouple to load by 10% as a good starting point for the linear to work into, until a more accurate means of
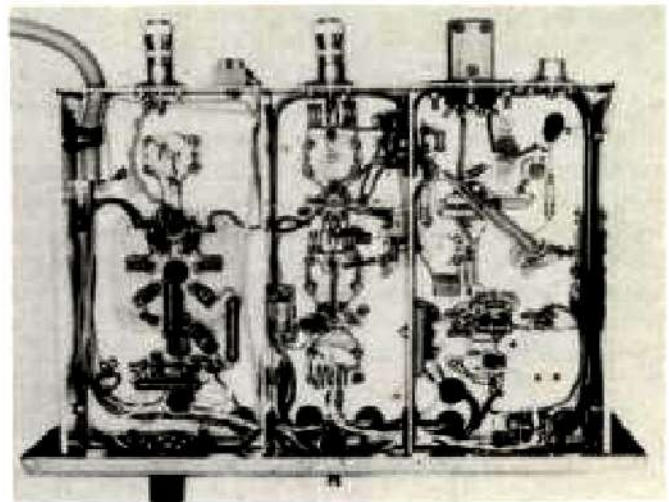
Fig.8.10. Under chassis view of the transverter.

checking linearity is available. If a QQV06-40A is used with 750v h.t. then adjust the stabilised bias supply to the grids to give an anode feed of 35 to 40mA standing current.
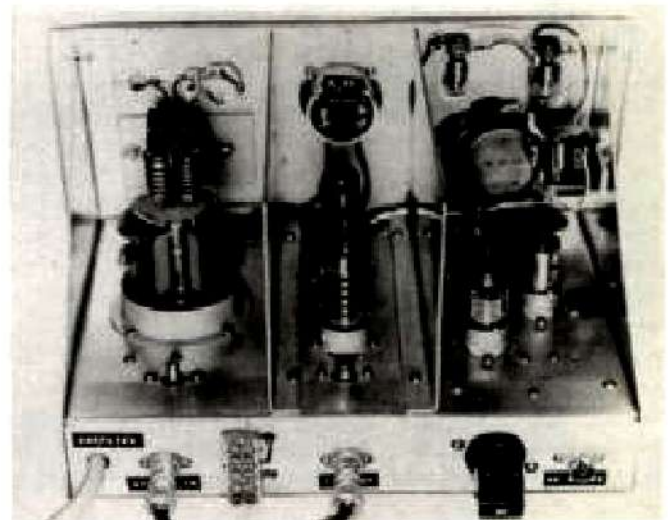
Fig.8.11. Above chassis view of the transverter.

Full tone or carrier insertion will drive the anode current up to 200mA on resonance and about 225mA off resonance - do not run at this level for long periods or the valve will be damaged. Having completed the c.w. test, then increase the anode current to an average of 120mA maximum, adjusting the drive after the s.s.b. exciter volume control has been set for normal working conditions. Adjust the s.s.b. drive after this has been done, on the transverter variable potentiometer control. The QQV03-10 buffer must be in Class A, so adjust input coupling for this. Again adjust couplings for AB1 operation of the linear. The looser the coupling the better will be suppression of any spurious frequencies. Each stage has more gain than is required, so reduce the power on the coupling coils but maintaining tuned circuit resonance within the pass band.

Other Bands

For those who intend to operate on 70 MHz the table gives useful crystal drive frequencies with 14 MHz s.s.b. drive. No change is needed apart from LC constants and the fact that the E180F stage is made into a buffer instead of a doubler. It has its grid circuit inductively coupled to the anode of the multiplier through a four turn coupling coil, and screening between the input and output circuits. For 70cm a transverter is quite a practicable design and some details of this are given later in this chapter.

Band	Conversion Frequency Crystal		S.s.b. Drive Frequency
70 MHz	l.f.	6.225 MHz	14 MHz
	h.f.	6.275 MHz	
144 MHz	c.w.	6.420 MHz	28 MHz
	s.s.b.	6.500 MHz	

DUMMY LOAD ATTENUATOR

To drive a transverter with s.s.b. such as described it is only necessary to have a p.e.p. output from the exciter of about one watt. This can sometimes be obtained from the penultimate stage of the exciter unit, but not always. Most low or medium power s.s.b. transmitters have a 6146 or perhaps a pair in the final, which means 50 or 100 watts p.e.p. output r.f. This power level is far too great for the transverter to accept and this has to be reduced. It is not possible just to turn down the audio gain control on the s.s.b. exciter, as this does not in any way attenuate the suppressed carrier, transmitter noise, spurious carriers or combination products which can drive the transverter at full output with the most undesirable results. For simplicity it is very much better to run the complete s.s.b. exciter or low power transmitter into its usual dummy load at its design power level and to take a portion of the load r.f. voltage and feed this to the transverter input attenuator potentiometer. By terminating the s.s.b. transmitter in its correct load and operating normally, all spurious and other carriers are the normal amount down compared to peak output. It is also more convenient when changing from h.f. to v.h.f. band operation. Also the s.s.b. transmitter does not have to be modified to fit links around small driver coils or switches in h.t. and heater supplies to the transmitter high power stages.
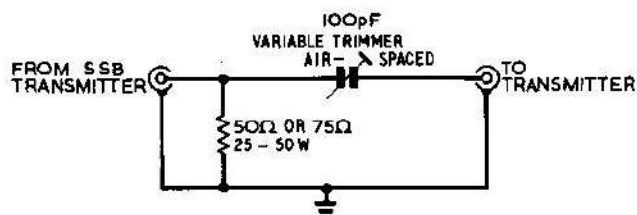
Fig.8.12. The dummy load attenuator.

If a commercial dummy load is not available then a satisfactory substitute can be made from a 50 or 75 ohm carbon resistor of 25 or 50 watt rating as described under power measurement. The 'hot' end of the load is taken to a 10/100pF variable capacitor and this is connected to the inner conductor of the coaxial lead which plugs into the transverter 's.s.b. input' socket. The correct setting of this capacitor will be that which produces about seven volts r.m.s. across the transverter input attenuator with its setting at maximum gain and the s.s.b. exciter producing its maximum rated power output into the dummy load.

For s.s.b. transmitters with a single 6146, or the equivalent, in the output stage, a suitable dummy load attenuator can be made with ten 500 ohm 3 watt carbon resistors in parallel and fitted into a large diecast box with its series capacitor and appropriate coaxial sockets for input and output r.f. Apart from the v.h.f. applications, the load attenuator is also useful for loading the driver stage of a large linear which has normal tuned input circuits and so only requires small driving power and much less than the exciter normally gives. It also provides a useful monitoring device for connecting to an oscilloscope or linearity detector when testing the h.f. band transmitter as well as other equipment.

THE VARIVERTER

Transmitters for v.h.f. and u.h.f. are usually crystal controlled. The use of crystals for common frequency working or a 'personal frequency' has something to commend it, but from an operating point of view it can be very tedious and timewasting to have to search over a band which is hundreds of kilohertz wide in the case of four metres and several megahertz for 2 metres and 70 centimetres.

Design Concepts

Perhaps the most obvious method of v.f.o. control would be to make what is in effect, a crystal substitute and plug this in to the crystal socket. This has several disadvantages, the most obvious one being that of stability. A fairly reasonable v.f.o. for 8 MHz may have a drift of 1 kHz per half hour. This would be perfectly adequate if the drift were not multiplied in the transmitter stages, however, as we know, to get to 2 metres an 8 MHz oscillator frequency has to be multiplied by a factor of 18 - that means a drift rate of 18 kHz per half hour. This of course is not acceptable, all the more so as there is a degradation of tone which is a frequency function as well. Another idea has been to use a Kaliatron oscillator on 72 MHz in a very well made box. Both these systems have the same difficulty when it comes to eliminating pick up in the receiver of the oscillator drive frequency on the band which is being used. In other words, it is not possible to

operate net frequency unless the v.f.o. is switched off, and if this is done, the drift becomes even more intolerable. A more workable system is, of course, to arrange a simple system of frequency shift, which is easily done with a varicap diode, but a different approach may be made for a v.f.o. with the stability that is suitable.

In the design of s.s.b. transmitters, it is necessary to mix the v.f.o. plus s.s.b. generated signal with a crystal oscillator conversion frequency to achieve the final frequency. This is a very old principle and one which is almost universal in its application to present day variable frequency drive generation. Its main advantage is that there is no multiplication of drift or fm hum noise. If the v.f.o. is stable to 100 hertz, then broadly speaking so will be the output on the final frequency. The same thing can be done to derive a stable v.h.f. output from a low frequency v.f.o. drive. To take practical figures: if an 8 MHz drive were to have a 1 kHz drift per half hour as in the previous example, and if the crystal conversion frequency drift were to be 100 hertz per half hour, then a v.h.f. drive with a drift of 1.1 kHz per half hour would result.

Variverter Principle

The word variverter has been coined from Variable Frequency Converter and it is basically a mixer, oscillator and amplifier arrangement which will change a crystal controlled drive at a frequency of 144-f MHz into a variable one covering 144-146 MHz, where f is a variable low frequency of 2 MHz range. To take a practical case: a crystal controlled oscillator operating on the third overtone of 7.3 MHz will give 21.9 MHz which when multiplied up by 6 will produce 131.4 MHz at the output of the chain. If a v.f.o. covering 12.6 to 14.6 MHz is added to 131.4 MHz, an output of 144.0 to 146.0 MHz will be available. The choice of these rather odd frequencies will become apparent a little later when the other benefits of the system are decribed. The Variverter is a "black box" which has a variable frequency element built into it for converting crystal drive to variable frequency drive.

Variverter Practice

From the block diagram of a modern crystal controlled v.h.f. transmitter, it is seen that it consists of, (a) a crystal oscillator and multiplier chain; (b) a low power r.f. amplifier and possibly (c) a high power final r.f. amplifier. The output of the multiplier chain is usually about a half watt and sufficient to drive the low level amplifier into Class C conditions. The coupling between the last multiplier and the low power amplifier is usually inductive or by means of a link line. Whatever system is used, the first step to be taken to use the Variverter method is to screen the

output multiplier anode coil from the low level amplifier grid and institute a new coupling circuit between the two stages, one which is link coupled to both, but the route is taken via a pair of sockets mounted at some convenient place on the rear of the transmitter chassis apron. These two sockets are connected together by a short double ended plug lead and thus the v.h.f. transmitter is not out of action whilst the Variverter itself is being made, but when it is ready it can be connected into the transmitter in place of the loop after changing the transmitter crystal frequency and retuning the multipliers to give an output frequency of 131.4 MHz.
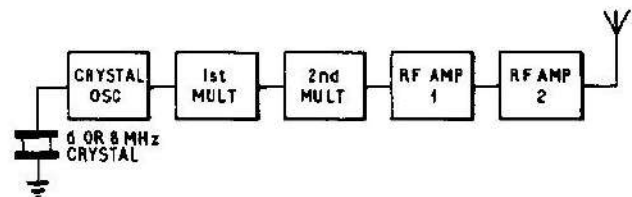
Fig.8.13. Block diagram of typical medium power 2 metre transmitter showing the r.f. link-up.

The 131.4 MHz is fed out of the transmitter into the Variverter, its output will be half to one watt on 144-146 MHz range, this is then fed back into the transmitter, to the low level amplifier grid input circuit.
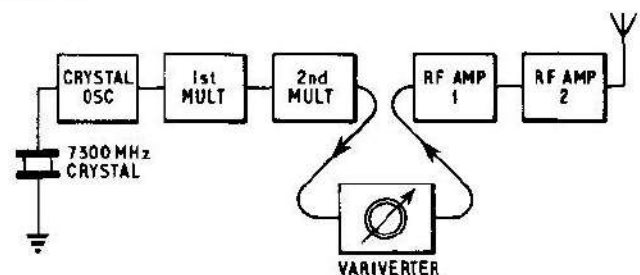
Fig.8.14. The same transmitter as Fig.8.13. but with the Variverter unit interposed.
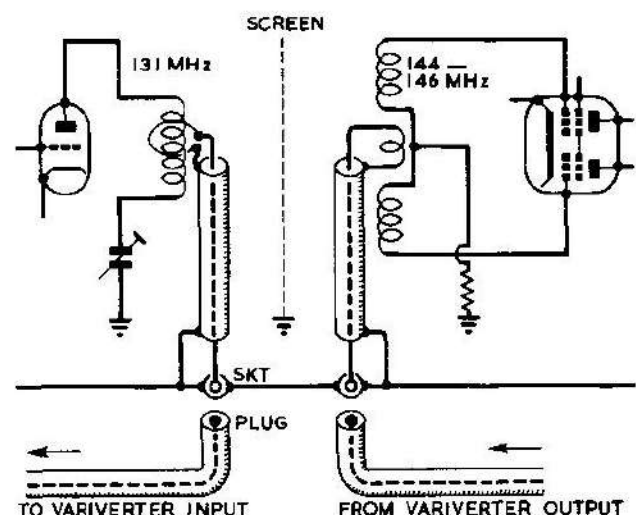
Fig.8.15. Illustrating the coupling between stages when using the Variverter unit.

Constructing the Variverter

It is best to choose a cabinet and chassis of reasonable dimensions to contain a receiver type power supply and three single valve stages. The three stages are (a) v.f.o., (b) balanced mixer and (c) buffer amplifier. The v.f.o. is of the conventional high C electron coupled oscillator type operating with a 6.3 to 7.3 MHz tuning range in its grid circuit and doubling in the anode with a medium impedance output link to the grid centre point of the balanced mixer stage input. A good slow motion dial is essential for a v.f.o., and preferably one with a fast and slow speed of motion to enable a quick change in frequency and accurate resetting to be effected with certainty.
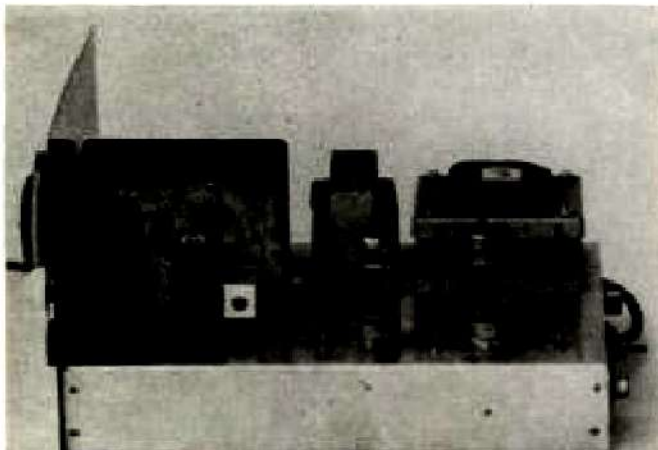
Fig.8.16. A general view of the Variverter.

The balanced mixer is a QQV03-10 double tetrode valve or two 5763, but the former is slightly more efficient than the latter. The output is inductively coupled through a 2 turn link direct to the grids of another QQV03-10 which operates under near class A conditions and this amplifies the power up to at least a watt of r.f. output which is more than enough to feed back to the transmitter r.f. amplifier stage. Filters are used to ensure that spurious frequencies are kept to a minimum, particularly those out of the band. The particular mixer configuration used effectively reduces the v.f.o. frequency and its harmonics at its anode circuit, and the tuned circuits and filter which follow the mixer ensure that unwanted products are attenuated to a level practical for amateur operation. Since the crystal chain final frequency and the unwanted mixer products may be radiated, it is essential to screen fully the circuits and preferably to feed the aerial through a band pass filter.

The circuits are quite straightforward and can be easily followed from the diagrams and photographs. If the Variverter is arranged to operate continuously then there will be little or no drift after the v.f.o. box and its components have achieved a steady temperature. With instant switch on the drift should not be more than 2 kHz in the first hour and this will settle to a few hundred hertz thereafter.
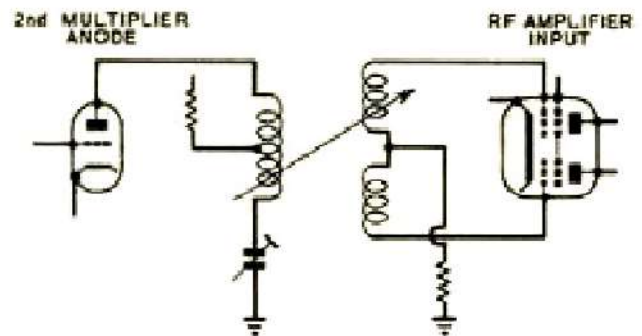
Fig.8.17. A common form of interstage coupling between last multiplier anode and first amplifier grid.

Adjustment of Mixer and Buffer

With any mixer it is important that it be operated correctly in order to prevent intermodulation products and consequent spurious radiations. The Variverter is no exception to the foregoing. Care must be taken to ensure that the mixer is not overdriven. Make sure that the v.f.o. drive is about 1/10th of the conversion frequency drive and couple each stage just enough to produce the drive or output needed.
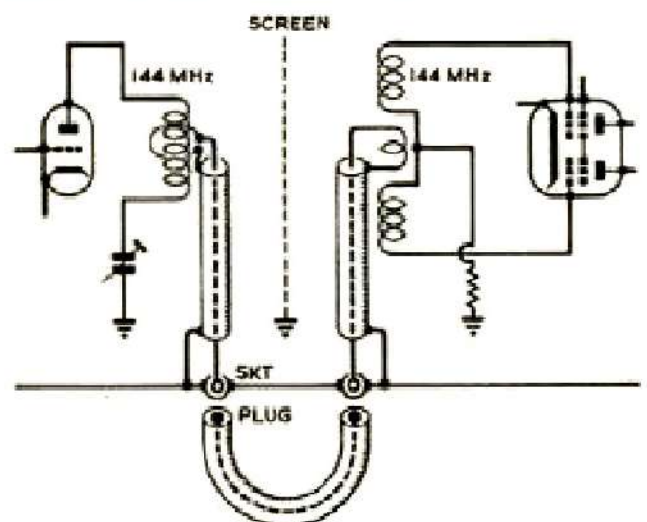
Fig.8.18. Showing a modified interstage coupling between the final multiplier anode and first amplifier grids incorporating an external link connector.

With the aid of a general coverage receiver, check the range of the tuning of the v.f.o. and its doubler output and adjust capacitors and coils to suit, check that the stabiliser tube is operating correctly and measure drift and tonal quality of note with the aid of a b.f.o. or frequency standard. The anode coils and the output filter tuned circuits should be set to 145 MHz initially with a g.d.o. Terminate the output in a small lamp load with moderate couplings throughout the unit. After this preliminary setting up, the adjustment should proceed as follows: disconnect the v.f.o. h.t. but leave all other h.t. connections made. Switch on the unit and measure anode currents without any drive
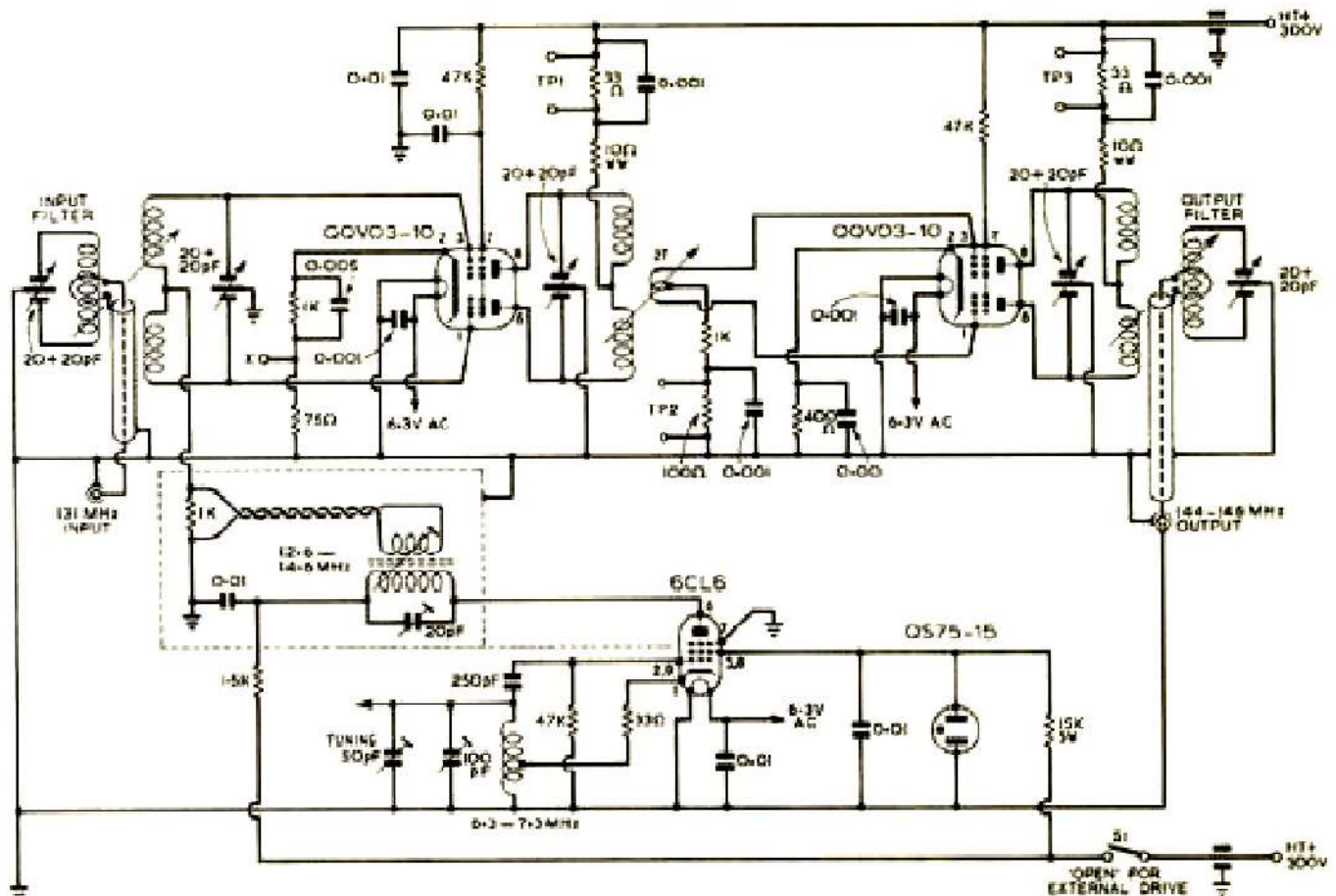
Fig.8.19. Circuit of the Variverter unit. Point X is for insertion of external 14 MHz drive for use with s.s.b., a.m., or c.w. TP 1-3 are for checking working conditions when lining up the unit.

connected to the mixer input. The mixer anode current will be about 15mA and the buffer amplifier about 40mA with an h.t. rail of 300 volts. Check to see that there is no self oscillation which will indicate its presence by variations in anode currents or illuminated load lamp in a severe case. It is unlikely that any self oscillation will take place, but if there is, the screening between stages should be checked for adequacy. Having passed the static and stability tests, drive from the transmitter at 131.4 MHz should be applied whilst watching the anode feed on the mixer. Trim the input filter and the mixer grid for maximum drive. Adjust the filter to grid coil coupling to set mixer anode current at 30mA maximum. Switch on h.t. to v.f.o. and set v.f.o. to 13.4 MHz. Watch rise of mixer anode current to see that it is not more than one or two mA. Reduce coupling to the v.f.o. anode circuit if it is more than this amount. Tune the mixer and buffer anode circuits and output filter and the lamp will light at about one watt level. Adjust the input coupling to the buffer under these conditions until no grid current shows in the centre tap resistor. Adjust output filter for maximum r.f. When all the foregoing have been completed the Variverter is in a narrow band working condition. To make it broadband, a little stagger tuning of stages is required and slight increase of

coupling to maintain a fairly even level of ½ watt of r.f. drive output. The lamp is removed and the output taken to the transmitter first amplifier input and the transmitter checked over for drive adequacy and general working conditions into a dummy load. Each of the two drive sources to the mixer should be switched off in turn to check that no signal whatsoever is radiated under this condition.
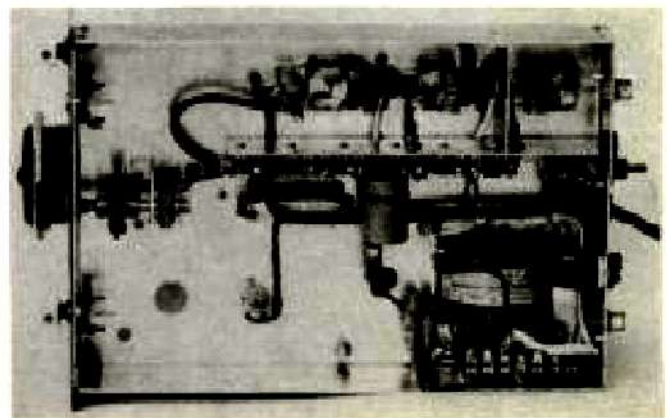
Fig.8.20. Showing the under chassis view of the Variverter.

All the normal transmitter switching can be used with the Variverter so that there is nothing to modify on the transmitter in this respect. If there is no crystal drive there is no output signal at 145 MHz and the station receiver functions normally. Switching on the crystal chain will produce a 2 metre signal from the mixer and this can be used for netting on to the channel desired before applying h.t. to the r.f. amplifier and radiating.

Additional Benefits

All that is required to radiate s.s.b. is to drive the balanced mixer with 14 MHz s.s.b. at about the one watt level instead of the v.f.o. in the Variverter, and out comes s.s.b. on 2 metres. The diagram, Fig. 8.19, shows the preferred method for doing this which does not cause complications if a.m. working is also required. The crystal frequency of 131.4 MHz was deliberately chosen with the eventual use of s.s.b. in mind. If 14.01 MHz is used from a s.s.b. exciter, add this to 131.4 MHz to make 145.41 MHz the s.s.b. channel internationally used in Europe and this country for two-way s.s.b. contacts. The Class A buffer also was adopted so that s.s.b. signals could pass through without distortion.
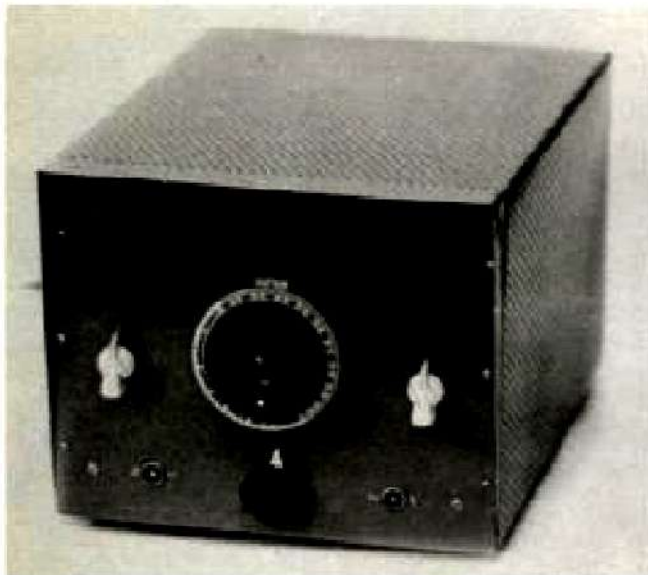
Fig.8.21. Front view of the Variverter.

Although the details given are for 2 metres, similar circuits can be used on 4 metres with suitable coils and tuning ranges. Crystals which can be used for 4 metres with 14 MHz drive are 6225 or 6250 MHz, operated in third overtone to help keep unwanted beats to the minimum.

A High Power Linear For 144 MHz.

The QQV06-40A is suitable as a linear amplifier for either 2 or 4 metres for up to 80 watts p.e.p. output under AB1 conditions. In order to achieve the legal limit of 400 watts peak r.f. output with linearity, it is necessary to use larger valves. The 4X150 series of tetrodes are the obvious choice because of their efficiency up to 500 MHz. Although these types of valves are of quite-old design, they are likely to be used for at least another decade.

Layout and General Design

The linear described was deliberately made physically large so as to prevent any difficulties being met with when it came to operation. A box construction with three separate compartments is used, the first compartment, which contains the grid tuning assembly, is made airtight and pressurised from a large air blower. Adequate cooling of the valves is essential for satisfactory performance.

The anode lines too must be efficient and as loss free as possible as they can aid in the general cooling if designed properly. Standard copper tube stock can be obtained which will just make a close fit over the anode cooling fin assembly of the 4X150 series valves, and this makes an excellent basis for an efficient anode circuit of reasonable size which is easy to couple to and also readily permits valve changing when this becomes necessary. The cooling air, after it has passed through the air flow socket and the actual anode itself, passes down the anode lines which get hot after conducting the heat away from the valve. The cooling air keeps the line temperature down effectively and this in turn keeps the anode cooler than it would be on its own. The short circuit bridge at the h.t. feed end of the lines must be a good fit with best quality copper used in a thoroughly well clamped arrangement. The anode tuning is effected by means of a flap of copper acting between two large plates mounted underneath the lines; the plates are separately adjustable to give a slight variation of anode to anode capacitance so as to allow the main variable tuning flap to be set at a practical spacing to cover the band. The h.t. end of the lines are supported by a p.t.f.e. plate mounted on a dividing partition, and the h.t. feed leads to the bypass capacitor and choke/resistor pass through this. These components are thus in a field free area which prevents any trouble from r.f. affecting the h.t. choke or getting on to the h.t. cable. The hot air from the lines passes out of the side of the box through a wire gauze outlet. This makes it very easy to check equality of heat dissipation from the two valves under working conditions with safety; no dangerous potentials are exposed on that side of the compartment.

Electrical Design

Push-pull operation of the two valves is employed for efficiency and ease of neutralisation.
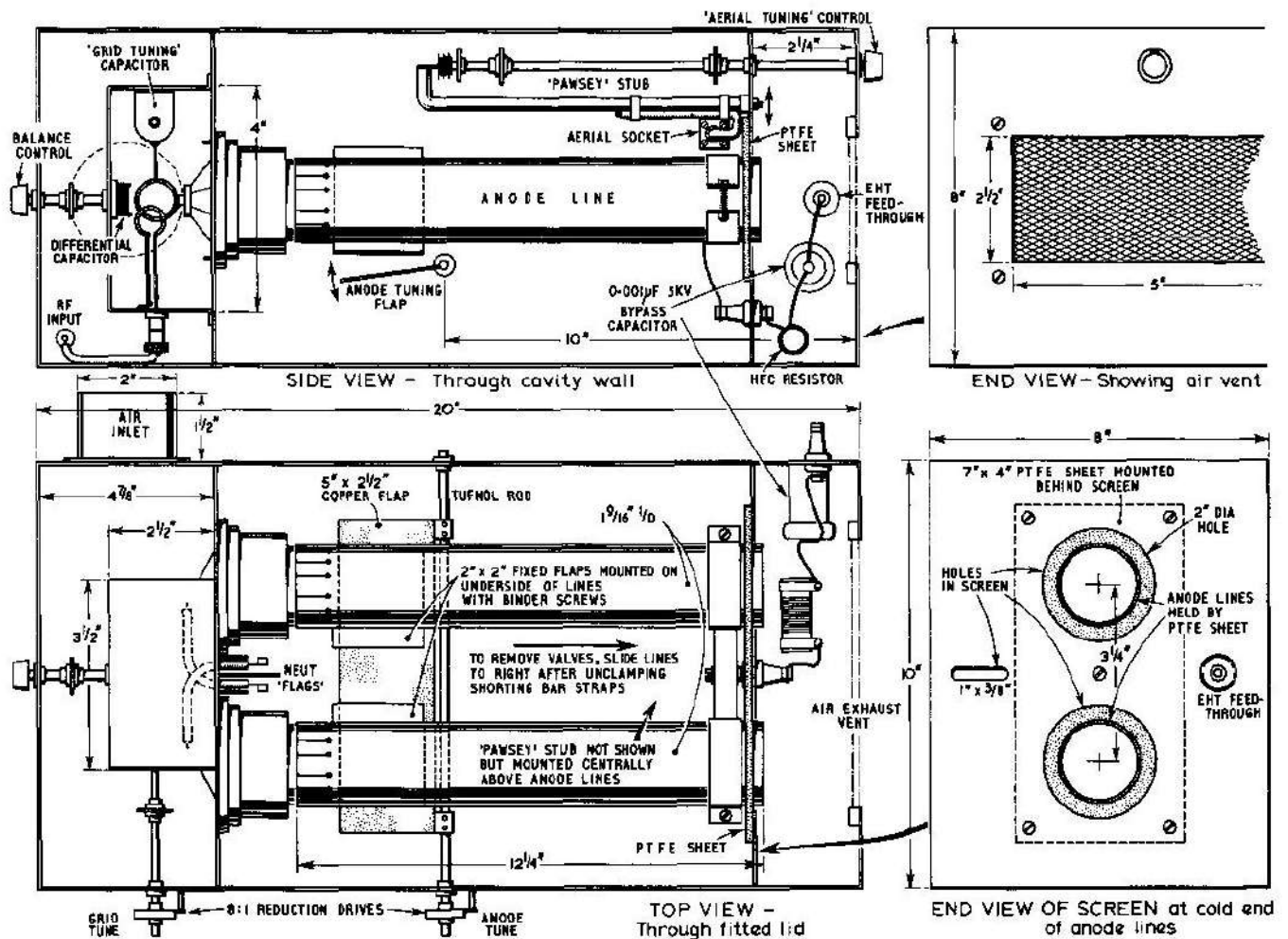
Fig.8.22. Interior of high power linear amplifier.
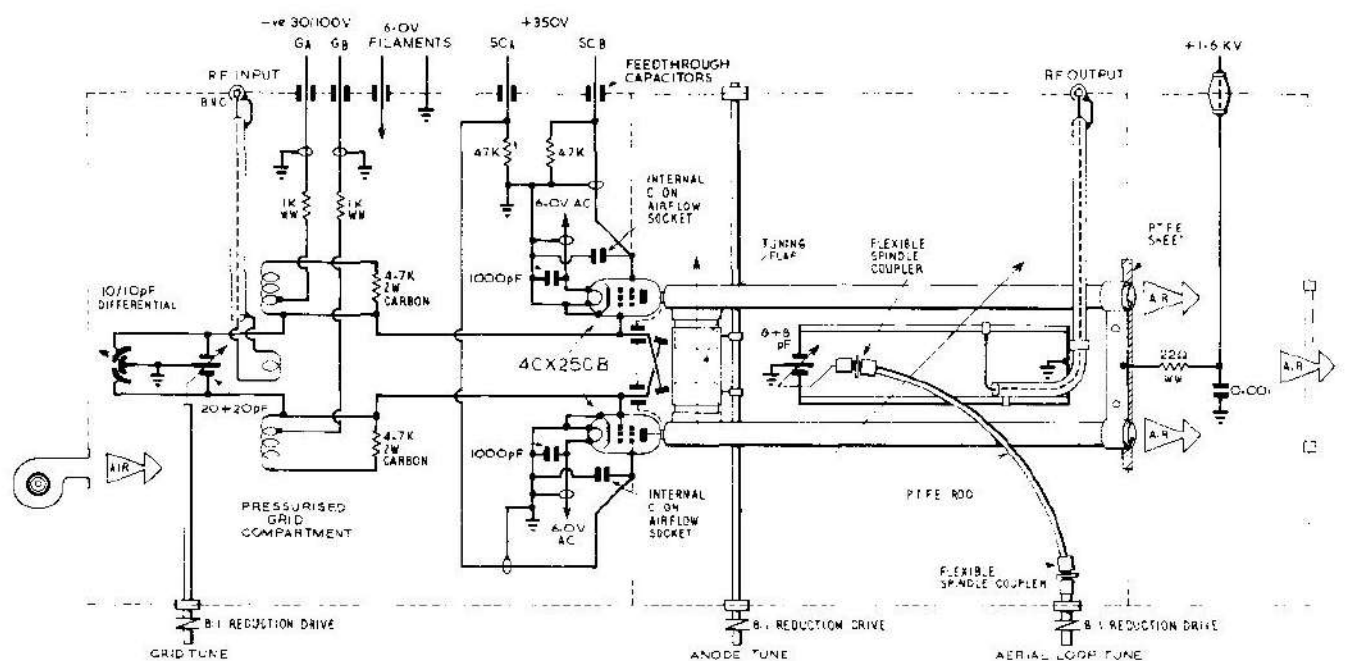
Fig.8.23. Circuit of the high power 2 metre linear amplifier.

The input is a split grid coil 'series tuned' arrangement with a drive balancing differential capacitor across the tuning condenser. Wirewound resistors are mounted at the r.f. dead point on each coil for connection to the bias to each grid. Separate grid and screen current metering is provided. The combined anode currents are measured for simplicity and to prevent loss by having to divide the anode short circuit with a capacity assembly. The cathodes are multi-pin and connected to earth on to the air sockets by thick wire or copper tape. All supply leads are screened up to the feedthrough capacitors mounted on the grid compartment rear wall. Preset coupling of the input link is used on the grid assembly and this link is connected by coaxial cable from a socket at the rear. Split stator tuning of the grid to grid circuit is employed. Feed through insulators for the neutralising leads are mounted close to the air sockets and connected to the two grid terminal soldering lugs on one side and to small flag-like plates on the anode side of the screening partition. The feedthroughs are electrically screened from 'seeing' the valve screen assembly by two semi-circular shields made from copper sheet. The anode lines are quarter wave long with the smallest amount of end capacitance practical to tune them. The aerial feeder is Pawsey stub connected and this stub is mounted above the anode lines. Tuning of the stub is by flexible cable drive to an end mounted split stator capacitor.

Alignment Procedure

With the valves in circuit, and no power on, the input and output can be checked by g.d.o. to enable coil and line adjustments to be made to give resonance on 145 MHz. A quick check on the neutralising adjustment can be made by coupling the grid dip oscillator to the anode lines whilst tuning through resonance on the grid - the plates are adjusted for the condition of 'no kick' when resonance is passed through, indicating little or no coupling between the input and output circuits.

The heaters are switched on after the blower has been running for a few minutes and the air flow checked on the end of the anode lines. No more than 6.0 volts should be permitted on the heaters and no harm will be done if it is kept down to 5.8 volts as a certain amount of cathode back-bombardment will take place when the amplifier is operational, and this will increase the cathode temperature and emission as a consequence.

After the heater voltage has been adjusted, the top covers are put on and screwed down. Apply grid voltage and check for grid heater contacts by observing that there are no reversed grid current meter readings. Apply a little r.f. drive to the grid circuits just sufficient to show some grid current, and adjust for balance with the differential capacitor in combination with the grid to grid tuning control. Swing the anode capacitor through resonance and see

that there is no dip of grid current over the whole anode tuning range, if there is, re-adjust the neutralising plates until no reaction is seen. Remove the r.f. drive and disconnect the screen supply voltage. Switch on the h.t. at about 1500 volts or less and observe the anode current, which should be zero under these conditions. Now apply drive to the grids just sufficiently for the start of grid current and observe the anode current meter which should still be at zero or just reading a little current at anode resonance. If there is some anode current, then note its value and switch off and isolate all dangerous supplies. Earth anode lines with an earthing wand before touching anything inside the compartments. Make a small adjustment to the neutralising flags one way and repeat the previous operation still without screen volts on the unit. If the anode current is now nearer zero, the neutralising has been done the right way. When no further improvement to the neutralising can be obtained, the amplifier should be checked under r.f. power conditions. Connect a load and reflectometer to the output r.f. socket and couple the Pawsey stub fairly tightly to the anode lines as a commencement. Without r.f. drive, switch on power to the amplifier, this time with screens connected to their stabilised supply at a low voltage. Set the grid bias for 100mA anode current. Leave things like this for a minute and then feel the air outlet and see that both valves are about the same dissipation, which they will be if their emission is good. Increase the screen voltage towards 350 in gradual steps. Still with no drive, make perfectly sure that the amplifier is stable by trying every combination of grid and anode tuning capacitor settings. The anode current must not move at any tuning point at any time. If this test is satisfactory and it must be before proceeding any further, apply a very little r.f. drive and tune the anodes for maximum in the forward power position of the reflectometer or load meter. Adjust Pawsey stub tuning for maximum power. Increase the drive up to grid current point and load up the anodes to about 450mA. Check power output if this can be done, and see that it is about 400 watts r.f. when the key is pressed. Overcouple by 10% to get a reasonably linear condition. Check on a local receiver, with its aerial input terminated in a resistive load, that there is no "hash" audible just off tune with different r.f. drive inputs which would indicate possible parallel mode instability. If all is well, then the aerial can be connected and the power gradually applied by means of the s.s.b. drive control. A reflectometer check on standing wave and r.f. power should be made before putting out a call on 145.41 MHz. The help of a local station should be enlisted to ensure that the transmission is clean and free from unwanted combination products, but do choose someone who knows what to look for and who is experienced in s.s.b. operating.

Making a linear as described is not for the novice, but for the amateur who has both technical and practical ability, it should not be difficult. The

voltages used are lethal, so absolutely rigid discipline must be exercised when getting the amplifier lined up.

The h.t. is left on at all times and is not switched by the control system. It is good practice, however, to remove the screen voltage during reception periods and to over bias the grids. This can be done through an auxiliary relay operated from the control circuit on the s.s.b. exciter. Make sure that the aerial changeover relay is very quick acting and that r.f. is not available until a few milliseconds after the relay has made contact. Power supplies for an amplifier such as this are conventional in design and this information can be obtained from the RSGB *Radio Communication Handbook*.

400 watts p.e.p. output and an aerial system of 13dB gain would mean an e.r.p. peak of 8kW, and that is a great deal of power to wield about on the band. For a local QSO, the s.s.b. transmitter drive control can be turned right down to achieve whatever power is adequate for a reasonable signal to be received. Linearity of the final will be very good, of course, under such conditions. Full power should only be resorted to for really long distance transmission or when fading is very deep. For satellite communication, high power s.s.b. is very useful, particularly if a synchronous satellite should ever be put up at its great height above the earth. Meteor scatter phone work and auroral bounce communication is quite feasible on s.s.b., but the full power is really required to do this at all successfully.

CHAPTER 9

AERIALS

In the v.h.f. and u.h.f. region there is a greater choice of aerial design available to the average amateur since the dimensions of the individual elements allow relatively elaborate arrays to be erected without too much trouble.

The actual choice will to some extent depend on the operator's interests and usually the array will be required to be fully rotatable. Because of the relatively small size of these aerials, high gain, multi element, narrow beam width aerials are extensively used, especially for long range working; alternatively very satisfactory general purpose performance can be obtained with a simple two-element quad. Height and good siting is of greater importance at these frequencies than at the lower frequencies, and in addition the length of feeder with its loss sets some limitation on the position of the aerial relative to the equipment.

Mobile aerials are very practical at 70 and 144 Mhz, and excellent performance can be obtained from whip and halo types respectively, using quite low power outputs.

For those who want only local contacts, simple omnidirectional aerials, such as a true slot, can be installed inside the roof space. At the highest frequency band covered in this Manual, 23cm, both dish and corner reflector type aerials are used extensively, and similar aerials are normally necessary for 2m and 70cm when long range work, such as moon bounce, is to be undertaken.

Power Gain and Beamwidth

A simple way to appreciate the meaning of aerial gain is as follows. Imagine the radiator to be totally enclosed in a hollow sphere, as indicated in Fig.9.1. If the radiation is distributed uniformly over the interior surface of this sphere the radiator is said to be isotropic. An aerial which causes the radiation to be concentrated into any particular area of the inside surface of the sphere and which thereby produces a greater intensity than that produced by an isotropic radiator fed with equal power, is said to have a gain. This gain is inversely proportional to the fraction of the total interior surface area which received the concentrated radiation.

