10 Repeaters

REPEATERS have been described as the second greatest development in amateur radio since the Second World War. The first was, of course, the advent of SSB which is firmly established as the speech DX mode on all bands.

What is a repeater? A repeater is analogous to the talk-through systems of the user services. In commercial two-way radio systems the base station and the mobile station are allocated different transmit frequencies so that any eavesdropper can only hear half the conversation. In commercial circles talkthrough is often a luxury in that it allows mobile-to-mobile communication rather than the more normal mobile-to-base traffic.

In the Amateur Service mobile-to-mobile contact is very common but the difficulty is that range is restricted – why not put a base station on the biggest hill in the county so that many mobiles can make contact through it? See Fig 10.1.

A repeater is a device which will receive a signal on one frequency, and simultaneously transmit it on another frequency. Careful design has meant that repeaters can receive and transmit on the same band, and this means that the same antenna can be used for both reception and transmission. In effect, the receiving and transmitting coverage of the mobile station becomes that of the repeater and, since the repeater is favourably sited on high ground or a tall mast, the range is greatly improved over that of unassisted, or *simplex* operation. The coverage areas of two mobile stations continually change shape, whereas the coverage of a repeater will stay constant, and can even be published (Fig 10.2).

The Amateur Service has specific bands and repeaters are confined to fairly small segments of these bands by international agreement so that the needs of all users can be accommodated in the limited amount of spectrum available. The repeater uses two frequencies simultaneously, one for transmit and one for receive.

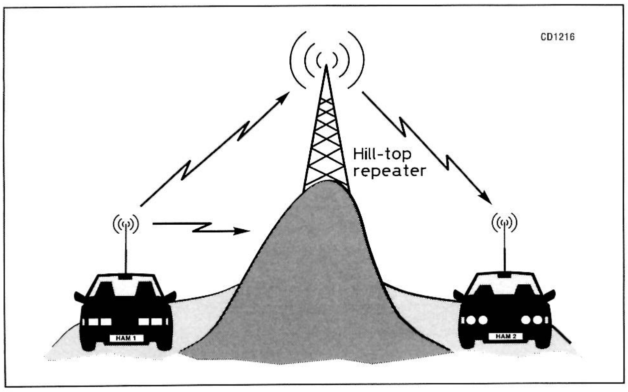


Fig 10.1. A repeater on a hill being used by two mobile stations

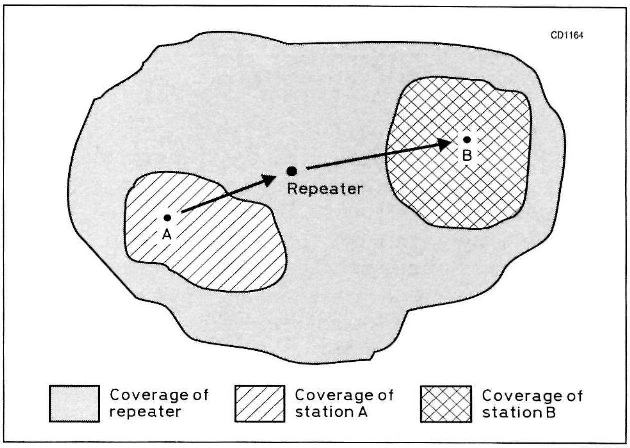


Fig 10.2. Repeater and simplex areas

Repeaters are operational in the UK on all of the VHF bands except for the 70MHz one because the small amount of space available means that it is not able to accommodate any. The difference between receive and transmit frequency is 500kHz on 50MHz, 600kHz on 145MHz, 1.6MHz on 432MHz and 6MHz on 1.3GHz. Channel spacings are 10kHz, 25kHz (soon to become 12.5kHz), 25kHz and 25kHz respectively. On the 50 and 433MHz bands the input frequency is higher than than the output frequency but in the other bands it is lower.

During the late 'sixties and early 'seventies FM came into widespread use on the VHF bands mainly because of its cheapness and the ease of construction of FM equipment compared with that for SSB. Surplus ex-private mobile radio (PMR) equipment was cheaply available and this was frequently converted from high-band PMR use to FM on the amateur 2m band.

The improvement to mobile communications system performance is quite dramatic because although the range is potentially less, the quality of reception is much better. This is because of the limiting effect of an FM signal which eliminates much of the ignition noise from surrounding vehicles. Although an individual may have satisfactorily suppressed his/her vehicle the problem is all those other unsuppressed vehicles around. It is also advantageous for the repeater in that when a receiver is close in frequency to an adjacent transmitter a lot of amplitude noise is generated which is easier to eliminate in an FM system by the nature of the limiting action of the FM detector.

Because of the limited spectrum available in the amateur bands it is customary to use a much narrower spacing between the receive (input) and the transmit frequency (output) compared to commercial systems, eg on the 430MHz band

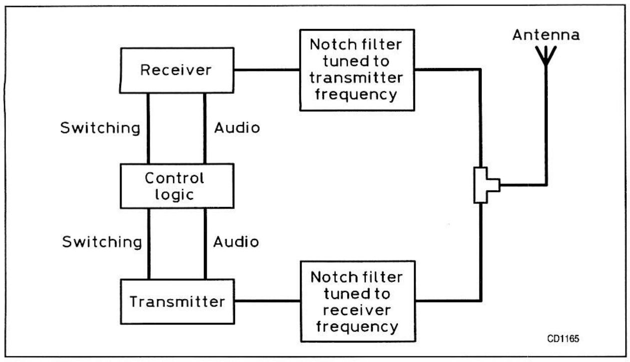


Fig 10.3. Block diagram of parts of a repeater

the spacing is 1.6MHz and on 144MHz band is 600kHz. There are nearly 300 repeaters in the UK, comprising approximately 80 VHF repeaters, 170 UHF repeaters, and 25 repeaters on the 1.3GHz band including 12 TV repeaters. There are also four TV repeaters operational on the 10GHz band. Some of these repeaters are extremely popular and are in use for many hours a day whereas the corresponding PMR system might only be in use for a few minutes per hour. This continuous service aspect of repeaters makes greater demands on power supplies and other components which have to be continuously rated rather than for intermittent use. Further information on these repeaters is available in the *Amateur Radio Operating Manual* [1], in a computer listing available from RSGB [2], in the *RSGB Yearbook* [3] as well as in the *Amateur Radio Diary* [4], all published by the RSGB. A list is also available on the Internet [5].

The fact that large numbers of amateurs use any repeater system means that the coverage area of the unit is very quickly established. This large number of users highlights another problem. How do you share out the available air time to all potential users on what is basically a single-channel device? Although many can listen, only one person at a time can transmit. In order to provide an incentive for short transmissions it is normal to provide a limitation to the talkthrough time permitted. After a given period of time, usually one to two minutes on the busiest VHF repeaters and five minutes on the UHF repeaters, the user *times out*, ie the repeater will no longer relay the input signal.

The simplest sort of repeater needs an antenna or antennas, a receiver, a transmitter, something to control it (usually referred to as the *logic*), and an arrangement of filters to enable it to receive and transmit at the same time. A further difficulty for the would-be repeater builder is that the DTI have stipulated that repeaters should not be triggered by a spurious transmission on their input frequency. Access to the repeater, ie the switching on of the transmitter, is accomplished by either a short tone burst of 1750Hz or by the transmission of a sub-audible tone. This is often called *CTCSS* which stands for 'continuous tone-coded squelch system'. The user has to transmit a tone which is below the audible range all the time. The CTCSS frequency used depends on the area and has the advantage of preventing a station accessing a repeater out of its area except in exceptional conditions. For example, all the London repeaters have the same CTCSS tone (82.5Hz) but the Brighton repeater (GB3SR) on the same frequency as the East London repeater (GB3EL) would not be accessed at the same time because its tone is 88.5Hz. For convenience of mobile users, repeaters that are using the sub-audible tone transmit an appropriate letter after their callsign in Morse code to indicate which tone should be used. The tones used and the geographic areas are co-ordinated by the RSGB's Repeater Management Committee. They are shown in Fig 10.4.

If the repeater receiver hears a valid tone on its input frequency it will relay the transmission. If either the carrier or sub-audible tone ceases then after a short time the transmitter will send a 'K' or 'E' in Morse which is a signal that the input of the repeater is clear and is an invitation for another user to make a transmission. If no further valid transmission is received then after a short period of time the repeater will close down. Also the repeater transmitter must identify itself in Morse code at intervals not exceeding 15 minutes. Often

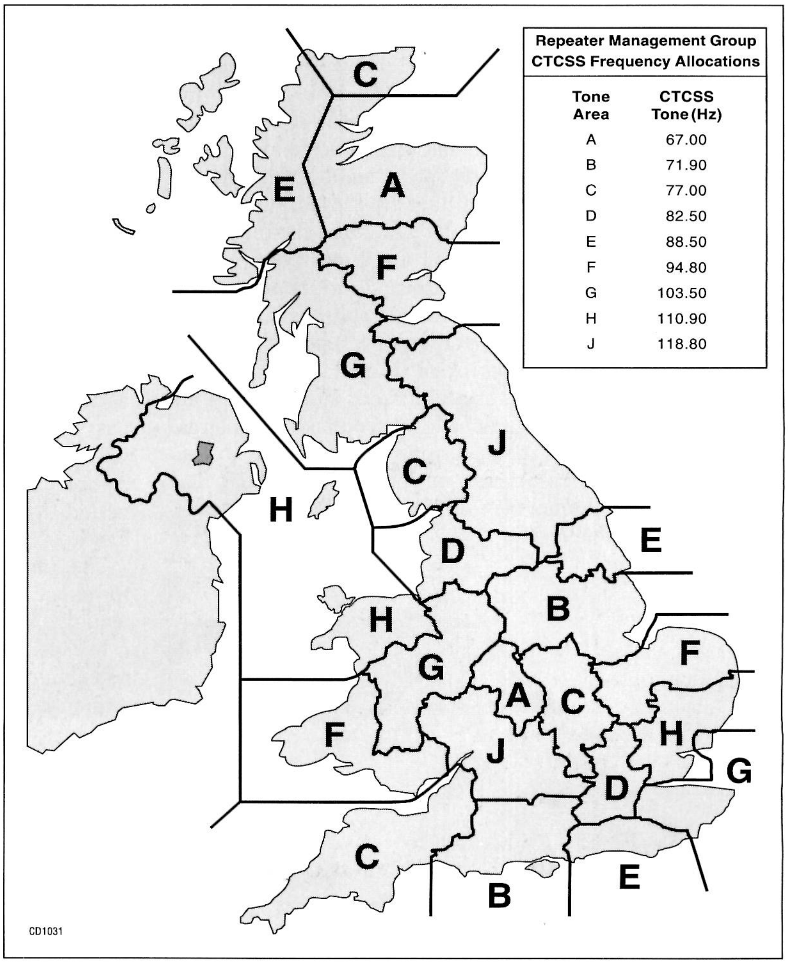


Fig 10.4. CTCSS tones in the UK

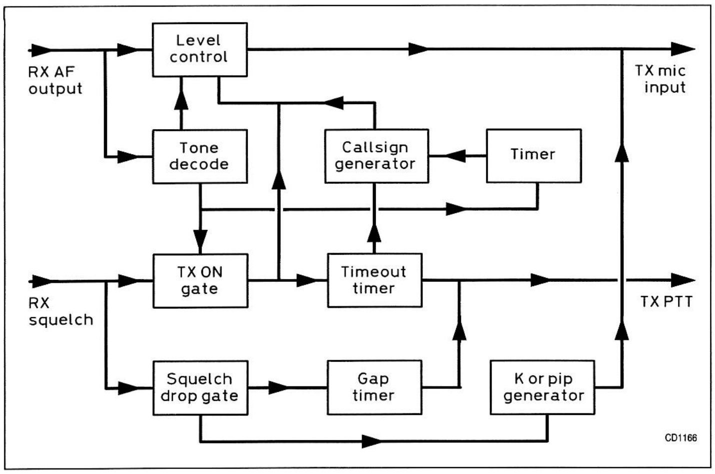


Fig 10.5. Block diagram of a simple logic system

repeaters respond with their callsign when accessed. Thus the repeater needs to be able to decide whether suitable conditions have been met before it can relay an incoming transmission, to know when to send an identifying callsign and when to close down.

As the repeater has to respond to these various situations it needs to be able to 'think' for itself especially as quite often it is remotely located. This means therefore that a logic system has to be built into it. This is often referred to simply as *the logic* by the repeater users. The control of the repeater can be accomplished by several sorts of logic systems. Some simple and very reliable circuits have been published in *Radio Communication* [6, 7]. There are other more sophisticated systems available and in operation such as the G8CUL and G1SLE logic systems. Many groups have built up their own logic systems.