The gain of an aerial is usually expressed as a power ratio, either as a multiple of so many "times"

A ...... ILLUMINATED AREA
ϕ ...... VERTICAL BEAMWIDTH (IN PLANE YZ)
θ ...... HORIZONTAL BEAMWIDTH (IN PLANE XZ)

DIRECTION OF MAXIMUM RADIATION

$$\text{POWER GAIN} = \frac{\text{SURFACE AREA OF SPHERE}}{\text{SURFACE AREA OF "ILLUMINATED" REGION}}$$
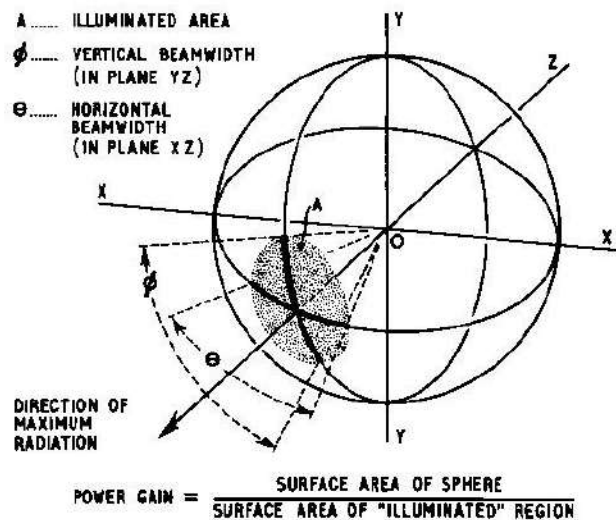
Fig. 9.1. Radiation from an aerial. An isotropic radiator at the point 0 will give uniform "illumination" over the inner surface of the sphere. A directional radiator will concentrate the energy into a beam which will illuminate only a portion of the sphere as shown shaded.

or in decibel units. For example, a power gain of 20 times could be represented as 13dB (i.e. $10 \log_{10} 20$).

The truly isotropic radiator is a purely theoretical concept, and in practice the gain of beam aerials is usually compared with the radiation from a single half-wave dipole fed with an equal amount of power. The radiation pattern of even a single half-wave dipole is markedly non-uniform, and in consequence the power gain of such an aerial compared with the hypothetical isotropic radiator is about 50 per cent (i.e. 3/2 times or 2dB), but since the half-wave dipole is the simplest practical form of radiator it is generally acceptable as a basis of comparision.
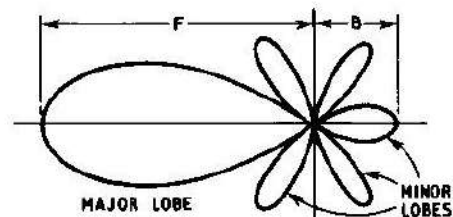
Fig. 9.2. Typical polar diagram of a v.h.f. aerial. The front-to-back ratio is represented by F/B.

The area of "illumination" is not sharply defined as shown by the shaded region in Fig.9.1 but falls away gradually from the centre of the area. The boundaries of the illuminated area are determined by joining together all points where the radiation intensity has fallen by half (i.e. 3dB): these are known as the half-power points. The gain of the aerial can then be determined by dividing the total surface area of the sphere by the illuminated area: e.g. if the total surface area were 100 sq. cm. and the illuminated area bounded by the half-power points were 20 sq. cm., the gain of the aerial would be five times or 7dB. The radiation in any particular plane can be plotted graphically, usually in polar co-ordinates; such a plot is called a polar diagram. A typical polar diagram is shown in Fig.9.2. The area of maximum radiation is called the major lobe. Spurious radiation occurs also in other directions and these areas, when small compared with the major lobe, are called minor lobes. All practical aerials exhibit such lobes and the aerial designer frequently has to compromise to obtain the optimum performance for any particular application. For example, an aerial may be designed for maximum front-to-back ratio, i.e. for minimum radiation in the direction opposite to the major lobe. To achieve such a condition it may be necessary to sacrifice some gain in the major lobe (or forward radiation) with a possible increase in other minor lobes (or side lobes) and thus the designer will need to consider all the implications before finalizing any particular design.

In practice the radiation from an aerial is measured in a horizontal and a vertical plane. The beamwidth is the angle between the half-power points in the plane under consideration. The vertical polar diagram is greatly influenced by the height of the aerial above ground; the higher the aerial the lower will be the angle of maximum radiation, and at the same time the effects of neighbouring objects such as houses will be minimized. The important requirement is to place the aerial well clear of such objects and this frequently means as high as can be safely achieved. Any aerial which has the property of concentrating radiation into any particular direction is said to possess directivity.

Bandwidth

The performance of an aerial array generally depends upon the resonant properties of tuned radiators such as dipoles or other frequency sensitive elements, and therefore any statement regarding its power gain or beamwidth will be valid only over a restricted frequency band. Beyond the limits of this band the properties of an aerial system may be entirely different. Hence it is useful to define bandwidth as that range of frequencies over which the power gain of the aerial array does not fall by more than a certain percentage as compared with the frequency at which maximum gain is obtained (e.g. a bandwidth of 15 MHz for a 50 per cent reduction in

power gain). Alternatively the bandwidth may be defined as the frequency band over which the standing-wave ratio of the aerial feeder does not exceed a prescribed limit (e.g. a bandwidth of 10 MHz for a standing-wave ratio not exceeding 2:1). The latter convention is the one generally used.

Capture Area or Aperture

Besides examining the action of a transmitting aerial array in concentrating the radiated power into a beam it is also helpful to examine the way in which the same aerial structure will affect the reception of an incoming signal. In this study it is convenient to introduce the concept of capture area or aperture of the aerial. This concept is frequently misunderstood, probably because it may appear to relate to the cross-sectional area of the beam (as represented by A in Fig.9.1): it is in fact related to the inverse of the cross-sectional area of the beam inasmuch as an aerial which has a high gain usually has a sharply focused beam (i.e. one small cross-sectional area) but at the same time the capture area of the aerial is large. The larger the capture area, the more effective is the aerial as compared with a simple dipole.

The actual size of the aerial system does not always give a reliable indication of the capture area. A high-gain array may have a capture area considerably greater than its frontal area as determined by its physical dimensions. The fundamental relationship between the capture area and the power gain of an aerial system is

$$A = \frac{G_1 . \lambda^2}{4\pi}$$

where A is the capture area and λ is the wavelength (measured in the same units as A) and G_1 is the power gain relative to an isotropic radiator.

A half-wave dipole has a gain of 3/2 relative to an isotropic radiator, and therefore this formula can be modified so as to give the capture area in terms of the gain of a half-wave dipole G_D instead of G_1 simply by introducing the factor 3/2, thus -

$$A = \frac{3}{2} \times \frac{G_D \lambda^2}{4\pi} = \frac{3 G_D \lambda^2}{8\pi}$$

Note that for a dipole, since $G_D = 1$, the capture area is approximately $\frac{\lambda^2}{8}$.

This formula shows that if the wavelength is kept constant the capture area of an aerial is proportional to its gain, and therefore if an increase in gain results in a narrower beamwidth it must follow that a narrower beamwidth, corresponds to a greater capture area (the term beamwidth being used here to signify both horizontal and vertical dimensions, i.e. in effect the cross-sectional area).

The formula also shows that for any given power gain the capture area is proportional to the square of the wavelength. For example, an aerial having a power gain of, say, 10 times relative to a dipole at 600 MHz (0.5m) would have a capture area one-sixteenth of that of an aerial having a similar power gain at 150

MHz (2m), and to achieve equal capture area the gain of the 600 MHz aerial would thus have to be 16 times greater than that of the 150 MHz aerial, i.e. 160 times relative to a dipole. This is unfortunate because it is the capture area of the aerial that determines its effectiveness in absorbing the incoming radiation: it means that as the wavelength is reduced it becomes increasingly important to design the aerial to have a higher gain.

These observations apply only to signals being received or transmitted in the direction of maximum gain. For directions other than the optimum the relationships become more complex.

Multi-radiator Arrays

High gain aerial arrays can be built up from a number of individual radiators such as half-wave dipoles. To achieve the maximum gain, the spacing of these radiators should be such that their respective capture areas just touch. Thus where the individual radiators are themselves high gain systems, such as Yagi-arrays, the centre-to-centre spacing of each radiating system needs to be larger, since the individual capture areas are greater.

Reciprocity Theorem

The theorem of reciprocity states that any particular aerial gives the same performance either as a transmitting or as a receiving system. Practical aerial designs are therefore worked out in terms of transmission because the characteristics are more easily determined in this way, and the resulting aerials are assumed to have similar properties.

Angle of Radiation: Polarization

The characteristics of propagation in the v.h.f. ranges are mostly determined by the influence of the troposphere, i.e. the part of the atmosphere extending from ground level up to a few thousand feet. There is little or no ionospheric propagation on frequencies higher than about 100 MHz and thus any energy which is radiated at more than a few degrees above the horizontal is wasted. Similarly to transmit to a particular point it is unnecessary to radiate a broad beam in the horizontal plane. Generally, therefore, the aerial designer tries firstly to reduce the vertical beamwidth to avoid wastage of power into space and secondly to reduce the horizontal beamwidth according to the required ground coverage. However, a narrow horizontal beamwidth can be a disadvantage because stations situated off the beam may be missed when searching. Thus a compromise between gain and beamwidth has to be made.

Radio waves are constituted from electric and magnetic fields mutually at right angles and also at right angles to the direction of propagation. The ratio of the electric component E to the magnetic component H in free space (E/H = Z) is known as the impedance of free space and has a value of about 377 ohms. When the electric component is horizontal, the wave is said to be horizontally polarized. Such a wave is radiated from a horizontal dipole. If the electric component is vertical, as in a vertical dipole, the wave is said to be vertically polarized.

It has been found by experiment that in the v.h.f. range horizontally polarized waves suffer less attenuation over long distances than vertically polarized waves, and this system is therefore often preferred. It has in fact been universally adopted for amateur communication in Great Britain and many other countries. Vertically polarized waves may be more suitable for special purposes such as short-distance or mobile communication and a simple ground plane or similar aerial can be used.

Sometimes the polarization is not exclusively horizontal or vertical and the radiation is then said to be elliptically polarized or, in the special case where the horizontal and vertical components are equal, circularly polarized. The effect of the addition of two components of the same kind (i.e. electric or magnetic) at right angles is to create a rotating field, the direction of which depends on the relative phase of the two components. Thus the polarization of the wave will appear to have either clockwise or counterclockwise rotation, a feature which is important in the use of helical aerials. A dipole will receive an equal pick-up from a circularly polarized wave irrespective of whether it is mounted horizontally, vertically or in an intermediate position. Horizontally or vertically polarized waves are known as plane polarized waves

AERIAL FEEDERS

Before considering aerial design it will be helpful to review the methods of conveying the power from the transmitter to the aerial. The feeder length should always be considered in terms of wavelengths rather than the actual length of the conductors. If the feeder length is short compared with the wavelengths, the loss caused by its ohmic resistance and by the dielectric conductance is unimportant as also is the effect of incorrect impedance matching. However, for v.h.f. operation the aerial feeder is usually many wavelengths long, and therefore both the loss introduced and the matching of the load to the feeder are of the utmost importance.

Two types of feeder, or transmission line, are in common use, namely the unbalanced or coaxial feeder and the balanced pair: the latter may be either of open construction or enclosed in polythene ribbon or tubular moulding. Each type has its own particular advantages and disadvantages, and these are summarised below.

In a coaxial cable, the radio-frequency fields are contained entirely within the outer conductor and hence there should be no r.f. currents on the outside. This enables the cable to be carried in close proximity to other cables and metal objects without interaction or serious change of its cable properties which might cause reflections and thereby introduce appreciable loss. Also there is no loss by external radiation.

The open-wire or balanced feeder has a radiation-loss which is dependent upon the ratio of the spacing of the wires to the wavelength and becomes more serious as the frequency is raised. The properties can also be severely changed by the close proximity of metal objects and the accumulation of ice or water on the separating insulation, and therefore much greater care must be taken in the routing of the feeder. Difficulties are often experienced when attempting to use this type of feeder with rotatable aerial arrays. It is, for instance, quite unsatisfactory to bind a ribbon feeder directly against a metal mast. When the feeder is kept well clear of metal objects, however, the loss tends to be less than that of coaxial cable unless the frequency is so high that the radiation loss is serious (i.e. 450 MHz and above).

Generally speaking, the use of open-wire feeders is restricted to bands of 144 MHz and below; coaxial cable is used for all frequencies up to about 3000 MHz. Above about 1000 MHz the loss in conventional types of flexible coaxial cable becomes prohibitive and rigid semi-air-spaced types of line are then used. Above 3000 MHz waveguides are usually necessary for feeder runs of more than a few inches.

A list of typical aerial feeder cable obtainable in Great Britain is shown in the table. Further information can be obtained from the various manufacturers. Complete details of r.f. cables made to Government specifications can be obtained from Defence Specification DEF-14-A, "Radio Frequency Cables" (H.M.S.O).

In the semi-air-spaced type of coaxial cable, the centre conductor is supported either on beads or on a helical thread; in some forms a continuous filling of cellular polythene is used as the insulator. This type of cable has a lower capacity per unit length than the solid type and hence the velocity factor (i.e. the ratio of the wave velocity in the cable compared with that in free space) is higher, being about 0.88–0.98 for helical and bead types and about 0.8 for cellular polythene types compared with 0.66 for solid cable. The use of semi-air-spaced centre conductors allows cables to be designed with less attenuation for a given size, or with the same attenuation for a smaller size, but unfortunately the bead and helical types suffer from the disadvantage that moisture can easily enter the cable; special precautions must therefore be taken to ensure a good watertight seal at the aerial end if such cable is used. Suitable material for this purpose is *Telcompound** or *Bostik* sealing strip. Cellular polythene has non-connecting air cells distributed throughout its volume, and consequently there is no

moisture ingress except at cut ends and then only for a very small distance. However, cellular polythene cables do suffer from another disadvantage; this is that the effective dielectric constant of the cellular polythene varies according to the number and volume of the air cells per unit length which unfortunately cannot be controlled with precision during manufacture. Thus the characteristic impedance may tend to vary along a length of feeder and although this may not matter at the lower end of the v.h.f. spectrum such cable is not recommended for frequencies above about 500 MHz.

It is also important to exclude water even from the outer braiding of any coaxial cable. If water has once entered a cable it is impossible to dry it out, and the loss in the feeder becomes progressively higher as the copper braiding corrodes.

Open-Wire Feeders

To obtain very low losses open-wire feeder line can conveniently be made from hard drawn 16 s.w.g. copper wire with separating insulators placed at 12" intervals. The insulators should be made from polythene and be shaped in the form of a disc with the centre removed, as shown in Fig.9.3. This ensures that in the places where the maximum electric stress occurs, i.e. between the conductors, the dielectric is air and not the solid insulating material, thus minimizing losses. To prevent excessive radiation loss
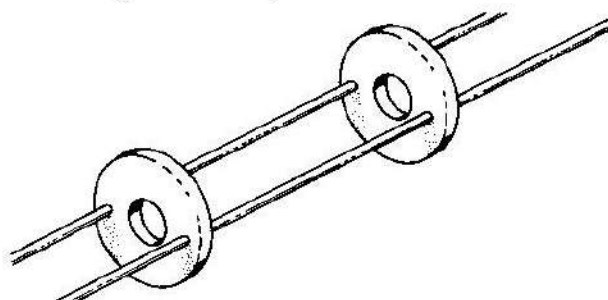
Fig. 9.3. Open wire line with disc insulators. The centre portions of the insulators are removed to minimize dielectric losses.

the characteristic impedance should not exceed about 300 ohms. Care must be taken to avoid sharp bends and also the close proximity of surrounding objects as already stated. The characteristic impedance of an open twin line is given by -

$$Z_0 = 276 \log_{10} (D/d)$$

where D is the centre-to-centre spacing and d is the diameter of the wire (measured in the same units). A chart of characteristic impedance in relation to the conductor size and spacing is given, see Fig.9.22.

Metal Insulators

A quarter-wave short-circuited transmission line presents a very high impedance at the open end and

---

CHARACTERISTICS OF TYPICAL BRITISH RADIO FREQUENCY FEEDER CABLES

Type of Cable	Nominal Impedance Z_0 (ohms)	Centre Conductor	Dimensions (in.) over outer sheath	Dimensions (in.) over twincores	Velocity Factor	Approximate Attenuation (db per 100 ft.) 70 MHz	145 MHz	430 MHz	1250 MHz	Remarks
Standard TV feeder ...	75	7/.0076	0.202	-	0.67	3.5	5.1	9.2	17	-
Low-loss TV feeder (semi-air-spaced) ...	75	0.048	0.290	-	0.86 approx.	2.0	3.0	5.4	10	Semi-air-spaced or cellular
Flat twin ...	150	7/.012	-	0.18 / 0.09	0.71	2.1	3.1	5.7*	11*	*Theoretical figures, likely to be considerably worsened by radiation
Flat twin ...	300	7/.012	-	0.405 / 0.09	0.85	1.2	1.8	3.4*	6.6*	
Tubular twin ...	300	7/.012	-	0.446	0.85	1.2	1.8	3.4*	6.6*	

R.F. CABLES-BRITISH UR SERIES

UR No.	Nominal Impedance Z_0 (ohms)	Overall diameter - inches	Inner conductor - inches	Capacity pF/ft.	Maximum Operating voltage R.M.S.	Approximate Attenuation (db per 100 ft.) 10 MHz	100 MHz	300 MHz	1000 MHz	Approx. RG. equivalent
43	52	0.195	0.032	29	2750	1.3	4.3	8.7	18.1	58/U
57	75	0.405	0.044	20.6	5000	0.6	1.9	3.5	7.1	11A/U
63*	75	0.853	0.175	14	4400	0.15	0.5	0.9	1.7	-
67	50	0.405	7/0.029	30	4800	0.6	2.0	3.7	7.5	213/U
74	51	0.870	0.188	30.7	15000	0.3	1.0	1.9	4.2	218/U
76	51	0.195	19/0.0066	29	1800	1.6	5.3	9.6	22.0	58C/U
77	75	0.870	0.104	20.5	12500	0.3	1.0	1.9	4.2	164/U
79*	50	0.855	0.265	21	6000	0.16	0.5	0.9	1.8	-
83*	50	0.555	0.168	21	2600	0.25	0.8	1.5	2.8	-
85*	75	0.555	0.109	14	2600	0.2	0.7	1.3	2.5	-
90	75	0.242	0.022	0	2500	1.1	3.5	6.3	12.3	59B/U

All the above cables have solid dielectric with a velocity factor of 0.66 with the exception of those marked with an asterisk which are helical membrane and have a velocity factor of 0.96.

This table is compiled from information kindly supplied by Aerialite Ltd., and B.I.C.C. Ltd and includes data extracted from Defence Specification, DEF-14-A (H.M.S.O.)

hence may be connected across an open-wire line without affecting the power flow in any way; such a device is called a quarter wave stub. The stub so formed may conveniently be used as a support or termination for an open-wire line as shown in Fig.9.4.
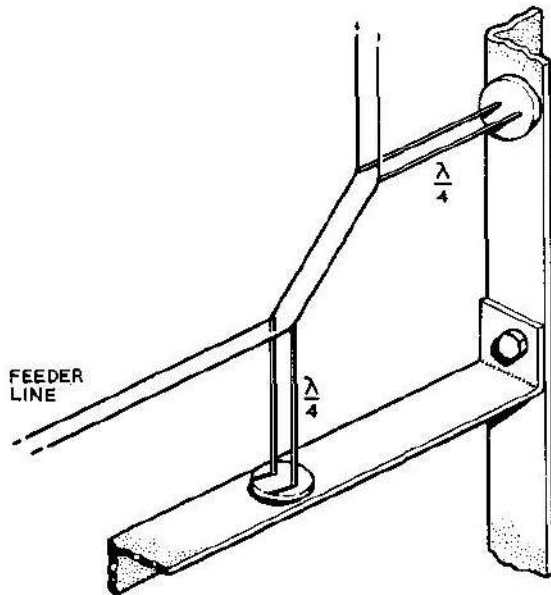
Fig.9.4. Quarter-wave closed stubs are often used for supporting open-wire feeders. When used in this way they are known as "metal insulators". The inductance of the short-circuiting conductors is minimized by making them in the form of large metal plates.

Since the stub must be resonant in order to behave as an insulator it can function only over a narrow range of frequency.

Balance to Unbalance Transformers

In most cases the aerial requires a balanced feed with respect to ground, and therefore it is necessary to use a device which converts the unbalanced output of a coaxial cable to a balanced output as required by the aerial. This device also prevents the wave which has been contained within the cable from tending to "spill over" the extreme end and travel back over the surface of the cable. Whenever this occurs there are two important undesired effects; firstly the re-radiated wave modifies the polar diagram of the attached aerial, and secondly the outer surface of the cable is found to have a radio frequency voltage on it.
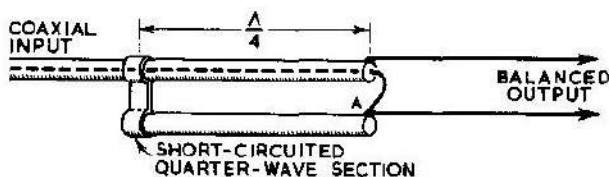
Fig.9.5. Quarter-wave open balun or Pawsey stub.

To prevent this, a balance to unbalance transformer (abbreviated to balun) is connected between the feeder cable and the aerial. The most simple balun consists of a short circuited quarter wave section of transmission line attached to the outer braiding of the cable, as shown in Fig.9.5. This is often known as a Pawsey stub. At the point A, the quarter-wave section presents a very high impedance which prevents the wave from travelling over the surface. The performance of this device is, of course, dependent upon frequency, and its bandwidth may have to be considered in the design.

Fig.9.6. Coaxial sleeve balun.

Several modifications of the simple balun are possible: for example, the single quarter wave element may be replaced by a quarter-wave coaxial sleeve, thus reducing radiation loss, see Fig.9.6. To prevent the ingress of water and to improve the mechanical arrangement, the centre conductor may itself be connected to a short-circuited quarter-wave line acting as a "metallic insulator" as shown in Fig.9.7. The distance d should be kept small, and yet the capacity between the sections should also be kept small since otherwise the quarter-wave section will not be resonant at the desired frequency. A satisfactory
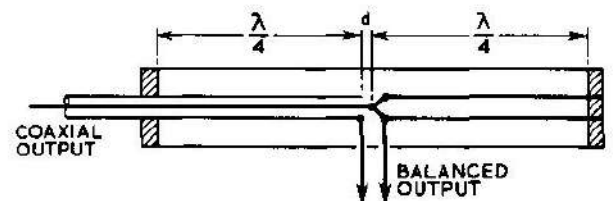
Fig.9.7. Totally enclosed coaxial balun. The right-hand section acts as a "metal insulator."

compromise is to taper the end of the quarter-wave line, although this is by no means essential. In practice, at a frequency of 435 MHz about ⅛" is a suitable spacing. The whole balun is totally enclosed, the output being taken through two insulators mounted in the wall.

A useful variation is that shown in Fig.9.8 which gives a 4:1 step-up of impedance. The half-wave loop is usually made from flexible coaxial cable, and allowance must therefore be made for the velocity factor of the cable when calculating a half-wavelength.

It may be inconvenient at frequencies above about 2000 MHz to mount the coaxial sleeve balun close to a dipole radiator. In this case the sleeve can be mounted a short distance back from the end of the line.
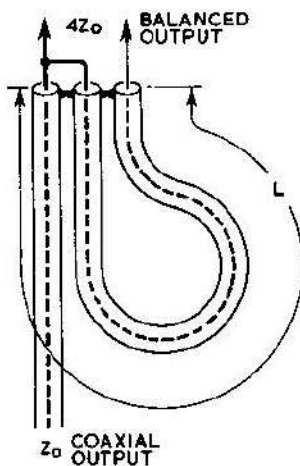
Fig.9.8. A coaxial balun giving a 4:1 impedance step-up. The length L should be $\frac{\lambda}{2}$ allowing for the velocity factor of the cable. The outer braiding may be joined at the points indicated.

The characteristic impedance of the balun element is not critical. A Pawsey stub may be constructed by attaching a piece of coaxial cable one quarter-wave long (the centre conductor being unused) to the braiding of the feeder cable at one end and the inner conductor at the other. The two sections should be spaced sufficiently to ensure an air dielectric between them. If the two pieces lie alongside one another the resonant length will be reduced and an inferior dielectric introduced. It is important to note that since it is the electrical characteristics of the outer surface that are being used, there is no need to allow for the velocity factor of the cable. Coaxial-sleeve baluns should have an outer-to-inner diameter ratio of between 2 : 1 and 4 : 1.

Although the type of balun illustrated in Fig.9.7 has a larger effective bandwidth than the other types described here, they are all suitable for the restricted frequency ranges in common use in the v.h.f. amateur bands.

IMPEDANCE MATCHING

For an aerial feeder to deliver power to the aerial with minimum loss, it is necessary for the load to behave as a pure resistance equal in value to the characteristic impedance of the line. Under these conditions no energy is reflected from the point where the feeder is joined to the aerial, and in consequence no standing waves appear on the line.

When the correct terminating resistance is connected to any feeder, the voltage and current distribution along the line will be uniform. This may be checked by using a device to explore either the magnetic field (H) or the electric field (E) along the line. One such device, suitable for use with a coaxial feeder, is a section of coaxial line having a longitudinal slot cut in the wall parallel to the line. A movable

probe connected to a crystal voltmeter is inserted through the slotted wall. This samples the electric field at any point, and the standing-wave ratio may be determined by moving the probe along the line and noting the maximum and minimum readings. The distance between adjacent maxima or between adjacent minima is one half-wavelength.

The fields surrounding an open line may be explored by means of an r.f. voltmeter, but it is much more difficult to obtain precise readings than with a coaxial line because of hand proximity effects and similar disturbances. Another device which measures forward and reflected waves is the reflectometer.

The term matching is used to describe the procedure of suitably modifying the effective load impedance to make it behave as a resistance and to ensure that this resistance has a value equal to the characteristic impedance of the feeder used. To make a complete load (i.e a load possessing both resistance and reactance) behave as a resistance, it is necessary to introduce across the load a reactance of equal value and opposite sign to that of the load, i.e. the reactance is "tuned out". A very convenient device which can theoretically give reactance values from the minus infinity to plus infinity, (i.e. pure capacitance to pure inductance) is a section of transmission line either of length variable between zero and one half-wavelength having an open-circuited end or alternatively of length a little greater than one half-wavelength having a movable short-circuit capable of being adjusted over a full half-wavelength. The short-circuited stub is to be preferred since it is easier to construct.

Although there is no need to make the characteristic impedance of a stub equal to that of the transmission line, it may be desirable to do so for practical reasons.
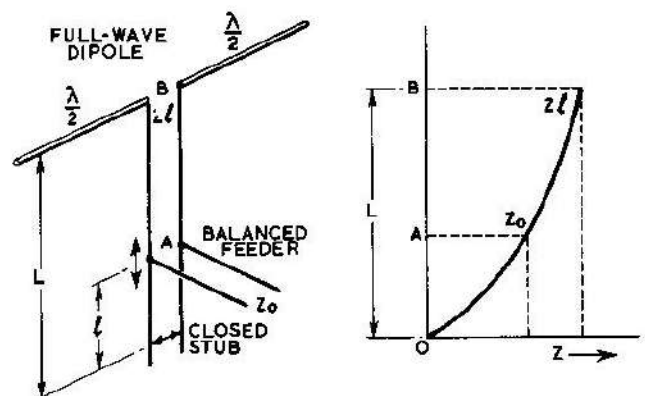
Fig.9.9. Stub matching applied to a full-wave dipole.

In addition to tuning out the reactance, a match still has to be made to the transmission line. The impedance at any point along the length of a quarter-wave resonant stub varies from zero at the short-circuit to a very high impedance at the open end. If a load is connected to the open end and the power is fed into the stub at some point along its length the

stub may be used as an auto-transformer to give various values of impedance according to the position of the feed point. This is shown in Fig.9.9. The distance L is adjusted to tune the aerial to resonance and will be one quarterwave long if the aerial is already resonant. The distance l is adjusted to obtain a match to the line. However, it is usually more convenient to have a stub with an adjustable short-circuit which can slide along the transmission line: see Fig.9.10.
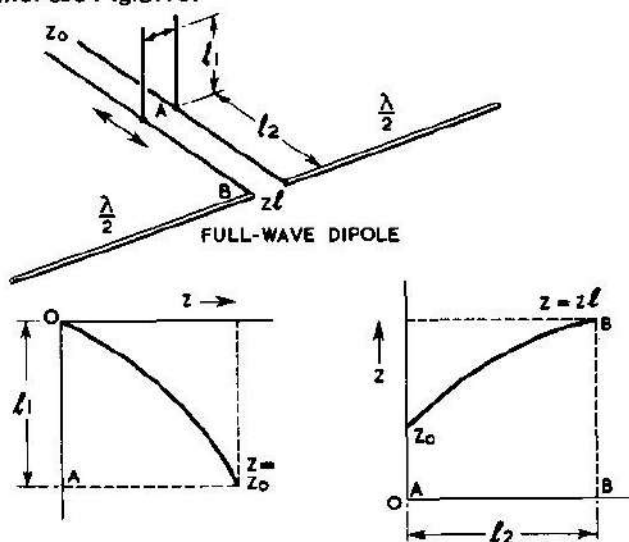
Fig.9.10. Stub matching with a movable short circuited stub.

In practice matching can be achieved entirely by the "cut and try" method of adjusting the stub length and position until no standing waves can be detected. The feeder line is then said to be flat. However, the frequency range over which any single stub matching device is effective is quite small, and where wideband matching is required some other matching system must be used.

Stub Tuners

On a coaxial line it is impracticable to construct a stub having an adjustable position. However, two fixed stubs spaced by a certain fraction of a wavelength can be used for matching purposes: see Fig.9.11. The spacing usually employed is $\frac{\lambda}{8}$ or odd multiples thereof. With this spacing independent adjustment of the short-circuiting plungers gives a matching range from 0.5 times the characteristic impedance of the transmission line (Z_0 upwards. As the spacing is increased towards $\frac{\lambda}{2}$ or decreased towards zero, the matching range increases, but the adjustments then become extremely critical and the bandwidth very narrow. The theoretical limit of matching range cannot be achieved owing to the resistance of the conductors and the dielectric loss; i.e. the Q is limited. To obtain the highest Q the ratio of outer-to-inner conductor diameters should be in the
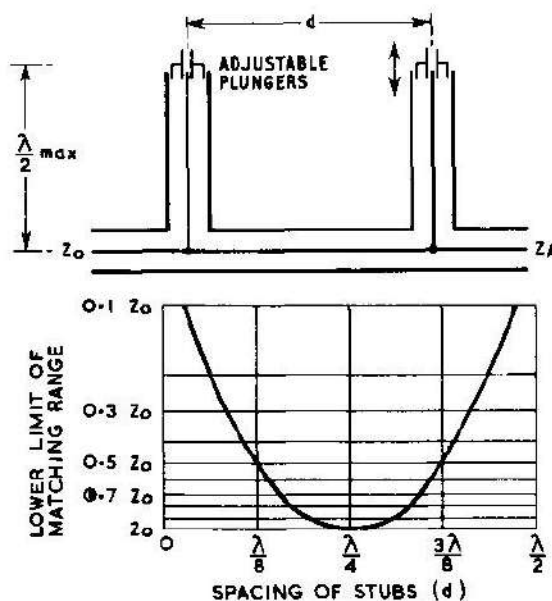
Fig.9.11. Two-stub coaxial tuner. The graph shows the lower limit of the matching range: the upper limit is determined by the Q of the stubs (i.e. it is dependent on the losses in the stubs). Z_0 is the characteristic impedance of the feeder.

range 2 : 1 to 4 : 1 (as for coaxial baluns). An important mechanical detail is the provision of reliable short-circuiting plungers which will have negligible inductance and also ensure low-resistance contact. They can be constructed of short lengths of thin-walled brass tubing, their diameters being chosen so that when they are slotted and sprung they make a smooth sliding contact with both inner and outer conductors.
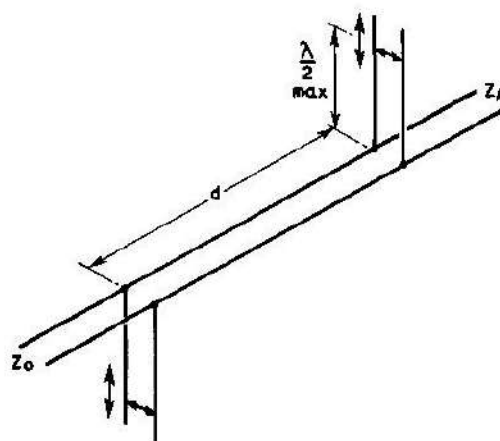
Fig.9.12. Two-stub open-wire tuner. With an open-wire line stubs should be mounted on opposite sides of the line, as shown, so as to avoid mutual coupling. The matching range can be seen from the graph in Fig. 9.11.

The two-stub tuner may be applied to open transmission lines if it is inconvenient to have a movable stub. In this case the stubs must be mounted

laterally opposite to each other to prevent mutual coupling, see Fig.9.12.

This type of tuner may, of course, be used for other purposes than to feed an aerial. For example, it will serve to match an aerial feeder into a receiver, or a transmitter into a dummy load. A greater matching
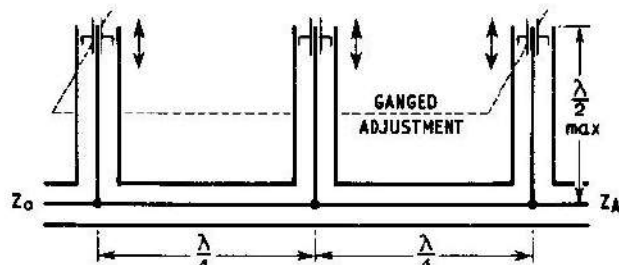
Fig.9.13. Three-stub tuner. This provides a greater matching range than a two-stub tuner. Z_O is the characteristic impedance of the feeder.

range can be obtained by using a three-stub tuner, the stubs being spaced at intervals of one quarter-wavelength apart, as shown in Fig.9.13. The first and third stubs are usually ganged together to avoid the long and tedious matching operation which becomes necessary when adjustments are made to three infinitely variable stubs.

Quarter-wave Lines

An impedance transformation can be effected by using a certain length of transmission line of a different characteristic impedance from the feeder. This may be used to match a load to a transmission line. A special condition occurs when the length of the
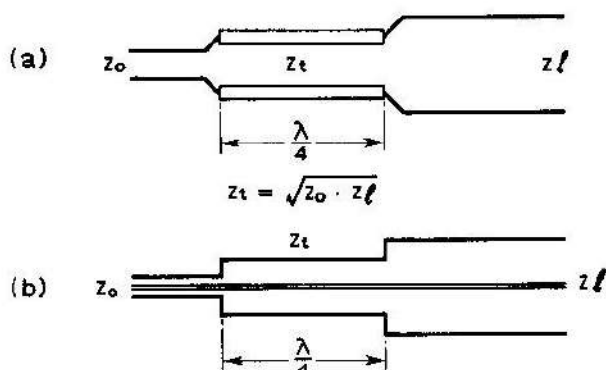
Fig.9.14. Quarter-wave transformers. (a) shows a construction suitable for open-wire lines, and (b) is the corresponding method for coaxial cables. Where a solid–dielectric section is used, due allowance must be made for the velocity factor.

section of line is an odd number of quarter-wavelengths and the following formula then applies—

$$Z_t = \sqrt{Z_o \cdot Z_l}$$

where Z_t is the characteristic impedance of the section of quarter-wave line and Z_o and Z_l are the feeder and

load impedance respectively. For example, if Z_o is 80 ohms and Z_l is 600 ohms—

$$Z_t = \sqrt{80 \times 600} = 251 \text{ ohms}$$

This matching section is useful for transforming impedance and is called a quarter-wave transformer: see Fig.9.14.

A section of tapered line can also be used to effect an impedance transformation, and an application of this principle is described later in this Chapter. Again a quarter wavelength section is only a special case, and to achieve a match in a particular installation the line lengths and the angle of taper should be varied until a perfect match is achieved. This form of matching device is often called a delta match.

TYPES OF AERIALS

The elements of v.h.f.–u.h.f. aerials are usually of halfwave or fullwave dipoles, the actual position and configuration of the elements will depend on the type, gain and directivity required. Each of the types described has its own particular merit and the final choice will have to be made by the amateur to suit his own circumstances.

Parasitic Arrays: The Yagi Array

By placing a reflector, usually a resonant element one half wavelength long behind a halfwave dipole, the radiation can be concentrated within a narrower angle. By adding further elements somewhat shorter than one half-wavelength, called directors, at certain spacings in the forward direction, a further gain can be achieved. Any aerial array which employs elements not directly connected to the feed line, i.e. parasitic elements, is known as a parasitic array. If the arrangement consists of a dipole with a reflector and two or more directors, it is known as a Yagi array: see Fig.9.17.

When compared with other aerial systems of similar size the Yagi array is found to have the highest forward gain, and it can be constructed in a very robust form. The effect of adding the reflector and director(s) is to cause the feed impedance of the dipole to fall considerably, often to a value of about 10 ohms, and the matching is then critical and difficult to obtain. This, however, may be overcome by the use of a folded dipole radiator. If the folded dipole has two elements of equal diameter, a 4 : 1 impedance step-up is obtained. By varying the ratio of the diameters, different impedance step-up ratios become available.

The length of the dipole or folded dipole required for resonance depends not only on the frequency but also to a lesser extent on ratio of the diameter of the element to the wavelength, the length required for resonance diminishing as the wavelength/diameter ratio is decreased.
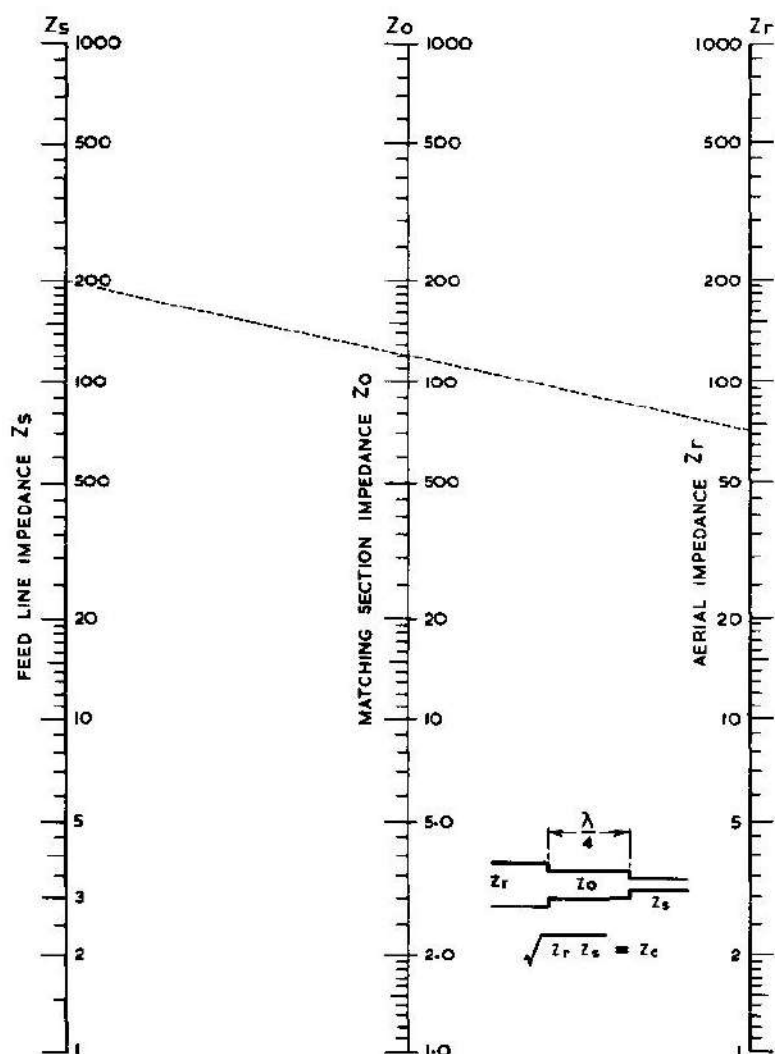
Fig.9.15. Matching section chart. This chart may be used to obtain the surge impedance of a quarter-wave matching section used as an impedance transformer from one real impedance to another. In the example shown, Z_r is 72 ohms and Z_s is 200 ohms, indicating that a quarter-wave matching section of 120 ohms is needed.

The forward gain is not appreciably affected by a variation of reflector spacing over a range of $\frac{\lambda}{8}$ to $\frac{\lambda}{4}$: under these conditions the forward gain is approximately at its maximum value. However, a considerable change in feed impedance takes place when this spacing is varied and this may be used as a convenient form of adjustment. The reflector is usually 0.5λ long although this should be reduced to about 0.475λ for the closer spacing.

The length of the directors is usually made about 0.43λ and the spacing approximately 0.25λ, but experiments have shown that where several directors are used, the bandwidth can be broadened by making them progressively shorter in the direction of radiation. The greater the number of directors, the higher the gain and the narrower the beamwidth. There is no advantage to be derived from using more reflectors spaced behind the first, but the front-to-back ratio may be improved somewhat by the use of additional reflectors as shown in Fig.9.17. These additional reflector elements should subtend a fairly wide angle at the farthest director to be effective. In practice, a trigonal reflector element/element spacing of about ¼λ is sufficient.

The gain obtainable from a Yagi array compared with a halfwave dipole is shown approximately by the empirical curve in Fig.9.18.

Resonant Lengths of Half-wave Dipoles

(Wavelength / Diameter)	(Value of Dipole length / Wavelength for resonance)	Feed Impedance (ohms)
50	0.458	60.5
100	0.465	61.0
200	0.471	61.6
400	0.475	63.6
1,000	0.479	65.3
4,000	0.484	67.2
10,000	0.486	68.1
100,000	0.489	69.2

The dimensions used in calculating the ratios must be in similar units (e.g. both in metres or both in centimetres). From *Aerials for Metre and Decimetre Wavelengths* by R. A. Smith.
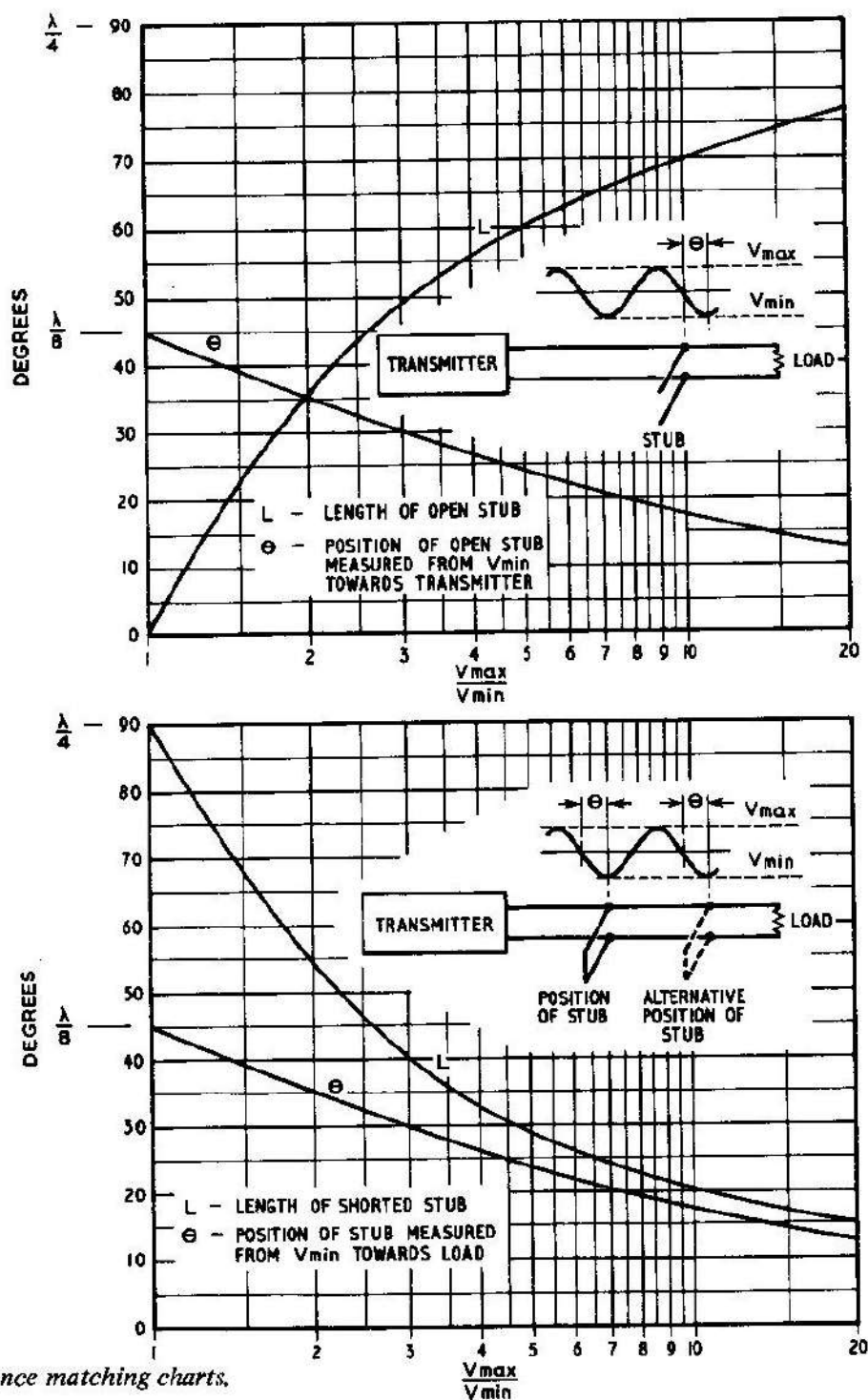
Fig. 9.16. Impedance matching charts.
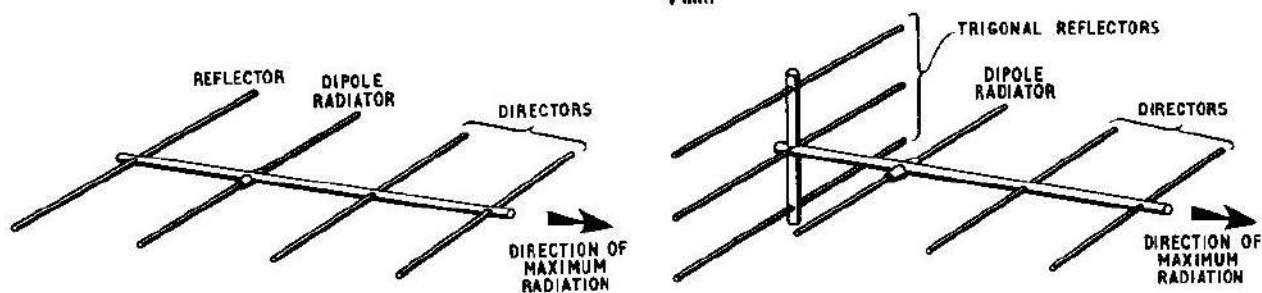
Fig. 9.17. A 4-element Yagi array with a single reflector and with trigonal reflectors. See table on page 9.12 for typical dimensions.

The bandwidth for a standing-wave ratio less than 2 : 1 is about 2 per cent for close-spaced beams and about 3 per cent for wider spacing. Element lengths, particularly those of the directors, are very critical (i.e. within fractions of an inch), and ideally telescopic rods should be used to enable fine adjustments to be made. Each change of element length necessitates a readjustment of the matching either by moving the reflector or in the matching device itself.

Typical element lengths for spot frequencies in the 4m, 2m and 70cm bands are given in the
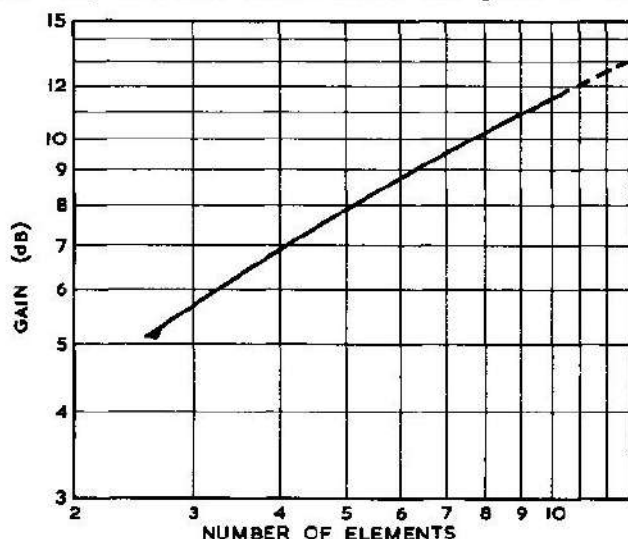
Fig.9.18. Chart showing typical gain plotted against number of elements for Yagi aerials.

accompanying table. The lengths are based on the assumption that the element diameter lies within the stated limits for the respective bands. Any departures from these diameter ranges will necessitate a change in the lengths of the elements; for a larger diameter the length will need to be decreased, and vice versa.

Typical Dimensions of Yagi Arrays—Driven Elements

	Length in Inches		
	70.3 MHz	145 MHz	433 MHz
Dipole (for use with gamma match)	79	38	12¾
Diameter range for length given	½-¾	¼-¾	⅛-¼

Folded dipole, (70 ohm feed)	Length in Inches		
	70.3 MHz	145 MHz	433 MHz
l length centre/centre	77½	38½	12½
d spacing centre/centre	2½	¾	½
Diameter of element	½	¼	⅛
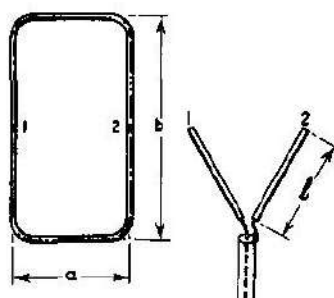
	Length in Inches		
	70.3 MHz	145 MHz	433 MHz
a centre/centre	32	14	5¾
b centre/centre	96	43¾	15 ⅛
Delta feed sections (length 70 ohm feed)	22½	12	4 ¾
Diameter of slot and delta feed material	¼	¾	⅛

Typical Dimensions of Yagi Arrays—Parasitic Elements

Element	Length of Element		
	70.3 MHz	145 MHz	433 MHz
Reflector	85½ in.	40 in.	13¼ in.
Director D1	74 in.	35½ in.	11¼ in.
Director D2	73 in.	35¼ in.	11 ⅛ in.
Director D3	72 in.	35 in.	11 in.
Succeeding directors	1 in. less	½ in. less	¼ in. less
Final director	2 in. less	1 in. less	¾ in. less
One wavelength (for reference)	168 in.	81½ in.	27¼ in.
Diameter range for length given	½-¾ in.	¼-¾ in.	⅛-¼ in.

Spacing Between Elements			
Reflector to radiator	22½ in.	17½ in.	5½ in.
Radiator to director 1	29 in.	17½ in.	5½ in.
Director 1 to director 2	29 in.	17½ in.	7 in.
Director 2 to director 3, etc	29 in.	17½ in.	7 in.

The above figures are based on a number of proved Yagi array designs. If the slot is used two sets of reflectors/directors are required one mounted above the other thus forming a stacked array. Match to the feeder can be effected by moving the radiator relative to the first director and the reflector relative to the radiator.

Stacked Yagis

The optimum stacking distance will depend on the gain of each individual Yagi array. Where the vertical and horizontal beamwidths of an individual array are approximately the same, i.e. with five or more elements, the optimum stacking distance will be approximately the same, both horizontally and vertically; this distance varies from about 0.75λ for an array of 4-element Yagis to about 2.0λ or more for 8-element Yagis. For practical reasons the spacing is usually less than optimum and there is a consequent

reduction in the total gain. Theoretically, the increase in gain obtainable by stacking two identical arrays in such a way that their capture areas do not overlap is simply two times (i.e. 3dB), but in practice this can sometimes be exceeded if there is a suitably favourable degree of coupling between the two arrays: for example, a pair of 3-element Yagi arrays could be made to yield an increase of 4.2dB. However, the increase usually proves to be less than the theoretical 3dB, and a figure of 2.2—2.5dB is all that can be ordinarily expected.

The feed impedance of a stacked array theoretically is the feed impedance of an individual Yagi array divided by the total number of Yagis employed, in practice the feed impedance of stacked Yagis is slightly less than this because of interaction between these Yagi arrays although this reduction is not so marked at the greater spacings.

Disadvantage of Conventional Yagi Arrays

Perhaps the most important disadvantage is that the variation of the element lengths and spacings causes inter-related changes in the feed impedance of a Yagi array. To obtain the maximum possible forward gain experimentally is extremely difficult because for each change of element length it is necessary to readjust the matching either by moving the reflector or by resetting a matching device. However, a method has been devised for overcoming these practical disadvantages by the use of a radiating element in the form of a skeleton slot, this being far less susceptible to the changes in impedance caused by changes in the parasitic-element lengths. This development is due to B.Sykes (G2HCG).
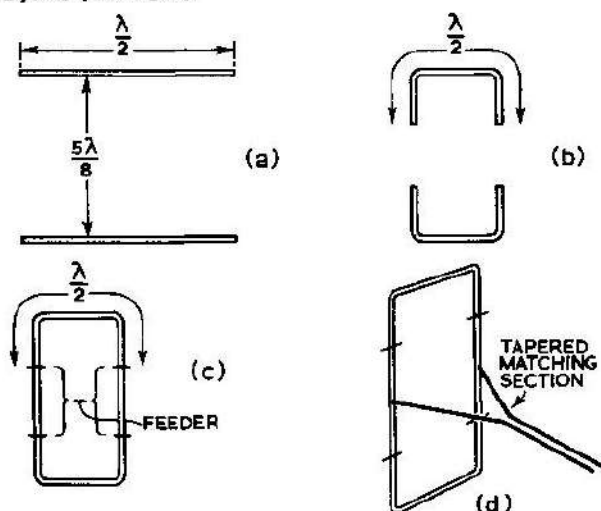
Fig. 9.19. Development of a skeleton slot radiator.

A true slot would be a slot cut in an infinite sheet of metal, and such a slot when approximately one half wavelength long would behave in a similar way to a dipole radiator. In contrast with a dipole, however, the polarization produced by a vertical slot is horizontal (i.e. the electric field is horizontal).

The skeleton slot was developed in the course of experiments to determine to what extent the infinite

sheet of metal could be reduced before the slot aerial lost its radiating property. The limit of the reduction for satisfactory performance was found to occur when there remained approximately one half wavelength of metal beyond the slot edges. However, further experiments showed that a thin rod bent to form a "skeleton slot" of dimensions approximately $\frac{5\lambda}{8} \times \frac{5\lambda}{24}$ exhibited similar properties to those of a true slot. The manner in which a skeleton slot functions can be understood by referring to the diagrams in Fig.9.19. Consider two half-wave dipoles spaced vertically by $\frac{5\lambda}{8}$.

Since the greater part of the radiation from each dipole takes place at the current antinode, i.e. the centre, the ends of the dipoles may be bent without serious effect. These ends may now be joined together with a high impedance feeder, so that end feeding can be applied to the bent dipoles. To radiate in phase, the power should be fed midway between the two dipoles. The high impedance at this point may be transformed down to one suitable for the type of feeder in use by means of a tapered matching section transmission line (i.e. a delta match). Practical dimensions of skeleton slot radiator are given in Fig.9.20.

$$S = \frac{6740}{f} \text{ inches}$$

$$W = \frac{2250}{f} \text{ inches}$$

$$[f = \text{FREQUENCY IN MHz}]$$

$$Z_0 = 276 \log_{10} \frac{W}{D}$$
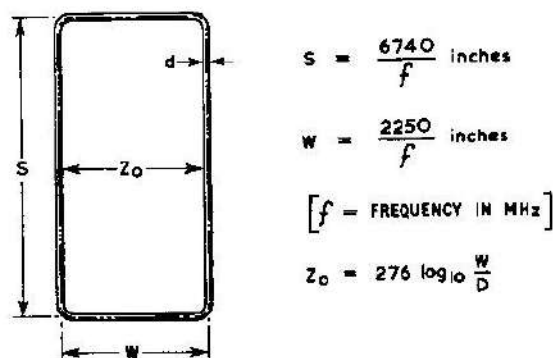
Fig.9.20. Dimensional relationships of a skeleton slot radiator. Both S and W may be varied experimentally from the values indicated by these formulae. For small variations the radiation characteristics of the slot will not change greatly, but the feed impedance will undergo appreciable change and therefore the length of the delta matching section should always be adjusted to give a perfect match to the transmission line.

It is important to note that two sets of parasitic elements are required with a skeleton-slot radiator and not one set as required with a true slot. One further property of the skeleton slot is that its bandwidth is somewhat greater than a pair of stacked dipoles.

Skeleton Slot Yagi Arrays in Stack

Skeleton slot Yagi arrays may be stacked to increase the gain but the same considerations of optimum stacking distance as previously discussed apply; in this case the centre-to-centre spacing of a pair of skeleton slot Yagi arrays should vary between 1λ and 3λ or more according to the number of elements in each Yagi array.

Each skeleton-slot Yagi array may be fed by 72 ohm coaxial cable, using equal lengths of feeder to some common feed point for the stacked array, and it would of course be desirable to use a balun at the point where the cable is attached to each array. A coaxial quarter-wave transformer can be used to transform the impedance to that of the main feeder. For example, if a pair of skeleton-slot Yagi arrays, each of 72 ohm feed impedance, are stacked, the combined impedance will be one-half of 72 ohms, i.e. 36 ohms; this may be transformed to 72 ohms by the use of a quarter-wave section of 52 ohm coaxial cable, allowance being made for its velocity factor. Larger assemblies of skeleton-slot Yagi arrays can be fed in a similar manner by joining pairs and introducing quarter-wave transformers until only one feed is needed for the whole array.

Stacked Dipole Arrays

Both horizontal and vertical beam widths can be reduced and gain increased by building up arrays of driven dipoles. This arrangement is usually referred to merely as a stack or sometimes as a bill-board or broadside array. Since this type of array is constituted from a number of radiating dipoles, the feed impedance would be extremely low if the dipoles were centre-fed. However, the impedance to earth of a dipole at its end is high, the precise value depending upon the ratio of its length to diameter, and it will therefore be more convenient to use a balanced high impedance feeder to end-feed a pair of collinear half-wave dipoles, a system called a full-wave dipole. The length for resonance and the feed impedance in terms of wavelength/diameter ratio is shown in the table.

Resonant Lengths of Full-wave Dipoles

$\left(\dfrac{\text{Wavelength}}{\text{Diameter}}\right)$	$\left(\dfrac{\text{Value of Dipole length}}{\text{Wavelength for resonance}}\right)$	Feed Impedance (ohms)
50	0.85	500
100	0.87	900
150	0.88	1100
200	0.896	1300
300	0.906	1500
400	0.916	1700
700	0.926	2000
1,000	0.937	2400
2,000	0.945	3000
4,000	0.951	3600
10,000	0.958	4600

The dimensions used in calculating the ratios must be in similar units (e.g. both in metres or both in centimetres).
From *Aerials for Metre and Decimetre Wavelengths* by R. A. Smith.

The full-wave dipoles are usually mounted with a centre-to-centre spacing, horizontally and vertically, of one half-wavelength and are fed in phase. Typical arrangements for stacks of fullwave dipoles are shown in Fig.9.21. Note that the feed wires between dipoles are one half-wavelength long and are crossed so that all the dipoles in each bay are fed in phase. The impedance of these phasing sections is unimportant provided that the separators, if used, are made of low-loss dielectric material and that there is sufficiently wide separation at the cross-over points to prevent unintentional contact.

To obtain the radiation pattern expected, all dipoles should be fed with equal amounts of power (as indeed would be desirable in any multi-radiator array), but this cannot be achieved in practice because the dipoles which are farthest from the feeder have a greater feeder loss than the nearest. However, by locating the main feed point as nearly symmetrically as possible these effects are minimized. Hence it would be preferable for the aerial shown in (a) in Fig.9.21 to be fed in the centre of each bay of dipoles; the feeder to each bay must be connected as shown to ensure that the two bays are fed in phase. If they were fed 180° out-of-phase the resultant beam pattern would have two major side lobes and there would be very little power radiated in the desired direction. The diagram (b) in Fig.9.21 shows two vertically stacked bays of full-wave dipoles fed symmetrically and in phase.
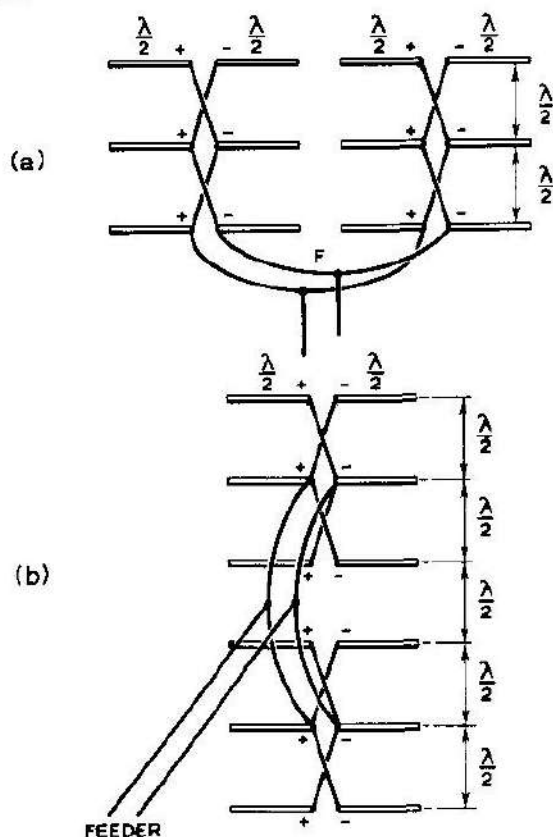
Fig.9.21. Typical stacks of full-wave dipoles. Note that the feedpoint F is equidistant from each bay of dipoles. For example of element lengths, see the table.

The spacing at the centre of each fullwave dipole should be sufficient to prevent a reduction of the resonant frequency by the capacity between the ends. In practice this spacing is usually about 1" for the 144 and 432 MHz bands.

Matching can be carried out by the use of movable short circuited open-wire stubs on the feed lines. The practical aspects of this operation are dealt with in a typical case later in this chapter.

As with the Yagi array, the gain can be increased by placing a reflector behind the radiating elements at a spacing of 0.1 — 0.25λ, a figure of 0.125λ being frequently chosen. For the 432 MHz band and for higher frequencies, a plane reflector made up of 1" mesh wire netting stretched on a frame can be used in place of the resonant reflector at a similar spacing. The mesh of the wire should be so orientated that the
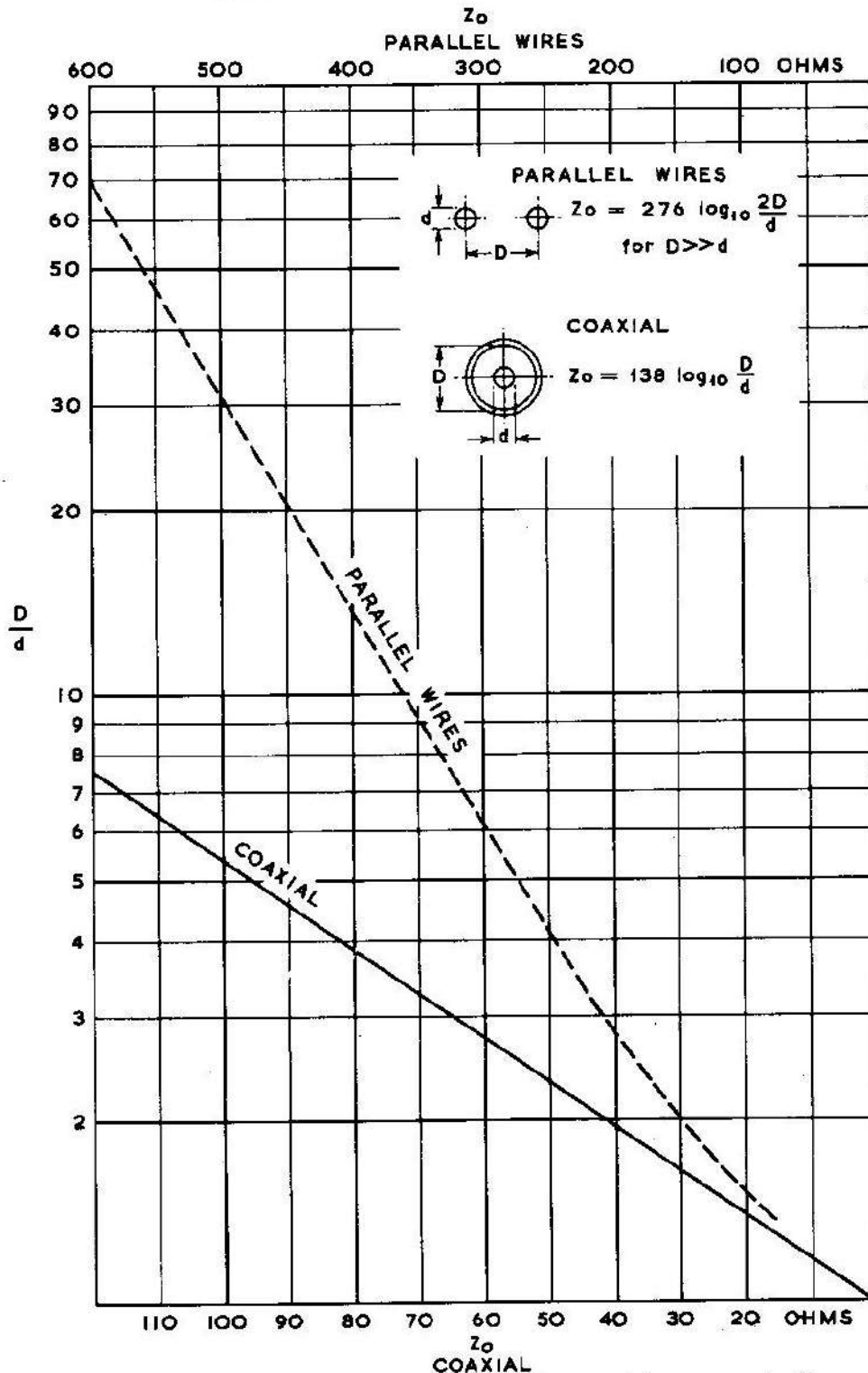
Fig. 9.22. Chart showing feeder line impedances for coaxial or open wire lines.

interlocking twists are parallel to the dipole. The wire netting should extend at least one half-wavelength beyond the extremities of the dipoles in order to ensure a high front-to-back ratio.

The half-wave sections of the full-wave dipole should be supported at the current antinodes, i.e. at their centres, either on small insulators or in suitably drilled wooden vertical members. Supports should not be mounted parallel to the elements because of possible influence on the properties of the aerial.

The bandwidth of this type of aerial is exceptionally large and its adjustments generally are far less critical than those of Yagi arrays.

For a stack having an adequate wire-net reflector, the horizontal beamwidth θ, vertical beamwidth ϕ and power gain G (compared with an isotropic radiator) can be calculated approximately from the following formulae.

$$\theta = \frac{51\lambda}{a} \qquad \phi = \frac{51\lambda}{b} \qquad G = \frac{4\pi ab}{\lambda^2}$$

where a and b are the horizontal and vertical dimensions of the reflector respectively, both being expressed in the same units as the wavelength.

These formulae are true only for an array which is large compared with the wavelength, but are suitable as a criterion for judging aerials of any type provided the equivalent aperture or capture area is known.

Skeleton Slots in Stack

Skeleton slots can be used to replace vertically disposed pairs of half-wave dipoles. As the optimum vertical dimension for a horizontally polarized skeleton slot is approximately ⅝ λ, it is no longer possible to use the vertical spacings shown for full-wave stacks. The slots are mounted vertically at a centre-to-centre spacing of one wavelength and fed through a tapered matching section, as for the skeleton-slot Yagi array, and are then connected to the phasing lines. Since the spacing between feed points is one wavelength there is no phase difference and it is unnecessary to transpose the phasing wires. The tapered matching sections should be adjusted to present an impedance of N times the desired feeder-cable impedance - where N is the number of skeleton slots employed. The impedance resulting from the connection of all the feed points will then equal the cable impedance.

A broadside array of skeleton slots may be built up by adding further slots horizontally at a centre-to-centre spacing of one half-wavelength.

Disadvantage of Multi-element Arrays

As the frequency becomes higher and the wavelength becomes shorter, it is possible to construct arrays of much higher gain although, as already described, the advantage is offset by the reduction in capture area. However, if the practice already described, namely that of using many driven or parasitic elements, either in line or in stack, is adopted, the complications of feeding become increasingly greater. Also as the frequency increases, the radiation-loss from open-wire lines and from phasing and matching sections likewise increases, and it is then difficult to ensure an equal powerfeed to a number of radiators. Preferably, therefore, the aerial should have a minimum number of radiating or other critical elements, such as resonant reflectors or directors. There are many aerials in this category but only those having immediate amateur application are described here.
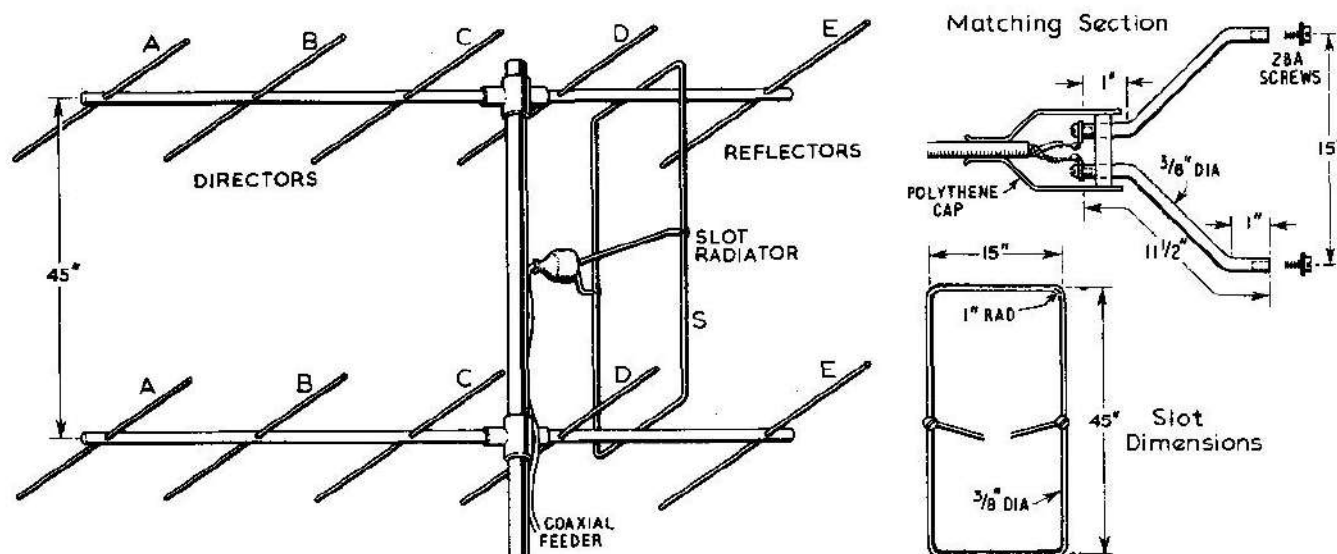
Fig.9.25. A 6 over 6 skeleton slot aerial.

Typical Yagi aerials for 144 MHz

Figures 9.23 and 9.24 illustrate the various standard forms using either a folded dipole, a split dipole or a slot radiator element. One of the standard types of balun will be necessary when the split or folded dipole is fed with coaxial cable.

All metal construction has an obvious advantage of being easier to make and easier to mount on to a mast or pole.

A 12-element Stack for the 70cm Band

Figure 9.26 illustrates the assembly of a twelve element stack arranged in two horizontal bays of three full wave dipoles having a gain of 13dB. With the dimensions indicated the centre frequency is 432 MHz. The radiating elements are constructed from thin-walled ¾" diameter brass tubing and the feeder and phasing wires are soldered on. The centres are supported on ½" polythene insulators mounted on
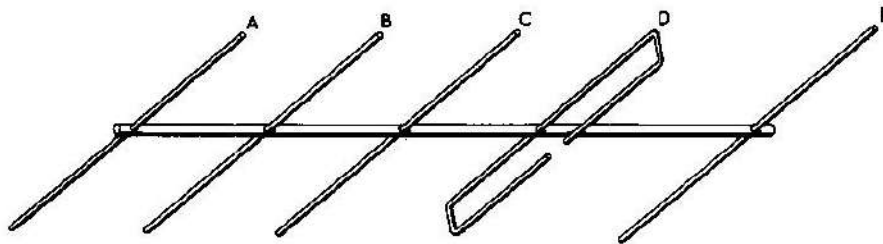
Fig.9.23. A typical 5-element Yagi aerial with folded dipole radiator for 144 MHz.

Element lengths		Spacing		Material
A	35"	A-B	12¼"	Elements ¼" aluminium
B	35½"	B-C	12¼"	Folded dipole: diameter
C	36"	C-D	12¼"	and spacing of upper and
D	38½"	D-E	20½"	lower section to suit
F	40"			impedance required
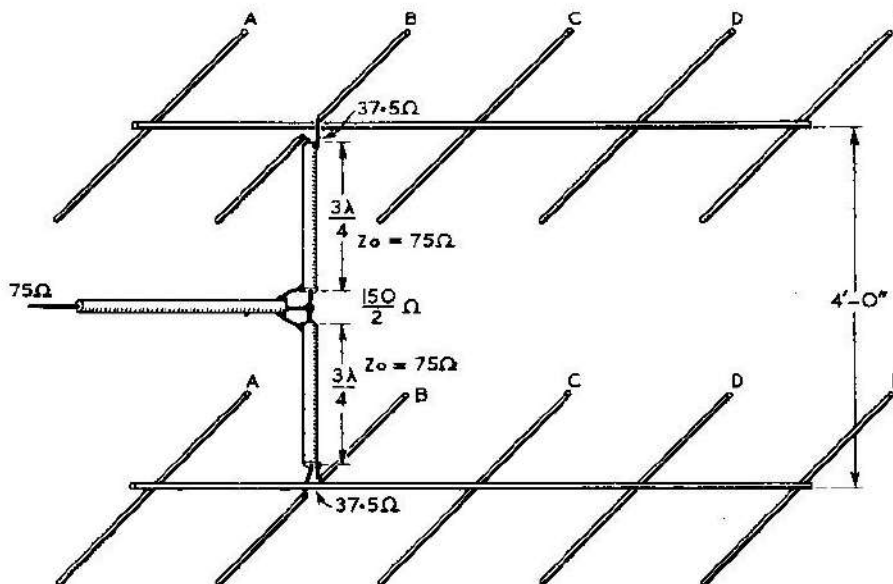
Fig.9.24. A 5 over 5 element Yagi showing matching arrangements.

Element lengths	Spacing
A 40½"	A-B 15"
B 19½" + 19½"	B-C 20"
C 36½"	C-D 20"
D 36"	D-E 20"
E 36"	

ELEMENT TO REFLECTOR
SPACING ____ 4½"

Two-stub tuner   Balun
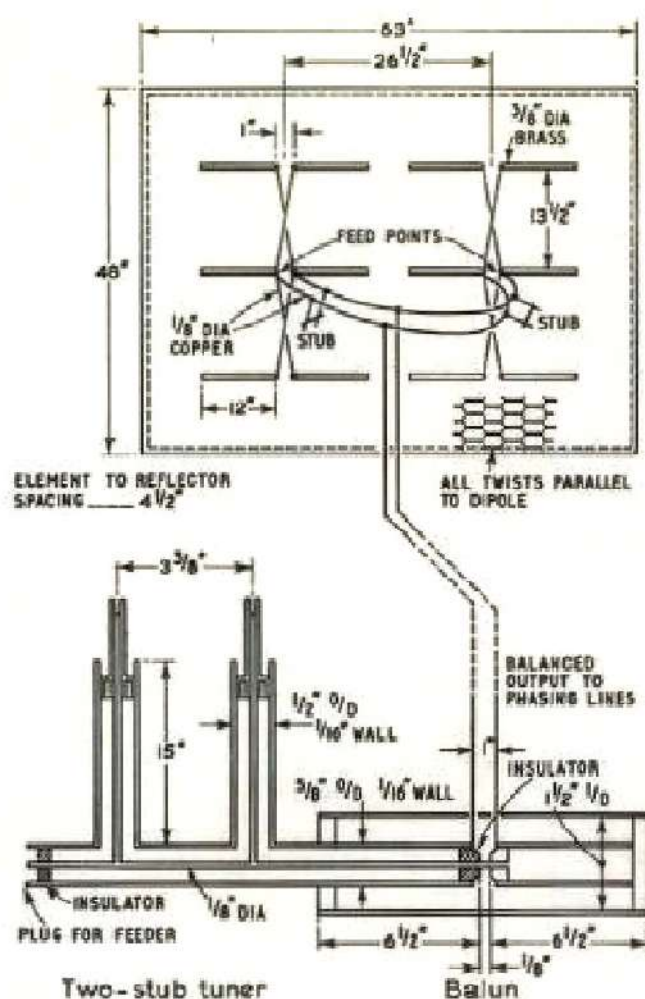
Fig. 9.26. A 12 element stack for 432 MHz, complete with two stub tuner and enclosed coaxial balun.

The remaining bay is then matched approximately to the open-wire feed line. This may be determined by using a very low wattage bulb (for example, 6 volt, 0.06A), the screwed body of which is held in the hand. The presence of standing waves is then examined by observing the glow in the bulb as it is slid along the feed line with its centre connection in contact with one of the feedline conductors. The procedure is then reversed, the resistor being placed so as to represent the bay that has been matched. Both bays are then reconnected and the final matching carried out by means of the two stub tuner using a slotted line or other matching device. If no special apparatus is available the array should be tuned for maximum gain as already described.

vertical 1" x 1" wooden members. The reflecting screen is made from 1" mesh galvanised wire netting mounted on a framework of aluminium-alloy angle. Each bay of three full wave dipoles is fed at the centre; the feed line taken straight through the reflector, and each of these feed lines is matched to the 300 ohm open-wire feed line by a movable short-circuited stub. The two 300 ohm feeders, one from each bay, are joined together to the output of a totally enclosed coaxial balun and a two-stub tuner. The aerial is fed with 72 ohm coaxial cable.

To match the aerial to the feeder, first one bay of the beam is disconnected and a 300 ohm 1 watt carbon resistor is substituted. The stub on this side should be set to exactly one quarter-wavelength long.

Fig. 9.27. A high gain 70cm aerial consisting of four 8 over 8 fed Yagi units arranged in a square formation.

HORIZONTAL POLAR DIAGRAM

THREE ELEMENT YAGI

HORIZONTAL POLAR DIAGRAM
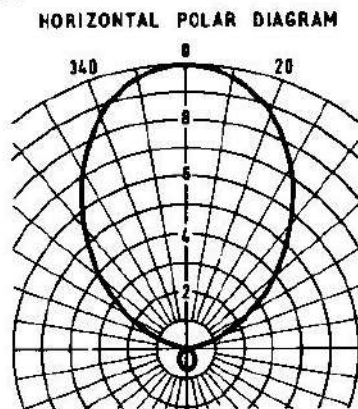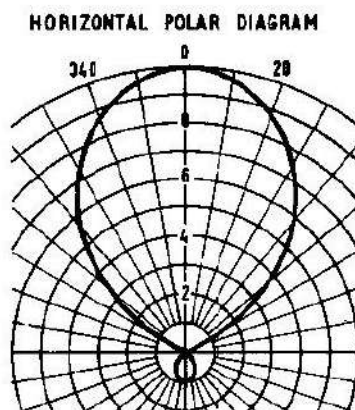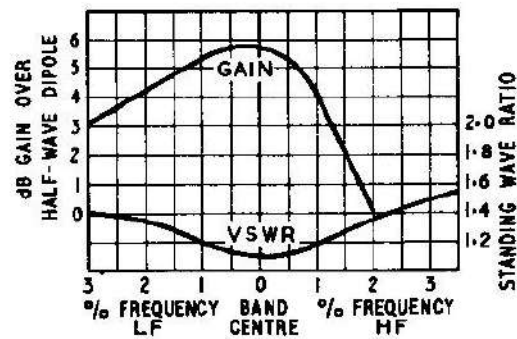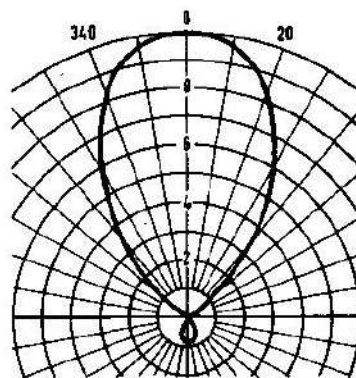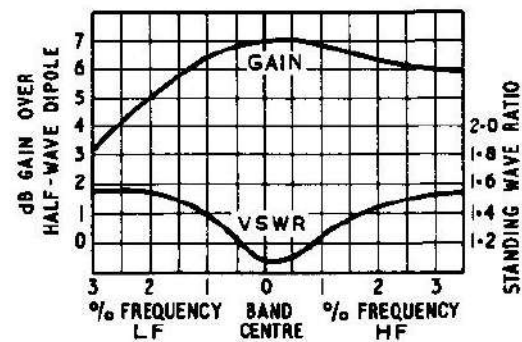
FOUR ELEMENT YAGI

FIVE ELEMENT YAGI
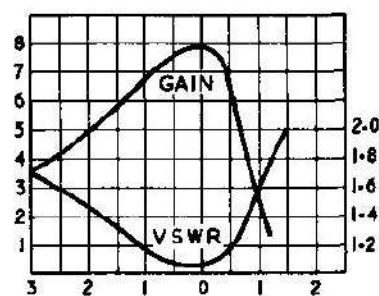
Fig.9.28. Charts showing polar diagrams and gain against v.s.w.r. of 3, 4 and 5 element Yagi aerials.
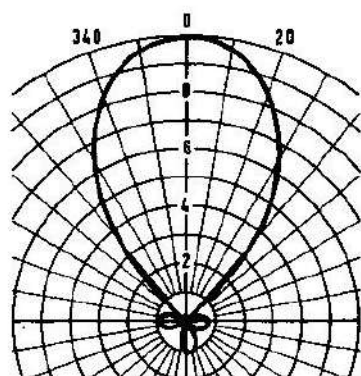
SIX ELEMENT YAGI
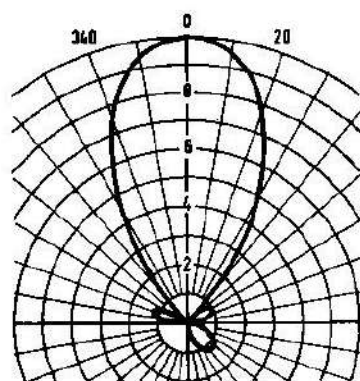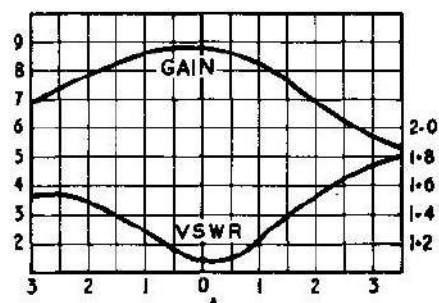
EIGHT ELEMENT YAGI
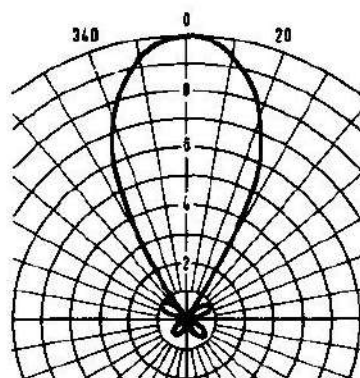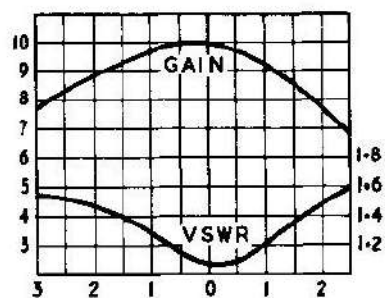
TEN ELEMENT YAGI
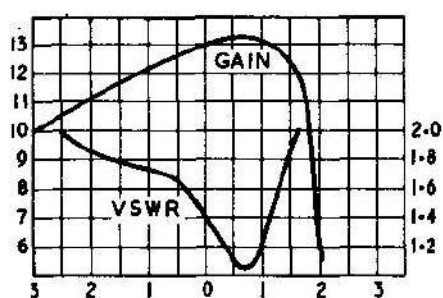
Fig.9.29. Charts showing polar diagrams and gain against v.s.w.r. of 6, 8 and 10 element Yagi aerials.

HORIZONTAL POLAR DIAGRAM

'2 OVER 2'
SLOT FED BEAM

HORIZONTAL POLAR DIAGRAM

'4 OVER 4'
SLOT FED BEAM
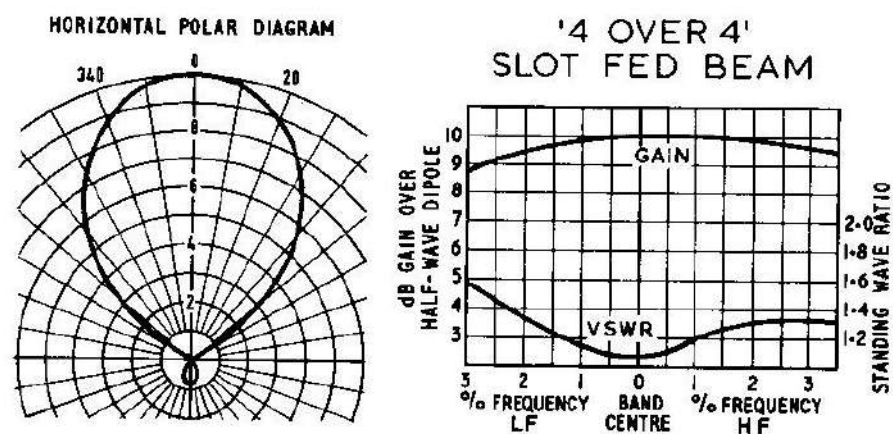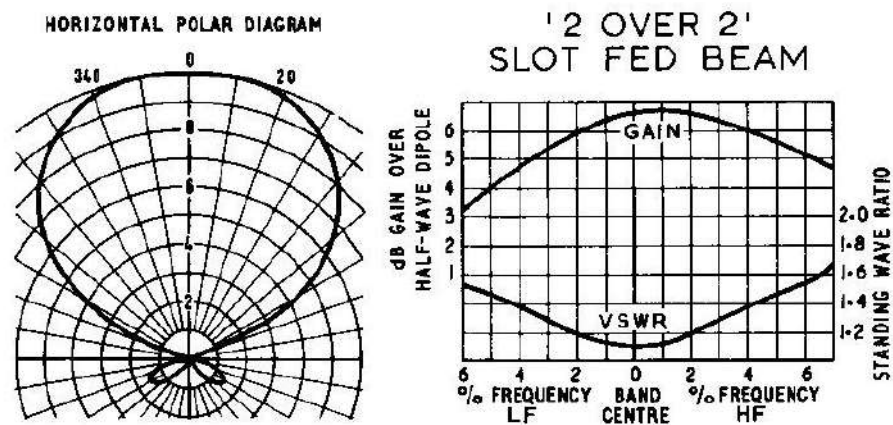
Fig. 9.30. Charts showing polar diagram and gain against v.s.w.r. of a 2 over 2 and 4 over 4 skeleton slot aerials.