Other more sophisticated techniques are beginning to appear which are microcomputer based and these can offer many other features such as remote control facilities. At least one group is using an elderly BBC microcomputer to control its repeaters including a very sophisticated interference rejection system and remote control.

HOW A SIMPLE LOGIC SYSTEM WORKS

The block diagram shows the simplest sort of repeater logic system which provides the basic minimum to cover the mandatory requirements for a working repeater as well as a user-friendly system that is simple to understand. Here is a general description of what it does. The receiver audio is fed to the toneburst decoder, producing an output which:

1. keys the transmitter;
2. feeds audio to the transmitter via the audio control;
3. starts the time out timer; and
4. activates the callsign timer.

When the user has finished talking the receiver squelch closes:

1. activating the 'K' or 'pip'
2. activating a timer determining the interval after which a callsign will be sent; and
3. resetting the timeout timer.

Thus the repeater will send a 'K' and may send a high-level callsign if there is no receiver audio present and close the repeater transmitter down. If a further transmission takes place after the 'K' then once again the timeout timer begins to time the length of the transmission. The callsign will be sent at a time governed by the callsign timer at low deviation so as not to interrupt the audio through the repeater. Many repeaters also send what is known as a *beacon callsign* when they are not in talkthrough mode which serves to notify listeners that they are within range of a particular repeater. This is relatively simple to incorporate as there is already the timer and the callsign generator available. Repeaters may send other information such as their location and signal strength of received signals, or even a busy tone to indicate that a user has timed out but is still transmitting. There are numerous possibilities including speech messages but before incorporating any of these ideas a prospective repeater builder must seek the advice of the Repeater Management Committee of the RSGB which handles repeater applications on behalf of the Radiocommunications Agency. For further information see reference [8].

Many repeater builders make some of these parameters variable so that they can tailor the repeater to the needs or wishes of local users. It has to be pointed out that the vast majority of UK repeaters conform to the pattern of logic outlined above and this has evolved after many years of experimentation by repeater builders and users. VHF repeaters tend to be very busy devices and the timeout is almost universally set at one or two minutes. It can all be summed up in the KISS acronym – "Keep it simple, stupid!" The system outlined is easy to understand and logical to use!

HOW A REPEATER TRANSMITS AND RECEIVES AT THE SAME TIME WITHOUT DESENSING ITSELF

Desensing or *desense* is a term referring to the problem of a receiver trying to listen in the same band as the local transmitter. A transmitter never transmits a single frequency but a range of frequencies distributed either side of the carrier frequency. This results in wide-band noise which is received by the adjacent receiver and prevents it from receiving any but the very loudest signals.

This desensitisation of the receiver is usually referred to as *desensing*. How is this problem overcome? The answer is to use very selective filtering, not only in the receiver but also in the transmitter. The filtering system is usually called a *duplexer* and the process as *duplexing* as the filters allow the transmitter and receiver to operate on their two separate frequencies simultaneously, ie duplex operation, as opposed to simplex operation where the receiver or transmitter cannot operate simultaneously.

A typical duplexer is made up of three cavity filters or *cavities* in the receiver input and three cavities in the transmitter output (see Fig 10.6). Each cavity is basically a very-high-Q filter which therefore has a very high loss either side of its resonant frequency. See Fig 10.7.

The three cavities in the receive leg are tuned to the transmitter frequency so as to give maximum rejection of the transmitter. This ensures that the receiver front-end circuits are not driven into overload by the strong carrier. The transmit

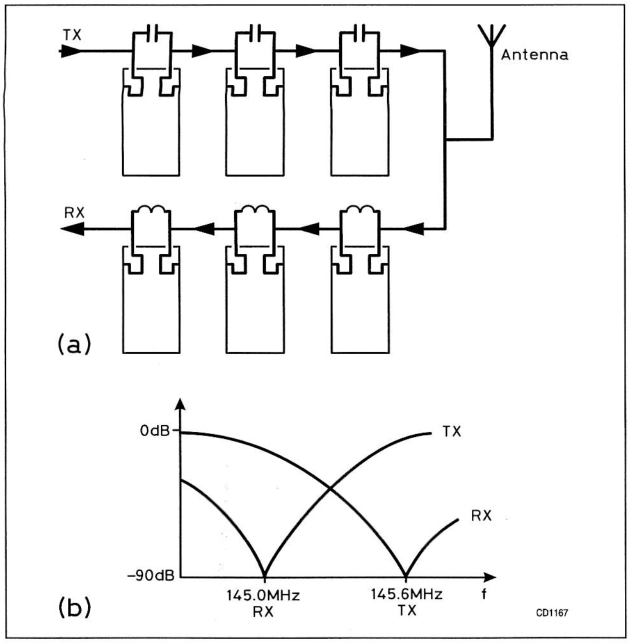


Fig 10.6. (a) Block diagram of duplexer, three-cavity receive/three-cavity transmit. (b) Response curves

cavities are tuned to the receiver frequency so that they attenuate the wideband noise generated by the transmitter as much as possible on the receiver input frequency. This enables the receiver to hear weak signals.

Fig 10.6(b) shows the response curve of each set of cavities at the receive and transmit frequencies. For the purposes of this diagram it has been assumed that the repeater receiver frequency is 145MHz and the transmitter frequency is 145.6MHz. However, there is a slight difference in the set of cavities used in the receive leg compared with those used in the transmit leg. Those in the transmit leg are wired up with a capacitor in parallel whilst those in the receive leg have an inductor in parallel. These additional components have the effect of skewing the filters' skirts so that the attenuation effects can be adjusted to allow little or no attenuation of the transmit frequency by the transmitter cavities and little or no attenuation on the receiver input frequency by the cavities in the receiver input. In practice this is usually adjusted on site to give the performance necessary. See Figs 10.8 and 10.9.

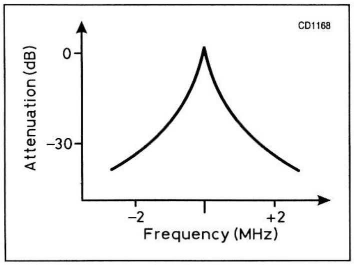


Fig 10.7. Response of a single cavity

As a typical receiver sensitivity is of the order of –130dBm and the transmitter generates noise on the receiver frequency at about –40dBm it is necessary to achieve an attenuation of about 90dB, ie –130 – (–40) = –90 in order for the receiver to be able to receive as well as when the transmitter is off.

In practice each cavity is capable of producing a rejection notch of 30dB so the repeater needs three in each leg to provide approximately 90dB. See Fig 10.10.

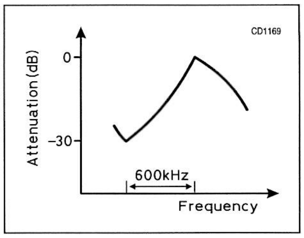


Fig 10.8. Response of single cavity with parallel capacitor

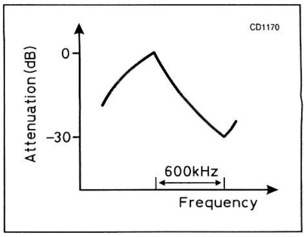


Fig 10.9. Response of single cavity with parallel inductor

USING REPEATERS

Using repeaters involves *duplex* operation where you transmit and receive on different frequencies, and most available transceivers have this facility built in. There is a user's code of practice. First, repeaters are primarily intended for *mobile* or *portable* users. It is definitely unacceptable to 'hog' them for long lengths of time. Always welcome newcomers to join in, encourage membership of the group, and remember your normal amateur codes of practice are as valid on repeaters as on simplex operation (use of callsign, courtesies etc).

Before attempting to transmit, ensure that:

(a) Your transmitter and receiver are on the correct frequencies (remember the repeater split).
(b) Your tone access (if fitted) is operating correctly.
(c) Your peak deviation is set correctly (some repeaters will not relay your signal if this is incorrect!)

Any adjustments you have to make should be done into a dummy load, *not on-air!*

Avoid using the repeater from your base station; it is really intended for the benefit of local mobile and portable stations. If you really do intend to try it from a fixed station, then use

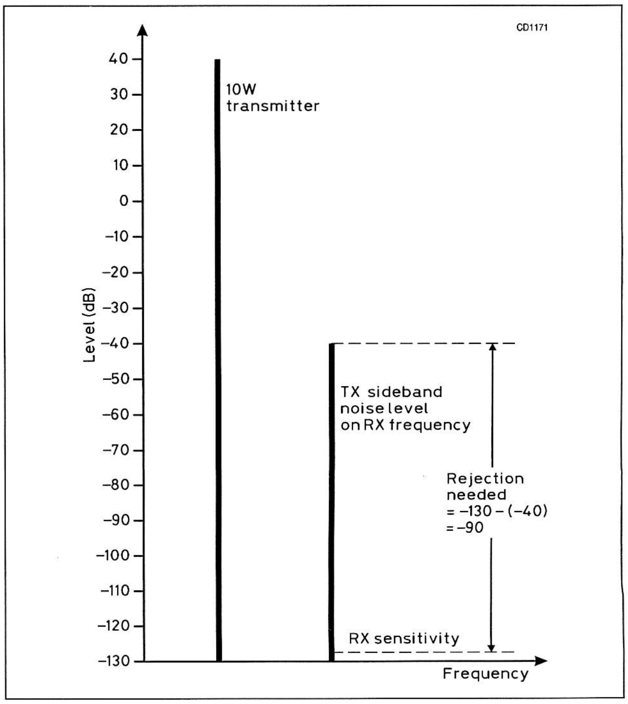


Fig 10.10. Transmitter-receiver levels for a repeater

the lowest power to get into the repeater (under 1W is acceptable in the majority of situations where you can hear the repeater well).

Always listen before transmitting. Unless you are calling another specific station, simply announce that you are "listening through", eg "GM8LBC listening through GB3CS". On the other hand, if you are responding to someone specific, try something like "GM0ZZZ from GM8LBC".

Once contact is established:

(a) At the beginning and the end of each over, you need give only your own callsign, eg "from GM8LBC".
(b) Change frequency to a simplex frequency at the first opportunity, especially if you are operating from a fixed station.
(c) Keep your overs short and to the point, or they may time out, and do not forget to wait for the 'K' or 'blip' if the repeater uses one.
(d) Do not monopolise the repeater when busy; others may be waiting to use it.
(e) If your signal is very noisy into the repeater, or if you are only opening the repeater squelch intermittently, finish the contact and try again later.

JOINING UP

Repeaters have initiated many newcomers to the amateur radio hobby. Repeater outputs can be monitored easily, either using widely available amateur equipment, or perhaps by a scanner. Repeater groups themselves have been able to join up prospective users and, through local meetings or contact, newcomers have been shown how to proceed. Many amateurs use repeaters in addition to their other amateur activities, perhaps using their local repeaters whilst travelling to and from work, then doing something entirely different in the evenings!

If you enjoy operating through your local repeaters then remember that they do cost money to run and this is done on a purely voluntary basis, sometimes by individuals but mostly by groups who would appreciate contributions towards the cost of running the repeater. Apart from the initial costs of the equipment there are the on-going costs of electricity and site rental. Site rentals for advantageous sites such as those of the BBC and NTL can cost several hundreds of pounds a year even at the favourable rates negotiated either by the local repeater group or by the RSGB.

References [1] to [5] give details of the repeater keeper who is the person to contact to offer your help to.

Amateur radio 'purists' have, particularly in the early years of repeater growth, shunned their existence, believing them not to be truly in the 'ham spirit'. Twenty-five years on, this is very much a minority view. The presentation of so many strands for devotees to find their niche is a strength of our hobby. Repeater builders are amongst the most technically competent and experienced people in amateur radio, and many are professionally employed in PMR or other professional communications. Repeaters are often co-sited with other major broadcasters or PMR users, and so have to be of a sufficient technical standard to co-exist.

Repeaters have sometimes become the target for abuse, generally by men with limited vocabularies (using profanities) or with strange voices, ('squeakies'). There have been no female abusers to our knowledge. The only way to treat the abuse in whatever form it takes is to ignore it completely. Any attempt to remonstrate with them encourages them since they know they have an audience. If possible, take a bearing on the repeater input frequency, make a tape recording of the abuser and note the time, date, frequency etc. This information should be sent to the Repeater Abuse Co-ordinator at RSGB Headquarters. Further information on how to deal with abuse problems is available from the same address.

If you feel there is a need for a repeater in your area then make sure you contact the Repeater Management Committee of the RSGB, c/o RSGB HQ and read reference [8] before you do anything else.