'6 OVER 6'
SLOT FED BEAM

'8 OVER 8'
SLOT FED BEAM
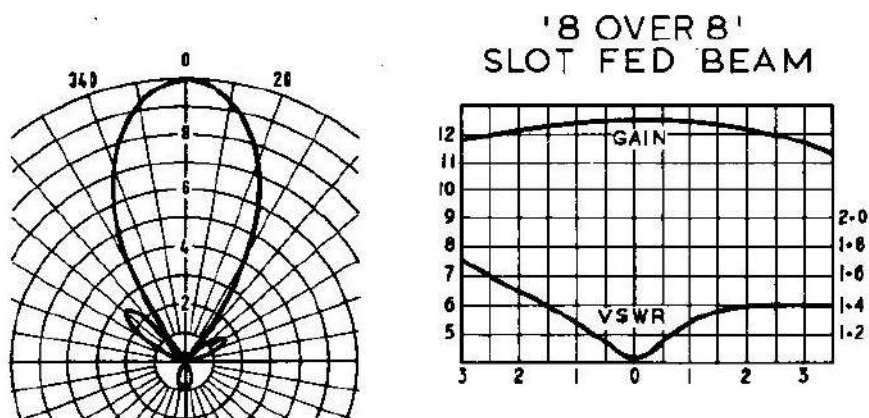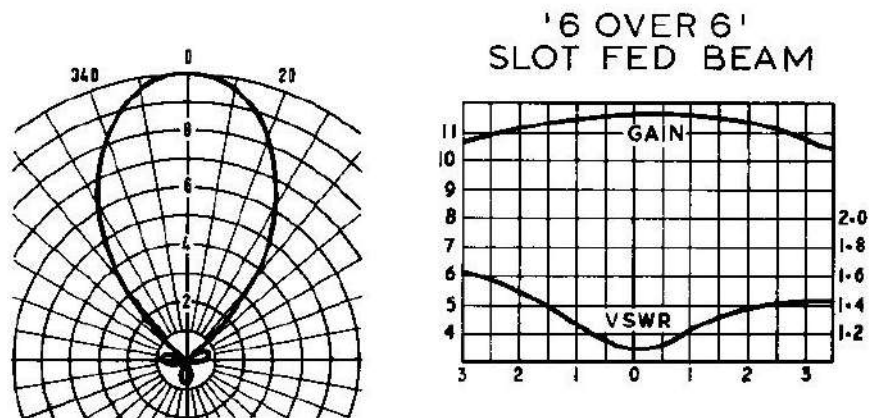
Fig. 9.31. Charts showing polar diagram and gain against v.s.w.r. of a 6 over 6 and 8 over 8 skeleton slot aerials.

THE QUAD

The quad aerial is particularly useful for locations where full size Yagi types might be objected to by neighbours.
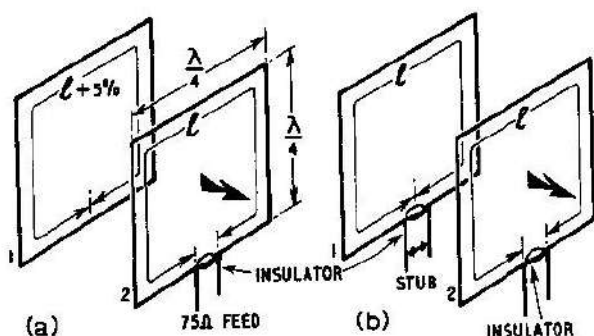
Fig. 9.32. Drawing showing quad aerial dimensions.

In spite of its small physical size a forward gain of 5½ to 6dB can be obtained and its front to back ratio is significant. There is no reason why directors should not be added to the basic two element array in the same manner as applied to the normal Yagi.

Typical dimensions for both 70 MHz and 144 MHz are given in the table. The actual spacing should initially be made adjustable and for a 144 MHz aerial the spacing between the radiator and reflector will be between seven and nine inches for a 72 ohm cable. Ideally a balun should be used to connect the cable to the radiator but this is not essential if the feeder is short and of low loss.

Band	Reflector 1 total length	Radiator 2 total length	Director (if used)	Approx. length of stubs if used reflector s/c	director s/c
70 MHz (a)	173 in.	165 in.	157 in.	-	-
70 MHz (b)	166 in.	165 in.	165 in.	8 in.	8 in.
144 MHz (a)	84 in.	80 in.	76 in.	-	-
144 MHz (b)	80 in.	80 in.	80 in.	4 in.	4 in.

The elements may be made of ⅛" to ¼" diameter preferably of solid aluminium so that the corners can be bent easily. If the vertical dimension of both elements are made the same, then two short cross pieces can be used to mount the aerial to the mast. These cross pieces may be metal so that the whole aerial with the exception of the feed point and the reflector stub (if used) can be very solidly built and will withstand high winds without damage. As with the other aerials quads may be stacked or built into a four square assembly.

A Four Square Quad Aerial for 144 MHz

Data for One Quad Aerial

The dimensions of the quad are shown in Fig. 9.35. Taking the "squared up" figure of 20¾" for each side of the driven element it will be seen that this approximates to 0.255 of a wavelength in free space at 144 MHz. The spacing between the two elements on a
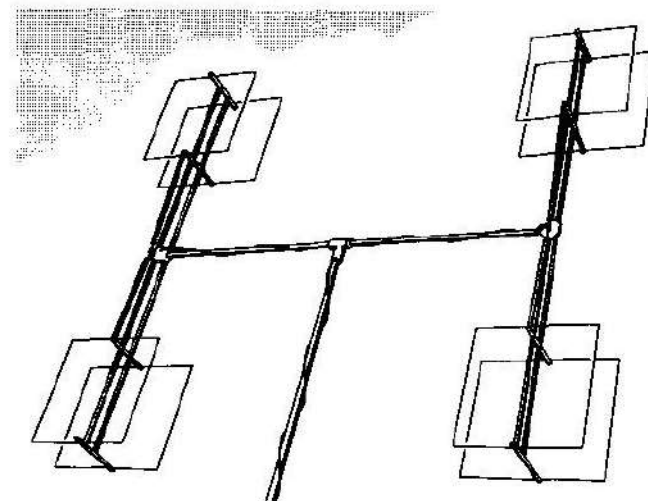
Fig. 9.33. Drawing showing a 144 MHz cubical quad array.

single unit only (see below) to provide a correct match into the 72 ohm feeder was found to be 7", which approximates to 0.08 of a wavelength. The spacing between the elements had a markedly critical effect on
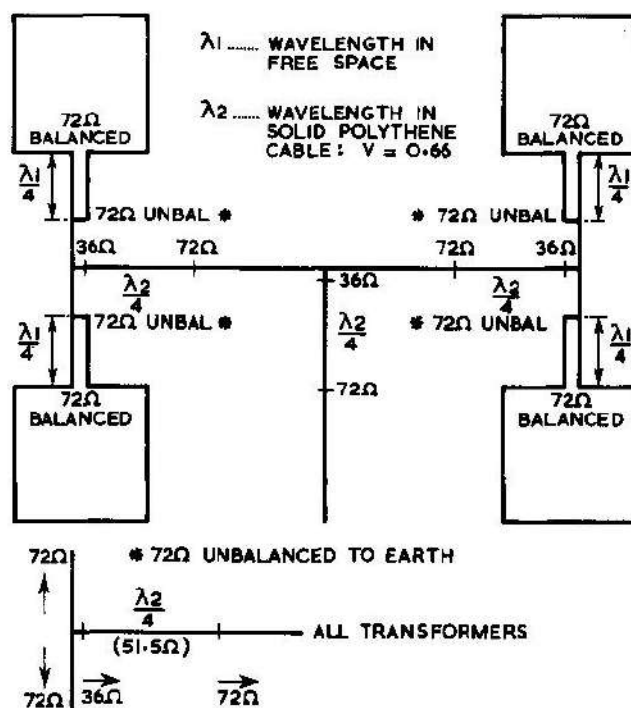
Fig. 9.34. The complete electrical system of the cubical quad array described.

the s.w.r. although the forward gain of the aerial was not affected in so drastic a manner.

The performance of the aerial was shown to have a forward gain of 5dB over a halfwave dipole and a back-to-front ratio of better than 20dB.

Two Quads

Two quad aerials mounted one above the other at a centre-to-centre spacing of $\frac{5\lambda}{8}$ (=65") and paralleled through a single quarterwave 51.5 ohm transformer. Performance figures of this type of aerial proved to be a forward gain of 8.2dB over a halfwave dipole with a back-to-front ratio of better than 10dB at 144 MHz. To obtain the 72 ohm input impedance for the double system it was found necessary to open out the spacing of the reflectors from the driven element on both quads to 9". This is necessitated by the fall in input impedance on a single quad due to the mutual coupling of the pair.
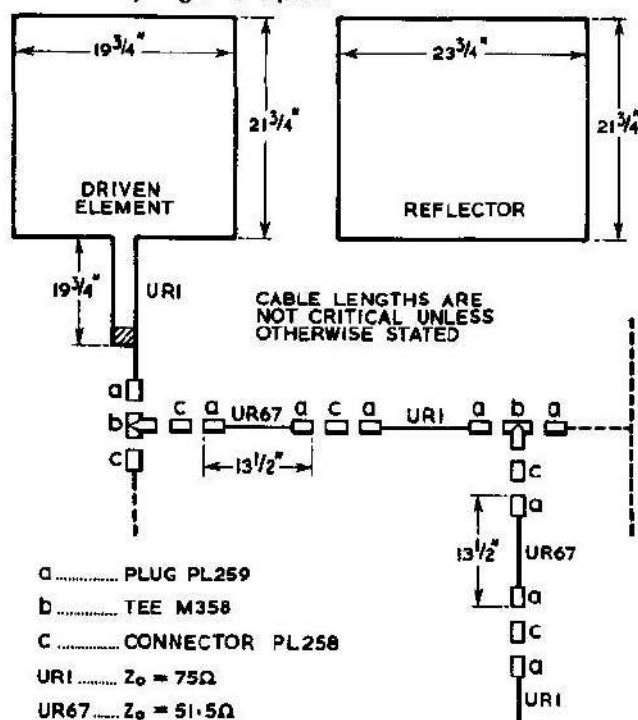
Fig. 9.35. Physical arrangement of the aerial system shown in Fig. 9.34. with details of matching sections of coax.

Four Quads

The final form of the aerial is shown in Fig.9.34 and 9.35. The layout was adopted because of the ease with which the feeder cable could be run and also to avoid a long length of unguyed mast. The spacing between the elements is 9" as in the case of the double quad while the vertical spacing between quads is 65". In the absence of the torsional loading problem, the horizontal spacing was set at 81½" centre-to-centre, i.e . a full wavelength in free space.
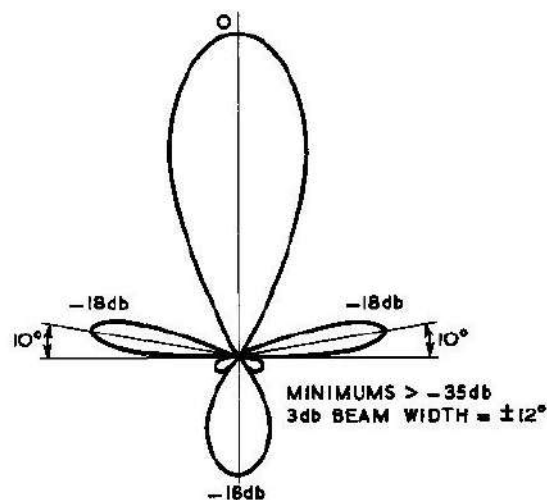
Fig.9.36. Horizontal polar diagram of the complete array.

The gain of this arrangement is 13.5dB over a halfwave dipole and a back to front ratio of 18dB. The polar diagram is shown in Fig.9.36.

The Corner Reflector

The use of an aperiodic plane reflector spaced behind a radiating dipole has already been discussed. If this reflector is bent to form a V, as shown in Fig.9.37 a considerably higher gain is achieved. The critical factors in the design of such an aerial array are the corner angle α and the dipole/vertex spacing S. The curves in Fig.9.38 show that as α is reduced, the gain theoretically obtainable becomes progressively greater. However, at the same time the feed impedance of the dipole radiator falls to a very low value, as can be seen from Fig.9.39. This makes matching difficult and hence a compromise has to be reached. In practice the angle α is usually made 90° or 60°; adjustments in a 60° corner are a little more critical although the maximum obtainable gain is higher. The final
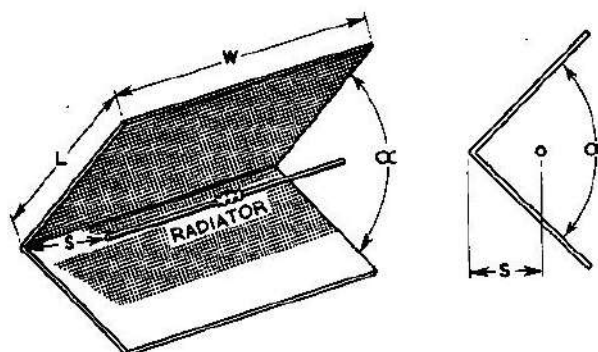
Fig.9.37. Corner reflector. The half-wave dipole radiator is spaced parallel with the vertex of the reflector at a distance S, its characteristics are shown in Fig. 9.38.
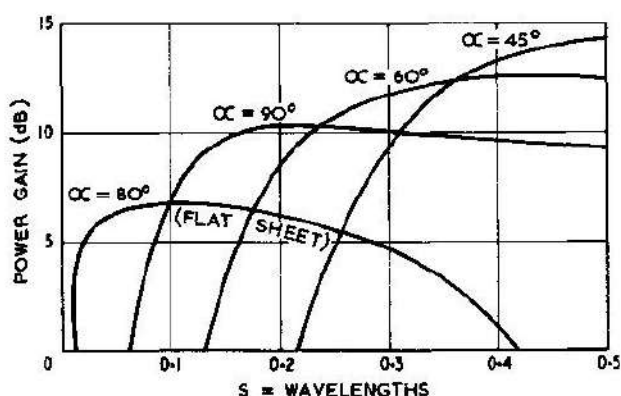
Fig. 9.38. Theoretical power gain obtained by using a corner reflector with a half-wave dipole radiator: see Fig. 9.37.

matching of the radiator to the line may be carried out by adjusting the distance S, which as seen from Fig. 9.38 does not greatly affect the gain over a useful range of variation but causes a considerable change in radiation resistance (see Fig. 9.39). A two-stub tuner may also prove helpful in making final adjustments.
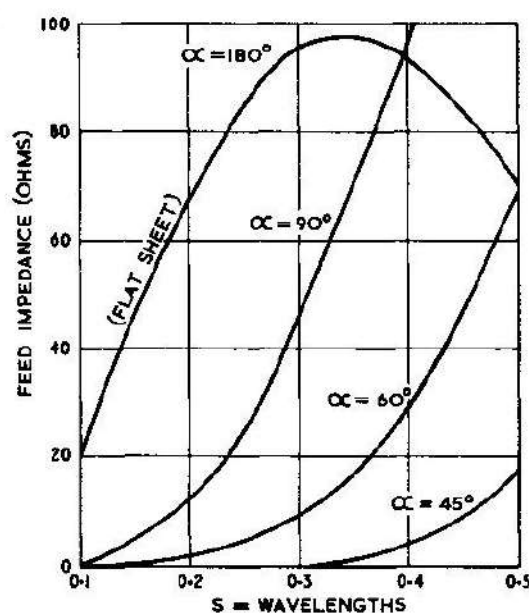
Fig. 9.39. Feed impedance of a half-wave dipole provided with a corner reflector: see Fig. 9.37.

The length of the sides L of the reflector should exceed two wavelengths to secure the characteristics indicated by Fig. 9.38 and 9.39, and the reflector width W should be greater than one wavelength for a halfwave dipole radiator. The reflecting sheet may be constructed of wire-netting as described previously or alternatively may be fabricated from metal spines arranged in a V-formation, all of them being parallel to the radiator: see Fig. 9.40. The spacing between adjacent rods should not exceed 0.1λ. See table below, Fig. 9.41.
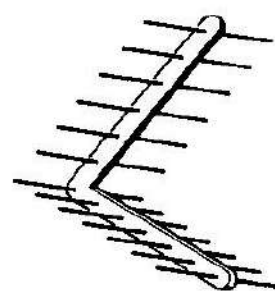
Fig. 9.40. The corner reflector can be modified by using a set of metal spines arranged in V-formation to replace the sheet metal or wire-netting reflector.

A useful approximation for the power gain G referred to a halfwave dipole is G = 300/α, where α is the angle between the sides measured in degrees.

The maximum dipole/vertex spacing S included in the curves shown in Fig. 9.38 and 9.39 is one half-wavelength. Spacings greater than this would require rather cumbersome constructions at lower frequencies, but at the higher frequencies larger spacings become practicable, and higher gains than would be suggested by Fig. 9.38 can then be obtained; see table below. This indicates that the corner reflector can become a specially attractive proposition for the 1296 MHz band, but the width across the opening should be in excess of 4λ to achieve the results shown.
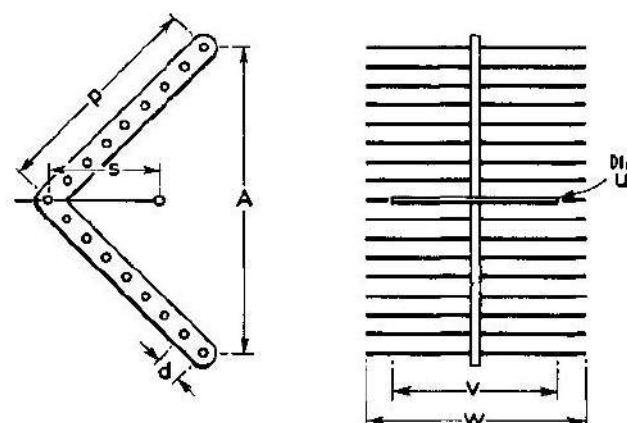
Fig. 9.41. Dimensions for a 60° corner reflector aerial system giving a gain of about 13dB. The feed impedance of the dipole radiator is 75 ohms. The apex may be hinged for portable work.

Dimensions in Inches

Band	p	s	d	v	w	A	u	λ
144	100	40	6	38	50	100	¾	168
433	35	13¼	1½	12¾	20	35	¼	27¼
1296	12	4½	½	4	8	12	⅛	9⅛

Trough Reflector

To reduce the overall dimensions of a large corner reflector the vertex can be cut off and replaced

with a plane reflector, an arrangement known as a trough reflector: see Fig.9.42. Similar performance to that of the large corner reflector can thereby be achieved provided that the dimensions of the trough do not exceed the limits indicated in the table.

CORNER/TROUGH REFLECTOR

Angle α	Value of S for maximum gain	Gain (dB)	Value of T
90°	1.5λ	13	1λ-1.25λ
60°	1.25λ	15	1.0λ
45°	2.0λ	17	1.9λ

This table shows the gain obtainable for greater values of S than those covered by Fig.9.38 assuming that the reflector is of adequate size. T is the distance between the plane reflector and the theoretical apex of the vertex.

The resulting aerial has a performance very little different from the corner reflector type and presents fewer mechanical problems since the plane centre portion is relatively easy to mount on the mast and the sides are considerably shorter.

The gain of both corner reflectors and trough reflectors may be increased still further by stacking two or more and arranging them to radiate in phase, or alternatively by adding further collinear dipoles within a wider reflector similarly fed in phase. Not more than two or three radiating units should be used since the great virtue of the simple feeder arrangement would then be lost.

THE PARABOLIC REFLECTOR

The principle of the parabolic reflector as used for light waves is well known. Such a reflector may be used for radio waves, although the properties are somewhat different because the ratio of the diameter of the reflector to the wavelength is not extremely large, as it is in the case of light waves. If a radiator is placed at the focus of a paraboloid and if most of the
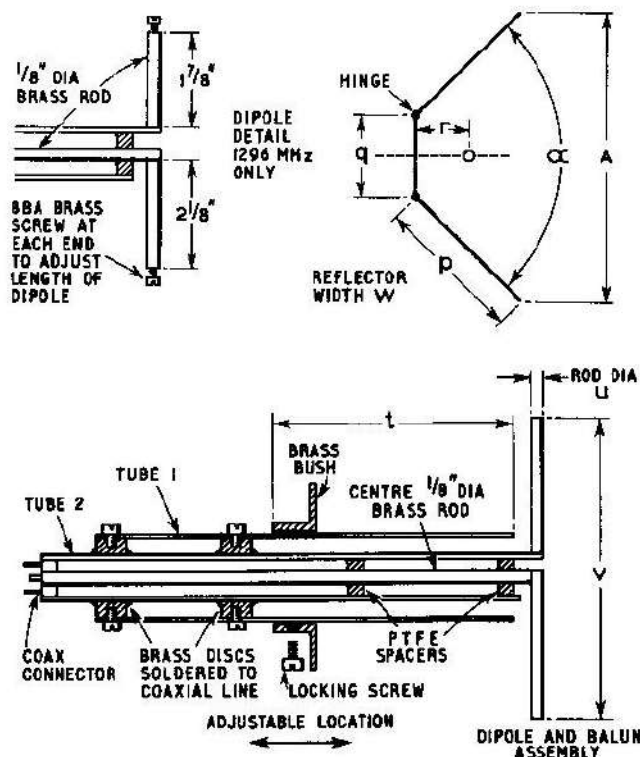
Fig. 9.42. Trough reflector and balun assembly.

			Dimensions in Inches					Tube 1	Tube 2	Gain
Band	A	p	q	r	w	t	u	v			
432	120	100	50	7½	60	6½	¼	12¾	1½" i.d. ⅟₁₆" wall	0.5" o.d. ¼" wall	17 dB
1296	45	36	9½	2½	20	2½	⅛	4¼	¾" i.d.	⅜" o.d. 20 s.w.g.	15 dB

energy is directed back into the "dish", a narrow beam will be produced. Assuming that the energy is
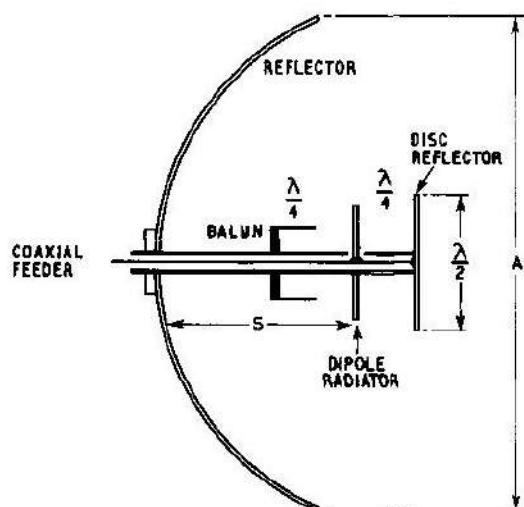
Fig.9.43. Typical design for a parabolic reflector and feeder arrangement. The dipole should be at the focus of the parabola. To achieve maximum gain the distance S should be made adjustable. The coaxial sleeve balun should be set back from the dipole.

uniformly distributed over the dish, the angular width of the beam θ will depend on the diameter of the reflector A approximately accordingly to the formula $\theta = 60\dfrac{\lambda}{A}$ where A is expressed in wavelengths. To avoid the obvious difficulties in making a parabolic reflector from sheet metal, a skeleton form of construction of wire netting stretched over wooden ribs may be found successful for frequencies up to 1500 MHz but as the wavelength becomes shorter the overall surface contour of the reflector must approach that of the true paraboloid more closely.
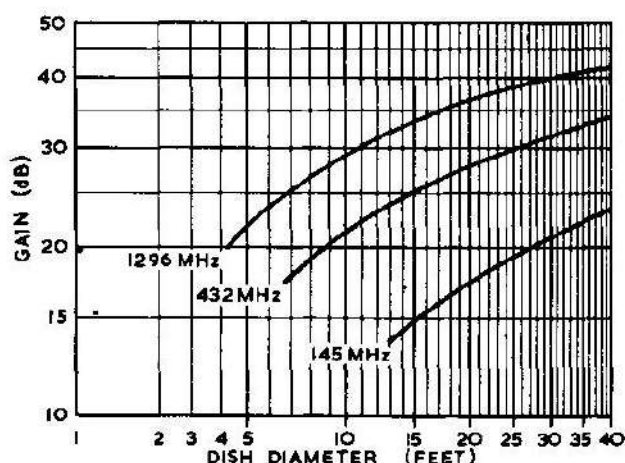
Fig.9.44. Chart showing gain in dB against dish diameter for the 145, 432 and 1296 MHz bands. From this it may be seen that to get a worthwhile gain on 145 MHz greater than with a Yagi aerial a dish diameter of 15 ft. is needed.

A suitable parabolic reflector for the 1296 MHz and 2400 MHz bands may be two to five feet in diameter. The radiator element is usually a half-wave dipole having a resonant reflector in the form of a disc one half-wavelength in diameter mounted one quarter-wavelength in front of the dipole to reflect the radiation back into the parabolic dish. The centre of
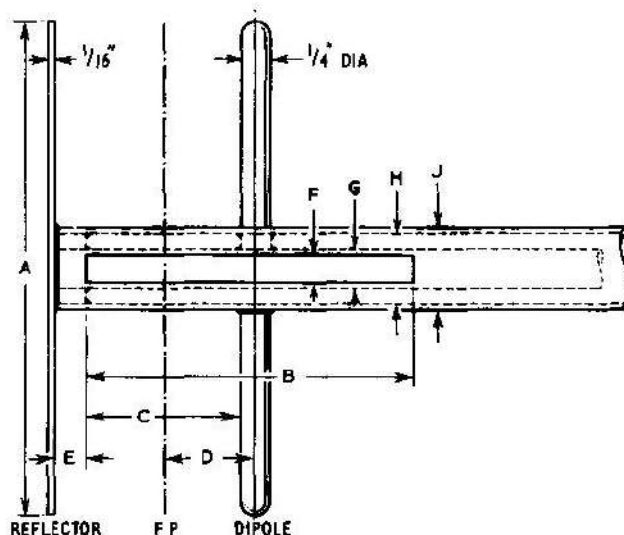
	A	B	C	D	E	F	G	H	J
23cm	4 9/16"	4 9/16"	2 9/32"	1 1/8"	11/32"	1/4"	0·27" FOR 50	0·622" OHMS	3/4"
13cm	2 1/2"	2 1/2"	1 1/4"	9/16"	11/64"	3/16	0·27" FOR 50	0·622" OHMS	3/4"

Fig.9.45. Details of dipole showing dimensions for 23cm and 13cm.

the dipole should be accurately positioned at the focus of the parabola. A balun transformer of the coaxial sleeve type may be mounted on the feed stem. A typical aerial assembly with a suitable feeder arrangement is shown in Fig.9.45. For this system a two-stub tuner will be found particularly convenient.

A 23cm Dish Aerial

The power gain given by a parabolic aerial is

$$\text{Gain} = \frac{4\pi A F}{\lambda^2}$$

where   A =   area of reflector aperture

F =   constant dependent on uniformity of phase and intensity over the reflector surface, typical value 0.6.

The overall power gain is proportional to the diameter squared (D^2). Doubling the diameter will increase the gain four times.

The beam width varies inversely with the reflector diameter and a perfect system is:-

$$\theta^\circ \text{ (beam width)} \approx 70 \frac{\lambda}{D}$$

where θ = beam width to the half power points
            D = the projected diameter of the reflector
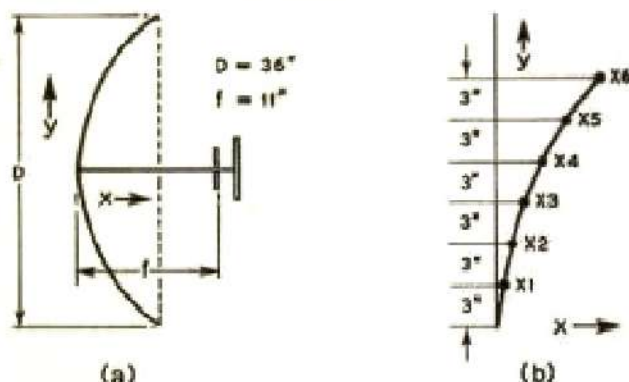
Fig. 9.46. *Drawing showing dimensions and construction of dish (a). The method of constructing a template for the fabrication of the dish is shown in (b).*

The position of the dipole or dipole - reflector with respect to the reflector is most important to make the maximum use of a given reflector. The dimensions for a three foot diameter dish are given in Fig.9.46 (a).

The general formula for a parabola is $Y^2 = 4fx$ from which a suitable template must be constructed and this is illustrated in Fig.9.46 (b).
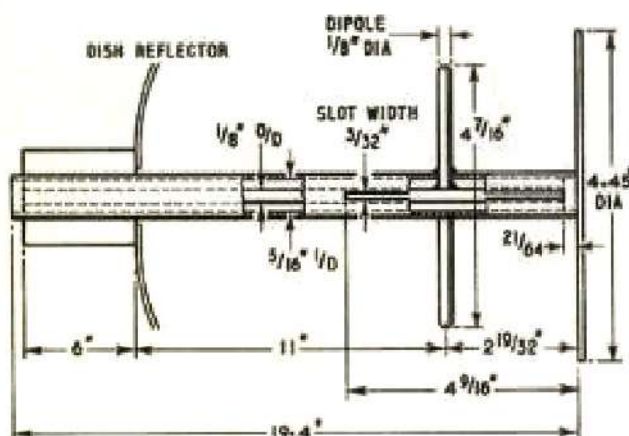
Fig.9.47. *Typical dish aerial assembly and feeder arrangement for 23cm.*

This may be set out for every 3" from the centre in the y plane and x is calculated from $x = \frac{y^2}{4f}$. This gives the following:

X1	X2	X3	X4	X5	X6
0.2"	0.82"	1.84"	3.28"	5.12"	7.36"

Construction of a Lightweight Dish

The framework is constructed by bending eight support struts to conform to the shape of the template and the inner ends of these are then brazed to a central tube. The centre tube is a 6" length of 2" outside diameter thick wall tube. Next, a 3 foot diameter ring is brazed to the outer ends of the struts.

The whole of the framework may be constructed of ¼" diameter galvanised wire provided all the joints are soldered or brazed.
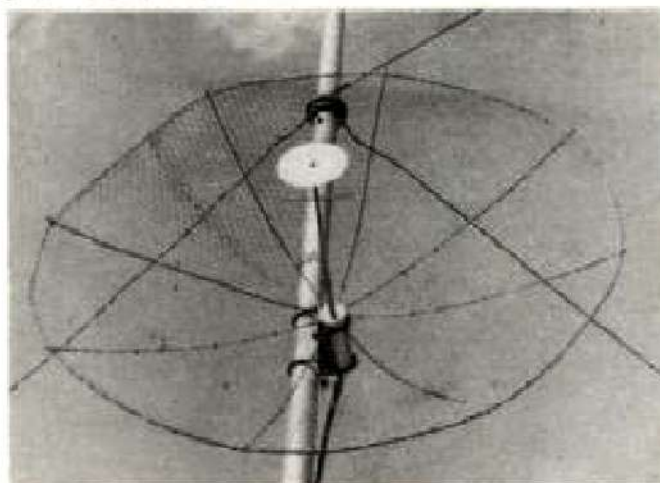
Fig.9.48. *This illustrates the complete dish and shows the method of attachment to the mast by means of a plate and two U-bolts.*

To this framework a covering of ½" mesh wire netting is attached by fixing it to the struts and outer ring with lacing wires and these should be soldered to make good joints. The whole assembly should finally be painted to prevent corrosion.

Radiator and Reflector Assembly

The arrangement of the radiator and reflector together with the coaxial feeder is shown in Fig.9.47. The coaxial feeder consists of a ¼" o.d. copper rod and a ⁷/₁₆" i.d. copper tube.

At one end of the outer tube a slot is made on both sides at right angles to the plane of the dipole – this forms the balun and its dimensions are important.

The dipole is of ¼" diameter copper rod, the upper half being jointed both to the inner and the outer of the coaxial line, the lower half only is connected to the outer tube.

The reflector disc, 4.45" diameter, is fixed to the outer end of the coaxial line so that is is 2"³/₃₂" from the dipole and both inner and outer are then connected together at the reflector. The coaxial feeder is supported in the centre of the dish reflector so that the radiator is 11" from the dish.

The method of fixing the whole assembly to the mast can be seen in the photograph and is by a plate attached to the central tube of the dish and with U bolts around the mast.

The Helical Aerial

Another simple beam aerial possessing high gain and wideband frequency characteristics simultaneously is the helical aerial: see Fig.9.49. When the circumference of the helix is of the order of one wavelength axial radiation occurs; i.e. the maximum field strength is found to lie along the axis of the helix. This radiation is circularly polarized, the sense of the polarization depending on whether the helix has a right or left hand thread.

If a pick-up dipole is used to explore the field in the direction of maximum radiation, the signal received by this dipole will show no change of amplitude as it is rotated through 360°, thus indicating true circular polarization. At any point to the side of the helix the wave will be elliptically polarized, i.e. the horizontal and vertical components will be of unequal strength.
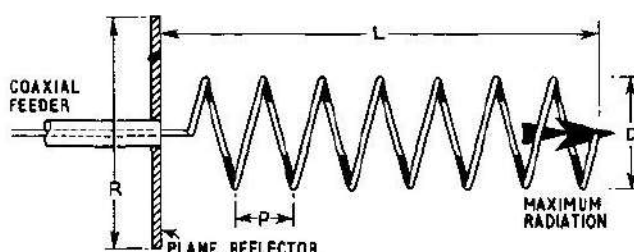
Fig.9.49. The helical aerial. The plane reflector may take the form of a dart-board type of wire grid to reduce wind resistance and weight. A gain of 12dB is obtainable with a 7-turn helix.

A helix may be used to receive the circularly polarized waves radiated from a transmitting helix, but care must be taken to ensure that the receiving helix has a thread of the same sense as the radiator; if a thread of the wrong sense is used, the received signal will be very considerably weaker.

The properties of the helical aerial are determined by the diameter of the spiral D and the pitch P and depends upon the resultant effect of the radiation taking place all along the helical conductor. The gain of the aerial depends on the number of turns in the helix. The diameter of the reflector R should be at least one half-wavelength. The diameter of the helix D should be about $\frac{\lambda}{3}$ and the pitch P about $\frac{\lambda}{4}$.

A helix of this design will have a feed impedance of about 140 ohms; this may be transformed to the feeder impedance by means of a quarter-wave transformer. A typical helical aerial having a 7-turn helix has a gain of approximately 12dB over a 2 : 1 frequency range. However, to achieve this gain fully it is necessary to use a circularly polarized aerial (e.g. a helix of the same sense) for reception. If a plane-polarized aerial, such as a dipole, is used there will be a loss of 3dB.

OMNIDIRECTIONAL TYPES

As mentioned earlier the general adoption by amateurs of horizontal polarisation in the United Kingdom and Europe as a whole, precludes the use of the simplest omnidirectional aerial - the vertical dipole, although this is sometimes used for mobile work on 70 MHz. In this case, it is used as a quarter-wave whip.

For general purpose use on the other high frequencies the aerial may vary from simple crossed dipoles or a single vertical slot, to halos both full or half size or an elaborate turnstile. The choice will depend on the operator's requirement but such an aerial is a worthwhile addition to a sharp pattern beam, as a means of hearing signals which may otherwise be missed by being in a minimum signal direction of the beam.

Simple Crossed Dipole

In this aerial two simple dipoles are mounted at right angles to one another and in the same plane.

They are connected together as shown, by a quarter-wave section of coaxial cable which is folded when in position. Allowance for the reduction in length by the velocity factor of the cable must of course be made. The dimensions given in the drawing
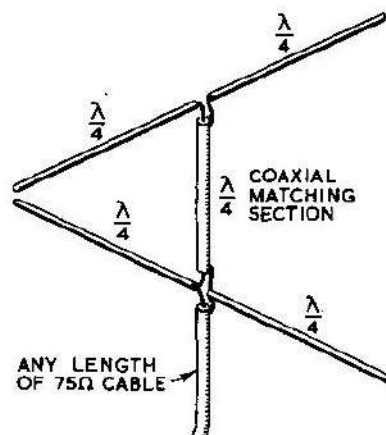
Fig.9.50. A simple crossed dipole aerial. For 2 metres the elements would be 19½" long of ¾" dia. tube. The coaxial matching section is 13½" long, allowing for the velocity factor of the cable.

are suitable for 144 MHz and suitably scaled elements and coaxial matching section may be used for other frequencies. The radiation pattern of an aerial of this type is almost circular.

Slot Type

This is a pure slot type aerial and differs from the simple rectangular type (skeleton) used in some Yagi aerials. It consists of a metal cylinder with a slot

running the entire length except for the short circuited ends. In the example illustrated the slot width is 1" wide and 58" long which is resonant just above 146 MHz and is tuned to the band by a small capacitor connected across the centre of the slot.
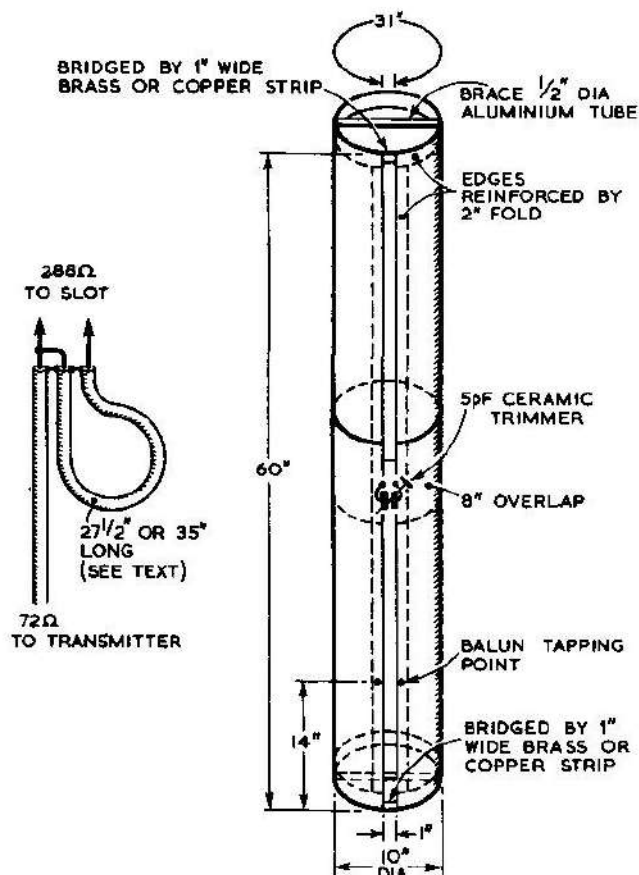
Fig. 9.51. A slot aerial for 2 metres. This shows the construction of the aerial and the balun.

Details of the construction, and the balun to feed this aerial, are shown in the drawing. The material for the cylinder may be of solid sheet, perforated sheet or mesh, and the material can be copper, brass, aluminium, zinc or even tinned iron. The losses would be lower with the higher conductivity materials, but at the powers likely to be used the high resistivity materials are unlikely to have any significant effect on the performance.

Construction

The cylinder described is made from perforated zinc sheet. Locally available supplies come in 36" wide coils of which two 35" lengths are required. These are spot soldered together, side to side, with an 8" overlap. The edges of the resulting large sheet are reinforced by folding to a width of 2" and spot soldering. The final size of the sheet is 60" x 31". This is then bent to cylindrical shape, leaving a 1" slot in the periphery which is bridged, top and bottom, with

1" wide strips of sheet copper (or tinplate) soldered into place. Each end of the cylinder is braced with a length of ½" diameter aluminium tube, flattened at the ends, bent at right angles for ½" at each end and bolted to the cylinder walls across the diameter. A small 5pF ceramic trimmer is soldered across the exact centre of the slot. The finished product is not robust, but this is not important if it is for indoor use.

The aerial is fed by 72 ohm semi-air spaced coaxial cable through a 4 : 1 balun. The feed point is about 14" from the lower end of the slot and the feeder, with balun tied to it, is run through the centre of the cylinder. The balun section is 35" long (27½" if solid dielectric cable is used, although 26" may be more suitable with some solid dielectric types).

The tuning of the aerial should preferably be done using a standing wave indicator of the type described in this Manual. The actual position of the connections to the slot and the adjustment of the tuning capacitor should be made for the optimum v.s.w.r. and a value of 1.2:1 is readily obtainable.

Reduced Size Slot

Another and more compact slot aerial, which has been called the 'Abe Lincoln' is claimed to have an almost circular radiation pattern.

In this form it is suitable for both mobile and fixed station operation, but if used on the roof of a
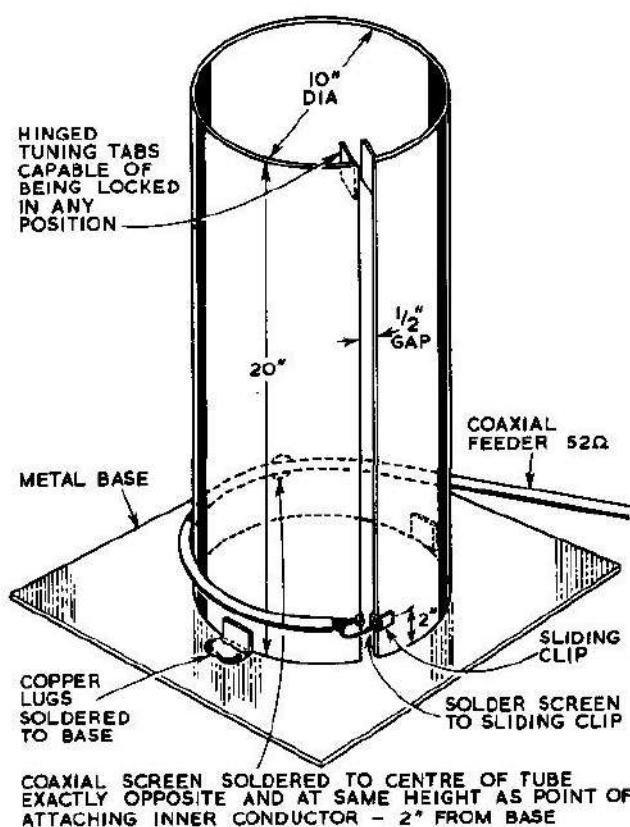
Fig. 9.52. Constructional details of the Abe Lincoln mobile aerial for 144 MHz.

car, the open slot and top should be filled in with suitable insulating material, such as perspex, to prevent noise produced by windage.

Although in the design illustrated, a metal base has been fitted, this is not essential, but it makes a mechanically stable unit. If the base is omitted then the bottom of the slot must be shorted with a suitably thick metal strap. Matching the feeder into the slot should be done with sliding contacts and using a s.w.r. bridge.

HALO AERIALS

These types are popular for mobile work, as they are light, easily made, and have an almost circular polar diagram. In fact there is a point of minimum signal in the direction of the side opposite to the gap.

Construction

The design is based on standard coaxial plugs and sockets which are readily obtainable on the surplus market. It was decided that it would be convenient to mount the halo on a coaxial plug, so that the whole assembly could be screwed on to a coaxial socket which would provide both the electrical connection and the mechanical support.

The general construction is illustrated in Fig.9.53 and to preserve mechanical rigidity at the weakest point of the aerial, i.e. in the centre of the radiating element, use is made of the gamma match for 70 ohm impedance, this permits the radiating dipole to be made in one complete length. The conventional series capacitor in the arm of the matching stub has been omitted on the grounds of mechanical simplicity. Mobile aerials are usually fed by a comparatively short feeder from the transmitter, and the slight increase in s.w.r. resulting from the residual reactance of the matching stub, does not introduce any further appreciable loss in radiated signal level.

The dipole proper is formed from a 39" length of ¼" 22 s.w.g. brass tube. This is normally supplied hard
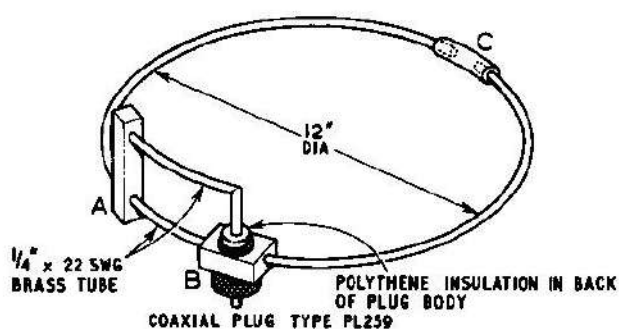
drawn and should be annealed before bending as follows: heat the tube to a dull red with a blowlamp and quench it immediately in cold water. It is important to anneal the tube uniformly along its length, otherwise it will not be possible to obtain a smooth bend when the circle is formed. The annealed length of tube should be bent round a mandrel approximately 12" in diameter.

When finished the ends of the tube are approximately 2" apart. The insulating spacer is formed from a 3" length of round polythene taken from a piece of ½" diameter coaxial cable. The centre conductor is removed and the ends are drilled ½" deep with a ¼" diameter drill. The ends of the brass tube are pushed into these holes, and due to the elasticity of the polythene, are gripped firmly without the need for any further clamping.
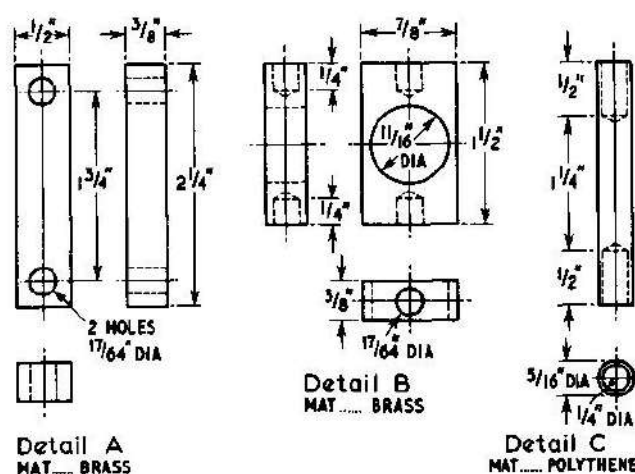
Fig.9.54. Details of the brass supporting blocks and polythene spacers.

The main support for the halo is provided by a brass block (Fig.9.54), which has a hole drilled through its centre into which is brazed the back nut of a coaxial plug type PL259. The brass tube of the circle is cut diametrically opposite the polythene spacer and the ends are brazed into either side of this brass block. The shorting bar for the feed point is also made of a brass block and is brazed into position 4½" from the centre of the dipole. The position of this tap is not critical and the dimension of 4½" has been found to be quite satisfactory for a match to 70 ohm cable. The conductor for the live side of the gamma feed is also made of ¼" brass tube. A 4½" length has one end brazed into the shorting bar and at the other end a further short piece of ¼" tube is mitred and brazed so that it projects down into the centre of the coaxial plug. The end of this tube is stopped with a 1" 6BA cheesehead bolt, brazed into it: this passes through the pin of the centre portion of the coaxial plug. A disc of ⅛" thick polythene is pushed over the ½" tube to locate it centrally in the free portion of the plug which is then screwed home into the back nut drawing the

Fig.9.53. Drawing showing the general construction of the halo aerial.

spacer up as it goes. The centre stud may then be soldered to the pin to complete the assembly Fig.9.55. The two locating pips on the free portion of the plug are filed flat so that when the plug is screwed on to a socket, the free portion of the plug (which carries the "hot" feed to the dipole) is also free to rotate.
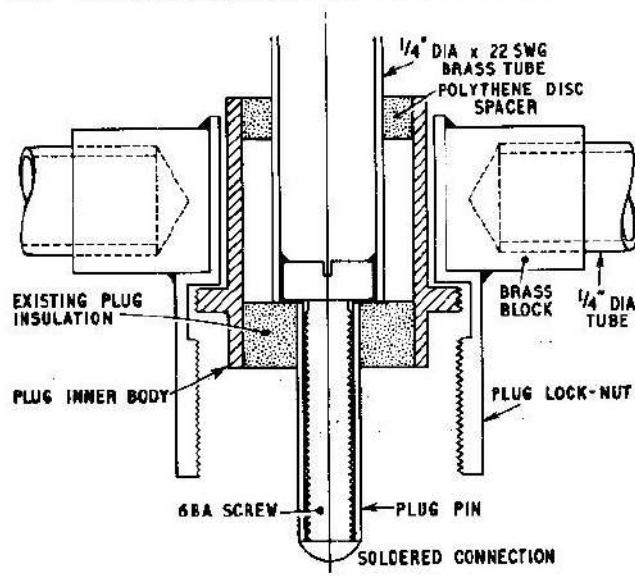
Fig.9.55. Method of assembly of the supporting plug.

Mounting on the Car

The mounting of the halo proper is detailed in Fig.9.56. An Amphenol tee connector, type M358, is screwed on to the halo plug and the incoming coaxial cable loaded into a plug, (type PL259) with ¼" adaptor, is screwed into the side of the tee fitting to provide the electrical connection. The bottom section
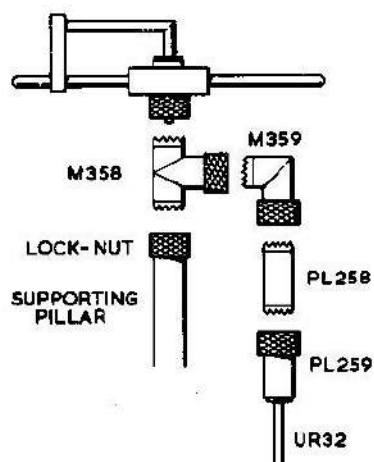
Fig. 9.56. Assembly of the coaxial fittings for the halo mobile aerial.

of the tee is supported mechanically by screwing it into the back nut of a further plug, type PL259. This back nut may be brazed to some convenient mounting to suit individual requirements. A ¼" thick disc of

polythene should be pushed into the bottom of the nut to prevent any possibility of the inner conductor of the tee connector coming into contact with the supporting block to which the back nut is fixed.

A Mini-Halo for 144 MHz

This aerial is a half size version of the standard halo but with capacity end loading of the elements.

The drawing, (Fig.9.57) is self-explanatory, but it is essential that the aerial should be of robust construction to avoid any whip. The construction is all brass: the elements and gamma match are made of ³⁄₁₆" external diameter brass tubing, while the main junction tube is ½" internal diameter and the shorting bar ⅜" x ¼" solid brass drilled and tapped 6BA. The ½" tube is surmounted with a ½" Perspex rod which should be a tight fit. The rod is drilled to take the centre core of the feeder, and a hole drilled at right angles takes the end of the gamma ³⁄₁₆" tube; this is drilled and tapped to anchor the feeder core. The screen is soldered to the inside of the ½" tube. Regarding the tubular capacitor, the inside sleeve is ⁵⁄₃₂" internal diameter filled for part of its length with polythene or p.t.f.e. The polythene core of thick
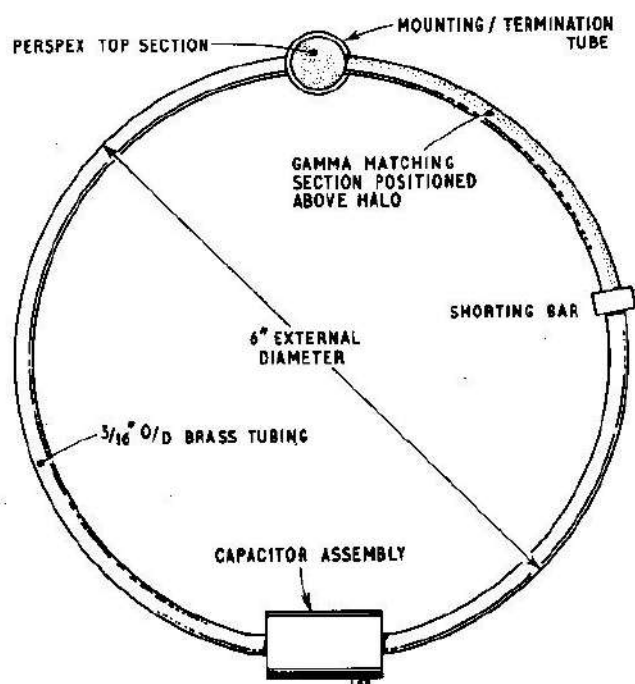
Fig.9.57. Plan view of the mini-halo for 144 MHz.

ex-government coaxial cable could be used, and this needs to be tapped to take one end of the element which has been threaded to suit ½" or possibly 3BA. The external tube would have to suit the external diameter of the inside tube depending on its gauge — possibly about ¼" internal size. The drain hole is important as this prevents moisture collecting, and should be positioned at the bottom. The collar

holding the other end of the element should be soldered in position. The element end (which has been tapped) is held in position with a 6BA screw and is a twisting fit. Before assembling, spring the elements apart to ensure that good contact is made between the capacitor and anchored end of the element.
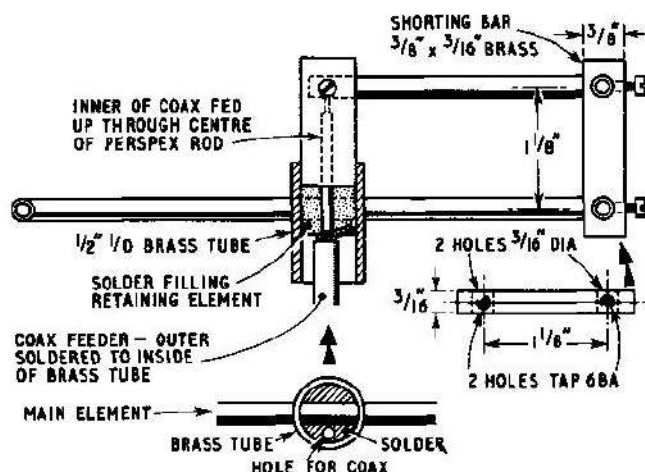
Fig. 9.58. Side elevation of the completed halo, with drilling details of the gamma match support and perspex mounting rod.

Tuning is accomplished by twisting the sleeve and is fairly critical, but once set it can be locked and forgotten. The value is between 2 and 3 pF. It will be found convenient to tune for maximum signal using GB3VHF if the location is suitable, or other steady carriers. The use of a signal strength meter is essential.
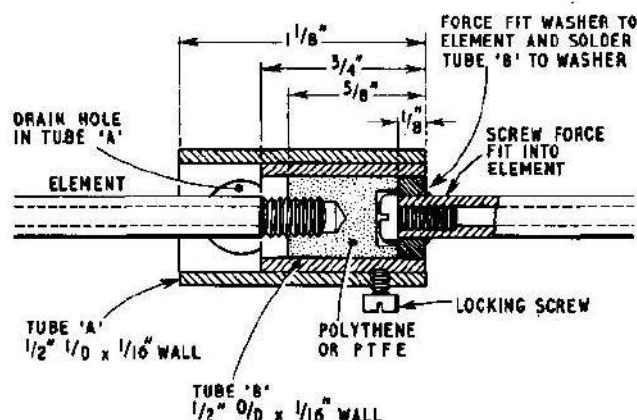
Fig. 9.59. A cut away view of the capacitor.

The importance of keeping moisture out of the capacitor was the reason for the particular form of construction and this was found to be quite satisfactory in wet weather. The gamma match is of the same dimensions as would be used on the full size version, approximately 4½" in length, and should be adjusted with a reflectometer for minimum standing waves.

A Clover Leaf Aerial for 144 MHz

This is an alternative aerial suitable for mobile use where the large size is not a disadvantage. It is about 4 ft across and will, of course, offer a greater wind resistance than the standard halo. The three leaf
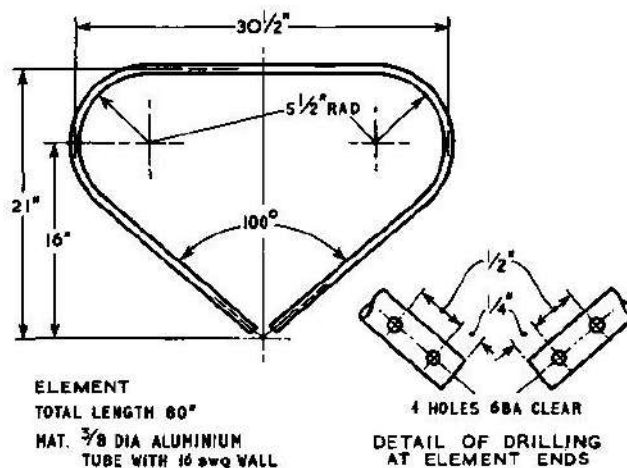
Fig. 9.60. Details of one element. Three identical elements are required.

elements are half-wave dipoles end fed with quarter-wave sections (like a folded dipole opened out) each set at 120° to one another and fed in parallel. To match these parallel elements to the 50 ohm cable used in this example, a matching stub is required to provide the required inductive reactance.

In Fig. 9.61 (a) the general arrangement of the aerial is shown and (b) is given to illustrate how the stub achieves the impedance matching needed, the centre of the stub will be zero impedance and this will increase to a maximum at the ends of the dipoles.

With the dimensions given the joint between the stub and the three dipoles will be 50 ohms. If a 75 ohm cable is to be used then the stub will have to be lengthened and the feed sections of the dipoles correspondingly shortened. An almost circular pattern
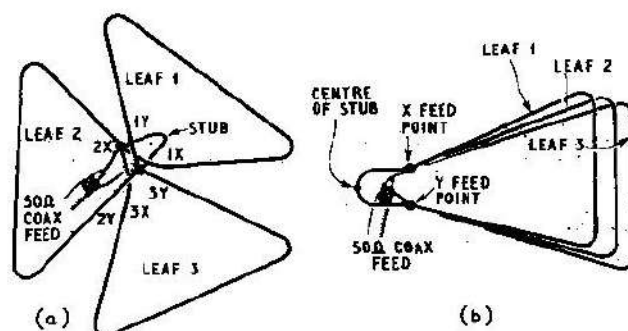
Fig. 9.61. The general arrangement of the clover leaf aerial. The feeder is connected to the two junction points shown.
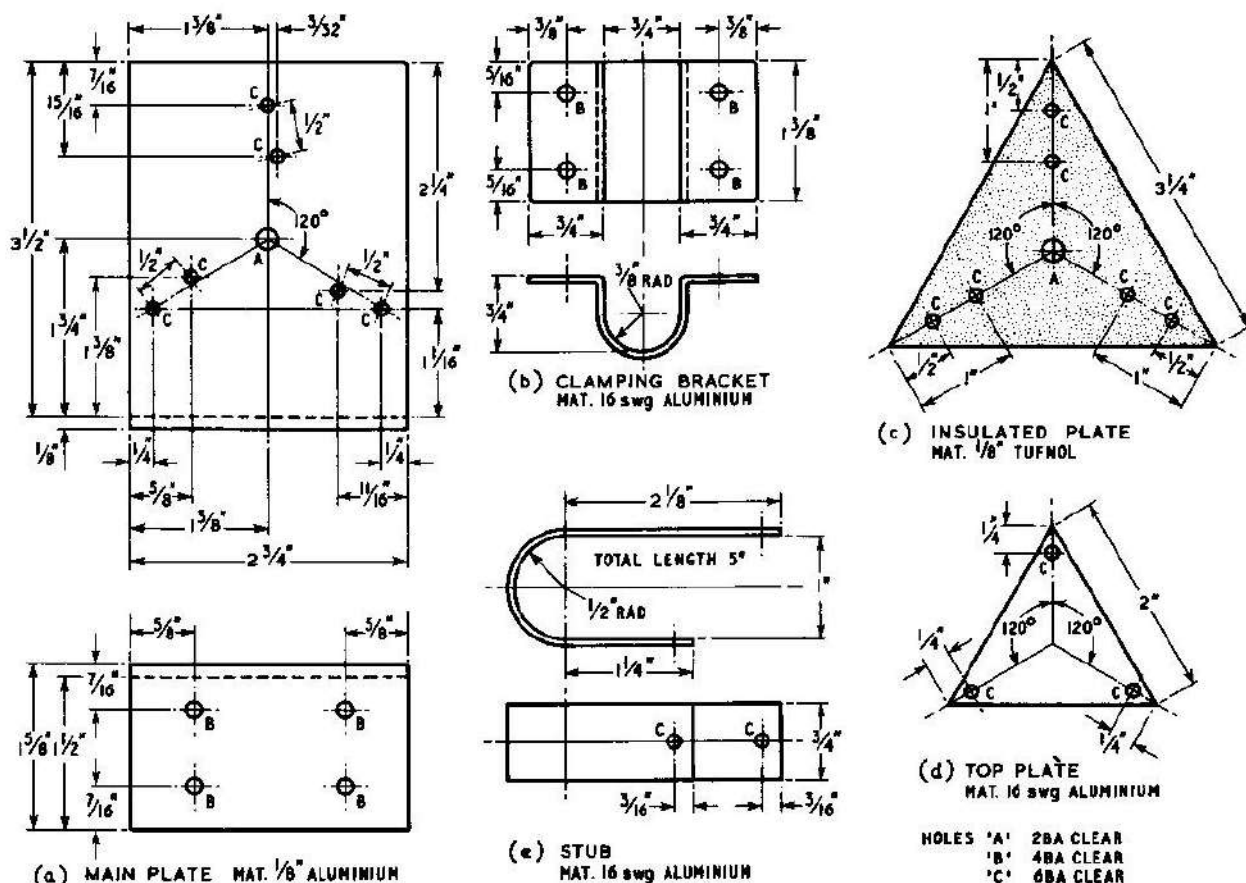
Fig.9.62. Constructional details of the centre assembly of the Clover Leaf aerial.

is achieved but slight nulls occur in line with the three gaps between the leaves. When correctly made a v.s.w.r. of about 1.2:1 should be obtained.
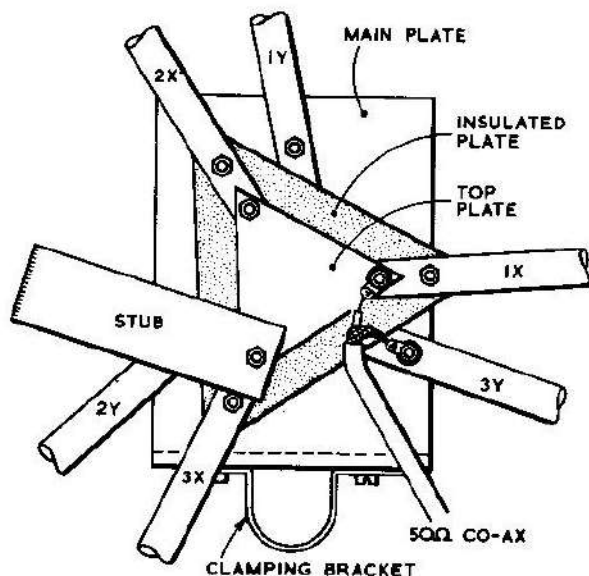
Fig.9.63. General arrangement of centre assembly.

An "OMNI-V" for 144 MHz

This aerial consists of a pair of halfwave dipoles. The centres of the dipoles are physically displaced to produce a quadrature radiation and the ends of each dipole are supported on a quarter-wave shorted stub. A pair of Q bars are tapped down the stubs to a point where the impedance is 600 ohms so that when the two units are fed in parallel they produce an impedance of 300 ohms at the centre. A 4:1 balance to unbalance coaxial transformer is fitted to the centre point of the Q bars so that a standard 75 ohm coaxial cable feeder may be used. The general arrangement is shown in Fig.9.64(a). Fig.9.64(b) shows how the aerial may be arranged to give a bi-directional radiation pattern.

Omni-directional (a)
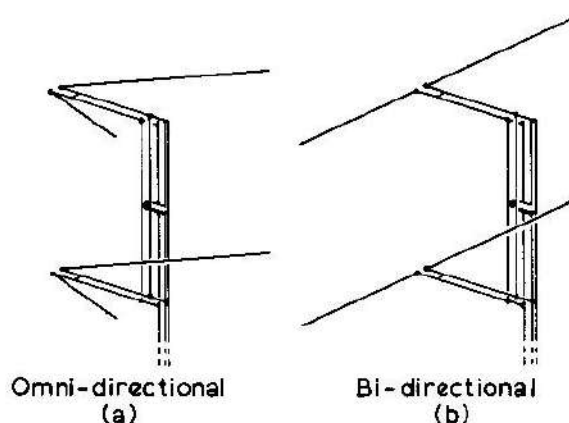
Bi-directional (b)

Fig.9.64. Formation of the omni-V aerial.
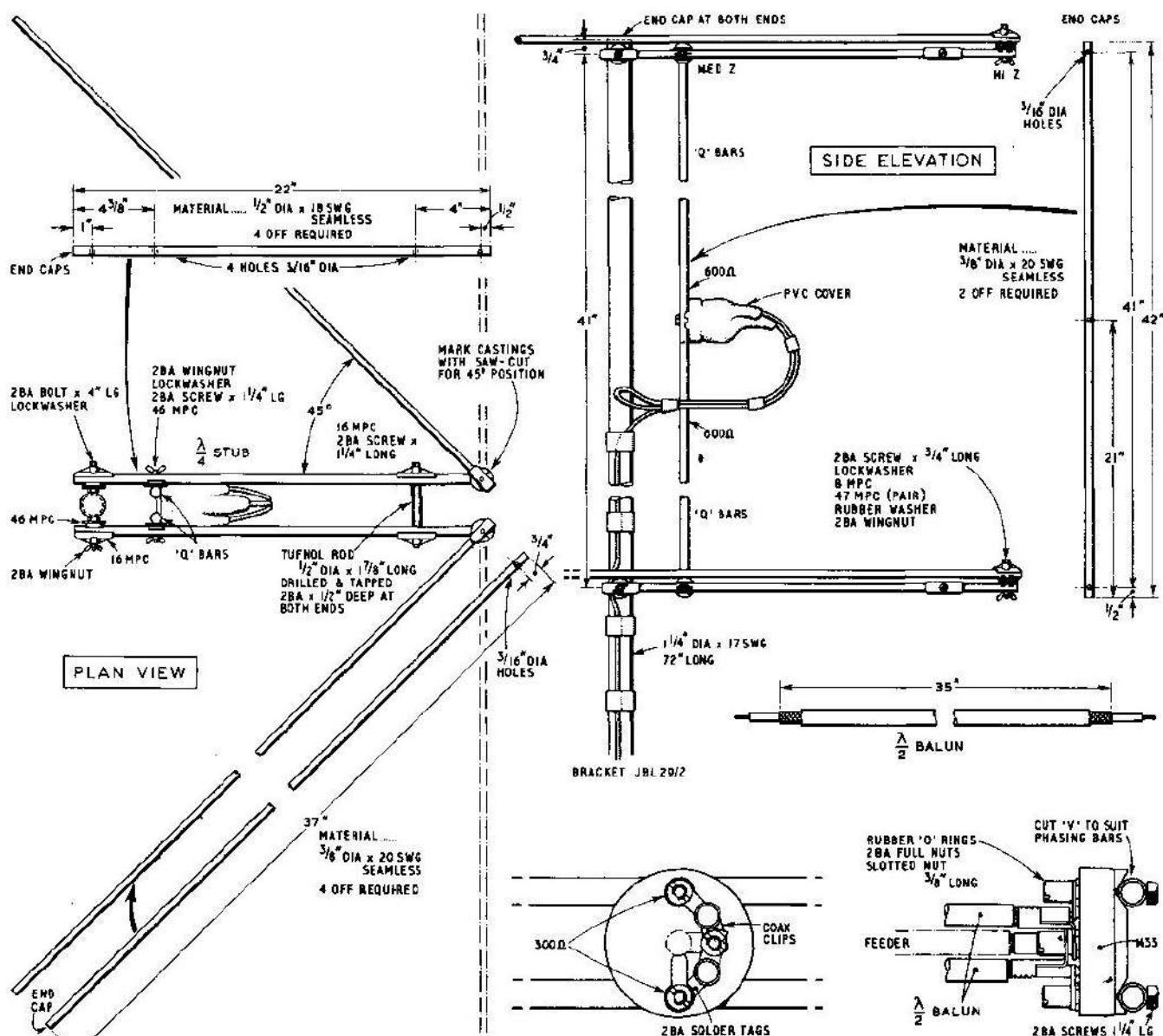
Fig. 9.65. The mechanical details of the construction of the omni-V.
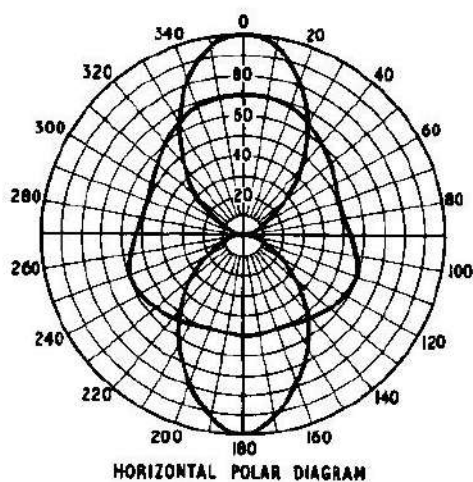
Fig. 9.66. The horizontal polar diagram for an average aerial, showing both the bi-directional and omni-directional charts.
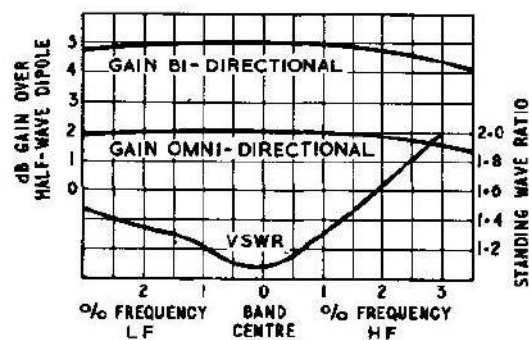
Fig. 9.67. Chart showing gain versus v.s.w.r. on the omni-V described.

Materials for Elements

The radiating efficiency of an aerial may be defined as the ratio of the powers radiated, to the power input to the aerial. The difference between the radiated power Pr and the input power Pin, is the power lost in dissipation by the aerial itself, Pd.
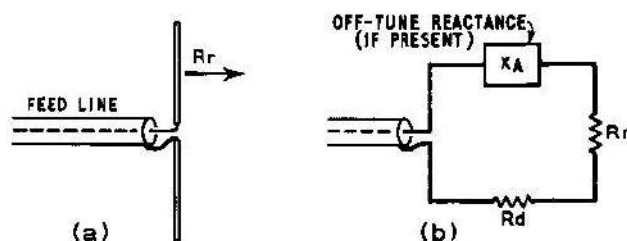
Fig. 9.68. An aerial and its equivalent circuit.

An aerial can be represented by the equivalent circuit shown in Fig. 9.68 (b). Power in such a circuit can only be lost in a resistive element, so that both the power dissipating elements are shown as Rf and Rd. Rr is proportional to the power radiated (effective radiation resistance) and Rd is proportional to the power dissipated in the aerial itself (effective loss resistance). When considering the relative efficiencies of the various materials for aerial elements it is the value of the effective loss resistance Rd which is important. Since most v.h.f. and u.h.f. aerials are made up of various elements of more or less uniform cross-sectional area, the "d.c." resistance may be calculated from,

$$Rdc = \frac{L}{A} \times \text{resistivity}$$
where L is the length
A is the area

and resistivity is the resistance per unit length or unit area of the material involved, and this is usually given in ohms per centimetre cube.

For example taking L as a halfwave length at 144 MHz and A as the area of ¼" diameter rod then typical values for Rdc are given below.

Metal	Resistivity (ohms/cm^3)	Resistance (Rdc)
Copper	1.72 x 10−6	0.0023 ohms
Aluminium	2.83 x 10−6	0.0037 ohms
Zinc	5.9 x 10−6	0.0077 ohms
Brass	7.5 x 10−6	0.0098 ohms

All these are negligible with respect to the radiation resistance Rr of a half-wave dipole, which for the chosen element diameter is about 65 ohms. This however, is only part of the effective loss resistance.

As the frequency of the current flowing in the material is increased from zero frequency or d.c. another factor, 'skin effect', modifies the current distribution in the cross section of the conductor, concentrating it more and more in the outer skin as the frequency is raised.

This, therefore, reduces the working area of the conductor and increases its effective resistance. At v.h.f. the skin carrying most of the current becomes quite thin, about 0.0005" and is proportional to the square root of the conductor resistivity. From this, it is obvious that the skin will be thicker for metals with higher resistivities, and since their r.f. resistance will be less relative to their d.c. resistance, than for the better conductors such as copper, this makes their use more attractive than might be expected.

The r.f. resistance, R'rf, for a current which is constant on the length of a halfwave element is approximately as shown below.

Metal	Skin Depth	R.f. Resistance R'rf	Ratio R'rf to Rdc
Copper	0.00022"	0.34 ohms	148
Aluminium	0.00028"	0.41 ohms	110
Zinc	0.00040"	0.60 ohms	78
Brass	0.00045"	0.68 ohms	69

The radiation efficiency and aerial loss in dB for the metals considered are as shown below:

Metal	R_L	Radiation Efficiency	Loss due to Aerial elements
Copper	0.17 ohms	99.75%	0.011dB
Aluminium	0.20 ohms	99.7 %	0.013dB
Zinc	0.30 ohms	99.5 %	0.022dB
Brass	0.34 ohms	99.4 %	0.026dB

These figures are for elements of ⅛" diameter, for elements of the more popular diameters, ¼" to ½", the r.f. resistance will be lower although for very thin walled tubes the d.c. resistance may be greater.

The following simple formulae can be used to calculate the skin depth of R'rf resistance.

For solid rod elements R'rf =
$$Rdc \times \frac{\text{Diameter in inches}}{4 \times \text{skin depth in inches}}$$

For thin walled tubes R'rf =
$$Rdc \times \frac{\text{Wall thickness in inches}}{\text{skin depth in inches}}$$

Skin depth at v.h.f. in inches = $2\sqrt{\frac{\sigma}{\mu F}}$

where
σ is resistivity in ohms per cm cube
μ is permeability taken as 1 for non ferrous metals
F is frequency in Hertz

Electrolytic Corrosion

The use of dissimilar metals in an aerial system is likely to cause considerable trouble due to electrolytic corrosion. Each metal has its own electro-potential and unless metals of similar potential are used, the difference will cause corrosion at the point of contact even when dry and when moisture is present this effect will be more severe.

If, for any reason, dissimilar metals must be used then considerable care should be taken to exclude moisture, the corrosive effects of which will vary with the atmospheric pollution.

The various metals can be arranged in groups as follows:-

Magnesium	Anodic
Aluminium	
Duralumin	
Zinc	
Cadmium	
Iron	
Chromium iron alloys	
Chromium nickel iron alloys	
Soft solder tin-lead alloys	
Tin	
Lead	
Nickel	
Brasses	
Bronzes	
Nickel copper alloys	
Copper	
Silver solders	
Silver	
Gold	
Platinum	Cathodic

Metals in each of the above groups may be used together with little corrosive action, but metals from different groups will suffer from this effect. Also, since the above is arranged in order, the greater the spacing in the list, the greater will be the effect.

The lower of the metals in this list will corrode those in the upper portion, for example, brass or copper screws in aluminium will corrode the aluminium very considerably, whereas with cadmium plated brass or copper screws there will be very much less corrosion of the aluminium.

AERIAL TEST GEAR AND ACCESSORIES

Tuning Adjustments

To tune up any aerial system it is essential to keep it away from large objects, such as buildings, sheds and trees, and the array itself should be at least two wavelengths above the ground. It is useless to attempt any tuning indoors since the change in the surroundings will result in completely different performance when the array is taken outside.

Undoubtedly the most effective apparatus for tuning up any aerial system is a standing-wave indicator or reflectometer. If there is zero reflection from the load, the standing-wave ratio on the aerial feeder is unity. Under this condition, known as a flat line, the maximum power is being radiated. All aerial matching adjustments should therefore be carried out to aim at a standing-wave ratio of unity.

If suitable apparatus is not available, the next best course of action is to tune the aerial for maximum forward radiation. A convenient device for this is a field strength meter comprising a crystal voltmeter connected to a half-wave dipole placed at least ten wavelengths from the aerial. When adjustments have resulted in a maximum reading on this voltmeter, the feed line may nevertheless not be "flat" and therefore some power will be wasted. However, if the best has been done with the resources available it is highly likely that good results will be achieved.

A REFLECTOMETER FOR V.H.F.

Introduction

When power at radio frequency is fed into a transmission line which is correctly terminated at its far end, this power is propagated along the line in terms of voltage and current waves and is all absorbed in the load at the far end of the line. This represents the ideal condition for the transfer of power from a transmitter to an aerial system: such a condition is rarely, if ever, achieved due to the impossibility of presenting the transmission line with an absolutely matched load. In practice, it is possible only to terminate the line with an aerial or load which approaches the perfect condition and under these circumstances a certain amount of power is reflected

at this mis-termination and is propagated back down the line again by means of further waves of voltage and current travelling in the opposite direction, to be either absorbed or re-reflected at the generator according to whether the generator impedance terminates or mis-terminates the line.
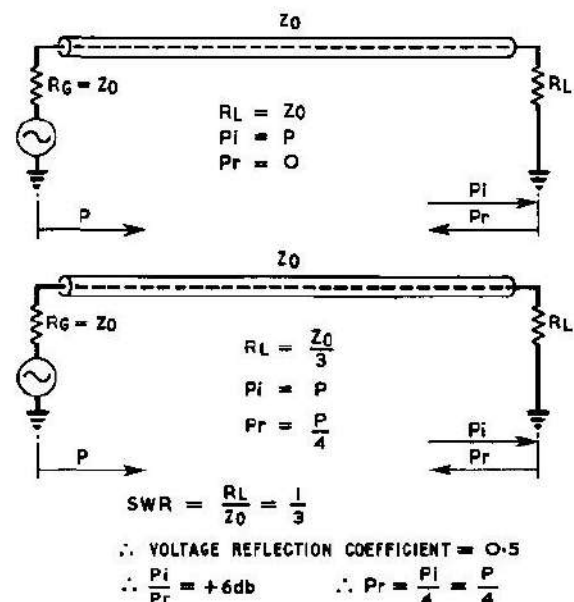
Fig.10.1. Effect of mis-termination on a transmission line in terms of the incident and reflected power at the load.

The amount of power reflected from the aerial or load mis-termination is directly proportional to the magnitude of the mismatch on the line, and therefore the mismatch on the line, or, in more practical terms, the standing wave ratio may be expressed in terms of the ratio of the forward or "incident" and the backward or "reflected" powers (Fig. 10.1).

If the s.w.r. = S, then the voltage reflection coefficient K is given by

$$K = \frac{1 - S}{1 + S} \dots \dots \dots \dots \dots \dots \dots \text{(i)}$$

Clearly if a device can be constructed which will differentially respond to power in terms of direction then such a device can be used directly to measure standing wave ratio, and the ratio M of incident to reflected power is given by

$$M = 20 \log_{10} 1/K \text{ dB} \dots \dots \dots \dots \dots \text{(ii)}$$

It can be shown that if a line, whose length is short compared with a wavelength, is introduced into the field of, and parallel to another line which is carrying power, then an amount of power is coupled into the secondary line, which is directly proportional to the magnitude of the power travelling in either or both directions on the main line. The configuration of
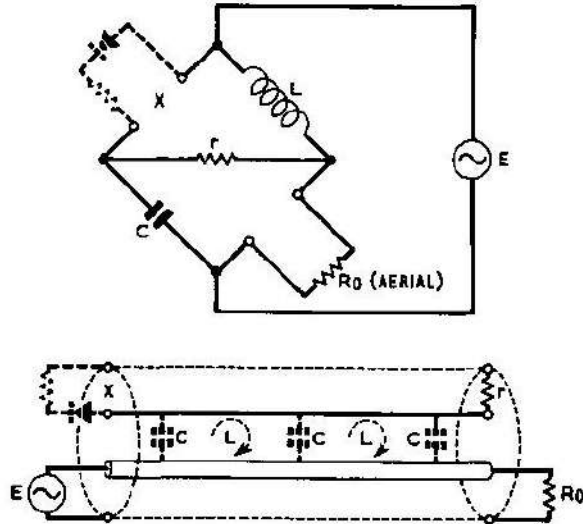
Fig.10.2. Maxwell Bridge representation of transmission line coupler.

main and sampling lines may be regarded as a Maxwell Bridge, the reactive arms of which are provided by the distributed capacitance C and mutual inductance L of the coupled lines, and the effective load on the bridge

is r (Fig.10.2). Then if $r^2 = L/C$, the bridge is effectively balanced at all frequencies, and now power from the generator E appears in the load r, but a proportion appears in the detector load. If two such subsidiary lines are coupled to a main transmission line carrying power and are respectively terminated at opposite ends, an output can be taken from each line which is respectively proportional to the incident and reflected power in the main line. This is the principle behind the reflectometer, Fig.10.3. The accuracy of such an instrument depends upon the correct termination of the sampling lines. Any mismatch on those lines will result in a standing wave along them, and consequently the r.f. voltages appearing at their output terminals will not be proportional to the forward and reflected powers. This parameter of performance♦ is termed the directivity of the reflectometer, and is measured as the ratio of the voltage developed on the "backward" sampling line, when the instrument is itself correctly terminated, to the voltage on this same line when the instrument is reversed. The directivity is usually expressed as a ratio in dB.

Design

Before the details of construction can be finalized, it is necessary to consider one or two design aspects of the instrument itself. It has already been shown how two voltages may be obtained which are proportional to the forward and backward components of power respectively. However, these
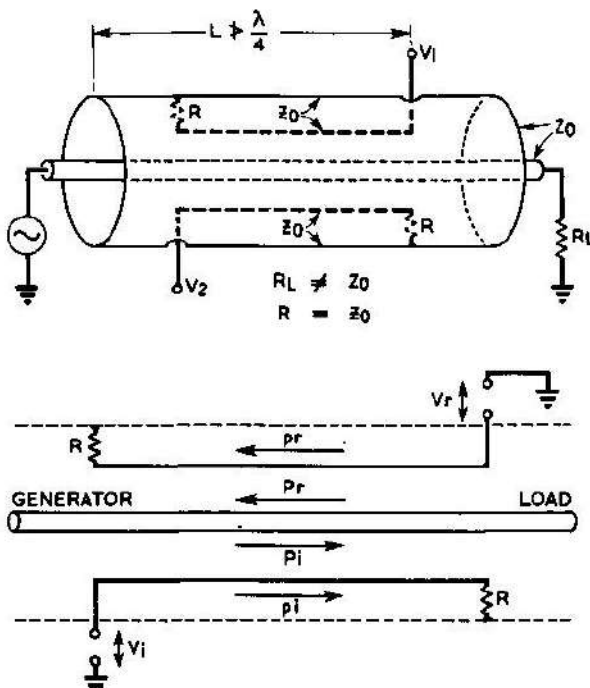
Fig.10.3. Arrangement of sampling lines to respond respectively to incident and reflected powers.
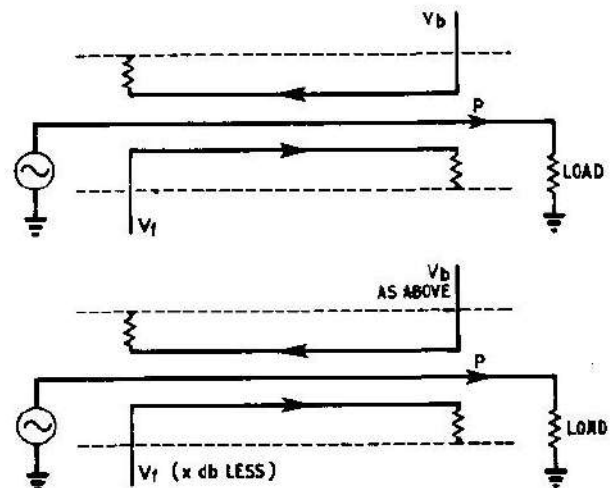
Fig.10.4. Instrument sensitivity and coupling ratios.
(a) sampling couplings equal.
V_f *gives meter f.s.d. ∴ s.w.r. Vb/Vf, say y dB . . . . . (i)*
y dB corresponds to a given meter deflection

(b) samplings couplings different by x dB.
V_f *gives meter f.s.d. less x dB for same power. S.w.r. ∝*
V_b/*meter f.s.d. ∴ for same deflection as (a),*
s.w.r. = x + y dB . . . . (ii)

10.2

voltages are still of a radio frequency nature, and it is necessary to convert them to d.c. before they can be used to deflect a conventional meter. If the "forward" voltage is arranged to produce f.s.d. on the indicating meter, then clearly the meter can be calibrated directly in s.w.r. by observing the deflection produced by the "backward" voltage, and making due allowance for any differences in coupling between the two sampling lines and the main line. This calibration will be valid independent of the actual transmitted power, since in each case the meter is adjusted for f.s.d. In practice it is easier to arrange for identical sampling lines, in which case the calibration of the meter becomes a simple question of the ratio of r.m.s. voltages applied at the rectifying diodes. This places an inherent limit on the sensitivity of the instrument at low s.w.r. but, provided that the relative couplings can be measured, it is possible to improve the overall sensitivity for a given power and meter sensitivity by arranging for an appreciably greater degree of coupling on the "backward" sampling line than on the "forward" and thus providing an immediate improvement of xdB in the lowest s.w.r. which can be

measured for a given deflection of the meter (Fig. 10.4). Care must be exercised that the coupling from either lines is not increased to the point where the presence of the sampling line distorts the electromagnetic field around the inner of the main line sufficiently to cause an effective change of Z_o of the main line and hence introduce an inherent s.w.r. in the instrument itself. As a general rule the coupling should not be greater than 30dB to maintain an inherent reflection coefficient of less than 3–4 per cent.

When the main line is carrying power which is subject to amplitude modulation, then the sampling voltage from the "forward" (and the "backward") line will also be subject to amplitude modulation at the same modulation depth. Since this voltage has already been rectified and arranged to deflect the indicator meter to full scale, then if this rectified (or detected) signal is once more rectified, a d.c. voltage will be obtained which is proportional to the audio frequency voltage modulating the carrier. This voltage can then be used to deflect the meter and this can be calibrated directly in percentage modulation. This calibration will also, to a first order, be independent of the transmitted power, since the meter has been adjusted for f.s.d. on the sampled detected carrier. In practice it is necessary to resort to full-wave rectification of the detected carrier, although this does not really provide sufficient d.c. voltage to cause large excursions of the meter reading under full modulation conditions, i.e. it is not possible to advance the meter to f.s.d. for 100 per cent modulation. It is recommended therefore that the "modulation meter" aspect of the instrument be regarded only as of an arbitrary quantitative nature.
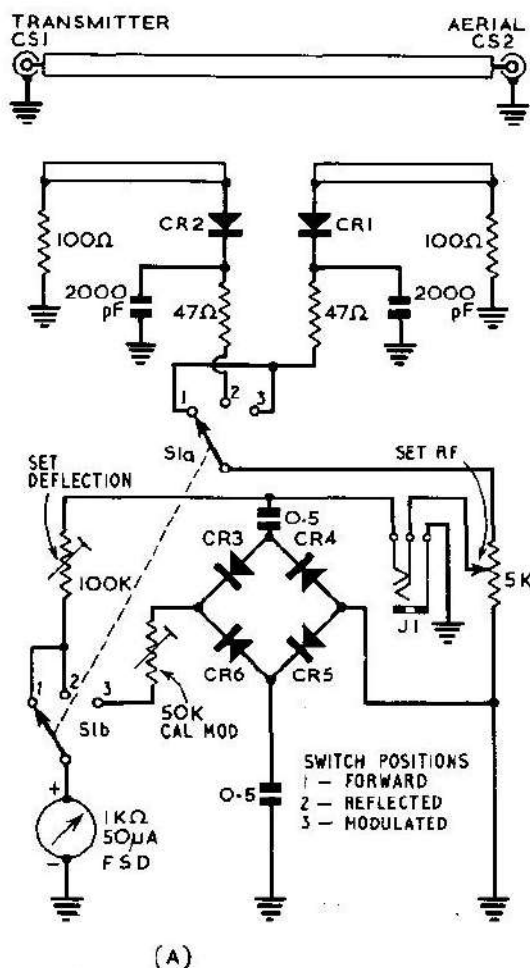
Fig.10.5. Circuit diagram of reflectometer. CR1, 2, GEX66; CR3, 4, 5, 6, GEX66 or GEX54. CR3–6 should be by-passed by 1000pF capacitors across each diode.
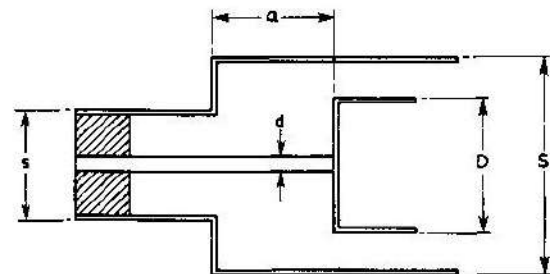
Fig.10.6. Optimum step length at a discontinuity Z_o = 138 $log_{10}s/d$ = 138 $log_{10}S/D$. Optimum step length a is a function of Z_o D/d.