REFERENCES

[1] *Amateur Radio Operating Manual*, 4th edn, ed R Eckersley, G4FTJ, RSGB, 1995.
[2] *The Repeater List*, a printout from the RSGB HQ computer containing the very latest information (updated monthly).
[3] *RSGB Yearbook*, 1998 edn, ed B Rider, G4FLQ, RSGB, 1997.
[4] *The Amateur Radio Diary 1997,* ed M Bamber, G0SHY, Bambers, 1996.
[5] An Internet link to RMC Online is given in the RSGB web site at www.rsgb.org.
[6] A J T Whitaker, G3RKL, *Radio Communication* 1980, pp34–42; A J T Whitaker, G3RKL, *Radio Communication* 1982, pp30–31.
[7] A J T Whitaker, G3RKL, *Radio Communication* 1983, pp882–885 and pp990–993.
[8] *The Guide to Repeater Licensing*, 1997 edn, RMC, RSGB, 1996.

11 Test equipment, methods and accessories

IN the VHF region and above there is still a fair amount of scope for experimentation and home construction as well as the use of the so-called 'black box'. There are many items of ex-commercial equipment that can be obtained at reasonable prices for adaptation.

In the case of a receiver, where low noise is of paramount importance, it is extremely difficult to adjust an input stage or preamplifier for the best signal-to-noise ratio unless a noise generator is used.

Details of these and other useful devices are described in this chapter. They are generally straightforward and, provided care and attention to detail is taken, satisfactory and reliable performance should be achieved. Readers are also referred to *Test Equipment for the Radio Amateur*, published by the RSGB [1], which gives details of other test equipment.

ANTENNA MEASUREMENTS

To keep the performance of any VHF/UHF station as near optimum as possible the antenna system should be properly tuned to start with and maintained in that condition.

To tune up any antenna 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 a completely different performance when the array is taken outside.

Undoubtedly the most effective apparatus for tuning up any antenna system is a standing-wave indicator or reflectometer. If there is zero reflection from the load, the standing-wave ratio (SWR) on the antenna feeder is unity (1:1). Under this condition, known also as a *flat line*, the maximum power is being radiated. All antenna matching adjustments should therefore be carried out to aim at a standing wave better than about 1.5:1 (some 4% reflected power). Many modern transceivers have SWR protection circuits built-in and these may cut in and start to limit power output. Consult the manual for your equipment if in doubt.

If suitable apparatus is not available, the next best course of action is to tune the antenna for maximum forward radiation. A convenient device for this is a field-strength meter comprising a diode voltmeter connected to a λ/2 dipole placed at least 10λ from the antenna. When adjustments have resulted in a maximum reading on the field-strength meter, the SWR may not be unity and therefore some power may be wasted. However, if the best has been done with the resources available it is highly likely that good results will be achieved.

Reflectometer for VHF

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 antenna system. Such a condition is rarely, if ever, achieved due to the impossibility of presenting the transmission line with a perfectly matched load. In practice, it is possible only to terminate the line with an antenna or load which approaches the perfect condition. Under these circumstances a certain amount of power is reflected at this mistermination 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 misterminates the line.

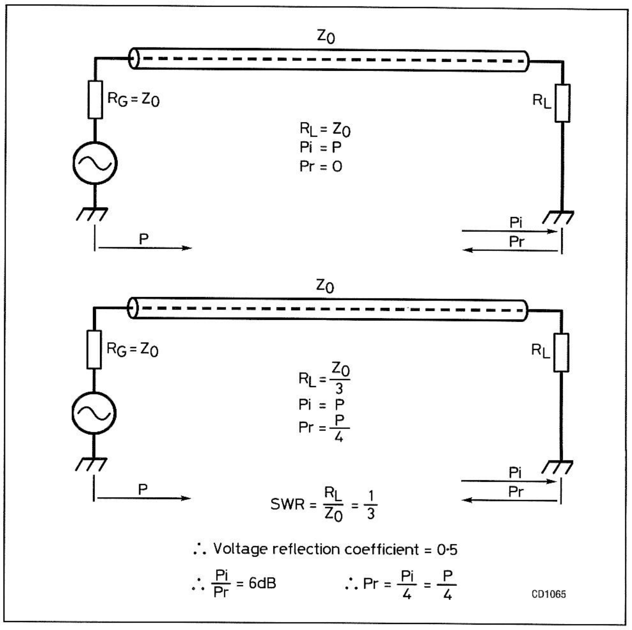


Fig 11.1. Effects of mistermination on a transmission line in terms of the incident and reflected power at the load

The amount of power reflected from the antenna or load mistermination is directly proportional to the magnitude of the mismatch on the line. 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 11.1).

If the SWR = S, then the voltage reflection coefficient K is given by:

$$K = \frac{S-1}{S+1}$$

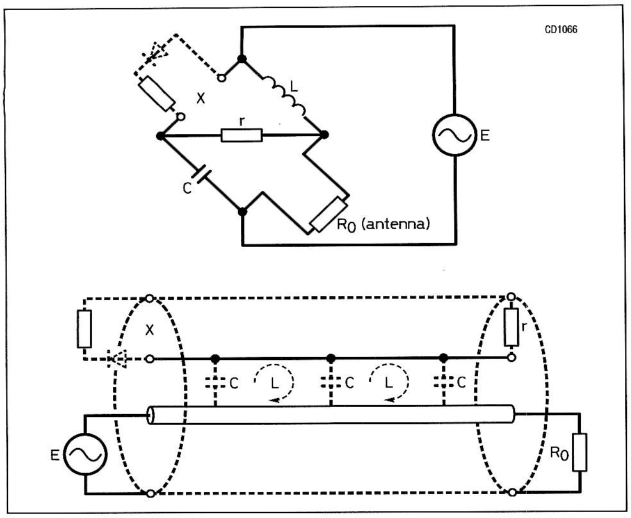


Fig 11.2. Maxwell bridge representation of transmission line coupler

Clearly, if a device can be constructed which will differentially respond to power in terms of direction, then it can be used directly to measure standing wave ratio, and the ratio M of incident to reflected power is given by:

$$M = 20 \log_{10} 1/K \text{ dB}$$

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 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 11.2). 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 11.3. The accuracy of such an instrument depends on the correct termination of the sampling lines. Any mismatch on those lines will result in a standing wave along them, and consequently the RF 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 the same line when the instrument is reversed. The directivity is usually expressed as a ratio in decibels (dB).

Design aspects

Before the details of construction can be finalised, it is necessary to consider one or two design aspects of the instrument

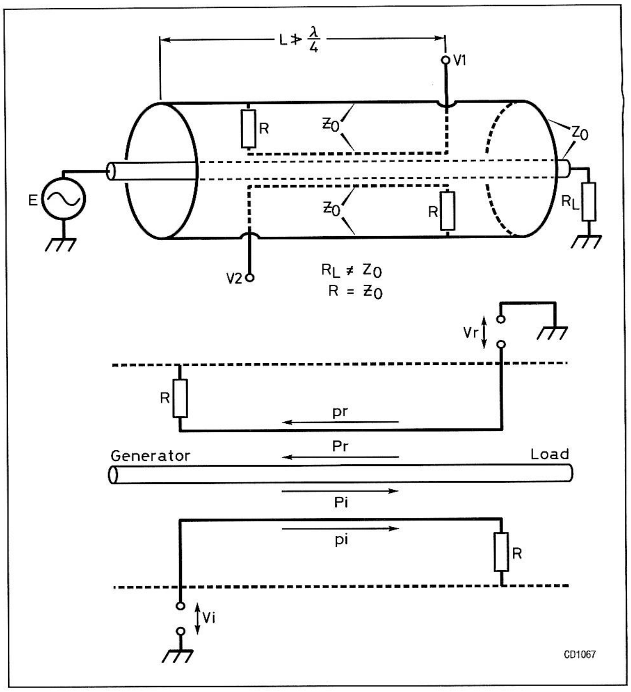


Fig 11.3. Arrangement of sampling lines to respond respectively to incident and reflected powers

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 voltages are still of a radio frequency nature and it is necessary to convert them to DC before they can be used to drive a moving coil meter.

If the forward voltage is arranged to produce a full scale deflection of the meter, then clearly the meter can be calibrated directly in SWR 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. The calibration will be valid independent of the actual transmitted power, since in each case the meter is adjusted for FSD.

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 RMS voltages applied to the rectifier diodes. This places an inherent limit on the sensitivity of the instrument at low SWR. However, 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 x dB in the lowest SWR which can be measured for a given deflection of the meter (Fig 11.4).

Care must be exercised that the coupling from either line 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_0 of the main line and hence introduce an inherent SWR in the instrument itself.

As a general rule the coupling should not be greater than

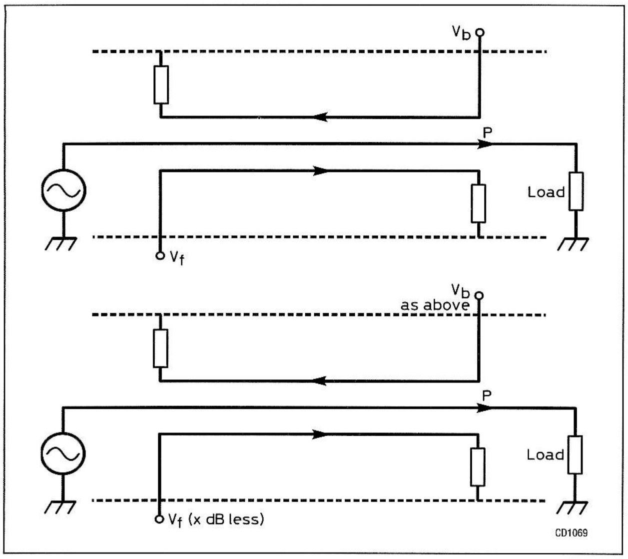


Fig 11.4. Instrument sensitivity and coupling ratios. (a) Sampling couplings equal. V_f gives meter FSD, so V_b/V_f (say *y* dB) which corresponds to a given meter deflection. (b) Sampling couplings different by *x* dB. V_f gives meter FSD less *x* dB for same power, SWR V_b/meter FSD, so for same deflection as (a) SWR = *x* + *y* dB

30dB to maintain an inherent reflection coefficient of less than 3–4%.

When the main line is carrying power which is subject to amplitude modulation, then the sampling voltage from the forward (and 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 meter to full scale, then if a rectified (or detected) signal is once more rectified, a DC 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 FSD 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 DC voltage to cause large excursions of the meter reading under full modulation conditions, ie it is not possible to advance the meter to FSD for 100% modulation. It is recommended therefore that the 'modulation meter' aspect of the instrument be regarded only as of an arbitrary quantitative nature.

A circuit diagram for a typical reflectometer is given in Fig 11.5. The diodes used *must* be capable of the frequency range the instrument is to cover.

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 lines 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 out conductors, ie

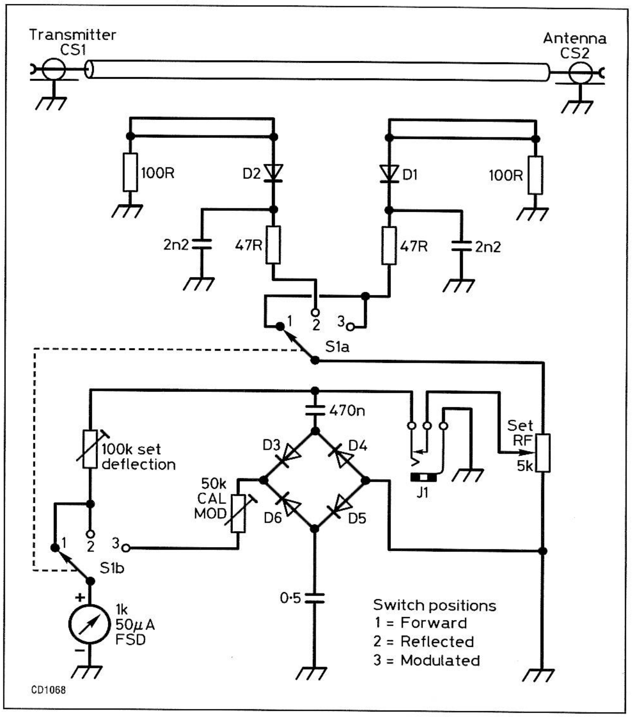


Fig 11.5. Circuit diagram of the reflectometer. D1, D2: OA91; D3–D6: 1N4148. D3–D6 should be bypassed by 1nF ceramic capacitors across each diode

this is a physical requirement. These discontinuities are of the right-angled step type (Fig 11.6), 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 die-cast box 114 × 89 × 55mm (formerly 4½ × 3½ × 2in), with a partition running the length of the box to form an almost square cross-section (51 × 51mm) into which the trough line is assembled. Any other spacing may be used but this complicates the calculation of Z_0 of the line.