The introduction of the instrument into a transmission line requires the use of plugs and sockets, and this in turn will lead to a discontinuity in the line at the ends of the reflectometer proper, due to the sudden transition from the relatively large inner of the instrument line to the inner of the coaxial fitting. The size of the inner conductor of the instrument must be large to maintain the line characteristic impedance while at the same time providing sufficient room to accommodate the sampling lines between the inner and outer conductors, i.e. this is a physical requirement. These discontinuities are of the "right

angled step" type, Fig.10.6, and there is an optimum arrangement of dimensions to provide minimum reflection at the step for any given characteristic impedance and inner conductors ratio. There is no simple arithmetical formula relating the step-length "a" to these parameters.

Construction

The design is based on a diecast box 4½" × 3½" × 2" deep, with a partition running the length of the box and fitted to form a 2" square cross section into which the trough line is assembled. Any other spacing may be used but this complicates the calculation of the Z_O of the line.

The characteristic impedance of a coaxial line with a cylindrical inner conductor and a square outer is given by

$$Z_O = 138 \log_{10} L/d$$

where L is the length of the side and d is diameter of the inner conductor and

$$L/d \text{ is } > 1.5$$

substitution of 2in in this formula gives d = 0.6in.

DIMENSION a = 0.055" ±0.005"
(CLEARANCE FOR 18 SWG
BRASS TAG - DETAIL B)

MATERIAL...... NOMINAL ½" BORE
COPPER WATER PIPE

TIN DIP

SOFT SOLDER
AND CLEAN

DETAIL B
MATERIAL......
18 SWG BRASS STRIP

ASSEMBLY OF
DETAILS A & B
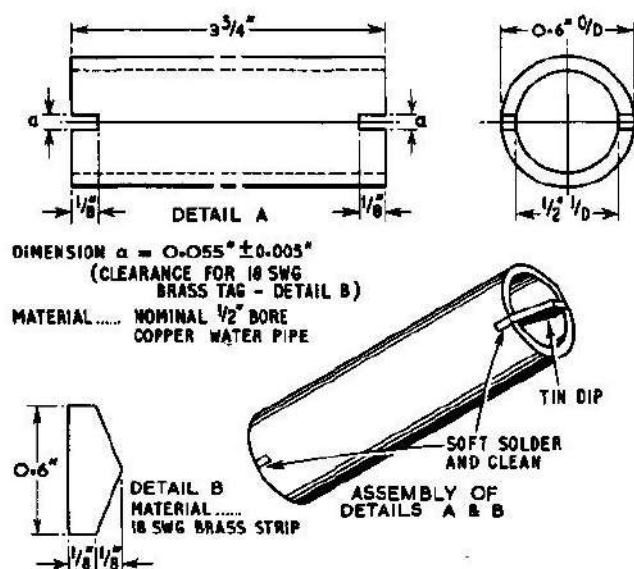
Fig.10.7. Construction of main line inner conductor.

The r.f. connections to the box are made using Amphenol type SO239 coaxial sockets; it is important to use the variety having nylon loaded bakelite insulation (yellow) to avoid distorting the inner of the socket when the line connection is being soldered. These sockets are mounted centrally at each end of the 2" square section of the box, and their spigots are cut down so that the overall dimension from the inside face of the box to the end of the spigot is ¼". The inner conductor, Fig.10.6 detail A, is slotted at each end for a depth of 1/8" and wide enough to accept 18 s.w.g. brass sheet as a tight fit; it is important to ensure that the slots at each end lie in the same plane. The small end pieces, Fig.10.6 detail B, are cut from

18 s.w.g. brass sheet and pushed into the slots at each end as shown. They are then soft soldered in position, the pointed end of each tab tinned, and the surplus solder cleaned off the outside to restore the cylindrical shape at the ends. This inner assembly may then be rested between the spigots of the coaxial sockets, and soft soldered in position, see Fig.10.9A.

The sampling lines are formed from a strip line of 18 s.w.g. brass lying parallel to the partition. The formula for the characteristic impedance of a strip line over an infinite plane is given as:

$$Z_O = 230 \log_{10} 4D/W \ldots\ldots\ldots (iii)$$

where D = distance from plane, W = width of strip and $0.1 < D/W < 1.0$. As already explained it is necessary correctly to terminate the sampling lines in order to preserve the directivity of the instrument, and a characteristic impedance of 100 ohms was selected, based upon the use of available 100 ohm 2 per cent tolerance ½ watt resistors as the terminating loads. This figure substituted in expression (iii) gives a value of D/W = 0.68. This provides a whole possible range of dimensions for the strip line and in order to achieve the required degree of coupling to the main line, a value of D = 0.25" and hence W = 0.375" was chosen by experiment. The sampling lines were made as long as conveniently possible, care being exercised to make them as near physically identical as possible.

The partition is made of 16 s.w.g. aluminium sheet and the sampling lines mounted in the positions shown in Fig.10.9. The spacing of the sampling lines may be trimmed by adjustment at the terminated end when the instrument is being set-up. The partition is
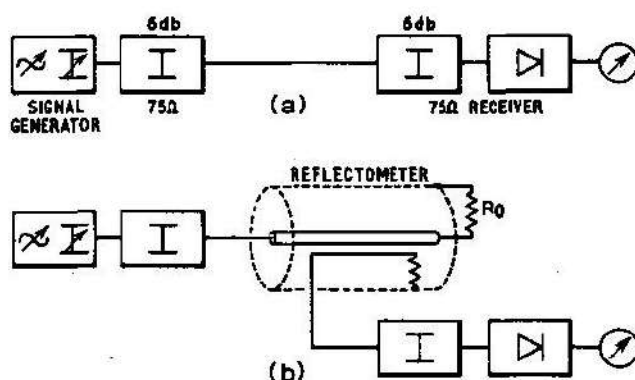
Fig.10.8. Insertion method for measuring coupling.

(a) Set a signal generator attenuator to give an arbitrary deflection K on receiver meter. Note signal generator attenuator setting, say x dB.

(b) Repeat with reflectometer in circuit and readjust signal generator to give same deflection K on receiver meter. Read attenuator setting, say y dB.

(c) Then coupling of main line to sampling line = x - y dB.

assembled with sampling lines, tag board on rear, and all components, before being fitted into the box. Connections from the other side of the partition to the various controls are made up as short flying leads to facilitate this assembly. The position of the various potentiometers and switches is not critical, and some alteration to the suggested layout may be necessary if different sized components are employed. Alternatively there is no objection to extending the d.c. outputs of the sampling line rectifiers to another chassis by means of a three core cable: it is of course, essential that the initial diode rectifiers are located immediately at the feedthrough bushes behind the partition, as shown in Fig.10.9.

Calibration

Accurate calibration of the reflectometer requires a signal generator with calibrated output, a receiver with some form of carrier level meter, and a load of known reflection coefficient suitable for direct connection to either end of the reflectometer test line (this load should be as near matched as possible). The procedure is then as follows:-

(i) Terminate the aerial end of the instrument, and measure the coupling of each sampling line in turn by the insertion method (Fig.10.8). Adjust the sampling line spacing for identical coupling.

(ii) Using the signal generator injecting directly into each sampling diode in turn (with sampling lines disconnected), calibrate the indicating meter in terms of dB relative to the injection voltage for f.s.d. This provides also a check on the match of the diode characteristics of each sampling circuit. These must be matched if the instrument is to read accurately at all transmitted power levels: two diodes at random from the box provided the results quoted for the prototype. The instrument is then calibrated directly in terms of the ratio of backward to forward voltages, expressed in dB, for all transmitted powers, provided it is always adjusted to f.s.d. on the forward position using the set r.f. control (The SET DEFLECTION control should be set, for any particular meter, to such a value as to allow the set R.F. control to function over the whole range of transmitted powers expected).

Many amateurs will, of course, not have available to them the necessary test equipment outlined above. However, this need not detract greatly from the appeal of the instrument, since, even without any calibration at all, the output from the "backward" line will always reduce as the s.w.r. on the main line is reduced; thus the reflectometer may be used quantitatively to indicate best s.w.r. when adjustments are being made to, say, an aerial system.

It is possible, without any test equipment other than a low power transmitter, to make some basic checks on the instrument as follows.
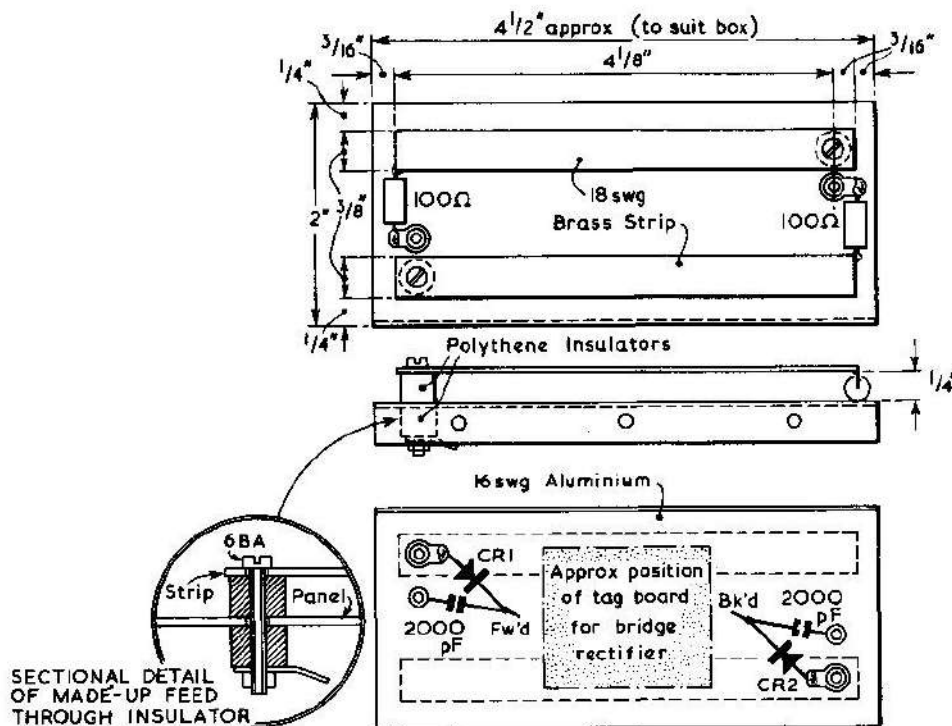
Fig.10.9. Arrangement of strip line on supporting partition.
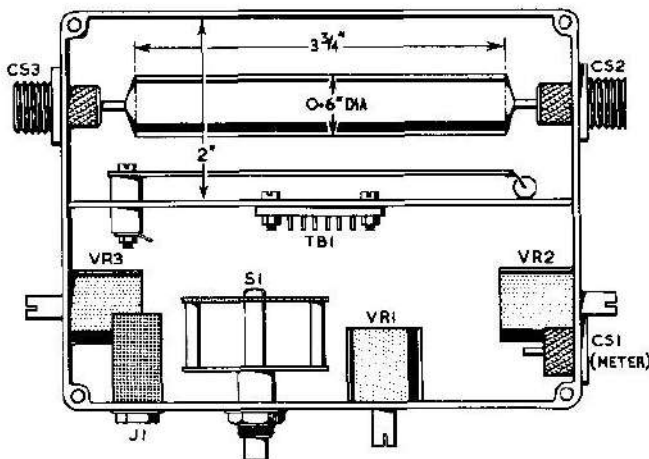
Fig.10.9A. General assembly of reflectometer.
VR1 = 50 K ohms. VR2 = 100 K ohms.
VR3 = 5 K ohms.

With an open circuit on the aerial end of the instrument, vary the power from the transmitter in steps, and take at each level, the "forward" meter readings with the instrument connected normally and then the "backward" meter readings with the instrument reversed. This will check the characteristic of the diodes, and also enable slight adjustments to be made to the sampling lines to equate the coupling. The latter adjustment should be carried out at the normal transmitter power only, for the best performance in practice. Care must be exercised when carrying out such checks, to avoid damaging the p.a. valve of the transmitter through excessive dissipation on no load. Provided that the dimensions given have been followed closely, the errors introduced due to stray differences in the final instrument should not be more than 2 or 3dB. Inspection of the calibration table shows that, for the lower values of s.w.r. such an error results in a very small error in s.w.r. this becoming increasingly worse as the s.w.r. gets larger. Therefore, an uncalibrated but carefully built instrument can be expected to indicate s.w.r. to an accuracy of ± 0.5 up to values of 2:1 becoming as poor as ±1.0 at 4:1. This should be quite adequate for most amateur uses.

The reflectometer described is designed for use in lines having a characteristic impedance of 72 ohms. Errors ensue from its use in lines of any other impedance.

Reflectometer Calibration Table

Meter Reading			
Forward	Backward	Level (db)	S.W.R.*
50	50	0	infinite
43	44	—2	8.8
37	36	—4	4.4
30	29	—6	3.0
23	21	—8	2.3

Reflectometer Calibration Table *(continued)*

Meter Reading			
Forward	Backward	Level (db)	S.W.R.*
18	17	—10	1.92
14	13	—12	1.67
11	10	—14	1.5
8	7	—16	1.37
6	5	—18	1.29
4	3	—20	1.22
3	2	—22	1.16
2	1	—24	1.13

*The s.w.r. column represents the conversion of "backward" meter readings for a "forward" reading of 50. For a given input level, the difference between lines was less than 1db over the full range. Zero level is equivalent to 1V r.m.s. in 100 ohms.

Power Limitations

The sensitivity of the instrument is such as to provide f.s.d. on a 50 µA meter for a carrier power of 5 watts. The upper limit is set by the dissipation in the resistors terminating the sampling line. These are rated at ½ watt and, since the forward line is dissipating power 32dB down on the incident transmitted power, the maximum transmitted power should not exceed 500 watts carrier.

Frequency Range

The performance of the instrument is constant over the 144 MHz band. The sensitivity will fall linearly with the decrease of frequency since the coupling lines are short. The impedance match of the instrument itself will deteriorate with increasing frequency due to the presence of the step discontinuities, and also the variations in the terminating loads on the sampling lines, which will become increasingly reactive.

AN ALTERNATIVE REFLECTOMETER

Another design of the unit just described, which will be found convenient for many purposes and is suitable for use at frequencies above 432 MHz.

Fig. 10.10 represents a section along a coaxial line through which a current I is flowing in the direction indicated on the inner conductor. A loop is inserted through the outer wall into the field inside. One end of this loop is terminated by a resistor R and

the other by an indicator M which may be any type of device, but usually a diode and microammeter. Current is induced into the loop, and the induced current, Im, will be in the opposite direction to that flowing in the inner conductor. In addition, since
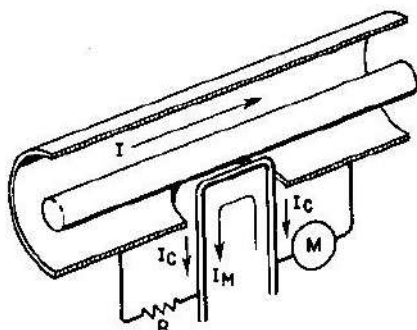
Fig.10.10. Illustration of the arrangement of the directional coupler into the sample section of the coaxial line.

there is capacitance between loop and the inner conductor a current Ic flows in each leg of the loop. In the resistance arm Im and Ic are additive whereas in the other arm they are in opposition.

By adjusting the capacitance C and the mutual inductance M, Im and Ic can be made to cancel, and the indicator will then read zero. If the current I in the inner conductor is reversed, Im reverses and the indicator will then read Im + Ic.
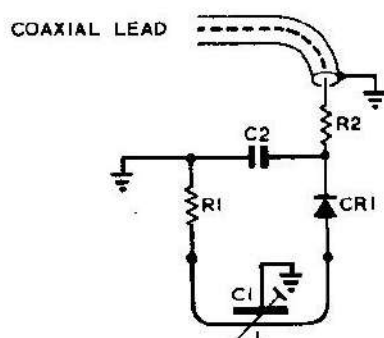
Fig.10.11. Circuit of coupler unit. R1, R2 = 70 ohms ½ watt.

C_1 = ¼" dia. disc on 6BA studding.
C_2 = 47pF ceramic.
D_1 GEX66 or similar
L = coupling loop 20 s.w.g. ½" wide, ⅛" spaced from mounting face.

With the aid of two loops it is then possible to read separately the two currents in the line, from which the reflection co-efficient K can be determined.

Construction of the Line

In order to make coupling to the inner conductor, it is necessary to have a section of the transmission line enlarged to enable the directional
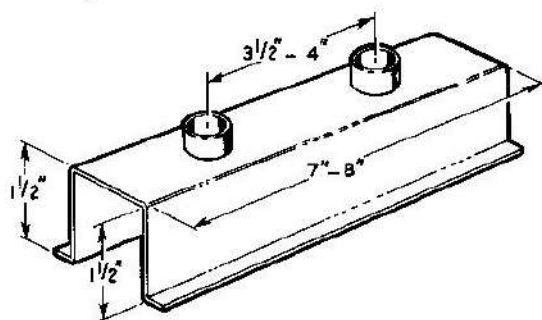
Fig.10.12. General arrangement of enlarged square section line showing position of coupler sockets.

couplers to be inserted. The enlarged section may be either of round or square cross-section but the latter is easier mechanically unless access to a lathe is available. The impedance of the new section of the line must, of course, be of the same impedance as the normal feeder line. The chart on page 10.8 shows the relationship between the inner and outer dimensions for any impedance between 20 and 200 ohms.
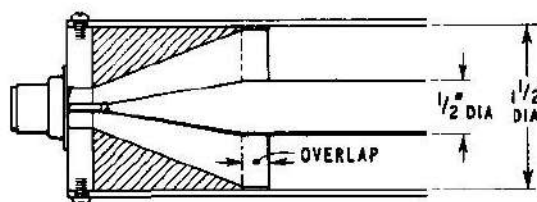
Fig.10.13. Showing end taper of both inner and outer of the section. The inner conductor may be shaped by turning or fitting a thin copper cone rolled and soldered to inner conductor. The outer taper can be made either by fitting blocks in the ends or folding thin sheets into the correct shape. If this latter method is used overlap should be allowed for adequate contact to the outer.

In Fig. 10.12 is shown a suitable general assembly arrangement. In this the bottom is closed by a suitable plate and the ends by plates fixed to all four sides of the square section. In order to preserve the impedance at each end of the enlarged line, tapers of constant D/d must be provided, this is shown in Fig. 10.13.

Detector Head Mounting

The detector heads and their sockets are made from telescope tubes of about ⅞" and 1" diameters, the sockets being ¼" long and spaced approximately 4" apart on top of the square section unit as shown in the drawing. They are soldered into holes cut to fit them and are slotted so that they may be clamped tightly to the detector heads. The clamps may be made from ¼" wide strip brass bent round and held together with a 6BA nut and bolt.
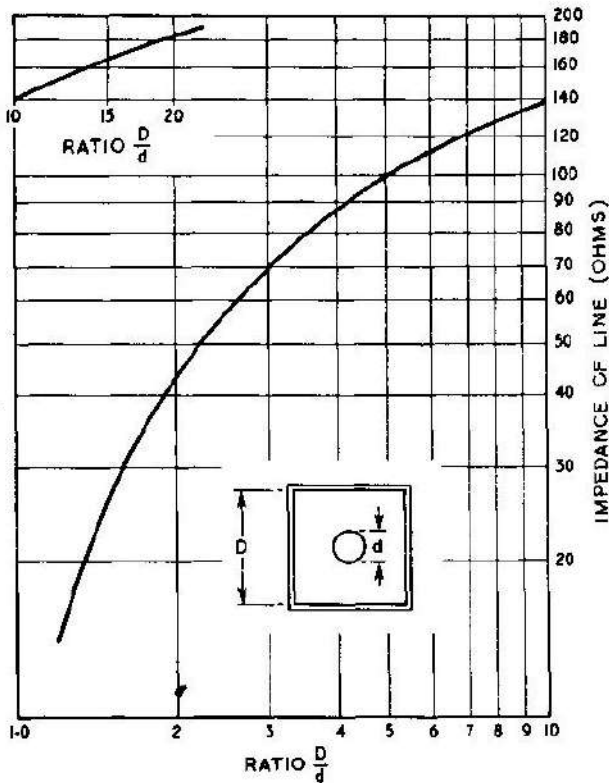
Fig.10.14. Chart showing the ratio of the internal dimensions of a square unit to the external diameter of the centre conductor for use in calculating the dimensions of line sections of required impedance.
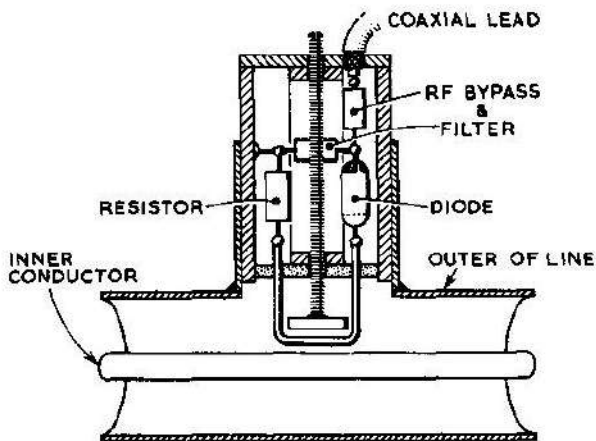
Fig.10.15 Practical design of a suitable directional coupler.

The Detector Heads

Two views of one of the two detector heads - which are identical - are shown in Fig.10.16. The heads are constructed on small frames made of ¼" wide thin brass strip bent into rectangles which fit snugly into the tubes. It is upon these "chassis" that all the components are mounted. The upper end of each head

is closed by a brass disc, the bottom end is made of resin board or similar material. These end plates are held in position by 8BA screws, which are also used for fixing the "chassis" inside the tube. It is important that the heads should be a good fit in their respective tubes so that the whole assembly can be easily pushed together.

The small trimmer condenser (C1 in Fig. 10.11) is made by screwing a piece of 6BA studding right through the rectangle and end plates. A ¼" diameter disc is soldered to one end, the other being slotted to take a screwdriver. The ¼" plate is then used to adjust the capacity between the loop and the "earthy" frame. The coupling loop is made from 20 s.w.g. wire and is ½" wide and ⅜" above the resin-board plate on which it is mounted. The loop is fixed through pairs of small holes ½" apart. Each end of the loop is passed through one hole, folded back through its neighbour and pinched to make a firm anchorage for each leg.
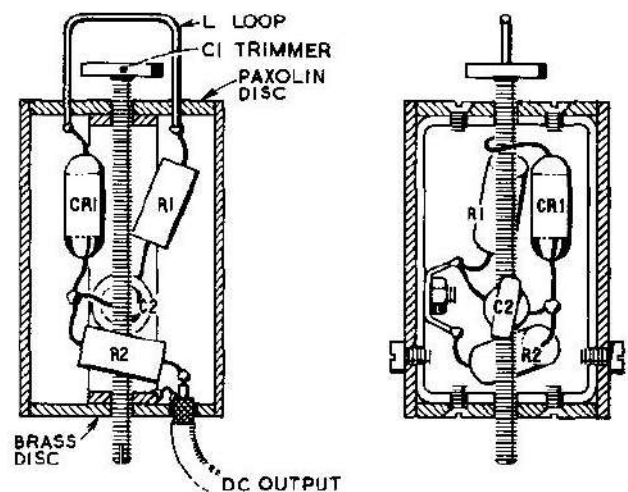
Fig.10.16. Two views at right angles to one another showing the construction of the detector heads.

It will be seen from the circuit of the detector head that one end of the loop terminates in a resistance R1, whilst the other feeds a germanium rectifier type GEX66 the output of which is decoupled by C2 and R2. Television type coaxial cable, terminated with coaxial connectors can be used to feed off this d.c. output. The leads are passed through holes cut in the brass end discs, but care should be taken to anchor the braid securely inside the head by bending a long soldering tag round it and lightly soldering. The tag can then be bolted to the frame.

Terminating Resistance

The initial setting-up process requires that a terminating resistance be used at one end of the unit. It must "look like" a resistance, even at 432 MHz. This means that it must be coaxial and of similar

10.8

impedance to the line. A long piece of coaxial cable, say 100 ft could be used for the purpose, but as such a length is unlikely to be commonly available, a special accessory will be described.
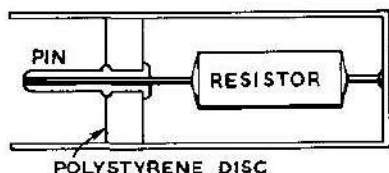
Fig.10.17. Construction of the terminating resistance as described in the text. The diameter of the tube is dependent upon the line impedance and the diameter of the resistor.

An ordinary composition resistor with wire ends, even if it is the correct nominal resistance, may be as much as 50 per cent in error at 432 MHz owing to the nature of the composition, and the high inductive reactance of the leads.

High stability resistors of the cracked carbon type have however, been found quite satisfactory: Welwyn type SA3623 or Painton type 74 are suitable. The resistor selected should have the same d.c. value as the required terminating impedance. It is used as the "inner conductor" of a concentric line. The diameter ratio of the line must, however, be made for a 30 to 40 per cent lower impedance than the required value, because a line in which the inner conductor is all resistance has a somewhat inductive impedance. The smaller outer diameter, by providing greater capacity, compensates for this effect. For 70 ohms the ratio is 2:1. Using the types of resistor specified above, the 70 ohm assembly (¾") diameter fits the Belling-Lee socket. The construction of these terminations is illustrated in Fig. 10.17. When made as shown, the terminations are within 2 to 3 per cent of the correct value at 432 MHz. However, the wire leads to the resistors should be as short as possible. In the 100 ohm type, the pin should be mounted in a polystyrene disc made to fit tightly in the tube.

Setting up Procedure

Setting-up the instrument is quite simple and only requires a low-power oscillator or transmitter that can be modulated, preferably with a continuous note. As the terminating resistor is rated at only one
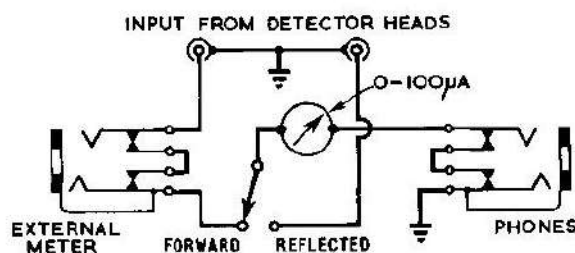
Fig.10.18. Circuit diagram of the meter switching for a mounted reflectometer.

watt, care should be taken not to overheat it. After setting up, full power may, of course, be passed through the Reflectometer.

The detector heads should be inserted so that the ends of the loops are about half-way between the inner and outer conductors of the line, using the clamps as depth controls. The terminating resistance is then plugged into one end and the modulated 432 MHz signal fed into the other.

The signal will be heard in headphones plugged into the 'phone jack and switched to the appropriate head. The head nearest the terminating resistor is next adjusted by simultaneously rotating it in its mounting and adjusting the trimmer until a sharp minimum is observed. This null point is quite well defined. The head should be securely fixed in position by its clamp. The input and output connections are then changed over, the 'phones switched to the other head and the procedure repeated.

This procedure can be carried out on 144 MHz but the settings may need slight adjustment for 432 MHz.

Use as a Power Meter

As the germanium detector has a reasonably good square law response, the d.c. output from the heads is proportional to the power in the line. Accurate calibration as a power meter is, however, a difficult task; nevertheless, it is possible to give a rough idea of what meter readings mean.

In the 1½" square 70 ohm line, if the loop is ¼" inside the line cavity the meter will read approximately 100μA for one watt or 10mA for 100 watts of r.f. power.

With the loop halfway between the inner and the outer these readings will be about doubled. As the voltage applied to the coupler is proportional to the frequency, the d.c. output will be proportional to the square of the frequency, i.e. when the instrument is used at 144 MHz the output will be about one tenth of that for 432 MHz.

THE MEASUREMENT OF R.F. POWER

There are a number of methods of power measurement capable of providing sufficient accuracy for all practical purposes, up to 1000 MHz. Where power monitoring is required as distinct from actual measurement the methods outlined can be used at still higher frequencies but with rather less accuracy.

Whenever possible the measurements should be made at relatively low impedance of say, 50 or 75 ohms. This is suitable for most transmitter or exciter outputs but at higher impedances some correction factors may become necessary.

In making power output measurements it should be remembered that *all* outputs including harmonics and spurious emissions if any are present, will be

included, and if it is required to measure only the power of the wanted signal, then some form of filtering will be required. For most purposes this is not necessary, but some thought should be given to this aspect when dealing with varactor diodes or other semiconductors.

THE THERMOCOUPLE POWER METER

Basically this is an extension of the familiar thermocouple in r.f. ammeters used for reading aerial current in l.f. transmitters.

The main component is a thermal converter which consists of a heating element, to the midpoint of which is attached a thermocouple. This is fixed by means of a suitable material to give good heat conduction, but still be an electrical insulator. Suitable units for use at v.h.f. and u.h.f. are normally enclosed in an evacuated envelope, the construction being such that stray reactive components are minimised. Their general form is similar to that of the old acorn valve, see Fig. 10.19.
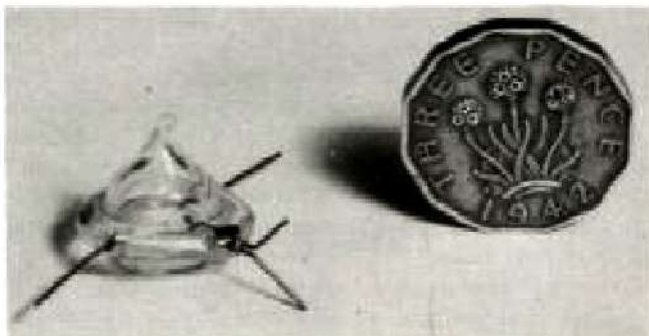
Fig.10.19. Typical thermal converter unit.

In operation, when an r.f. current passed through the heater element a thermal e.m.f. is produced in the thermocouple, the magnitude of which will depend on the value of the r.f. current. For calibration, if the r.f. current is replaced by a d.c. or low frequency a.c. (mains frequency) current and its value adjusted to give the same thermal e.m.f. it can be shown that the r.f. current has an r.m.s. value equal to that of the calibration current. Therefore, since the current can be measured accurately, this, together with the load resistance, will enable an accurate measurement of the total r.f. power to be made.

Thermal converters are available in a wide range of heater ratings and typical types are shown in the following table.

Table of Typical Thermal Converters

Current Range	Heater Resistance	Thermocouple Resistance	Thermal e.m.f. (open circuit)
mA	ohms	ohms	mV
10	25	8	7
10	50	8	12
25	10	4	7
25	25	4	16—18
50	3	4	7
50	8	4	12—15
100	1.5	4	7
100	2	4	12—15
500	0.3	4	7
500	0.45	4	12—15
1000	0.15	4	7
1000	0.25	4	12—15

Available from Ormandy and Stollery Ltd,
3, Victoria Place,
Brightlingsea.
Essex.

The actual value of the heater resistance will, of course, vary with the temperature of the element and hence the current flowing. Therefore, in order to avoid a variation of the load resistance and as a result a variation of the matching impedance seen by the source, it is desirable to provide a padding resistance

(b)
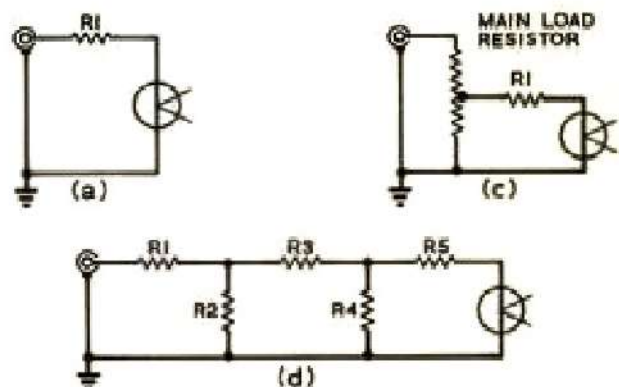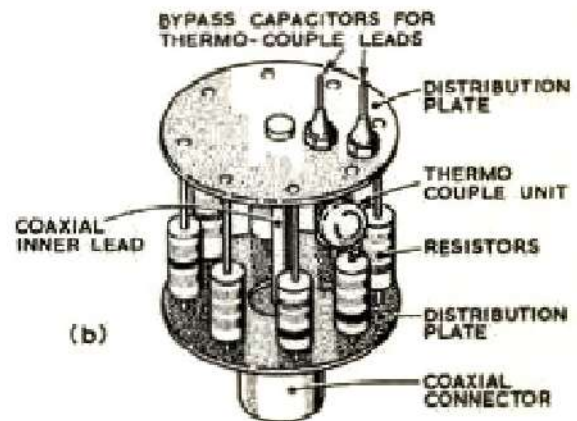
Fig.10.20 Methods of use of thermal converter for higher powers.

to swamp these changes. As this type of instrument is very sensitive, for most amateur purposes some such arrangement will be needed to accommodate the power level to be measured. Suitable arrangements are shown diagramatically in Fig. 10.20.

In Fig. 10.20 (a) RI is chosen so that the combination of the heater element resistance and the external resistance (R1) is equal to the required load resistance.

When higher powers are to be measured or improved high frequency performance is needed the arrangement shown in Fig. 10.20 (b) should be adopted. In this the load resistance is made up of a number of small resistors arranged in parallel in a circular formation and connected to two plates, one fastened on to the outer of the coaxial connector and the other to the inner. The measuring thermocouple is connected in series with one of the resistors or in some circumstances the thermocouple can be connected in series with the centre connection for all the resistors.

Another alternative arrangement suitable for higher power is shown in Fig. 10.20 (c), in this case a single large resistor is mounted in a suitable housing and the thermocouple connected to a tapping near the earthy end, such as about 5 to 10% of the total value. At higher frequencies still the use of a multi-section attenuator will generally provide a better match, this is shown in Fig. 10.20 (d). In this the components for the individual sections are chosen to provide constant dissipation as opposed to the usual requirement of constant attenuation. The construction of this type of attenuator should be such as to maintain a constant impedance, the individual resistor units should be of the film type, R1, R3 and R5 being rod type and R2 and R4 of the disc type.

The accuracy of this sytem will be limited mainly by the thermal converter whose reactive components and skin effect will be increasingly significant above 500 MHz. The connections to the thermal converter should, of course, be as short as possible, but bends or soldered joints should not be closer to the vacuum seal than about four to five mm. and the thermocouple leads should be decoupled by feedthrough capacitors.

The thermal e.m.f. will be only a few millivolts (see the table on page 10.10) and therefore, in order that a robust type of indicating instrument can be used, a transistor d.c. amplifier should be employed.

The Thermistor Bolometer Power Meter

Thermistors and Bolometers are made in a wide range of type and form depending on the precise purpose for which they are required. A typical form is shown in Fig. 10.21. The small bead which is the active component, is usually a compound of metallic oxides possessing a negative temperature coefficient of resistance. In use it is preferable to embody it in a constant resistance, made to avoid changes in the impedance match which would result in errors.
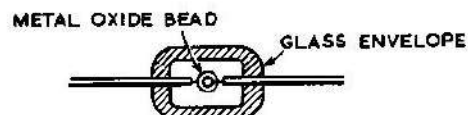
Fig. 10.21. Construction of a microwave thermistor.

In Fig 10.22 it will be seen that the thermistor R2 forms one arm of a conventional Wheatstone bridge, the resistors R3 and R4 are equal and form the ratio arms, while R1 is made equal to the desired thermistor resistance in a typical arrangement. The current to the bridge is made variable by VR1 and is indicated by M1. The magnitude of the current is adjusted until the resistance of the thermistor is driven *down* to the required value, as shown by the null indicator which may be either a centre zero meter or a
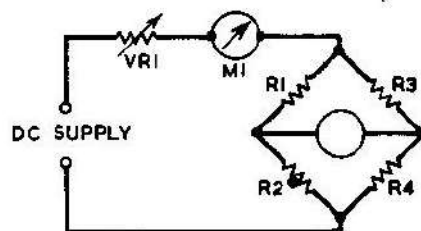
Fig. 10.22. A simple thermistor bridge.

d.c. amplifier connected to a less sensitive meter.

If now an r.f. current (power) is applied to the thermistor, its temperature will rise and the bridge become unbalanced. If the current in the bridge arms is now adjusted until balance is restored, the change in d.c. power will be the same as the applied r.f. power, assuming the conversion to heat is the same in both cases. The substituted power may be calculated from

$$P = (I_1{}^2 - I_2{}^2) R^2$$

where $P =$ the substituted d.c. power
$I_1 =$ current through the thermistor at balance with no r.f. applied.
$I_2 =$ current through the thermistor at balance with r.f. applied.

The value of R3 and R4 should be made high compared with R1 and R2 and the current may be read directly from M1. Lower values of R3 and R4, should be of simple ratio to the values of R1 and R2 in order to simplify calculations.
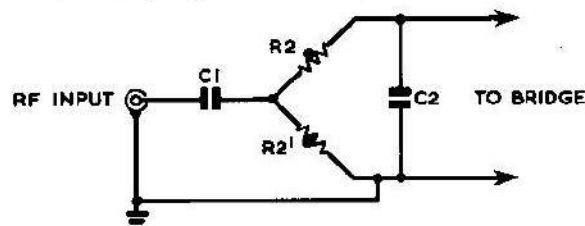
Fig. 10.23. A coaxial thermistor mount.

Typical thermistors have an operating resistance of 50 to 500 ohms and for amateur purposes the method shown in Fig. 10.23 is preferred.

In this case r.f. power is applied to the two thermistors R2 and R2^1 via a capacitor C1 which provides d.c. isolation for the substitution power. For purposes of r.f. R2 and R2^1 are connected in parallel across the input by the capacitor C2, which together with C1 must have a low reactance compared with the impedance of the line carrying the r.f. power, together with low loss at the operating frequency. The resistors R2 and R2^1 are in series as far as the d.c. substitution current is concerned. Most thermistors are of low power rating, usually in the region of 10mW, and should be used with the attenuator shown in Fig. 10.20 when power in the 1 watt range is required.

Bolometers are similar in construction to thermistors, a fine wire replacing the metal oxide bead. However, they have a positive temperature coefficient of resistance and may be used in similar circuits to those used for thermistors. Their temperature coefficient is, however, lower than thermistors and they are therefore less sensitive.

For amateur purposes a bulb may be used in place of a proper bolometer and a festoon type is very suitable.

Lamps as r.f. loads for power measurements.

Probably the most inexpensive method of measuring r.f. power output is to use a two lamp comparative method, where one lamp is connected to the equipment and the other to a variable source of power. It is essential that both the lamps should be of the same type and power rating and it is desirable that before use the consumption data is recorded for both lamps so that allowances may be made for any differences. The actual difference between two lamps of the same type and rating is unlikely to be sufficient to make calibration essential for practical purposes but is necessary for those who wish to be as accurate as possible.

As may be seen from the typical curves shown in Fig. 10.24 the light output from a lamp varies rapidly with changes of voltage or current and it is important therefore that direct comparison of the output is made. This is best arranged by fitting the two lamps side by side into a box having a ground glass screen over the lamps and with a dividing panel between the two lamps. The two lamps will then each illuminate the two separate halves of the ground glass screen.

Matching is easier using a ground glass screen instead of viewing the lamps directly and this is particularly true at the higher filament temperatures. When the lamps are enclosed in the box there is also much less interference by stray room lighting. A suitable arrangement is shown in Fig. 10.25. A variable power supply for the standard lamp is required and either a voltmeter and ammeter or a voltmeter and a
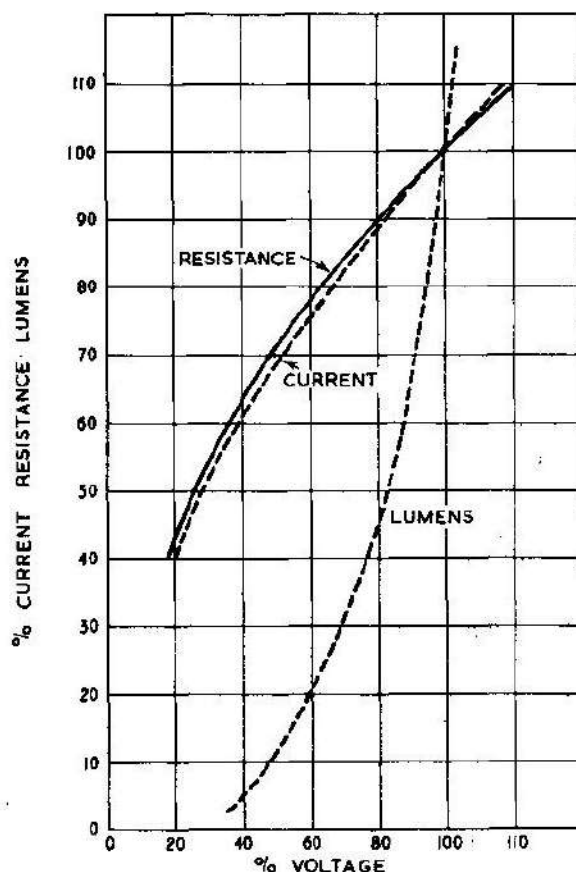
Fig.10.24. Characteristics of gas filled lamps.

calibration curve of the lamp is necessary for making measurements.

If a Variac or other iron-cored device is used to control the applied a.c. voltage instead of a resistor it will be necessary to take account of wave-form error in moving iron voltmeters and ammeters which read true r.m.s. values. This point may be a little academic in view of the inherent inaccuracies of this form of power measurement, but it may be of interest to those seeking the highest possible accuracy.
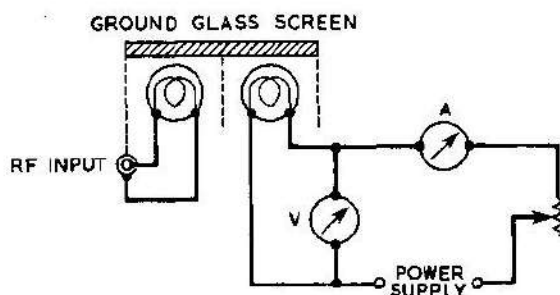
Fig.10.25. Diagram of a suitable two lamp comparator.

Alternative Single Lamp Method

In this case the lamp is supplied with power either from the equipment or a measured source, the

light output being observed by means of a selenium photo cell and micro-ammeter or a photographic exposure meter. The procedure is then as follows:-

(i)   First supply r.f. power to the lamp from the transmitter.

(ii) Take a reading of the light output using the indicator (selenium photo cell and microammeter or exposure meter) at a suitable and measured distance.

(iii) Next, supply the lamp with a measured source of power, with the indicator at the same distance as before, increasing the light output until the indicator reads the same as in step (ii). The actual power output can then be computed from the calibration data of the lamp being used.

Lamps cannot, of course, be regarded as non-inductive and some care is necessary in making connections to keep the inductance as low as possible, particularly when using them at v.h.f. and u.h.f. It is worth noting that the screw and single contact type cap is more readily fitted to coaxial plugs or sockets than the double contact bayonet type.

It is difficult to get a lamp which will be equal to the normal load resistance of 50 or 70/80 ohms. And it should be remembered that its characteristic varies considerably with temperature.

Nevertheless the output indicated by this means is reliable, but retuning of the output is likely to be required when they are substituted.