The characteristic impedance of a coaxial line with a cylindrical inner conductor and a square outer is given by:

$$Z_0 = 138 \log_{10} L/d$$

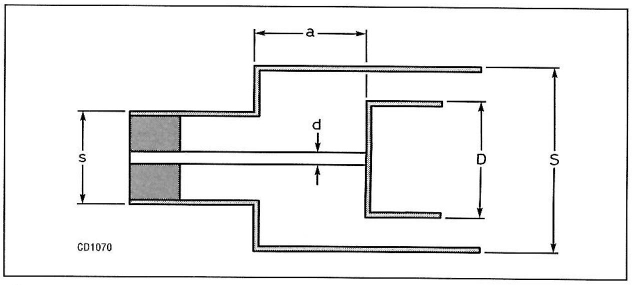


Fig 11.6. The characteristic impedance Z_0 is given by 138 $\log_{10} s/d$ which is also 138 $\log_{10} S/D$. The optimum step-length *a* is a function of Z_0 and D/d

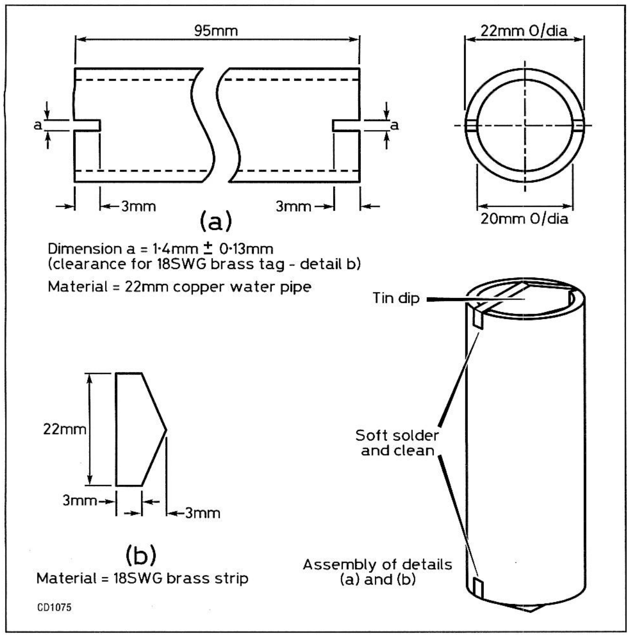


Fig 11.7. Construction of main line inner conductor

where L is the length of the side and d is the diameter of the inner conductor. It is assumed that $L/d > 1.5$. For a 50Ω line, substitution of 51mm in this formula gives $d = 22$mm.

The RF connections to the box can be made from N-type, BNC, TNC or SO239 type connector, the last being a non-constant impedance type. When soldering to these it is wise to insert a plug into them – this will hold the middle pin central as prolonged soldering usually softens the surrounding insulation.

The sockets are mounted centrally at each end of the 51mm 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 about 6.5mm. The inner conductor detail (Fig 11.7, detail a) is slotted at each end for a depth of 3mm and wide enough to accept 18 SWG (1.4mm) 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 11.7, detail b) are cut from 18 SWG (1.4mm) brass sheet and pushed into the slots at each end as shown and soldered in position. The pointed end of each tab is then tinned, any surplus solder being removed in order to keep the cylindrical shape at the ends. The inner assembly may then be rested between the spigots of the coaxial sockets and soldered into position (Fig 11.8).

The sampling lines are formed from a strip line of 18 SWG (1.4mm) brass lying parallel to the partition. The formula for the characteristic impedance of a strip line over an infinite plane is:

$$Z_0 = 230 \log_{10} 4D/W$$

where D is the distance from the plane, W is the width of the strip and the ratio D/W has a value between 0.1 and 1.0. As already explained, it is necessary to terminate the sampling lines correctly in order to preserve the directivity of the instrument, and a characteristic impedance of 100Ω is used, based upon the use of available 100Ω, 2% tolerance 0.5W resistors as the terminating loads. This figure substituted in the above expression gives a value of D/W of 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 $D = 6.5$mm and hence $W = 9.5$mm was chosen by experiment. The sampling lines were made as long as conveniently possible, care being exercised to make them as near as physically identical as possible.

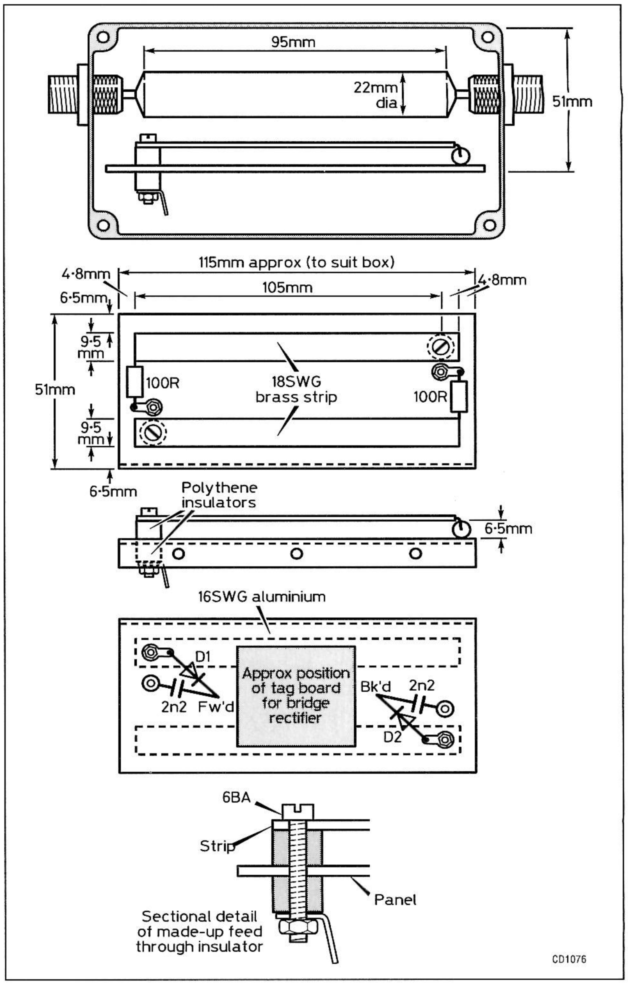


Fig 11.8. Arrangement of strip line on partition and tag board

The partition is made from 16 SWG (1.6mm) aluminium sheet and the sampling lines mounted in the positions shown in Fig 11.8. 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 assembled with sampling lines, tag board on the rear, and all components, before being fitted in 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. An alternative and neater solution would be to mount most of the components on a PCB instead of the tag board, but keeping D1, D2, C1 and C2 in approximately the positions shown.

The position of the various potentiometers and switches is not critical, and some alteration to the suggested layout is permissible. Alternatively there is no objection to extending

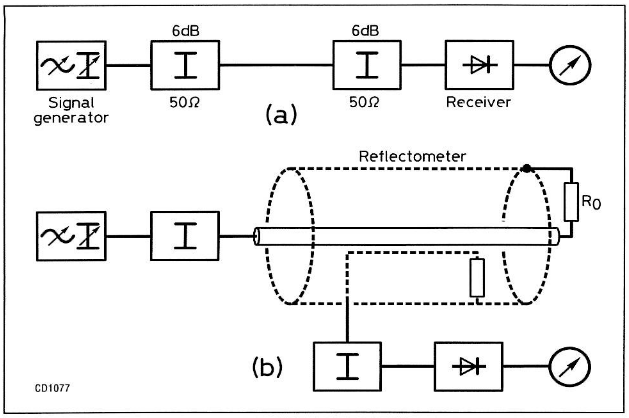


Fig 11.9. Insertion method for measuring coupling. (a) Set a signal generator to give an arbitrary deflection *K* on receiver meter. Note the signal generator attenuator setting, say *x* dB. (b) Repeat the exercise with reflectometer in circuit and readjust the signal generator to give the same deflection *K* on the receiver meter. Read the attenuator setting, say *y* dB. Then the coupling of the main line to the sampling line is *x* – *y* dB

the DC outputs of D1 and D2 via a three-core cable to another position. D1 and D2 must be fitted as per Fig 11.8.

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.

First terminate the antenna end of the instrument and measure the coupling of each sampling line in turn by the insertion method (Fig 11.9). Adjust the sampling line spacing for identical coupling.

Then, using the signal generator injecting directly into each sampling diode in turn (with sampling lines disconnected), calibrate the indicating meter in terms of decibels relative to the injection voltage for FSD.

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 decibels, for all transmitted powers, provided it is always adjusted to FSD on the forward position using the SET RF control. (The SET DEFLECTION control should be set, for any particular meter, to such a value as to allow the SET RF control to function over the whole range of transmitted powers expected.)

Many amateurs will, of course, not have the necessary test equipment outlined above available to them. 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 usually reduce as the SWR on the main line is reduced. Thus the reflectometer may be used qualitatively to indicate best SWR when adjustments are being made to, say, an antenna system.

It is possible, without any test equipment other than a low-power transmitter, to make some basic checks on the instrument as follows.

Table 11.1. Reflectometer calibration figures (72Ω)

Meter reading			
Forward	Backward	Level (dB)	SWR
50	50	0	∞
43	44	–2	8.8
37	36	–4	4.4
30	29	–6	3.0
23	21	–8	2.3
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

With an open-circuit on the antenna 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 characteristics of the diodes, and also enable slight adjustments to be made to the sample lines to equalise 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 output PA device 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 2–3dB. Inspection of the calibration table shows that for the lower values of SWR such an error results in a very small error in SWR, this becoming increasingly worse as the SWR gets larger. Therefore, an uncalibrated but carefully built instrument can be expected to indicate SWR 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 SWR of column of Table 11.1 represents the conversion of backward meter readings for a forward reading of 50. For a given input level, the difference between the lines was less than 1dB over the full range. Zero level is equivalent to 1V RMS in 100Ω.

Power limitations

The sensitivity of the instrument is such as to provide FSD on a 50μA meter for a carrier power of 5W. The upper limit is set by the dissipation in the resistors terminating the sampling line. These are rated at 0.5W and, since the forward line is dissipating power 32dB down on the incident transmitted power, the maximum transmitted power should not exceed 500W carrier.

Frequency range

The performance of the instrument is constant over the 144MHz band. The sensitivity will fall linearly with 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

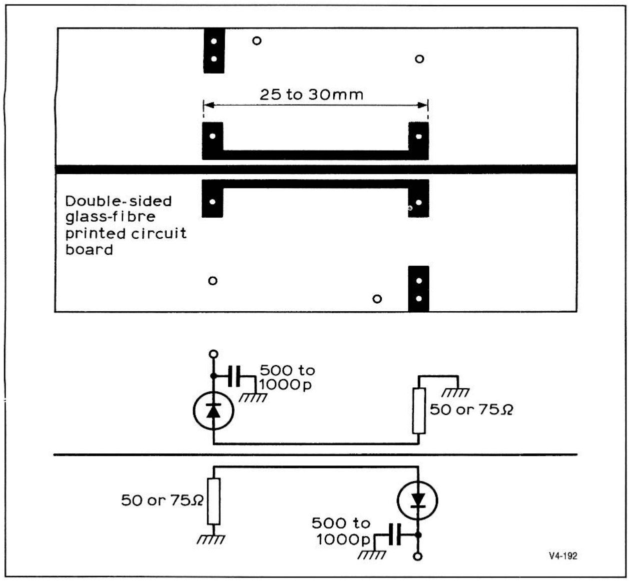


Fig 11.10. Printed circuit directional coupler

also the variations in the terminating loads on the sampling lines, which will become increasingly reactive.

Printed circuit directional coupler

A simple method of constructing a directional coupler is by use of a printed circuit board as shown on Fig 11.10. The line impedance can be made in accordance with the design information given in 'Tuned circuits' in Chapter 3. The coupling lines may be of any convenient length to suit the meter in use but they should be short compared with $\lambda/4$.

The diodes should be a signal type such as a silicon Schottky barrier type (eg BAT83, 1N6263 etc) or a germanium type (eg OA47, OA91 etc) which have low forward voltage drop and suitable for the frequency range concerned. The terminating resistors should be as far as possible non-inductive types of good stability. The bypass capacitors may be disc, plate or feedthrough type, the latter having the advantage of providing a terminal for connection to the meter. It is important that the actual value should be suitable for the frequencies to be used.

General-purpose directional coupler

A reliable directional coupler can be made employing readily available items as an alternative to the above PCB type and without recourse to machine tools. A short section of air-spaced coaxial line is used with the coupling loops inserted into the line through the slots in the outer tube.

The general arrangement is shown on Fig 11.11. The whole unit is assembled on a piece of single-sided, copper-clad PCB made to fit a die-cast box, say, 92 × 38 × 31 mm (eg Eddystone 27969P). The size is not too important but should be kept small. The outer of the coaxial line is made from a piece of 8mm copper tube with its ends opened out by cross-sawing and slitting. This is attached to the copper side of the PCB. The inner conductor is made from a piece of brass or copper rod/tube – for 50Ω use 3.5mm rod and for 75Ω use 2.5mm.

The coupling loops are made from 24 SWG (0.5mm) brass or copper strip mounted on four small stand-off insulators. The spacing between them and the inner line should be equal and adjusted so that each provide the same readings when used either way it is connected. This should be carried out with the output socket connected to an appropriate dummy load.

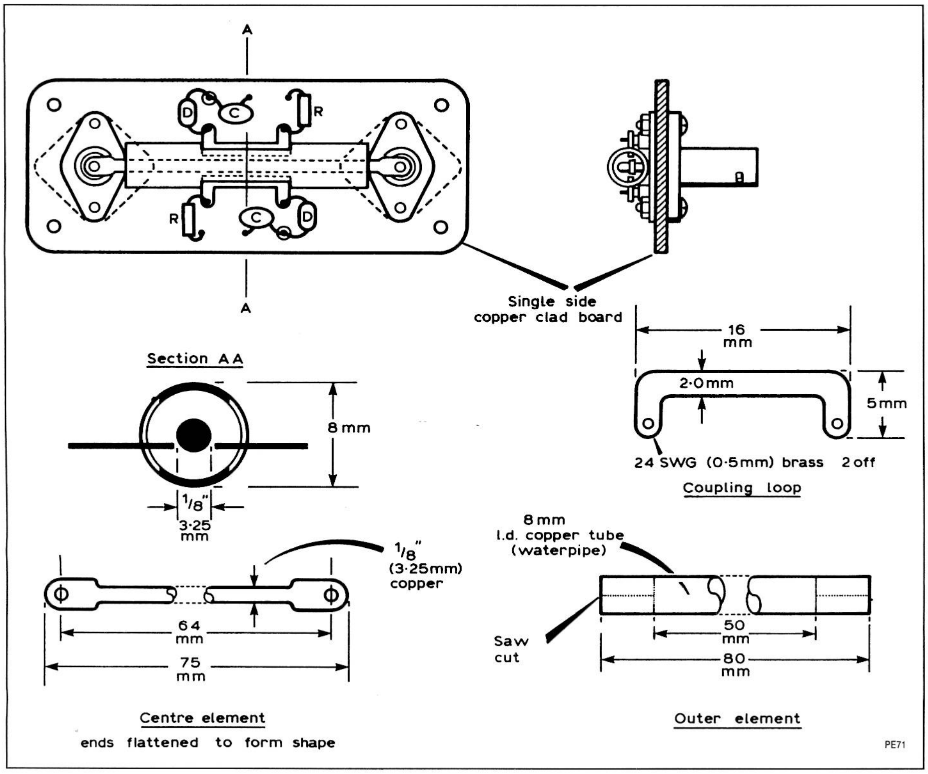


Fig 11.11. General-purpose directional coupler

Resistive VSWR bridge

Measurement of VSWR below about 450MHz can often be accomplished more conveniently by the use of a resistance bridge rather than a slotted line which is cumbersome at VHF. This method is also suitable when using other test equipment of only low power.

As already discussed in this chapter, VSWR measurements are most frequently associated with antenna and feed systems yet it could also be that the input impedance of an amplifier or similar equipment needs to be examined. It becomes more important as frequencies are increased and where impedance discontinuities give rise to unexpected losses.

A home-constructed bridge

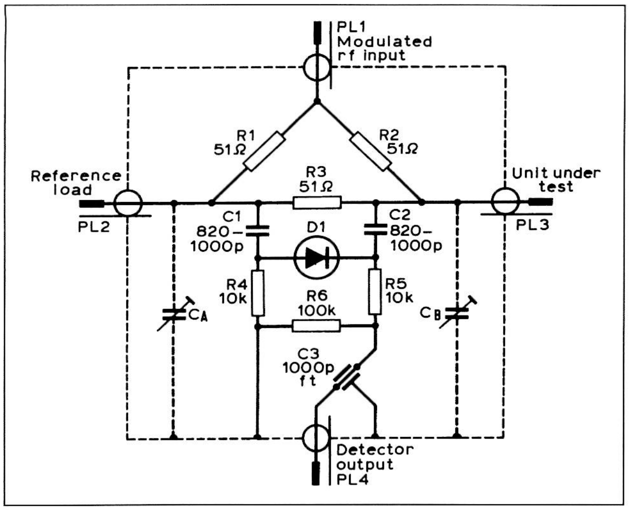


Fig 11.12. Circuit diagram of VSWR resistive bridge

can be built without difficulty, but for best performance some care in detail is needed, notably in the choice of matched resistors R1 and R2. Their installation is as far as possible identical both in respect of their lead length (which should be kept to a minimum) and their location (their relation to the ground plane and connections).

The circuit diagram of the bridge is shown in Fig 11.12, while in Fig 11.13 two methods of construction are shown, both based on a small die-cast box. In method (a) the BNC connectors are attached to the sides of the box and the circuit components fitted to a single-sided, copper-clad, glassfibre PCB. The PCB is positioned so that it is at the level of the insulation projecting from the BNC connectors so that the important resistors R1, R2 and R3 rest on the copper ground plane when soldered in position. If necessary stand-offs can be used for the junctions of components.

In method (b) the whole assembly is fitted to the lid of the die-cast box with the copper-clad PCB fitted within the raised edge of the lid. In this form it is easier to make good

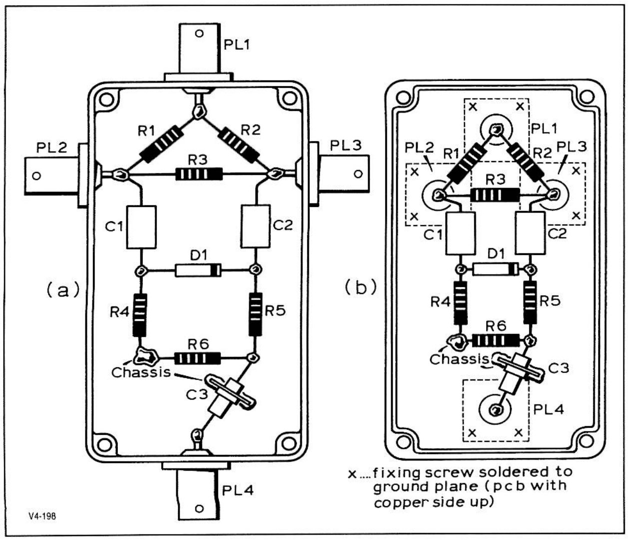


Fig 11.13. Two alternative methods of construction of the bridge

connections between the board and the four connection sockets, which may be soldered in position.

In order to test the bridge, a modulated RF signal is connected to port PL1 and an audio detector to PL4. Then, with two reference loads of the same impedance (in this case 50Ω) connected to PL2 and PL3, the output detector should read a very low value. If there is any significant output, a small capacitor (such as a ceramic plate soldered close to either PL2 or PL3 as indicated by CA or CB) should be added and adjusted to obtain balance and reduce any residual signal. Once substantial balance has been obtained, the next step is to remove one of the two reference load resistors from PL2 or PL3 – the detected output should rise by some 30dB. If the reference loads to PL2 and PL3 are interchanged no difference should be detected.

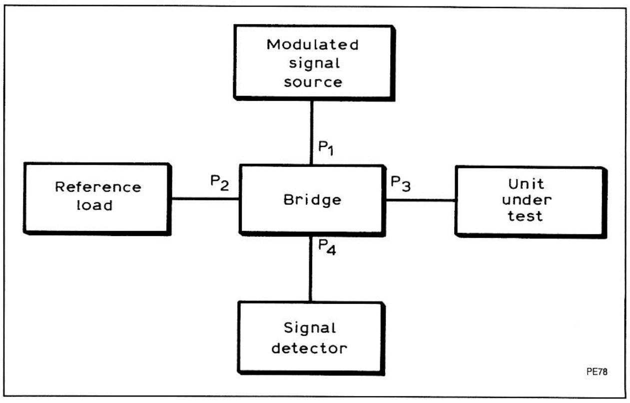


Fig 11.14. General set-up of the bridge

For a 50Ω bridge, it is useful to have one or two fixed-value mismatch loads available such as a 75Ω one for 1.5:1 VSWR, a 25Ω one for 2:1 and a short-circuit for infinite VSWR.

Operation of the bridge is readily appreciated from Fig 11.14. If identical reference loads are connected to PL2 and PL3, signals will be equal and in phase so that no output should be observed at PL4. When the unit under test at PL3 is different from that of the reference load a difference signal will be observed at PL4 which is proportional to this difference.

The bridge resistors R1 and R2 may be any appropriate value such as 50Ω or 75Ω; 50Ω is of more general application.

MEASUREMENT OF RF POWER

The UK Amateur Licence requires that you should be able to measure output power. Power is normally first assessed when running into a resistive dummy load which presents the correct load to the transmitter. Such a load is required in any case to permit non-radiating adjustments to be made to the transmitter.

Power measurements when running into a dummy load can be made directly using RF voltmeters, oscilloscopes etc, taking into account their frequency limitations. The use of reflectometers is preferable for monitoring power when coupled to an antenna system as they also give an indication of what is happening on the feed system – these are covered earlier in this chapter.

Definitions of power

The following information is taken from the *Amateur Radio Licence Terms and Limitations Booklet* BR68 [2]. Only the relevant paragraphs have been included.

Notes to the Schedule

(a) *Maximum Power* refers to the RF power supplied to the antenna. Maximum power levels will usually be specified by carrier power. For emissions having a suppressed, variable or reduced carrier, the power will be specified by the *peak envelope power* (PEP) under linear conditions.

(e) Interpretation

(i) *Carrier power:* The average power supplied to the antenna by a transmitter during one radio frequency cycle taken under the condition of no modulation.

(iv) *Mean power:* The average power supplied to the antenna by a transmitter during an interval of time which is sufficiently long relative to the lowest frequency encountered in the modulation taken under normal operating conditions.

(v) *Peak envelope power (PEP):* The average power supplied to the antenna by a transmitter during one radio frequency cycle at the crest of the modulation envelope taken under normal operating conditions.

The effect of modulation on power output

This section deals with the basic measurement of power, without specifying measuring equipment, of either basic transmitters or an amplifier. For measurement of modulation parameters the reader is referred to later in this chapter.

In a carrier-wave situation (CW) or with a frequency-modulated signal, the output is of constant amplitude and so it is relatively easy to measure the output power. Key the transmitter and determine the RMS voltage (V_{RMS}) of the resulting carrier across a dummy load (R). The power is given by:

$$P = V_{RMS}^2/R \quad \text{watts}$$

If the signal is amplitude modulated (double sideband with carrier) then the overall output power increases. The power is divided between the sidebands and the carrier component. With 100% modulation the output power increases to 1.5 times the unmodulated condition – the power contained in each of the two sidebands is one quarter that in the carrier. It is suggested that for this form of modulation the carrier power is measured (ie with no modulation) as described above. This value can be multiplied by 1.5 to give the maximum output power available.

With single sideband modulation, no power is output until modulation is applied. The output envelope is non-sinusoidal in appearance. The normal method for measuring output power is by observation of the modulation envelope and determination of the peak envelope power – this is the parameter defined by the licensing authority. Equipment for making these measurements is described in the following sections.

Dummy loads

A dummy load is a resistor (or group of resistors) which has the same resistance value as an antenna system. It should be purely resistive and so should provide an SWR of 1:1. The dummy load is normally constructed so that it provides minimal radiation when a transmitter is operated into it. Transmitters should always be set up into dummy loads before connecting them to the antenna system.

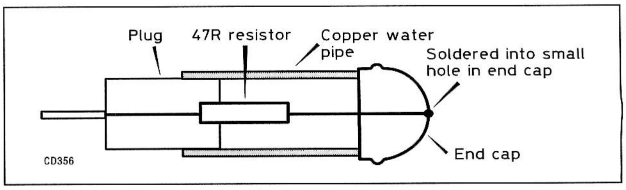


Fig 11.15. Typical construction of low-power dummy load

A resistor, no matter what type, will always have associated with it inherent inductance and capacitance, and the way it is mounted will also affect these values. The ideal resistor is one which has no associated capacitance and inductance, and is also one which does not change its value appreciably with frequency and power dissipation. This is unfortunately difficult to arrange in the real world, and the best one can do is to choose a resistor which minimises these adverse effects. In practice, impedance presented by the dummy load changes with frequency and hence will not provide an SWR of 1:1 – this effect is more pronounced as frequency increases. This is why any dummy load which is purchased should have some information included with it concerning frequency range and expected SWR values.