Dummy Loads

The production of a reliable dummy load does not present any real problem and it may be of two general forms:

(a) one large resistance

(b) a multiple of small, one or two watt resistors arranged in a cylindrical form.

In either case the whole unit should be enclosed in a metal screen.

In the case of (a) the resistor element forms the outer of the coaxial line and to preserve its impedance the centre conductor from the free end to the coaxial socket should be of appropriate diameter. This may be calculated from:

$$Z_0 = 138 \log_{10} \frac{D}{d}$$

where D is outside diameter of resistor element and d is diameter of inner conductor

Example, for a 75 ohm 1" diameter resistor the inner conductor should be 4.67mm or approximately 0.3".

In the case of (b) a number of parallel connected resistors to make up the wanted line impedance which might be 50 or 70 ohms, can be assembled as shown in Fig. 10.26. As described earlier under power measurement, a measuring device may be connected in series with any one of the resistors to indicate power, but alternatively a bulb may be used instead as an indicator.
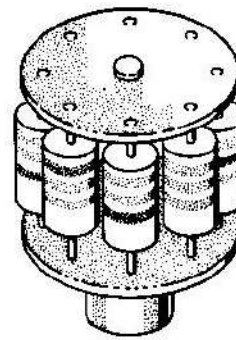
Fig.10.26. A multi-resistor dummy load.
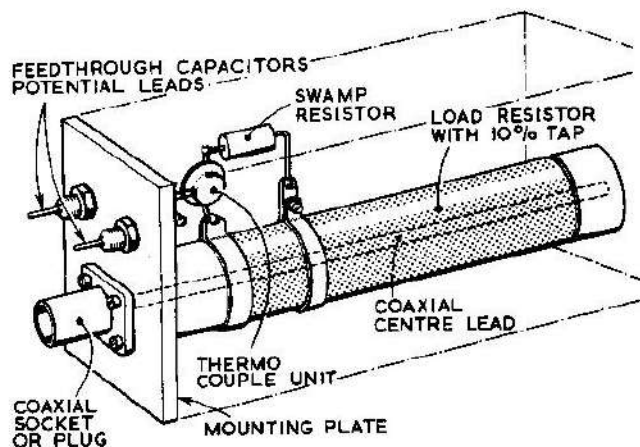
Fig.10.27. A suggested mechanical arrangement for a power meter using a thermal converter connected to a tap on the load resistor.

In either form, these dummy loads may be considerably increased in power rating by immersion in a suitable liquid. Transformer oil is recommended, but ordinary oil may be used.
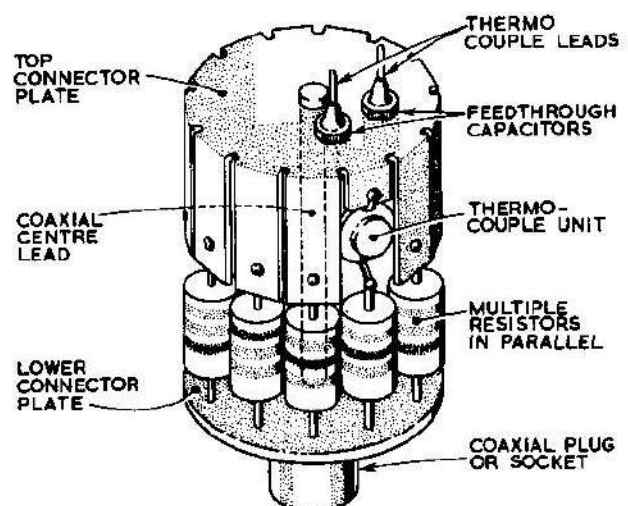
Fig.10.28. An alternative arrangement for a power meter using a number of resistors in parallel with the thermal converter unit connected to one of the resistors.

An R.F. Bridge

This is a Wheatstone type bridge suitable for use on frequencies up to 432 MHz. It requires care in the construction to ensure absolute symmetry of the component layout together with the use of miniature components and matched pairs where necessary. This will then give a bridge which has an accuracy of the order of 1% of full scale deflection which is good enough for practical purposes, and should fulfil most amateur needs.

Fig. 10.30. Component layout of r.f. bridge.

Fig. 10.29. Circuit of an r.f. bridge.

R1 R2 100 ohms ± 1% metal oxide type
R3 R4 4.7 K ohms ± 1% metal oxide type
*R5 R6 Terminating resistor consisting of two 150 ohms in parallel for 75 ohms,
C1, C2 0.001 μf ceramic disc,
C3, C4, C6, C7, C8, C9 0.01 μf ceramic disc,
CR1 GEX66 or CV2290,
VR1 50k ohms miniature variable resistor,
VR2 2.5k ohms miniature variable resistor,
Z known—75 ohm resistor fitted into a coaxial plug;
*If 50 ohms is required then R5—R6 will be 2 x 100 ohms—Z known = 50 ohms.

The circuit of the bridge shown in Fig. 10.29 is given in the same form as the component layout shown in the general drawing Fig. 10.30 and is self-explanatory.

The whole unit, which uses a separate meter, is built in to the smallest size standard diecast box.

Operation

The method of operation with this bridge is to start with the value of Z known and adjust the length of the stubs, or aerial matching device until a balance is achieved.

This unit will be found ideal for cutting coaxial cable to ¼ and ½ wave lengths, where Z known is either an open circuit or a short circuit. The accuracy that can be achieved will be better than 1% of full scale deflection of the meter.

GRID DIP OSCILLATORS
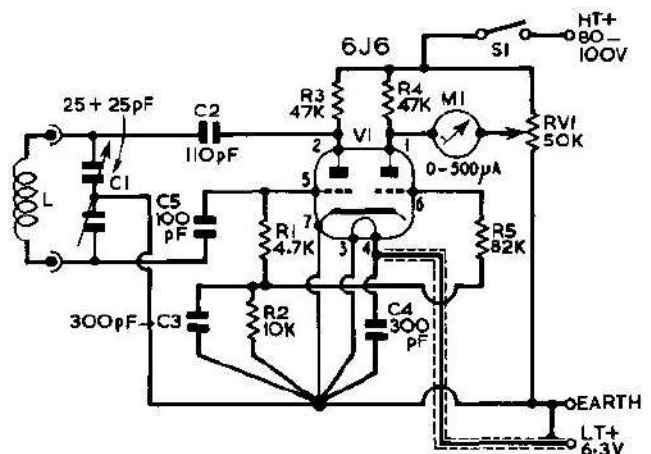
Fig. 10.31. Circuit diagram of the g.d.o. for 70-160 MHz. As explained in the text, the range can be expanded with additional coils to cover 23 to 220 MHz.

A grid dip oscillator, or g.d.o. as it is usually called is an essential piece of v.h.f. - u.h.f. equipment. It depends for its function on the movement of a meter connected into a tunable oscillator. When this oscillator is loosely coupled to an unknown tuned circuit, the meter will indicate when the g.d.o. tuning passes through resonance.

Construction

The first of the examples to be described covers 23 to 220 MHz and is suitable for the exciter stages and amplifiers of transmitters down to the 2m band. As shown in the circuit a 6J6 (ECC91) is used as the oscillator and the second half of the valve is used as a d.c. amplifier so that this, in fact, gives a rise in grid current instead of a dip. Only brief constructional

Coil B — 18 SWG ENAMELLED WIRE — 3/8" I/D
Coil A — 16 SWG ENAMELLED OR BARE WIRE
Coil F
UX5 PIN BASE (807)
1st TURN
1 3/8" DIA

Coil C .... 2½ TURNS CLOSEWOUND
D .... 3½ TURNS CLOSEWOUND
E .... 7½ TURNS SPACED ONE WIRE DIAMETER

Fig. 10.33. Construction of the coils.

(a)   (b)
L
C2 C1 C5
RV1
S1
V1
METER
HT & LT INPUT SOCKETS

(c) SUB-CHASSIS DETAILS
C2 C3 R2 C4 R1 R5 R3 R4

Fig. 10.32. Front panel and arrangement of components within the diecast box.

details are given as a guide and the frequency range curves are typical only. Calibration of individual units is necessary for reliability. CI is a split stator capacitor, rather than a butterfly type, to give a 180 deg. scale.

HAIRPIN F
1½ TURN COIL A
6½ TURN COIL B
2½ TURN COIL C
3½ TURN COIL D
7½ TURN COIL F

Fig. 10.34. Approximate calibration curves for the coils shown in Fig. 10.33.

A Grid Dip Oscillator
For The 70cm Band

This is a single range g.d.o. covering a frequency range of 390 to 450 MHz suitable for the alignment of 70cm amplifiers, and local oscillator chain final multipliers for i.f.'s up to about 40 MHz below the signal frequency. The tuned circuit consists of L and C and is positioned so that the maximum amount of the inductance projects from the screened box to allow easy coupling to the circuit being adjusted. The
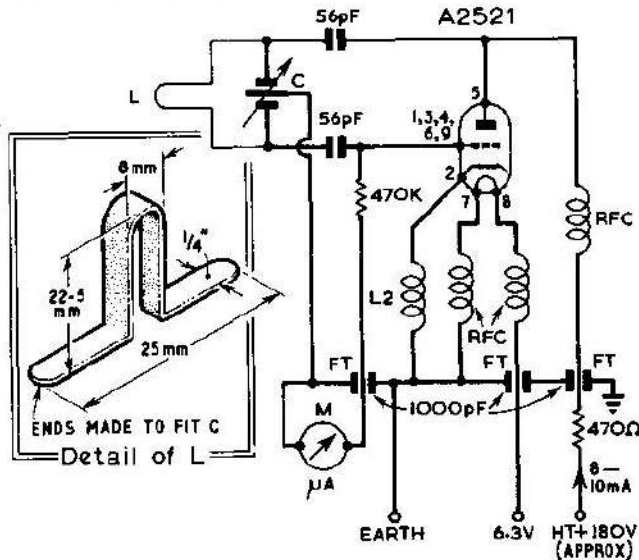
Fig. 10.35. A g.d.o. for 70cm.
C is split stator capacitor with a 6.5pF swing.
r.f.c. = 18 turns 26 s.w.g. on ½ watt resistor.
L2 = 11 turns 26 s.w.g. on ½ watt resistor.

valve used as the oscillator should be one of the high performance types, and in the prototype an A2521 was used. Alternative types such as an EC88/6DL4, A1714/CV408 and 6AM4 may be used if allowance is made for the different base connections and/or inter electrode capacities.

Calibration is best carried out using Lecher wires if a calibrated source is not available and with care a reliable result can be obtained, certainly better than 0.5%.

A Transistor Dip Oscillator

This oscillator, covering 3.5 to 150 MHz is suitable for tuning up all stages of receivers and transmitters for 4m and 2m, as well as the exciter stages of transmitters for the other bands.

In order to provide consistent operation during the life of the battery the oscillator and its d.c. amplifier are fed by a fixed voltage set by a zener diode. The circuit is basically a grounded emitter oscillator followed by a d.c. amplifier. A switch (S1) is included to cut off the oscillator and allow the unit to be used as a sensitive absorption wavemeter.

The construction of this unit is illustrated in Fig. 10.37. A box of 6¾" x 2⅛" x 1¾" with a flanged lid is a convenient size as a container. As with all units of this type, care should be taken to ensure that the oscillator circuit is wired rigidly and with as short leads as possible so that the highest frequency is attained without difficulty. The coil plugs and sockets are of the coventional three pin battery type but any suitable alternative may, of course, be used.

COIL DETAILS FOR THE TRANSISTOR DIP OSCILLATOR

Coil	Frequency Range MHz	No of turns	Wire Size s.w.g.	Dia.	Length	Value of Cf in pF	Type
A	3.4-8.0	27	22	⅞"	⅞"	2200+680	(a)
B	6.7-16	13	22	⅞"	½"	3,300	(a)
C	13.5-34	6	22	⅞"	⅞"	2,200	(a)
D	33-85	3	18	¾"	½"	330	(b)
E	50-150	'U'	18	see 10.38(c)	-	68	(c)
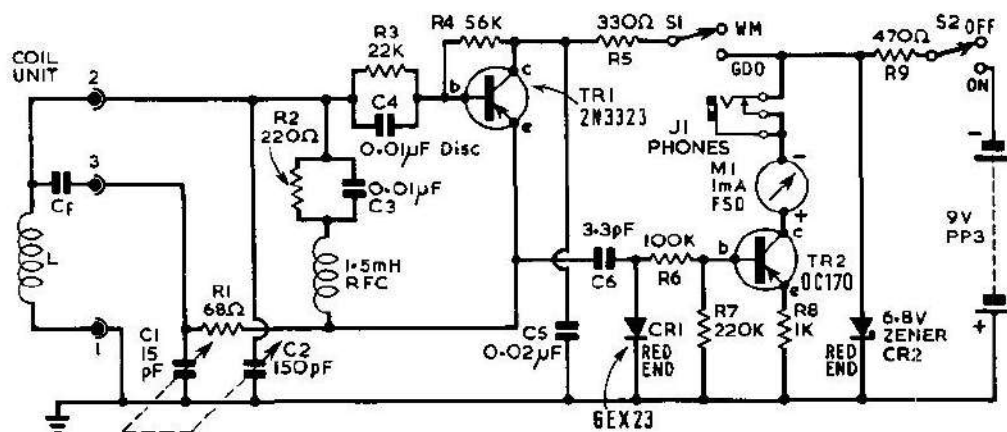
Fig.10.36. Circuit of the transistor dip oscillator.
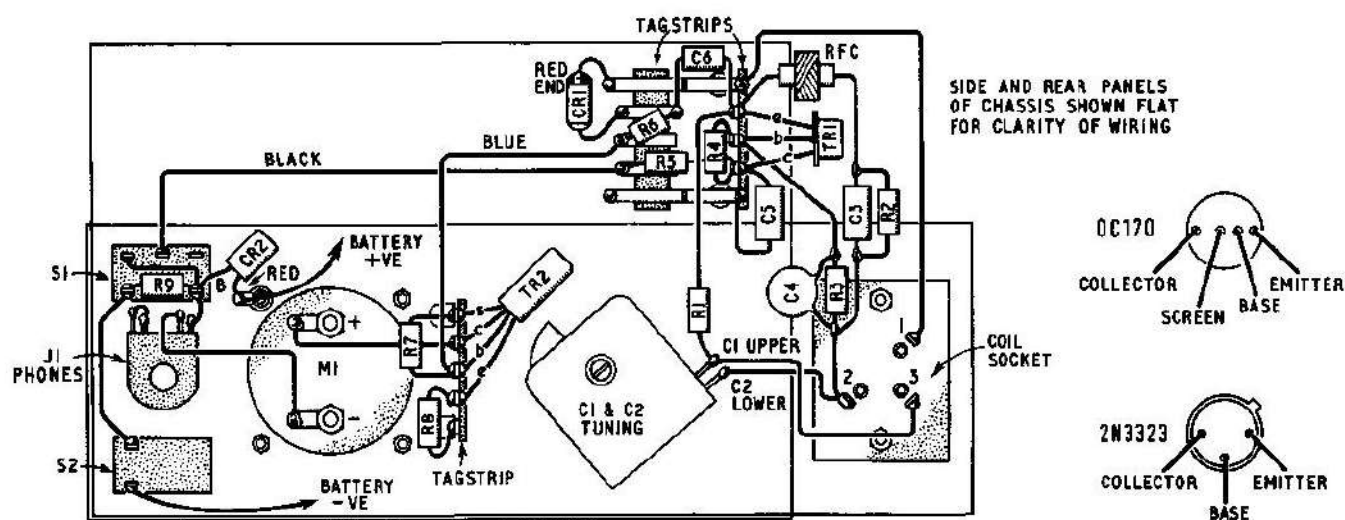
Fig.10.37. Wiring and component layout for the dip oscillator.
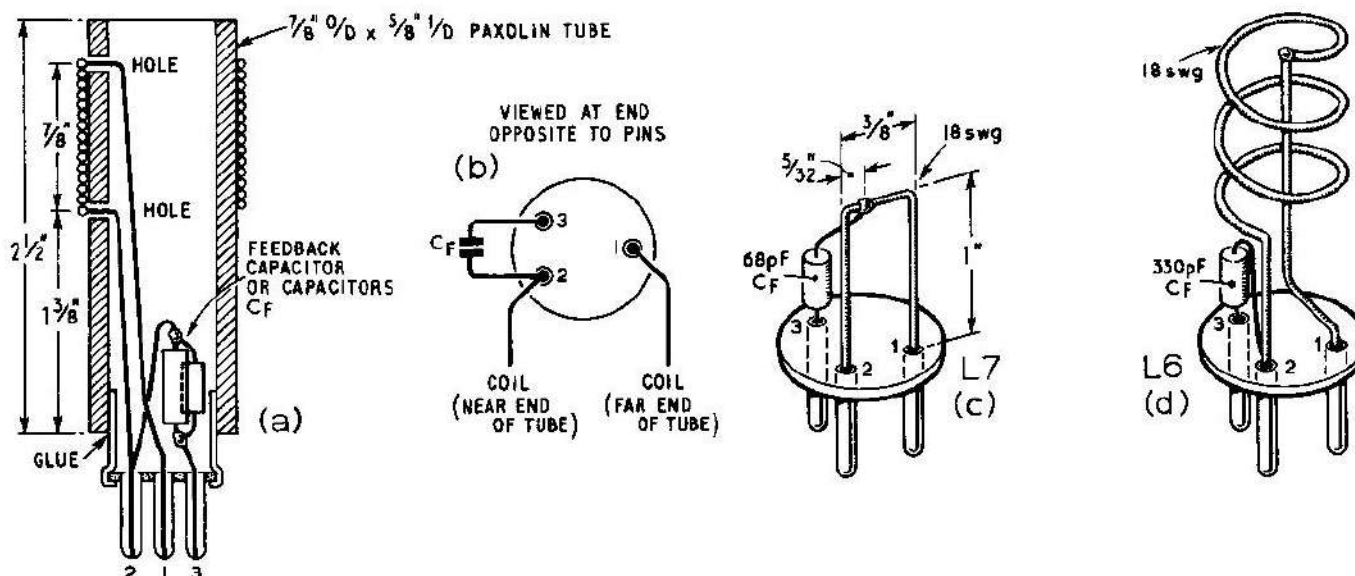
Fig.10.38. The coil assemblies. The five lower ranges are wound on paxolin tubes, while ranges 6 and 7 are air wound and self-supporting.
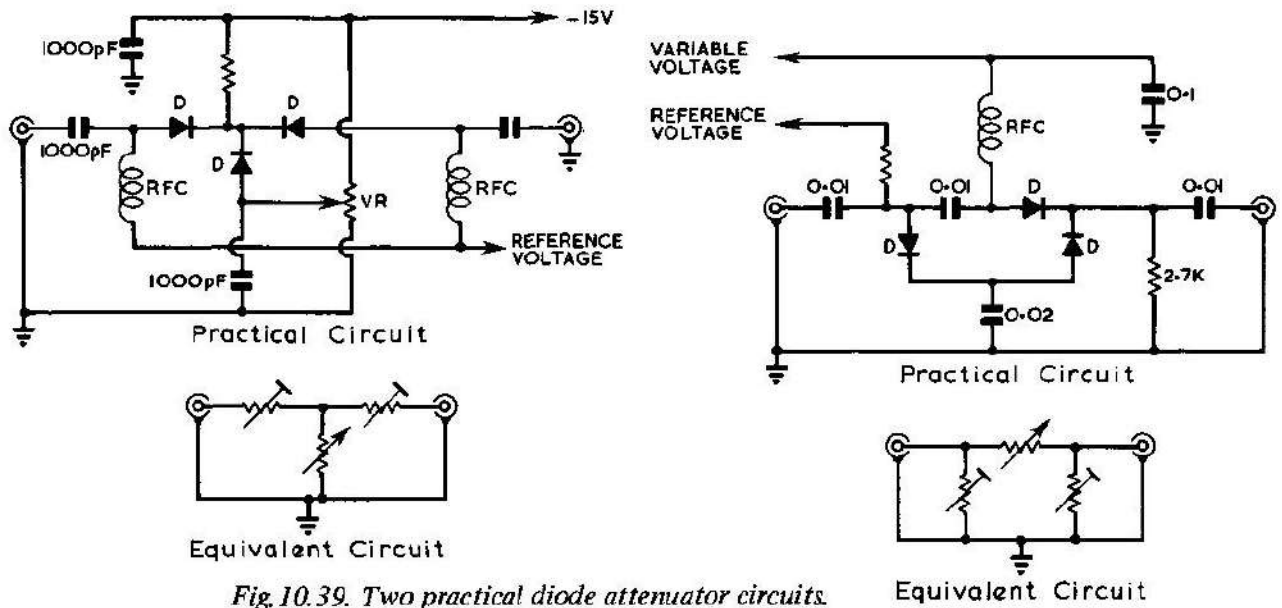
Fig. 10.39. Two practical diode attenuator circuits.

Diode Attenuators

Another application for diodes of the type normally used for switching purposes is in attenuator circuits. The effect is achieved by using a variable resistance to bias the diodes to a suitable value to obtain the degree of attenuation required.

The reference voltage must be set to a suitable value depending on the diodes in use and the value of the preset resistor arms in the particular circuit.

The resistance of the diode will vary from a few ohms when forward biased, to several thousand ohms when reverse biased.

Attenuators of this type are suitable for a number of applications, such as in the receiver input to prevent cross modulation from strong local signals, or the adjustment of exciter power to a required level.

Attenuation of from one or two dB up to 30 or 40dB can readily be obtained from either of the circuits shown, by the suitable setting of the variable resistor VR in the left-hand drawing, or the voltage in that on the right.

Noise Generators

The most satisfactory method of assessing the performance of a receiving system, is to measure the noise factor. For the two lower frequency bands, 4 and 2 metres, a simple thermionic noise diode such as
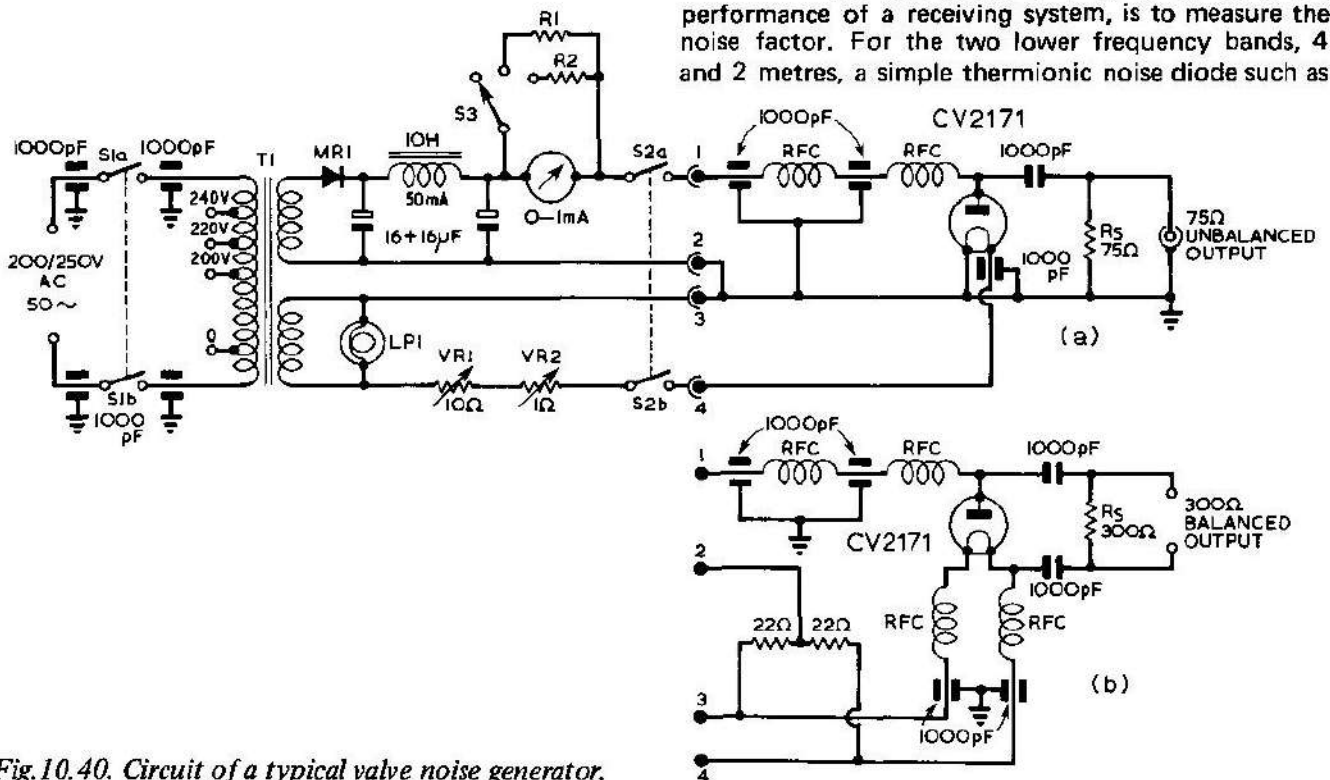
Fig. 10.40. Circuit of a typical valve noise generator.
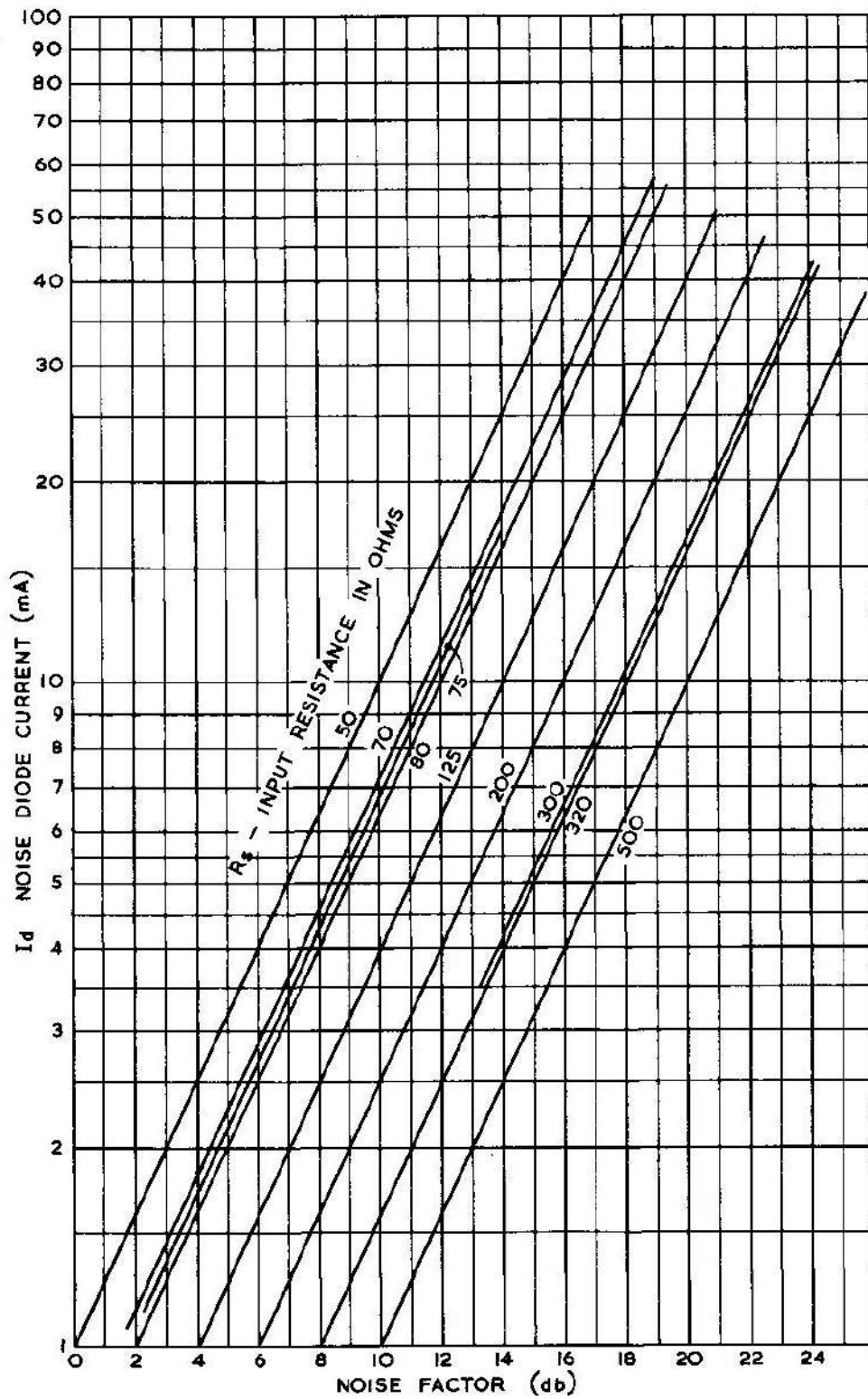
Fig.10.41. Chart showing noise factor for various values of source impedances.

the A2087/CV2171 can be used. Although at 70cm the same arrangement may be used to give an indication, a more suitable type is the coaxial diode type CV2341. This diode may be used at higher frequencies but for accurate readings it is necessary to standardise it by reference to a gas discharge source. The circuit given is for a complete noise generator showing separate outputs for (A) coaxial, (B) balanced, receiver inputs. In these interchangeable heads the source resistor Rs should be suitable for the input impedance of the receiver, or the Z_0 of any interconnecting coax.

In use, noise output from such a generator can readily be calculated but for convenience a chart is given showing the noise factor for various diode currents at selected typical source resistance values. The noise diode head units are best constructed so that they can be plugged directly into the receiver (or converter) aerial input socket.

The method of making measurements is to adjust the receiver gain controls until the amplified noise signal produced by the source resistance Rs and the receiver noise produce a suitable deflection on a meter connected to the output This may conveniently take
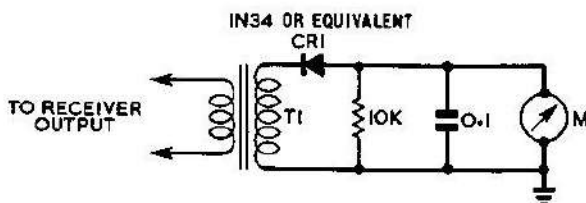
Fig.10.42. Circuit of a suitable output indicator. T1 is a small audio output transformer. CR1 is a IN34 or equivalent.

the form of an a.c. voltmeter connected across the loudspeaker. A suitable deflection might be 0.5 volt. This initial setting up is done with the diode filament switched off, and any a.g.c. in the receiver chain disabled.

Having set the 'zero' conditions, the noise diode filament is switched on and slowly increased in temperature by reducing the variable resistance in its circuit until the output power of the receiver is doubled. If a voltmeter is being used as an indicator then the volts will need to rise by $\sqrt{2}$ times. Now read the diode anode current needed for this operation. The noise factor can be calculated from the following formula taking the normal room temperature as 290 deg. K.

$$NF \text{ (in dB)} = 10\log_{10}\frac{20 \text{ Id Rs}}{1000}$$

where   Id   =   diode anode current in milliamps

            Rs   =   source resistor in ohms

If the temperature is other than 290 deg.K, this may be corrected by the formula:

$$NF \text{ (at } 290^{\circ}k) = NF \text{ at different temperature} \times \frac{290-T}{290}$$

The output indicator could be calibrated so that the meter M is scaled with 3dB (2 to 1 power) points. This will enable the output to be set at various points and if tests are made at several levels it will enable some assessment of the linearity of the receiver to be made. Care must be taken though, that no overloading takes place anywhere in the receiver chain. Some improvement in linearity may be obtained by switching on the b.f.o. as this sometimes improves performance of some types of detector.
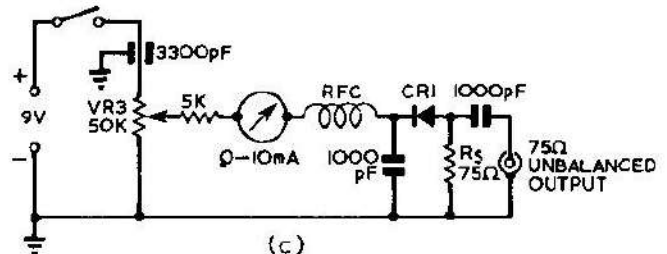
Fig.10.43. A silicon diode noise generator. The diode could be IN21, GD8, or equivalent. M is 0–10mA. It should be noted that this type of noise generator is suitable only for comparative noise measurements.

A circuit is given for a simple type of noise generator, using a silicon diode. This type is only useful for comparative measurements or as a noise source. It cannot be used to measure the noise factor of a receiver.

100 kHz Crystal Marker for the v.h.f. range.

The circuit consists of a conventional transistor 100 kHz crystal oscillator followed by a tunnel diode harmonic generator amplifier which provides marker signals up to and beyond the 2m band.
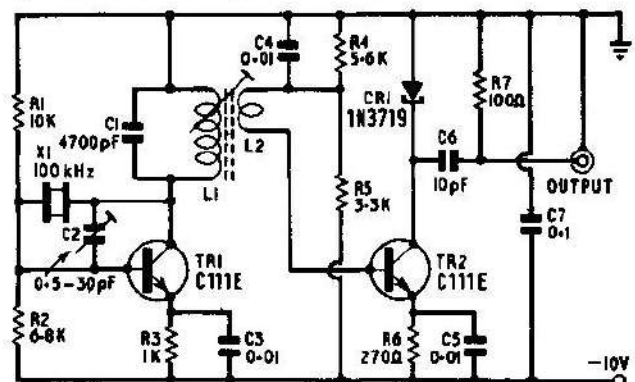
Fig.10.44. A 100 kHz crystal marker.

L1 = 500 μH. L2 = ¼ of number of turns of L1.
X1 = 100 kHz crystal of parallel resonance type.

The adjustment of the unit is very simple and consists of two operations only,
    (a) tune L1 for maximum oscillator, and
    (b) adjust C2 for zero beat with BBC Droitwich On 200 kHz.

Note that a 10 hertz error at 100 kHz will result in approximately 1.5 kHz at 145 MHz. If long term stability is desired then a crystal oven will be needed, but for most purposes this is not required.

POWER SUPPLIES

As the power handling capabilities of transistors is increased, the need for a suitably stabilised power supply becomes necessary especially for fixed station operation. Unlike valve equipment for this type of apparatus the need is for high current at low voltage, and this calls for an entirely different approach to the problem. The unit described is a typical example.

Circuit Principles

Unsmoothed d.c. is produced by a bridge rectifier and is stabilized by a series emitter follower driven by a grounded emitter amplifier in a feedback circuit. Control of the output voltage is achieved by supplying the base of the amplifying stage from an adjustable point on a potential divider between the output terminal and a d.c. reference rail. The use of an auxiliary reference rail permits adjustment to be continued down to zero volts output, without loss of loop gain in the amplifying circuit. This reference rail is produced by use of an overwind on the mains transformer in conjunction with a halfwave rectifier and series stabilization by an emitter follower. No feedback is used in this auxiliary supply, the emitter follower being driven by a Zener voltage reference diode.

By the use of an external reference facility, leads and plugs between the supply and a remote piece of apparatus may be included inside the feedback loop, with a considerable reduction in supply impedance.

An overload trip facility is provided, the current sensing element being a one ohm resistor in series with one of the supply rails. The trip circuit comprises a Schmitt trigger circuit and amplifier which effectively short circuits the main feedback amplifier in the event of an overload, thereby reducing the output of the unit to zero.

Parts list for the stabilized power supply

C1,	450µF	25V d.c.
C2,	1000µF	12V d.c.
C3,	4500µF	40V d.c.
C4,	50µF	12V d.c.
C5,	450µF	35V d.c.
C6,	25µF	25V d.c.
C7,	100µF	25V d.c.
D1-4,	BYZ10	
D5,	OA10	Mullard
D6,	OAZ202	
D7,	OAZ207	
F1,F2,	½A anti-surge fuse	
M1,	0-1A d.c.	
M2,	0-20V d.c.	
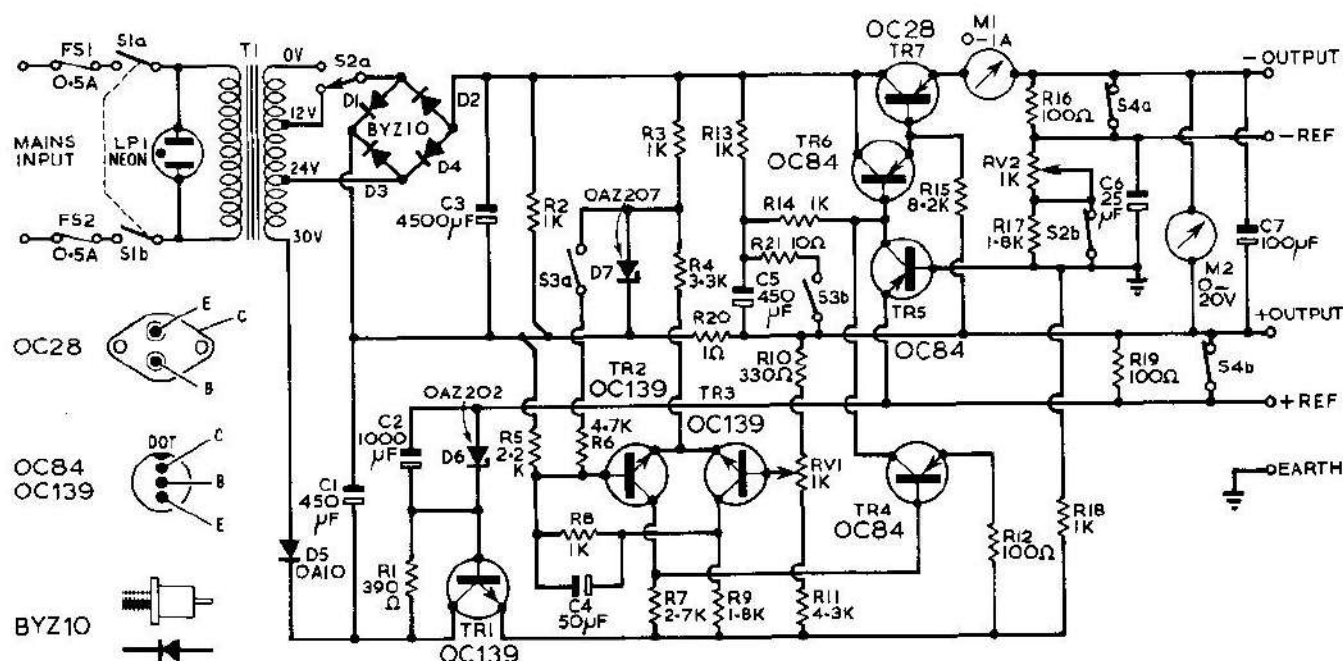
Fig.10.45. The circuit of the 20V 1A stabilized power supply. S2, the voltage range selector, is shown in the 0–10V position.

Parts list for the stabilized power supply—*continued*

R1,	390K ohms ½ W.
R2,	1K ohms 1 W.
R3,	1K ohms 1W.
R4,	3.3K ohms ½ W.
R5,	2.2K ohms ½ W.
R6,	4.7K ohms ½ W.
R7,	2.7K ohms ½ W.
R8,	1K ohms ½ W.
R9,	1.8K ohms ¾ W.
R10,	330 ohms ½ W.
R11,	4.3K ohms ½W.
R12,	100K ohms ½ W.
R13,	1K ohms 1 W.
R14,	1K ohms ½ W.
R15,	8.2K ohms ½ W.
R16,	100 ohms ½ W.
R17,	1.8K ohms ½ W.
R18,	1K ohms W.
R19,	100 ohms ½ W.
R20,	1 ohms 3 W.
R21,	10 ohms ½ W.
RV1,	1K ohms W.W.
RV2,	2K ohms W.W.
S1,	2 pole on/off mains switch.
S2,	2 pole 2 way rotary switch. (may be combined with S1).
S3,S4,	2 pole 2 way slide switch.
T1,	30V 2A Mains transformer.

TR1,	OC139	
TR2,	OC139	
TR3,	OC139	
TR4,	OC84	} Mullard
TR5,	OC84	
TR6,	OC84	
TR7,	OC28	

All resistors are ±10% types.

Circuit Details

The mains transformer taps on the secondary are such that two voltage ranges can be accommodated to minimize power dissipation in the emitter follower.

A d.c. voltage is built up across the main reservoir capacitor C3 by the bridge rectifier D1 to D4. BYZ10 diodes were used here in the prototype but there is no reason that BYZ13 diodes or a similar type should not be used if these were found more readily available.

The d.c. reference line makes use of the 30V overwind on the transformer, halfwave rectification being used to charge the reservoir capacitor C1. The series emitter follower stabilizer is TR1, the base of

this being held some 5.5V above the positive output rail by the Zener diode D6. The smoothing capacitor C2 reduces the ripple on the auxiliary line to an acceptable level (about 10mV).

The main emitter follower stabilizer TR7 is driven by a second emitter follower TR6, R15 being inserted to prevent the output of the unit rising with no load applied, due to collector base leakage in TR7. The potential on the base of TR6 is controlled by the grounded emitter amplifier TR5, d.c. for this amplifier being derived from the reservoir capacitor through R13 and R14. Splitting the d.c. supply path permits a smoothing capacitor C5 to be inserted, this capacitor reducing the ripple in the supply to the small signal amplifying stage. A second important function of C5 is in the standby mode. When S3b is closed, C5 is shorted through a 10 ohms resistor and the base voltage of TR6 is reduced to near zero. Because of the forward base emitter voltages of TR6 and TR7, the output of the unit is exactly zero. Recovery from this condition after opening S3 is regulated by the time constant R13 C5. The slow potential build up caused in this way prevents premature firing of the overload trip circuit due to current surges in the output of the unit.

The emitter of the grounded emitter amplifier is returned to the voltage reference diode used in the auxiliary rail stabilization circuit. The base of the amplifier is connected to a point on a potential divider between the negative output reference and the auxiliary reference rail. Adjustment of RV2 changes the position of the tapping point on the divider and, since the base emitter potential of TR5 remains substantially constant, this results in a change in output voltage. In order to change voltage ranges, a fixed resistor, a little less than the maximum value of RV2 is switched into the divider. C6 improved the loop gain of the amplifier at high frequencies, the value of this loop gain at 1 kHz being approximately 34dB.

The ammeter is placed inside the feedback loop so that the effect of its impedance is reduced by the amplifier, the error in reading due to the drain in the divider being negligible if a 1A instrument is used.

A small electrolytic capacitor C7 is placed across the output terminals of the unit, this capacitance having most effect at frequencies above those at which the feedback amplifier operates i.e. above 10 kHz.

The shunt resistor R2 across the main reservoir capacitor is designed to pass a greater current than is flowing in the auxiliary reference circuit. Without this resistor, when the drain on C3 exceeded the current in the auxiliary circuit, the voltage across C1 would change by about a volt, causing a small change in the reference level and hence the unit output. As it is, the effect of increasing load does modify the reference slightly and this may give rise to a negative d.c. output impedance. This negative impedance is not of sufficient magnitude to support oscillation on any practical load.