The best type of resistor to use is that made from carbon. Unfortunately it is becoming increasingly difficult to obtain power ratings in excess of 2W from distributors but tubular carbon resistors of higher power ratings will often be seen at rallies. *Never use wirewound resistors for RF.* However, these may be adequate for measuring AF power.

A low-power dummy load can be made from a single 47Ω resistor with surrounding shield as shown diagrammatically in Fig 11.15, and this is obviously easier for those with mechanical skills and some ingenuity. To increase the power dissipation it would be possible to make the metal container a tight fit around the resistor. However, this may pose problems if conduction can occur from the resistor case. Providing a small clearance can be ensured around the resistor, then the space could be filled with heatsink compound which is thermally but not electrically conducting. Alternatively one could fill the case with cooling oil and/or put fins onto the outside of the case. The use of the metal shield prevents unwanted radiation and also provides a low-inductance path.

To increase the power rating it is possible to use resistors in parallel – Fig 11.16 shows a typical arrangement. These should, if at all possible, be encased in a metal shield to prevent unwanted radiation, possibly a perforated shield to permit air

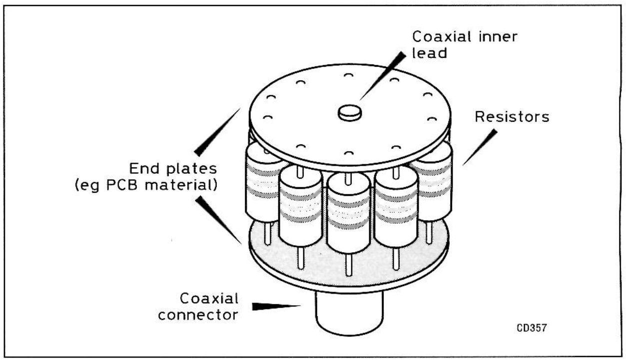


Fig 11.16. A multi-resistor dummy load

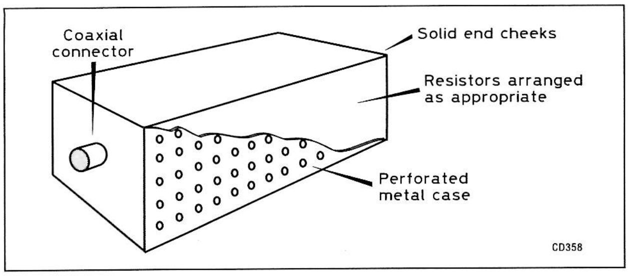


Fig 11.17. Possible construction of a higher-power dummy load

flow – see Fig 11.17. It might even be possible to build an RF probe and/or attenuator for measurements as suggested elsewhere in this chapter.

The characteristics can be improved by arranging the resistors in a coaxial manner. Ideally the pitch circle of the resistors and centre coaxial conductor should be carefully calculated but, as this arrangement tends to be very short compared to a wavelength, non-adherence has little effect until the higher frequencies are reached. The characteristic impedance can be calculated using:

$$Z_0 = 138 \log_{10}(D/d)$$

where D is the pitch diameter of the resistors and d the diameter of the inner coaxial connector. Typical arrangements to make approximately 50Ω are shown in Table 11.2.

The overall power rating is the sum of the power ratings of each resistor used.

To obtain higher power ratings it may be possible to place the load resistors in a perforated screened container and air blow them. This could be accomplished by placing a thermal switch on the resistors and using it to switch a fan on once the temperature has risen above a certain point.

Sometimes large tubular carbon resistors come onto the surplus market, eg from firms such as Morganite. These can make excellent dummy loads. Try to form them in a coaxial manner with the feed up the centre. Again, air blowing can be used to increase the power dissipation.

When using a dummy load, remember that it may be possible to dissipate a much higher power for a short period of time, providing a long cool-down period is allowed between applications of power. A commercial dummy load may often be provided with advice on this method of use.

An interesting development is the production of power film resistors in TO-126 and TO-220 packages with power ratings of 20W at +25°C case temperature when mounted on a heatsink of some 6°C/W. These resistors show good characteristics up to at least 300MHz. Typical of these are the MP820/821 from Rhopoint Ltd. Lead impedance is estimated to be 0.39 to 0.47nH per millimetre. These are also produced by other manufacturers such as Welwyn and Meggitt CGS.

Table 11.2. Resistor values for 50Ω dummy loads

Resistance (Ω)	No. in parallel	Approximate value
100	2	50
150	3	50
390	8	49
560	11	51
1000	20	50

For the type quoted above, the equivalent series inductance of the internal resistance film is about 7nH and has a shunt capacitance of about 1pF. These values will be effected to some extent when the device is mounted on a heatsink. Table 11.3 gives typical design guidelines and assumes the leads are terminated 2.5mm from the body of the resistor.

Table 11.3. Typical impedances of MP820/821 resistors

R (Ω)	10MHz	100MHz	500MHz
10	10 + j0.43	10 + j4.3	10 + j21.7
25	25 + j0.4	25 + j4	24.8 + j20
50	50 + j0.3	50 + j2.8	48.4 + j14.3
75	75 + j0.09	74.8 + j0.9	69.9 + j5.3
100	100 – j0.2	99.6 – j1.9	91 – j6.6
120	120 – j0.5	119.3 – j4.6	105 – j17.6
150	150 – j1	148.7 – j9.6	122.5 – j35.8

These figures are courtesy of Rhopoint Ltd

Use of RF voltmeters and/or probes

You can obtain an RF voltmeter, eg as surplus equipment, or make a probe as suggested in the next section. In fact, the commercial instrument may well use a probe. However, the measuring equipment *must* cover the frequency range in which the power measurements are being undertaken. If a peak reading voltmeter is being used do not forget to convert the peak voltage to RMS voltage by dividing by √2 before using the formula given in the earlier section 'The effect of modulation on power output'. *Don't forget to take into account any attenuators used.* This method of measuring power should be used for carrier power only. Fig 11.18 shows the basic arrangement for these measurements.

RF diode probe

This device allows the scope of a DC voltmeter to be extended to measure AC voltages in the VHF range, and by careful construction probably higher. The probe essentially rectifies the AC immediately and then only has to pass a DC voltage to the meter. The diode is often the limiting factor; to get high-speed operation the diode junction must be narrow and hence this reduces the breakdown voltage. Using a BAT46 Schottky barrier diode the maximum input voltage is about 35V RMS; with the 1N914/1N4148/OA91 it is about 45V RMS. Using a Schottky or germanium diode the forward voltage drop is of the order of 0.2 to 0.3V but with a silicon type it is about 0.6V. The probe should be mounted in a small metal cylinder which is well screened and the resulting DC signal fed via a coaxial cable to the DC meter – see Fig 11.19.

A typical circuit is shown in Fig 11.20 with component values for feeding a 50μA meter movement. The advantage of arranging the capacitor and rectifier in this manner is that the capacitor also acts as DC blocking.

An alternative when the circuit is to be fed into a high-input-impedance DC voltmeter such as an electronic analogue

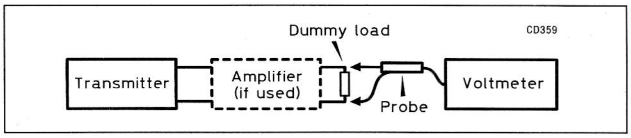


Fig 11.18. RF power measurement using probe and voltmeter

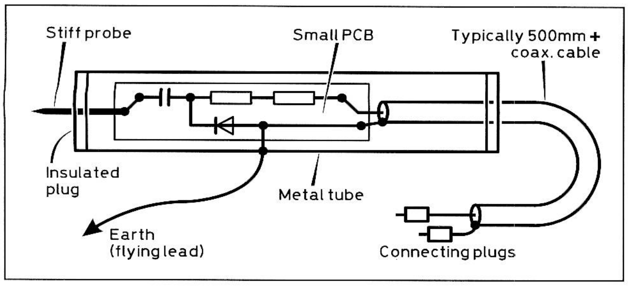


Fig 11.19. Typical construction of an RF probe

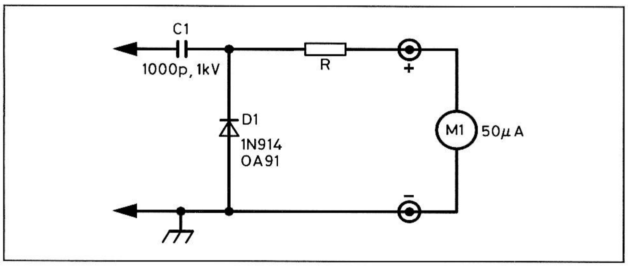


Fig 11.20. RF probe circuit. For R = 270kΩ + 12kΩ, the meter scaling is 0–10V, and full-scale power in 50Ω is 2W. For R = 820kΩ + 27kΩ, the meter scaling is 0–30V, and full-scale power in 50Ω is 18W

meter or a digital meter is shown in Fig 11.21. The input resistance of the meter should be such that it is 10 times the value of R2. Also, the value for the series resistor (R1) should be approximately 41% of the combined resistance R2 in parallel with the meter input resistance. This then allows the meter to read the RMS value of the RF signal. The values shown are suitable for a meter with an input resistance of at least 10MΩ. For a meter of input resistance of 1MΩ, reduce the values of R1 and R2 by a factor of 10.

To measure higher voltages and hence higher power levels, one suggestion is to use a resistive divider across the load – using resistors suitable for the frequencies encountered. Fig 11.22 shows a divide-by-10 unit suitable for a 50Ω system. Remember, the actual voltage is 10 times the value as read on the meter.

An RF millivolt probe

The RF diode probes previously described are limited to voltages in excess of about 1V as a result of the diode forward voltage drop. A method to extend measurements down to a few millivolts is to use the IC transistor array CA3046 which has a minimum gain-bandwidth product of 300MHz. The concept is to amplify the RF signal before detection.

The suggested arrangement is shown in Fig 11.23. The 14-pin DIL device should be mounted in a small screened case with a probe for the RF input in the usual manner. Every effort should be made to keep stray capacitance to a minimum and no IC socket should be used. The input impedance should be about 50kΩ in parallel with 3pF. With the arrangement of two symmetrical DC Darlington pairs the maximum offset voltage will be less than 1mV.

The working range will be from about 1mV to 4V and the device is intended as an add-on unit for

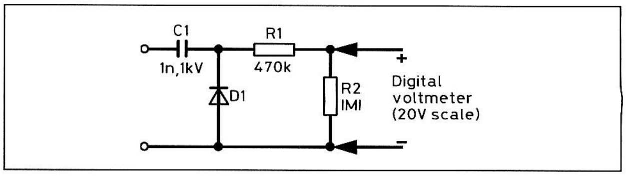


Fig 11.21. RF probe for digital voltmeter

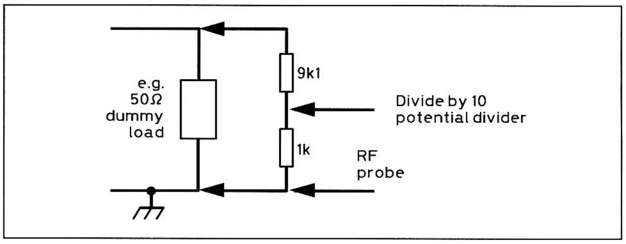


Fig 11.22. Suggested method for higher voltages

a voltmeter. Calibration can be carried out in the several-volt region. Useful measurements should be possible to frequencies in excess of 100MHz.

Alternative ICs that could be considered for the probe are the SL560C (at least 300MHz) and the MAR series (up to 1000MHz) by Mini Circuits. However, these are low-input-impedance types. Good VHF/UHF constructional techniques should be used.

QRP wattmeter

The wattmeter described here will read up to a maximum of 3W and a frequency well in excess of 30MHz; it should therefore be useful for the 50MHz and 70MHz bands and, with careful construction, the 144MHz band. It is in essence a peak-reading voltmeter with internal 50Ω dummy load and is not designed to read standing wave ratios. Sufficient information is given for the design to cope with varying meters and full-scale power levels.

Circuit description

The circuit of the complete unit is given in Fig 11.24. Resistor R1 forms the dummy load, D1 provides rectification, C1

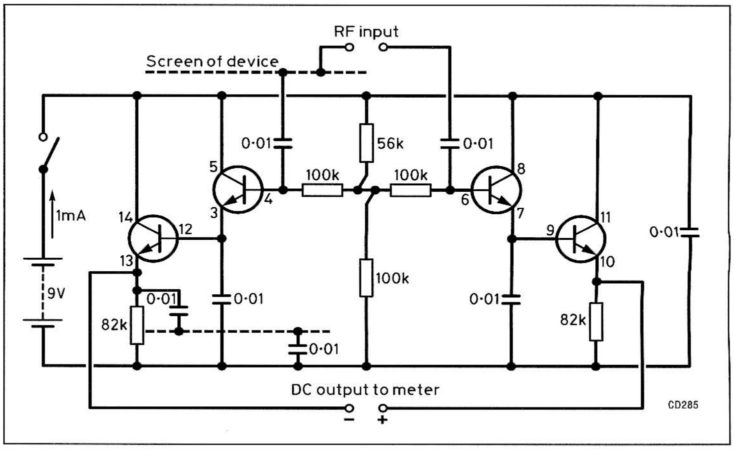


Fig 11.23. Peak-reading RF millivoltmeter probe. All capacitors are disc type. The numbers refer to pins on the CA3046 which is a 14-pin DIL device

smoothing, R2 is to limit the current through the meter M and C2 provides RF decoupling.

The dummy load should be made from carbon resistors and be of adequate rating to cope with 5W. As a minimum, use 1W resistors – three of 270Ω and two of 220Ω will give a load of 49.5Ω. Another arrangement would be four of 330Ω and three of 390Ω, giving an equivalent resistance of 50.5Ω. Other arrangements are of course possible.

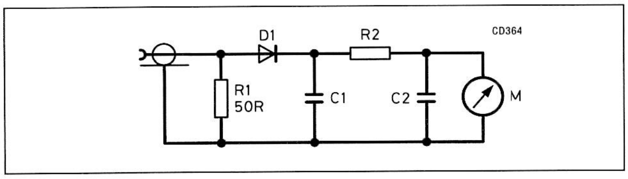


Fig 11.24. Circuit diagram of QRP wattmeter. C1 and C2 are 1000pF ceramic

The diode needs a little consideration – it must be capable of high-speed working, have a minimum PIV of 23V and as low a forward drop as possible in order to minimise errors at the low-power end. Although the ubiquitous 1N914/1N4148 is more than adequate, a lower forward voltage drop can be obtained from a Schottky diode such as the BAT85. For higher power levels use the 1N914. Both capacitors should be of the ceramic type with at least a 30V rating.

Resistor R2 can be calculated to cope with various sensitivity meters. Use a meter of between 50μA and 1mA sensitivity. Neglecting the meter resistance, R2 is given by:

$$R_2 = \frac{V}{I_{FSD}}$$

where V is the peak value of the rectified sine wave and I_{FSD} is the sensitivity of the meter. V is calculated from:

$$V = \sqrt{(100P)}$$

where P is the power being measured.

Thus for a 100μA meter and full-scale deflection for 5W, R = 223.61kΩ (220kΩ + 3.6kΩ) and the meter resistance of about 1kΩ is negligible compared to this. Assuming the meter has a linear scale, then the current corresponding to a given power is given in Table 11.4.

Construction

It is suggested that the whole unit is mounted in a metal box with some ventilation for the dummy load. The circuit from diode to meter should be kept as far away as possible from any circuits carrying RF, and be shielded if at all possible. Use a BNC or SO239 socket for connection.

Table 11.4. Power indicated by current readings for the QRP wattmeter

Power (W)	Current reading (μA)
0.1	14
0.5	32
1.0	45
2.5	71
5.0	100

Below 0.1W the forward voltage drop of the diode becomes significant.

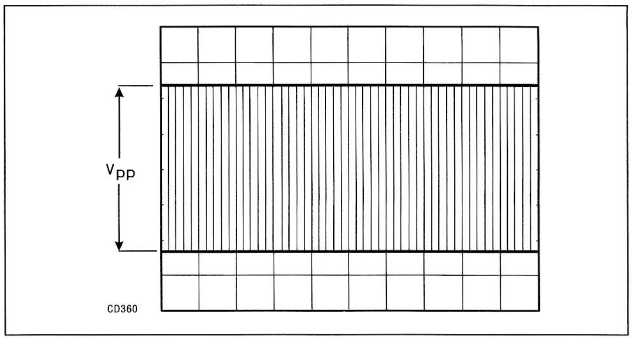


Fig 11.25. Oscilloscope display for carrier only

Use of the oscilloscope for power measurement

This requires the use of an oscilloscope with a timebase or a CRT monitor. It should be stressed, however, that oscilloscopes working into the VHF range are expensive. This section deals solely with the measurement of power, eg voltage display on an oscilloscope. An oscilloscope/CRT can also be used for some modulation measurements. At 100W (CW) the peak-to-peak voltage across a 50Ω dummy load is 200V and at 400W PEP, the maximum peak to peak voltage that will be measured is about 400V! *You have been warned.*

As with the RF voltmeter, the most straightforward measurement is of carrier power which is obtained from the key-down condition for CW operation or the constant amplitude of a frequency-modulated signal. Connect an oscilloscope instead of a voltmeter (Fig 11.18) bearing in mind any frequency or voltage limitations of the oscilloscope and probe. Measure the peak-to-peak amplitude V_{pp} (Fig 11.25) across the known dummy load R. The average power is then calculated from:

$$P_{avg} = \frac{V_{pp}^2}{8R} \text{ watts}$$

The same physical connections are made across the dummy load with the oscilloscope for PEP measurements, but the transmitter should be driven by a two-tone oscillator – see Fig 11.26. The output of the oscillator should be fed into the microphone socket and be of amplitude equivalent to that from the microphone.

Set the timebase on the oscilloscope to be in the audio range and a waveform similar to that shown on Fig 11.27 will be obtained. Measure the peak-to-peak voltage V_{pp} at the peak of the envelope (as shown) – the power is given by the same formula as above.

The input capacitance of an oscilloscope can start to have an appreciable effect at 30MHz – the reactance of 25pF is 212Ω at 30MHz, and obviously affects the readings. It may then be better to use a divide-by-10 probe that will decrease the parallel capacitive loading to about 12pF. This still represents

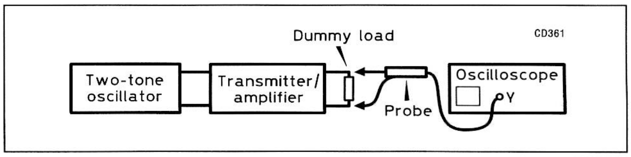


Fig 11.26. RF measurement for SSB work

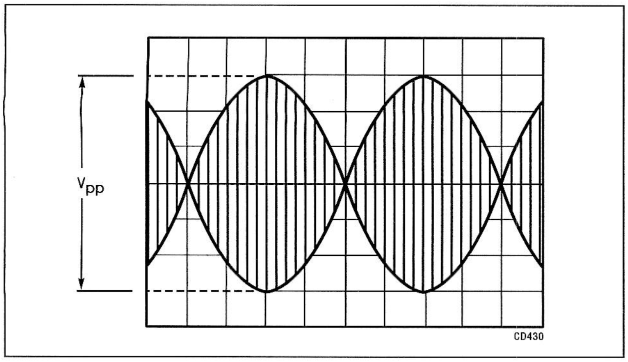


Fig 11.27. Two-tone test display

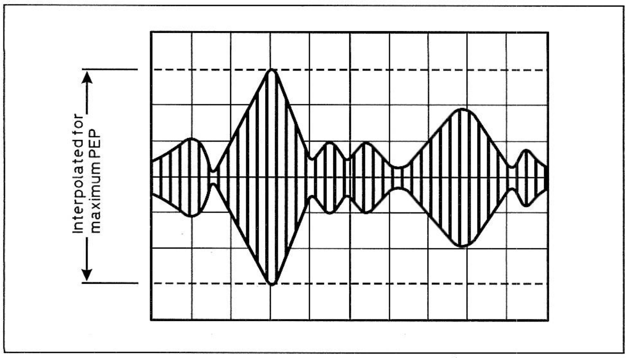


Fig 11.28. Speech waveform and interpolated maximum PEP

a capacitive reactance of 442Ω at 30MHz and the voltage read from the screen will be lower than in reality. To reduce these effects some high-quality oscilloscopes have 50Ω inputs – use these at higher frequencies *but remember that these will have significant loading on a circuit* and may have to matched into the latter.

If the same oscilloscope is continually used to monitor output power on SSB, then note should be made on the graticule or display of the positions corresponding to various power levels. The peak of the speech modulated waveform should then never exceed the maximum permitted level – see Fig 11.28.

If it is possible to feed the Y signal directly to the plates, then the capacitive loading is much smaller, the readings are therefore more accurate and it will be possible to use the oscilloscope to higher frequencies.

REPRESENTATION OF AN ANTENNA SYSTEM USING CIRCUIT COMPONENTS

An antenna system (including feed cable if necessary) represents an impedance at the feed point to whatever is driving it. This can either be considered as being made of a series circuit or a parallel circuit as shown in Fig 11.29.

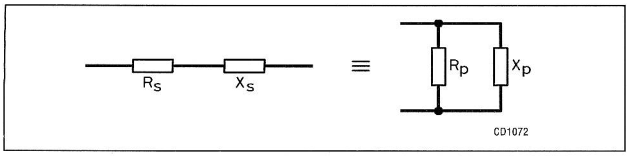


Fig 11.29. Series and parallel equivalence

It is possible to convert between these circuits, the equations being given below. Please note that these values are only true at one particular frequency.

$$R_p = \frac{R_s^2 + X_s^2}{R_s}$$

$$X_p = \frac{R_s^2 + X_s^2}{X_s}$$

$$R_s = R_p \times \frac{X_p^2}{R_p^2 + X_p^2}$$

$$X_s = X_p \times \frac{R_p^2}{R_p^2 + X_p^2}$$

For optimum power transfer, the resistive part should equal the source resistance and the reactive part should cancel with the source reactance – in effect the condition for resonance. Thus it is important to be able to make these measurements at the frequency of concern. Remember also that power can only be dissipated in a resistive element.

Thus, if serious work is to be undertaken on antennas, it is important to determine feed-point impedances. Commercial equipment to perform this function is quite expensive but the following two circuits will give a good indication of conditions.

Also, don't forget that what the transmitter 'sees' is an impedance represented by the antenna and the associated transmission line. If, and only if, the transmission line is a multiple of half-wavelengths (taking into account cable velocity factor) will the feed point impedance be that of the antenna. Ideally measurements should be made directly at the antenna terminals if at all possible.

An RF bridge

This is a Wheatstone-type bridge suitable for use on frequencies up to the 70cm band (430–440MHz). It requires care in 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 of the meter which is good enough for practical purposes and should fulfil most amateur needs.

The circuit of the bridge shown in Fig 11.30 is given in the same form as shown in Fig 11.31 and is self-explanatory.

The whole unit, which uses an external meter, should be built in the smallest-size die-cast box. The component list is given in Table 11.5.

Table 11.5. Component list for RF bridge

R1, 2	100R ±1% metal oxide
R3, 4	4k7 ±1% metal oxide
R5, 6	100R
VR1	50k miniature pot
VR2	2k5 miniature pot
C1, 2	1000p ceramic disc
C3–9	10n ceramic disc
D1	OA91, CV2290, BAT85 (low-voltage-drop diode, suitable for frequency range)
Z_{known}	50R, fitted in coaxial plug

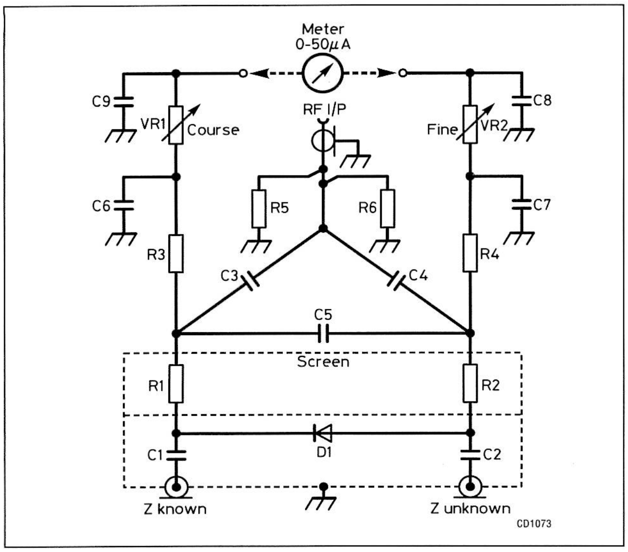


Fig 11.30. Circuit of the RF bridge

Operation

The method of operation with this bridge is to start with the value of impedance known and adjust the length of stubs, matching device etc until a balance is achieved.