The Schmitt 'trigger circuit used in the current overload trip is TR2 and TR3. The emitter supply for this pair is derived from a Zener reference diode D7 in order that the performance should remain unaffected
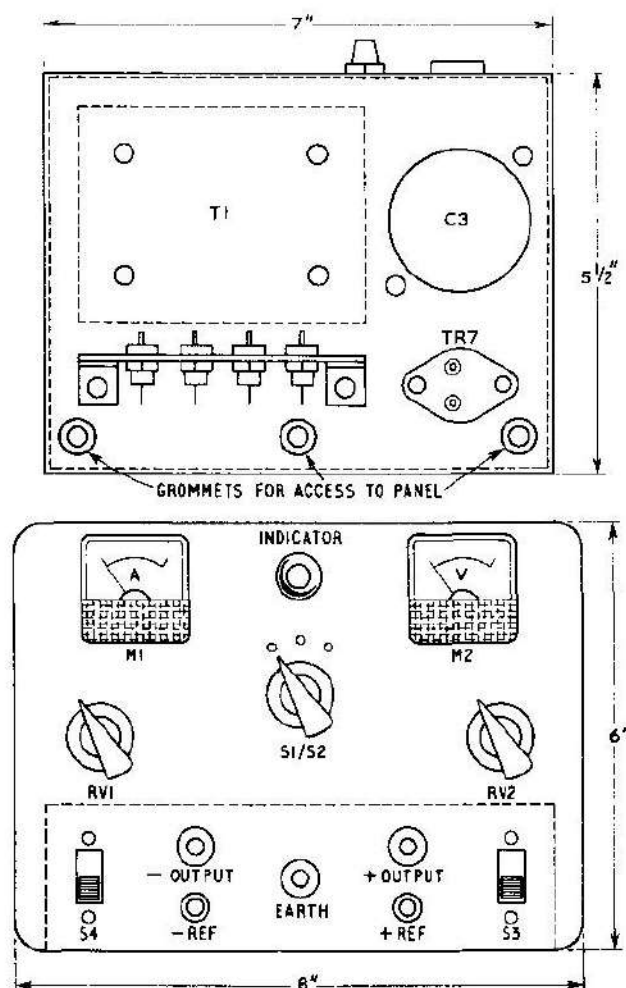
Fig.10.46A. A suitable panel and chassis layout for the power unit.

by a change in output range. In the normal state, TR3 is bottomed and TR2 cut off. As the load current is increased the base of TR3 is taken more negative until this transistor begins to cut off. As the collector voltage rises towards the reference rail, TR2 is pulled on by the coupling effect of C4 and R8. Regeneration completes the switching cycle and TR2 becomes fully bottomed, the base of TR4 being pulled negative. TR4 starts to conduct, the collector current being sufficient to pull the base of TR6 to zero with TR5 cut off.

When the output has fallen to zero, the overload current will be zero, but the hysteresis voltage in the trigger circuit is sufficient to hold the unit off. The trigger can only be reset by closing S3a which turns TR2 off after a short time determined by C4 and the magnitude of the hysteresis voltage. This short interval is sufficient to allow C5 to discharge through R21 and S3b, thereby avoiding a pulse of output when selecting

the standby mode. The current level at which the trip fires can be adjusted by RV1 in the base circuit of TR3.

If, for sake of simplicity, it is required to construct this unit without the current trip facility, all components associated with TR2, TR3 and TR4 could be omitted without affecting the performance of the unit in any other respect. It would, however, be advisable to put a fuse in the output circuit; a 2A fuse in series with the ammeter would afford sufficient protection to TR7 in most cases.

Construction

The BYZ10 diodes need no conduction cooling and may be mounted on a paxolin strip or some other insulating material. The OC28 emitter follower needs an excellent heat sink if the unit is not to fail under low voltage, high current, conditions. If a chassis similar to that specified is used, the transistor may be bolted to this, a mica or anodized aluminium spacer being used to insulate the transistor from the chassis. If some other chassis form is to be used, it should be ensured that an insulated heat sink with a thermal resistance of less than $3°C$ per watt is provided for the OC28.

Full details of a suitable layout for the front panel and chassis are shown in Fig.10.46A.

The low power sections of this unit lend themselves to printed circuit or pin board construction. All the low power stages can be accommodated on a board 5" x 3½" which will fit underneath the main chassis.

Testing

When the unit has been checked thoroughly for any mechanical faults, it may be tested as follows. Turn RV1 to its maximum position (slider nearest auxiliary rail); RV2 to minimum (shorted); S4 to Reference Internal; S3 to Normal and S2 to 0 — 10V d.c. Switch on the mains supply and check that, on increasing the setting of RV2, an output voltage reading is obtained. If no such output is forthcoming, measuring the voltages across C3 and C1 should give an indication of whether the fault is in the supply circuit or stabilizing circuit. To eliminate the influence of a faulty current trip circuit, it may also be found useful to disconnect the collector connection of TR4.

After a voltage has been obtained on the output terminals, increase RV2 to maximum and check that the output is at, or above 10V. Now change the 10V to 20V range and check that RV2 has its expected range.

Return S2 to the 10V range and set the output voltage to 6V d.c. Now reduce the trip level potentiometer slowly to zero. If the output falls to zero the potentiometer should be very near the low end of its range when the trip fires. Increase RV1 to

maximum again and, if in the last test the overload trip was fired, reset the trip by switching S3 momentarily to STANDBY. The output level should be observed to climb slowly back to 6V as C5 charges through R13.

Connect a load of about 10 ohms to the output of the unit and adjust the output voltage until the load current is 0.6A. Now reduce the setting of RV1 until the overload trip fires. RV1 should be near mid range at this time. Reset the trip and RV1 and increase the output to 1A to ensure that the maximum setting of the trip level is sufficient. Any slight discrepancies in the range of RV1 can be removed by adjusting R10 and R11 until the correct range is obtained. Only after the current trip range has been fully checked should any short circuit test be attempted as, if the current trip is faulty, an overload will destroy TR7.

Should any difficulty be experienced in obtaining the expected performance from the unit, the voltages shown in the table may be used as a guide in fault finding.

	e	b	c
TR1	+5.4	+5.6	+9.4
TR2	+0.7	+0.5	+5.3
TR3	+0.7	+0.8	+0.7
TR4	+5.4	+5.3	-10.5
TR5	0	+0.1	-10.6
TR6	-10.3	-10.6	-12.9
TR7	-10.3	-10.3	-12.9

Specimen d.c. voltages present in the power supply while it is supplying 10V at 0.5A. All voltages were measured with an Avometer Model 8 on the 25V d.c. range and are with respect to the positive rail.

Operation

Operation of the unit is generally straightforward but a few notes may help to explain the facilities provided to their best advantage.

Current trip. This should generally be set at about twice the anticipated peak load current, though if only short circuit faults are expected, it may be left at its maximum setting.

Voltage Range At the overlap point between ranges, it is generally better to use the low range than the high, since a great deal of power is wasted in the OC28 if the high range is used near its minimum voltage setting.

Standby Switch. The standby switch is operated momentarily to reset the current trip and may also be used while making adjustments to an experimental circuit under test. When returning to normal, the output will build up slowly to prevent the overload

trip firing due to a current surge into a capacitative load. This slow build up may also give enough time to spot any fault under test before any damage is done.

External Reference Switch. The external reference facility may be used in situations where a low source impedance from the power supply is vital to prevent oscillation or some other unfortunate condition. This is often the case where the unit being powered is separated from the supply by a long supply lead and, perhaps also some plugs and sockets which all add to the supply impedance.

In this case, a pair of leads should be brought back to the power supply from the supply rails in the unit under test, these leads being inserted in the appropriate reference sockets on the power supply. If S4 is now opened, the feedback loop will extend to include the supply leads and hence their effective impedance will be reduced by a factor of up to 50 at frequencies over which the feedback amplifier operates.

A TUNNEL DIODE PROTECTED POWER SUPPLY FOR TRANSISTOR EQUIPMENT

Transistors, unlike the thermionic valve, are very sensitive to voltage surges. The normal mains supply system is subjected to many switching and other surge producing effects and transistor equipment should be protected by some fast acting device to reduce these transient surges. This power supply will do just that as well as providing a stabilised output adjustable between 18 and 35 volts.
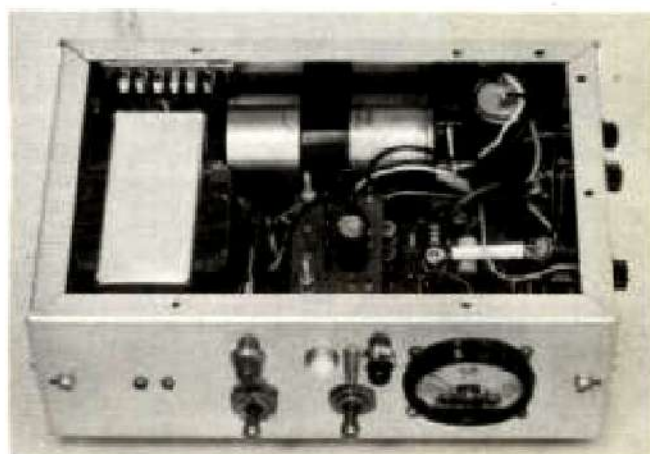
Fig.10.46. The complete tunnel diode protected power unit.

Input

The input transients can be dealt with very simply; a capacitor and resistor across the transformer secondary (Fig.10.47).
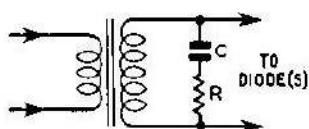
Fig.10.47. Input transient suppressor device.

The capacitance necessary to suppress the switch-off transient is given by:

$$C = 400 \; \frac{\text{transformer total power rating}}{(\text{diode p.i.v})^2} \; \mu F \text{ for stalloy}$$

cores. When the transformer is switched on, voltage surges may occur due to oscillation of the transformer leakage inductance with the surge suppression capacitor. This can be damped to a safe value with a resistance in series with the capacitance:

$$R = 100 \; \sqrt{R_1/7C} \text{ ohms,}$$

where R_1 is the load resistance and the frequency is 50 hertz.
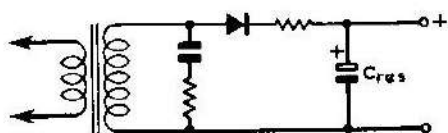
Fig.10.48. Voltage surge suppression following rectifier.

The reservoir capacitance is an effective short-circuit when the unit is switched on, so some series resistance before the reservoir is essential (Fig.10.48).

The resistance can either be external or internal to the transformer to limit the surge to a safe value (say, 10 amps for a 1 amp rectifier).

Output

In Fig.10.48 the output impedance will be high so the output voltage varies considerably with the current drawn. There are plenty of stabilizing circuits to reduce the output impedance and one which is not too complex is given in Fig.10.49.
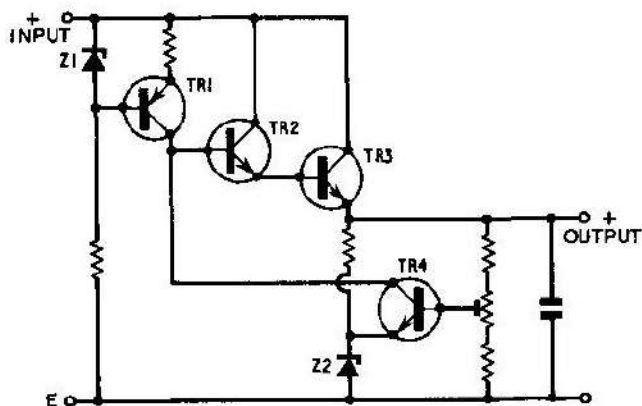
Fig.10.49. One form of stabilizing circuit which will reduce the output impedance.

If a heavy output current arrives TR3 could very well be blown up, and possibly TR2 as well. In other words, a current operated trip should switch the supply off *before* that kind of thing happens. A mechanical switch would be far too slow, so we are left with electronic switches.
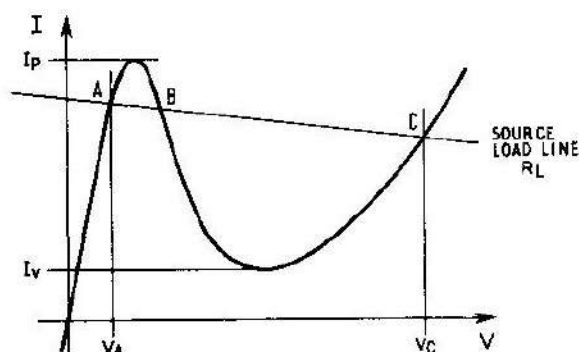
Fig.10.50. This shows the two points A and C where the diode is stable.

The normal kind of switch either relies on a fairly heavy voltage drop (s.c.r's) or a negative supplementary rail (Schmitt, Eccles-Jordan) for reliable working. It was to get round these disadvantages that a tunnel diode was used. A tunnel diode can switch in 1 mμS but the switching range is small. Amplified it performs perfectly.
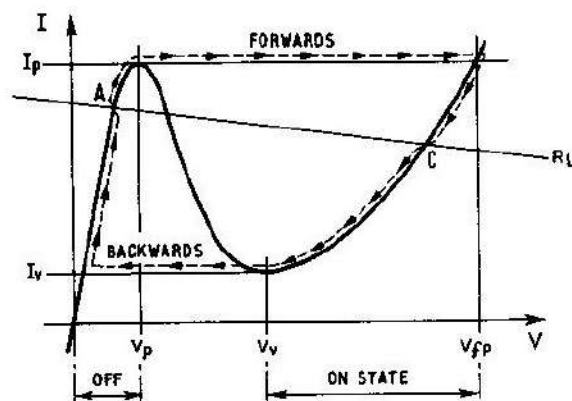
Fig.10.51. Illustrating the tunnel diode as a switch.

Switching using Tunnel Diodes

In an article on tunnel diodes mention was made of driving a diode with a source impedance greater than that of the diode's own negative resistance *(RSGB BULLETIN,* February 1965, p.97). In fact, if a tunnel diode is driven from a current source (as it sees it), there can only be two places on the V/I curve where the device is stable; e.g. in Fig.10.50, A and C. B is unstable, as it is on the negative resistance part of the characteristic. If the diode were biased to point A and it received a positive going pulse, provided that the pulse lifted it above the top of the curve, Ip it would switch to C; and the only way it could be

brought back again would be to have a negative going pulse going below the valley current. Or, alternatively, just switch the supply off. The sequence of events is shown in Fig.10.51.

For a 1mA tunnel diode (IN2929, IN2940, JK 19A, etc) Vp is about 50 to 60mV and Vfp is about 0.5V. The time taken to switch, once Vp has been reached, is easy to find approximately. With a CR combination, the time constant in seconds for the capacitor to charge up to 63 per cent of the supply voltage is equal to the capacitance multiplied by the resistance in other words, t = CR. R can be replaced by $\dfrac{V_{fp} - V_p}{I_p - I_v}$ and C can be taken as the junction capacitance. With the IN2940,

$$I_p = 1mA$$
$$I_v = 0.15mA$$
$$V_{fp} = 500mV$$
$$V_p = 60mV$$
$$C = 5pF$$

$$\text{making } t = \frac{(V_{fp} - V_p)C}{I_p - I_v}$$

$$= \frac{440.5.10^{-12}}{0.85}$$

$$= 2.5 \text{ m}\mu S$$

However 2.5 mμS is too short and the way to slow it down is to increase the capacity across the diode. With 0.05 μF the switching speed drops to 25 μS, which is just about what is needed.
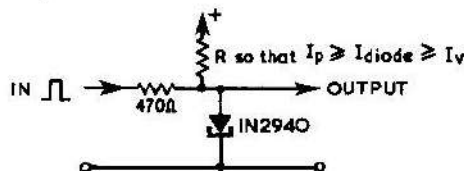
Fig.10.52. Circuit of the tunnel diode switch.

The switching range, A to C or else V_p to V_{fp} with a germanium tunnel diode, cannot be less than 200mV. This can be amplified by hanging a transistor directly coupled stage on to the diode. Now the minimum voltage that could be across "C" in Fig.10.47 is 350mV (V_v) and the maximum, 500mV, which is not enough to drive a silicon transistor. With
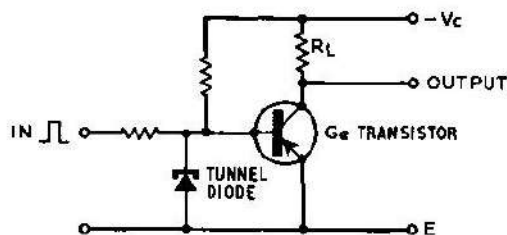
Fig.10.53. The addition of a germanium transistor completes the switch, which may trip in a range of only 1 per cent.

a germanium transistor, however, things are different. When it is switched on, the major part of the tunnel diode bias current will be diverted to the base of the transistor and this then switches on.

The stability of this kind of switch depends mainly on how stable the bias current is and how stable the diode peak current remains (I_p). With good design it is possible to have reliable tripping in a range as small as 1 per cent.

Complete Circuit

Fig.10.54 shows a ½ amp. 28 volt power supply with fairly low output impedance. It switches off in 25 μSat any current which may be desired, from O to a surge current of 0.8 amp without even a flicker on the meter. The output voltage can be adjusted with VR15 to between 18 and 35 volts, and the voltage drops by 0.1 volt on full load.

COMPONENTS LIST FOR Fig. 10.54

C1	0.47 µF 100V (Mullard)
C2	2500 µF 64V (Mullard)
C3	0.05µ ceram 30V
C4	0.05 µ ceram 50V
C5	3.2 µF 64V (Mullard)
CR1	SJ203K (AEI)
CR2	SJ203K (AEI)
CR3	SJ203E (AEI)
CR4	SJ203E (AEI)
F	½ amp
M	½ amp (low resistance)
R1	390 ohms ½W
R2	2.7 ohms 3W
R3	6.8K ohms ½W
R4	2.7K ohms ¼W
R5	5.6K ohms ½W
R6	180 ohms ¼W
VR7	10K ohms min preset var.
R8	1 ohm 3W
VR9	300 ohms min preset var.
R10	1 ohm ¼W
R11	6.8K ohms ½W
R12	1K ohm ¼W
R13	6.8K ohms ½W
R14	3.9K ohms ½W
VR15	1K ohm min p/s var.
R16	1K ohm ¼W
R17	10 ohms 10W
T1	40V ¼ amp tapped 6V (sec., d.c. res. min. 3 ohms)
TD	1N2940 (G.E.)
TR1	AF118 (Mullard)
TR2	TR6 2N696 (G.E.)
TR3	2S301 (Texas)
TR4	2S019 (Texas)
TR5	2S013A (Texas)
Z1	MR56H (AEI)
Z2	MR33H (AEI)
Z3	MR62H (AEI)
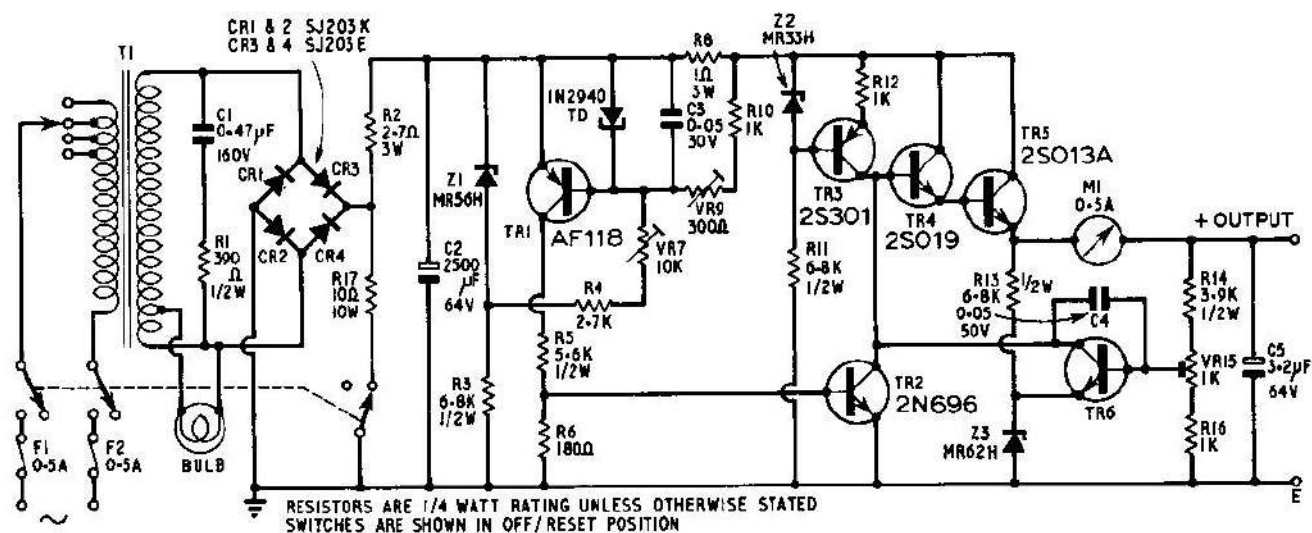
Fig. 10.54. Complete circuit of the tunnel diode power supply.

The switch operates off a 1 ohm resistance in the main h.t. rail. VR7 controls the normal standing bias in the diode (about 0.75mA) from a 5.6 volt zener diode (diodes of this range have the lowest temperature coefficient), and VR9 is a fine control for tripping. One practical point; the 0.05 μF capacitor (C3) across the diode junction should be ceramic and the lead length between it and the diode should be as short as possible (¼" at most). The d.c. amplifier used here is an AF118 because it is cheap and can stand the h.t. voltage. When the switch is off, the collector-emitter voltage of the AF118 is as near full h.t. as makes no difference. When the switch is operated, the transistor turns hard on ($V_{ce} = 0.5$) thus turning TR2 on (0.7 volt across the emitter base junction). This brings the bases of TR4 and TR5 (a Darlington pair) down to ground level. The output voltage sinks to 5mV across 62 ohms, and 1.1 volts open circuited. Resetting is by switching the unit off and waiting for the reservoir capacitor to discharge through R17. The switch controlling R17 is ganged to the main power switch and should be a "break-before-make" type.
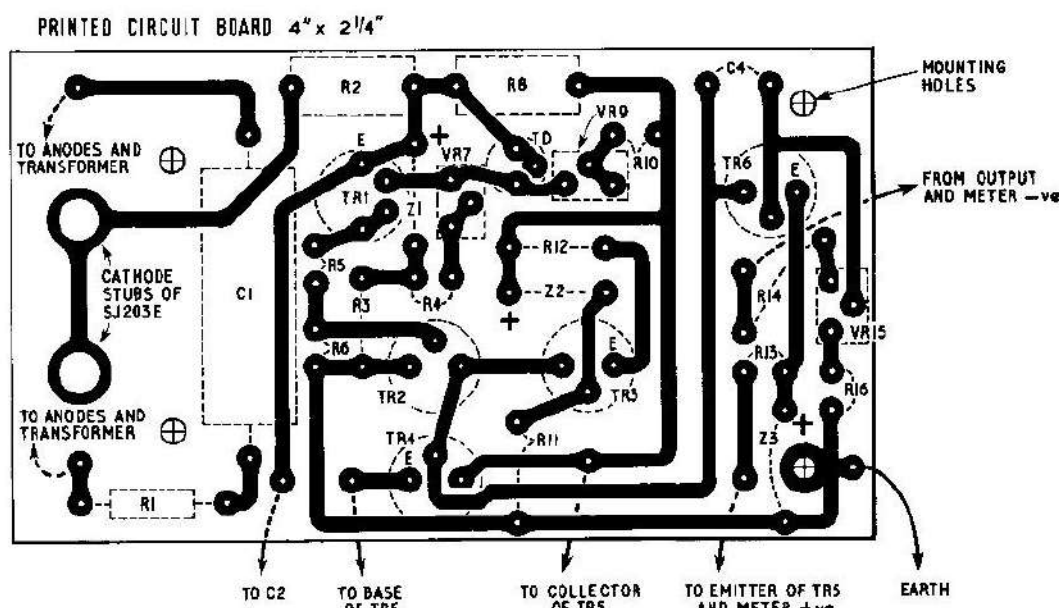
Fig. 10.55. A suggested printed circuit layout.

Resetting could be done by having a shorting button across the tunnel diode (and a protective fuse in the emitter circuit of TR5). While this method of resetting would work, no mechanical fuse can act fast enough to protect TR5 (and quite likely the load as well) from burning out if the unit still has got a short circuited load connected across it.

If different cut-out times are needed C3 can be changed or switched as shown below.

Time μS	C3 μF	Time μS	C3 μF
10	0.02	100	0.2
25	0.05	250	0.5
50	0.1	500	1

500μS (or ½ millisecond) is the longest cut-out time constant which can safely be used, and 10 or 25μS should be sufficient protection, say, for overlay transistors or other high current low impedance devices.

It would have been possible to have saved a transistor in this circuit (TR2) by having the switch in the negative line but transistors of this kind are cheap, and it was thought worth while doing it this way, so as not to break the earthline.

For higher currents, say 1.5 amps, the sensing resistor should be reduced to 0.3 to 0.5 ohms, the emitter resistance (R12) belonging to TR3 should be reduced to 560-620 ohms and, more important, TR4 (2S019) should have a heat sink. 1.5A is the maximum that the final power transistor (2S013A) must pass. Any n-p-n transistors can be used in the stabilizer section provided they can stand the voltage and current. For negative h.t. lines, n-p-n transistors should be changed for p-n-p types and vice versa and the zener and tunnel diodes should have their connections reversed.

A Low Power Unit Suitable for use with Converters and Similar Equipment

This small unit will deliver 6.3V at 1.5A and 150V at 30mA, which is adequate for the average
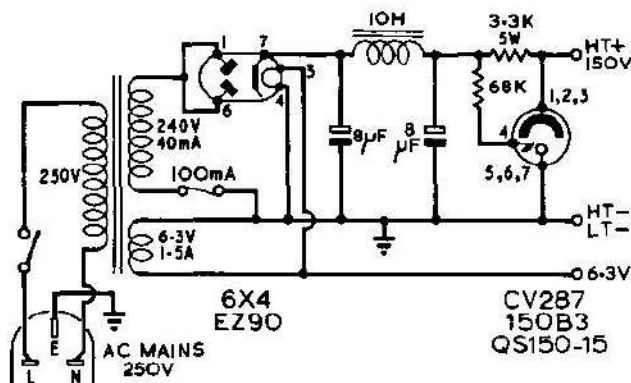
Fig.10.56. A simple power supply, suitable for use with valve converters.

converter. A suitable circuit is shown in Fig.10.56 and the photograph illustrates the general layout when the unit is built into a No.650 standard Eddystone diecast box which measures 3½" x 4½". If desired a suitable relay can be incorporated so that the h.t. supply may be interrupted while the station is transmitting.

COAXIAL RELAYS

In most v.h.f. and u.h.f. installations, coaxial cable is employed for the transfer of r.f. power from the transmitter to the aerial, and from the aerial to the receiving system. For rapid changeover of the aerial between the receive and transmit modes, a relay is required which, ideally, has been specifically designed to function with this type of cable. These are known as coaxial relays.

Coaxial relays are constructed so that they exhibit a characteristic impedance which matches that of the coaxial cable with which they operate. When the impedance of the relay and that of the cable are identical, no mismatch point will be introduced into the transmission line to degrade the standing wave ratio and produce reflections in the feed system. In
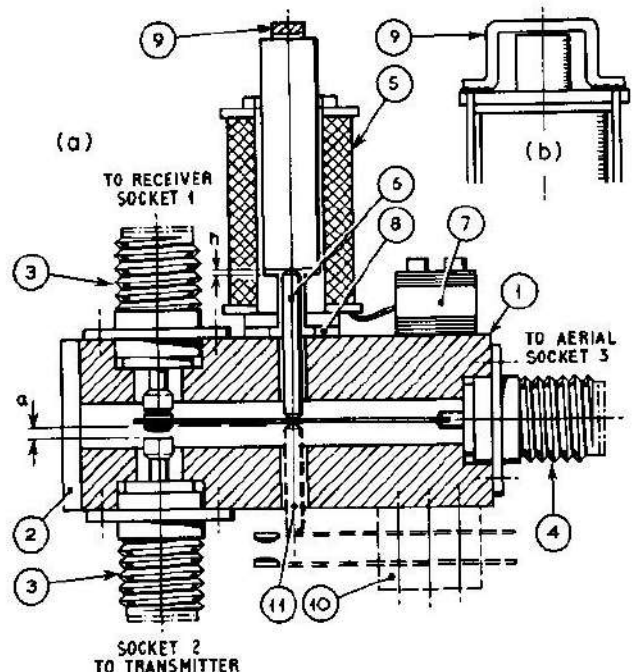
Fig.10.57. A general cross-section view of the assembled relay.

practice, absolute perfection in matching is rarely, if ever, attained, due in the main to the fact that the impedance quoted for coaxial cable is normally a nominal figure and subject to some variation.

The relay to be described may be constructed for coaxial line impedance of 52 ohms, 60 ohms or 75 ohms, the impedance being determined by the dimensions of the cavity containing the switching arm,

and the size of the switching arm itself. Full dimensions are given, and while these are expressed in millimetres, they may be converted into inches by use of the following formula:

$$\text{Inches} = \frac{x}{25.4}$$

where x is the dimension given in millimetres.

One particular feature of the relays is their good crosstalk damping level, and this is of special interest in receiving systems employing high conductance triodes which can be quite easily damaged by excessive r.f. leakage during transmission periods. Specific information on the crosstalk level is given later.

The form of the relay is such that it can be mechanically linked to other contact sets and so be built up into a complete master control unit. When giving thought to such an arrangement, however, the possible need to arrange sequential switching should be remembered, and as will be seen, this is related to the r.f. power that the coaxial relay is expected to carry.

Mechanical Description.

Fig.10.57 shows the cross section through the relay.

Three coaxial sockets (3) (4) are fitted to a rectangular brass block which has been drilled out both horizontally and vertically to produce a cross-shaped cavity. Socket 3 (4) which is fitted to the righthand end, carries a spring switching strip fitted with double contacts, these being positioned precisely in line with the stubs on sockets 1 and 2 (3). The stubs of sockets 1 and 2 (3) are fitted with contacts and these will mate with those on the spring arm.

The spring strip attached to socket 3 (4) is formed so that it is normally in contact with the stub of socket 1. Pressure on the actuating rod (6) will move the spring strip from its rest position in contact with socket 1, over to the contact on socket 2. The actuating rod (6) is powered by the solenoid (5) and its associated plunger. The extent of the movement of the plunger/actuating rod is shown as (h) on the drawing.

A second actuating rod (11) may be fitted opposite the primary actuating rod (6). This allows switching of other contact strips (10) which may be built on to the coaxial relay unit and employed for control switching purposes.

After the position of the spring switching strip, and the contact separation (a) have been adjusted, the front of the brass block (1) is closed by fitting the cover (2) and the leads from the solenoid terminated on the connecting block (7).

Design Impedance

Coaxial cable usually employed has an impedance of either 52 ohms, 60 ohms or 75 ohms. For the reasons already covered, the characteristic impedance of the relay, sockets and cable should all be the same, and this means that the relay must be designed to operate with the particular cable concerned in respect of its quoted impedance. The section of an asymmetrical h.f. conductor corresponding to the relay is shown in Fig.10.58.

The flat spring strip of width (f) and thickness (s) lies centrally in the circular conductor, this being formed by the horizontal bore through the block (1). The characteristic impedance Z with air dielectric (K=1) can be calculated from the formula given in Fig.10.58 or derived from the graphs. The dimensions required to produce specific impedances are given in the table associated with Fig.10.59.
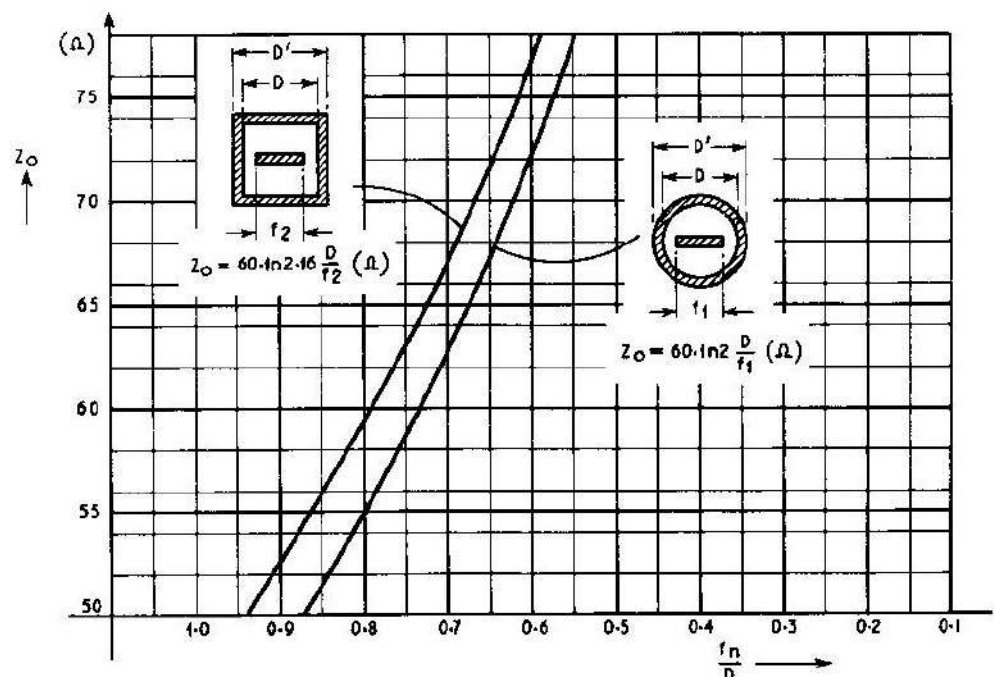
Fig. 10.58. The characteristic impedance of a conductor having circular or rectangular cross-section assuming an air dielectric. (K = 1).

Z (Ω)	f (mm)	D (mm)
52	6·7	8
60	5·9	8
75	4·6	8

$$z\,(\Omega) = 60 \cdot \ln 2 \frac{D}{f}$$
where $s \ll D$

Fig.10.59. Conductor dimensions for different impedances.

Coaxial Sockets

The sockets shown in the description are the types UG58A/U and UG568/U.

The normal h.f. coaxial sockets with a square mounting flange have, on their mounting side, one or more cylindrical steps. This means that the relay body (1) has to be counter-bored accordingly. As the dimensions of these steps differ according to the manufacturer, the dimensions of the counter-bore will have to be adjusted according to the exact type of socket employed. It is particularly important that the end of the outer conductor of the socket makes good electrical contact with the base of the counter-bored hole.

* SEE TEXT

DIMENSIONS ARE IN MILLIMETRES

DIMN mm	VERSION 'A'
a	18·25
b	8·0 *
c	16·0 *
d	3·5 – 9·5 *
e	>3·2 *

Fig.10.60. The relay body component parts.

The N series is rated for use on frequencies up to 10,000 MHz, but in practice is not generally used above 3000 MHz. The insulation may either be Teflon or polystyrene according to the manufacturer. They are waterproof and can handle up to 1500 volts, and may be obtained for 70 ohms impedance. The C series has the same characteristics and is available in 50 ohms or 70 ohms impedance.

Connectors are available for matching and interconnecting one series with another.

The construction when N or C type sockets are employed can be simplified by ignoring the stepped shoulders on these sockets. The diameter of the cavity bore is made constant from one end of the block to the other. That is, the diameter (b) on the lefthand side of the section A—B is increased to the same as that of (c) on the righthand side (Fig.10.60).

Making this bore a constant 9.5mm throughout its length will clear the shoulders of the sockets. If this course is adopted, the impedance of the relay is adjusted by correct selection of the size of the switching arm. This will be found in the table associated with Fig.10.60.

ITEM 9   LOOP
14 swg COPPER WIRE
BENT TO SHAPE AS SHOWN
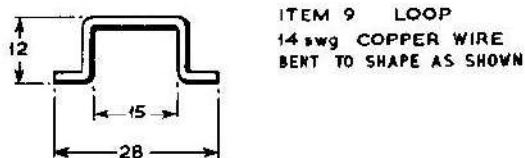
Fig.10.61. The plunger retaining loop.

Ideally, the main cavity containing the switching arm should be drilled about 0.2mm undersize, and then finished with a reamer. The final surface must be smooth and free from burrs, particularly at the intersections with cross drilled holes.

The hole (e) which accommodates the actuating rod should not exceed 3.2mm in diameter, and if it is required that the coaxial relay shall operate other contact sets driven by the same energizing system, then this hole should be drilled right through the block.

4 HOLES 3·2mm DIA
COUNTERSUNK FOR M3
3mm BRASS SHEET

ITEM 2
(a)

PLASTIC (TROLITUL
OR SIMILAR)
ROD
ITEM 6
(b)

SPRING
CONTACTS

ITEM 4b or 4d
(c)
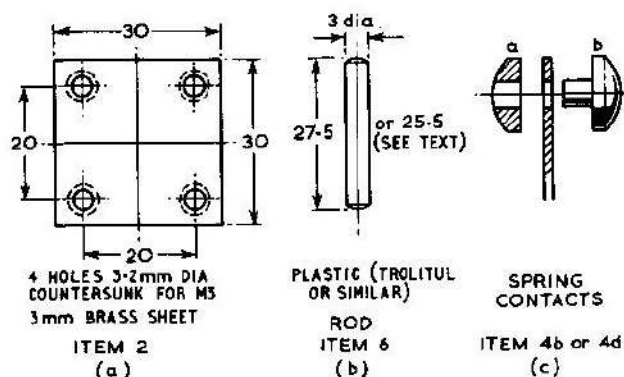
Fig.10.62. The relay body and plate, the plastic actuating rod, and the method of assembling the spring contacts.

The main cavity bore housing the switching arm should be silver plated.

The solenoid is secured to the body of the relay by two countersunk brass screws which pass right through the body of the relay block, and this is illustrated in section X-Y in Fig.10.60 the screws being designated M3.

The solenoid specified has only a very small hole in its base, and this has to be increased to 3.2mm in order to accommodate the 3mm actuating rod. If it is made larger than 3.2mm, the performance of the solenoid will be affected.

Prior to plating the main cavity, the solenoid should be secured to the main block, and the alignment of the drilling of the base of the solenoid to that of the hole in the block which accommodates the actuating arm, checked. Any adjustments that are found to be required should be made before the cavity is plated.

COUNTERSUNK FOR M3
ITEM 8 ANGLE
NOT REQUIRED IF DRILLED
AS SECTION X-Y ON ITEM 4

SECTION A - B
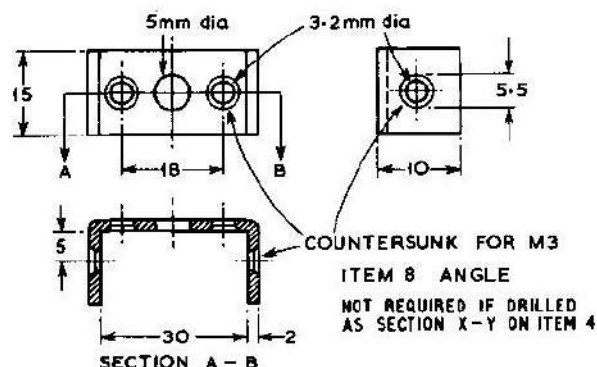
Fig.10.63. The U-support for the solenoid. This item is not required if the solenoid is mounted direct on the relay body.

After assembling the relay and checking the contacts, the wire loop (9 of Fig.10.61) is soldered to the magnet frame as shown in Fig.10.57(b) to prevent the plunger falling out of its housing.

The rod (6) shown in Fig.10.57 and in Fig.10.62(b) should be 25.5mm long, but if the solenoid is mounted on an angle piece (8) in Fig.10.63 and Fig.10.57 instead of being secured directly to the main body of the block, it will have to be increased to 27.5mm. The rod is made from 3mm diameter polystyrene stock as is that for coupling in other contact sets if they are fitted.

Assembly

Fig.10.64 provides details of the manner in which the contact is fitted to the spring arm, while Fig.10.62 (c) illustrates how the contacts are prepared prior to being fitted.

The shaft of one of the contact rivets is removed, and the back filed flat. This rivet is then drilled centrally with a hole equal to the diameter of the shaft of the other rivet. A sandwich is made consisting of
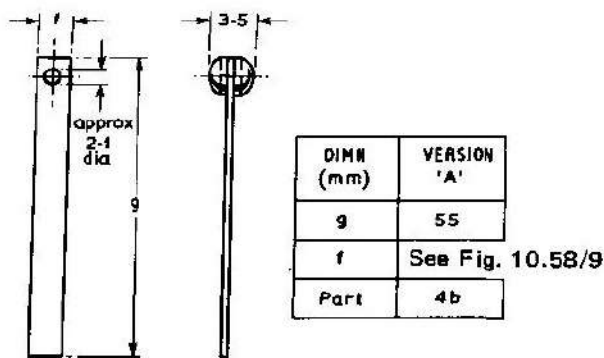
Fig.10.64. The switch spring dimensions.

DIMN (mm)	VERSION 'A'
9	55
f	See Fig. 10.58/9
Part	4b

the stemmed rivet passing through the spring arm—previously drilled to take the stem as shown in Fig.10.62 (c) and the drilled rivet head slipped over the stem. The shaft which protrudes through the drilled contact head is marked at about 0.3mm proud of the dome of the head and carefully cut at this point. The shaft is now dressed back with a light hammer until a hemispherical head is produced with the spring strip firmly clamped. Provided that the rivetting has been correctly done, soldering will not be necessary. Finally, a small radius is filed on each corner of the spring strip adjacent to the contacts.
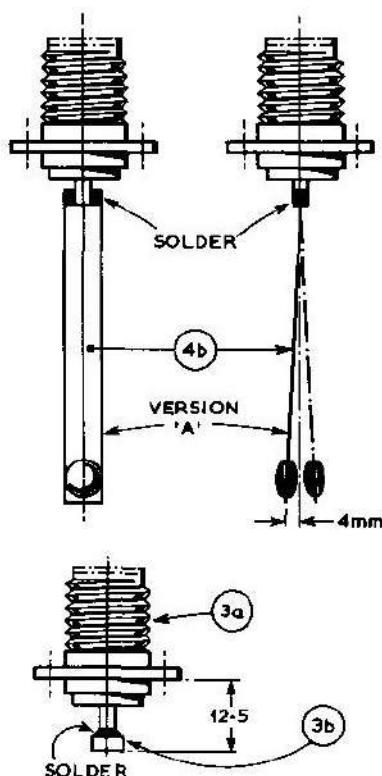
Fig.10.65. Sockets used at 3 and 4 Fig.10.57 showing contacts attached.

Before soldering either the spring arm or the contacts to the sockets, a plug should be mated with the socket concerned. Under the heat of soldering, the insulation may well soften and cause displacement of the centre conductor, and the plug will assist in retaining this conductor in its correct position.

Fig.10.65 shows how the contacts are fitted to the sockets identified as (3) in Fig.10.57. The smooth contact rivet (3b) is inserted in the end of the socket's inner conductor, clamped, and then lightly soldered all round, taking care that no solder reaches the contact surface.

Prior to fitting the contact spring switch arm, the end opposite to that carrying the contact set should be provided with a small slot - using a coping saw — of sufficient size to accept the centre spigot. A small jig, made from wire, should be devised to hold the coaxial socket and the spring arm in their correctly related positions while the soldering is undertaken. Without such an arrangement it will be very difficult to position the spring centrally to the spigot on the socket.

After the spring arm has been fitted to the socket, but before the socket is fitted to the body of the relay, a small set should be made in the spring arm so that it deviates by about 4mm, see Fig.10.65. This will produce a contact pressure of about 50 grammes, which is correct.

Once the relay has been assembled, it is checked for contact pressure and contact register. The "unenergized" position should present no difficulty since this is determined by the angle of the set placed in the arm. However, with the solenoid energized, the switching strip should not be bowed too far by the actuating rod (6). Adjusting the length of this rod will vary the contact pressure, and consequently the amount of "bow" in the switching strip. Finally, the cover (2) should be screwed into position.

Performance

With the relay terminated in a 60 ohms impedance, the following figures were obtained, the relay itself, of course, having been designed for 60 ohms operation.

Frequency	Reflected power	Crosstalk level
145 MHz	0.1 per cent	−40dB
435 MHz	0.7 per cent	−32dB
600 MHz	1.0 per cent	−30dB
1000 MHz	4.0 per cent	−23dB

On 145 MHz and 432 MHz the relay proved itself capable of handling more than 200 watts of r.f. power. Its performance in this respect on higher frequences has not so far been checked.

When switching appreciable r.f. power it is not recommended that this should be done by the relay directly. Rather, sequential switching should be employed, the power being applied after the switching action has taken place. Failure to observe this requirement could lead to flash-over and arcing with serious damage to the contact surfaces.

10.32

INDEX

INDEX

JESSOP

VHF-UHF MANUAL

RSGB