This unit will be found ideal for cutting coaxial cable to quarter- and half-wavelengths, where the known *Z* is either an open-circuit or a short-circuit.

RF noise bridge

The circuit described here is from reference [1]. It is a useful circuit for measuring the *R* and *X* components of an impedance or antenna system at a given frequency. It also allows a modulated signal to be obtained, if desired, by pulsing the supply to the noise generator. Such modulation may aid detection of the balance point, especially if an AM receiver is used. The circuit consists of a wide-band noise generator

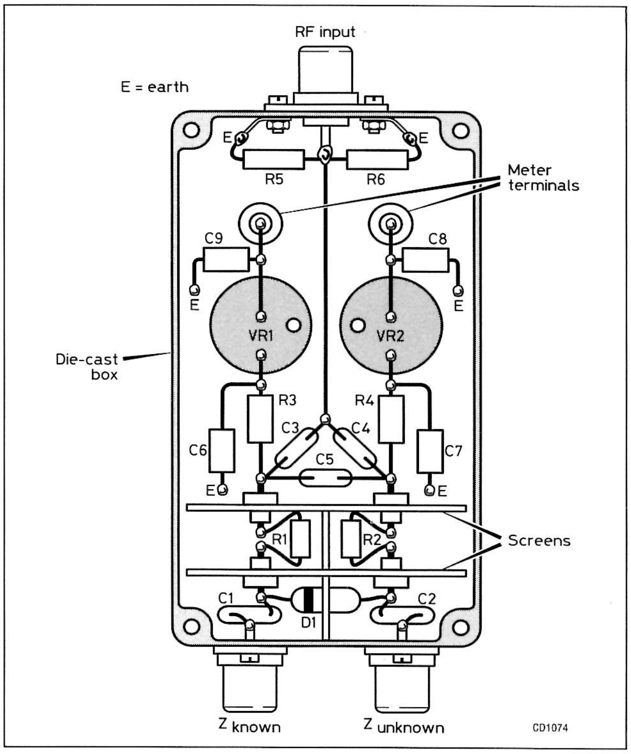


Fig 11.31. Component layout of the RF bridge

followed by a bridge for making the measurements. The bridge allows the measurement of the parallel components of an unknown impedance to be measured. The circuit requires 9V DC at about 25mA.

Circuit description

The circuit is shown in Fig 11.32. The white noise is generated by the zener diode D1 operating at low current. It may be possible to maximise the noise by suitable choice of the

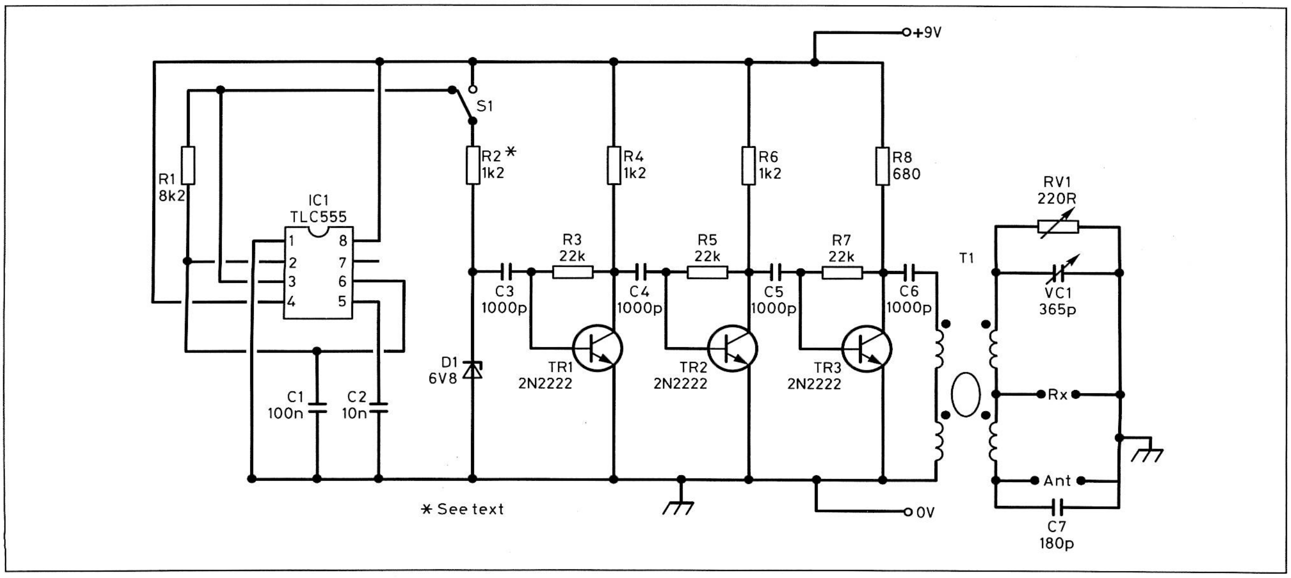


Fig 11.32. Circuit diagram of modulated RF noise bridge

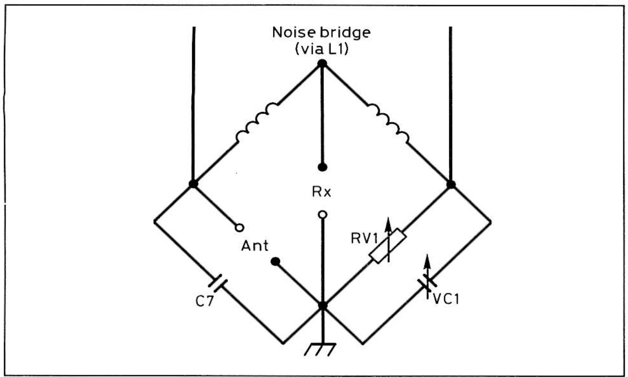


Fig 11.33. Diagrammatic representation of noise bridge

zener diode and R2. The frequency range of this noise should extend up to at least 200MHz. The noise source is followed by a three-stage wide-band amplifier to raise the noise level to the order of 100µV, and this enables a receiver to be used as a null indicator.

The noise output from the amplifier is applied to a quadrifilar-wound toroid which forms the transformer T1. This provides two arms of a bridge circuit which has a variable resistor and capacitor in the third arm to obtain a balance against the impedance (eg antenna system) in the fourth arm. The bridge circuit is shown diagrammatically in Fig 11.33.

When the noise across the RV1/VC1 arm equals the noise across the antenna/capacitor combination, the bridge is said to be 'balanced', and this occurs when the received noise signal is at a minimum. The values can be obtained from the settings of RV1 and VC1. The inclusion of C7 allows an offset to be used so that inductive reactance can be measured. The mid-point setting of VC1 is equal to zero reactance. If a noise bridge is only required to measure the resistive part of the antenna impedance then omit C7 and VC1.

Timer IC1 is in astable mode and runs at about 850Hz with a 50% duty cycle, and this can be used to provide current for the zener circuit via S1, so modulating the noise source. The zener diode can be alternatively fed from the constant-voltage power supply line.

Construction

The component list is given in Table 11.6. The toroid transformer consists of a dust-iron core, type T50-6, which is wound as follows. Cut four lengths of 26 SWG enamelled copper wire about 120mm long, twist them together and then thread them through the toroid to give 14 turns and evenly spaced to cover the circumference. Divide the turns into two pairs, each pair consisting of two windings connected in series, the end of one winding connecting to the start of the other – be careful. Check that the two pairs are insulated from each other. Endeavour to keep the lead lengths in the bridge as short as possible and symmetrical. The variable resistor RV1 should be of high quality and with a carbon, cermet or conductive plastic track – *not wirewound!*

When constructing the circuits, ensure that the noise generator and amplifiers are well away or screened from the bridge transformer and measuring circuit. The potentiometer case should not be earthed and, if it has a metal spindle, this should be isolated from the user and should not contact ground.

Table 11.6. Components list for the RF noise bridge

R1	8k2
R2, 4, 6	1k2
R3, 5, 7	22k
R8	680R
RV1	220R pot (see text)
C1	100n, 50V ceramic
C2	10n, 50V ceramic
C3–6	1000p, 50V ceramic
C7	180p, silver mica
VC1	365p Jackson type 01 gang
D1	6V8 zener, 400mW
TR1-3	2N2222
IC1	TLC555
S1	SPCO switch
T1	T50-6 dust iron core, 4 windings, each 14t or 596100001 ferrite core, 4 windings, 6t each

Resistors are 0.25/0.5W, ±5% unless stated otherwise

The complete circuit should be mounted in a screened box such as a die-cast type with appropriate connectors – eg UHF type or BNC. In order to avoid coupling into the measuring circuit of noise by way of current in earth loops, the earthed side of the noise source should not be joined to the general chassis earth of the bridge but should be taken by an insulated lead to the frame of the variable capacitor.

As in all high-frequency measuring circuits, lead inductance should be kept to an absolute minimum and, where any lead length more than a few millimetres is unavoidable, copper foil at least 6mm wide should be used. All earth returns should be taken to the capacitor frame. Capacitor C7, which should be silver mica, can be soldered directly across the UNKNOWN socket.

A suitable PCB pattern and component layout is given in Appendix 1.

Calibration

Connect a test resistor (of a carbon type) across the UNKNOWN socket with the receiver tuned to 3.5MHz. Adjust RV1 and VC1 to give a null. The value of RV1 is at the position equal to the test resistor and the capacitor should be at approximately the mid-mesh position or the zero reactance condition – mark these positions. Repeat with different values of test resistor up to 220Ω in order to provide a calibration scale for RV1. Repeat this operation with known values of capacitance in parallel with the test resistor, up to a maximum value of 180pF. Mark the corresponding null positions on the VC1 scale with the value of this capacitance. Repeat this procedure at 28MHz to check the accuracy of the bridge. If the layout has been carefully attended to there should be little difference in the null positions.

To calibrate VC1 for negative capacitance values (ie inductance) it is necessary to temporarily place given values of capacitance in parallel with VC1. Gradually decrease the value of these capacitors (CT) from 150pF towards zero, obtaining null positions and marking the VC1 scale with the value of –(180 – CT) pF, ie if 100pF is substituted then the negative C value is 80pF.

Using the noise bridge

For work on an antenna, a noise bridge should ideally be connected across the antenna terminals. This is usually not practical, in which case a noise bridge should be connected to the antenna by a length of line which is a multiple of a half-wavelength at the frequency of interest (taking into account the velocity factor of the cable).

Connect the impedance to be measured to the UNKNOWN socket, switch on the noise generator and tune the receiver to the frequency at which the test is to be made. Use RV1 and VC1 to obtain a minimum noise reading on the receiver S-meter. The values must now be converted to circuit components. The value recorded from RV1 is the resistive part of the impedance. The value from VC1 is the parallel reactive component of the impedance and, depending on the sign, is either inductive or capacitive. If it is positive, then the value of shunt capacitance is read directly from the VC1 scale. If it is negative, the VC1 reading represents the value of the shunt inductance and must be calculated as below.

If a negative value of capacitance (C) is obtained this can be converted to an inductance value using the formula:

$$L = \frac{1}{4\pi^2 f^2 (180 - C)}$$

where f is in megahertz and C in picofarads.

This can be accomplished with the following BASIC program:

```
10  REM Noise Bridge Inductance Calculation
20  CLS
30  C=180
40  INPUT "Negative C Value in pF",CV:
    CV=ABS(CV)
50  INPUT "Working Frequency in MHz",F:
    F=F*1000000
60  K=1/(4*3.14159^2*F^2)
70  L=K*1/((C-CV)*1E-12)
80  L=L*1000000
90  PRINT "Inductance in uH is ",L
100 END
```

ATTENUATORS

The need for good attenuators capable of working at frequencies up to several hundred megahertz or higher often arises. These are relatively easy to construct out of standard resistors and can be put to a variety of uses. For example, attenuators at RF can be used as pads between interacting stages, eg varactor multipliers, or to follow noise or signal sources to bring their output close to 50Ω. At IF they can be used for calibrating attenuators, since their attenuation is fairly predictable at lower frequencies. They may also be used for calibrating S-meters etc, and as a reference for noise measurements, eg Sun and ground noise. An attenuator might also be useful between a transmitter and transverter.

Design and construction of simple attenuators

Attenuators are normally made from T or pi networks – see Fig 11.34. For this exercise it is assumed that load and source impedances (R_0) are equal and resistive. The design of these is covered by the following formulae.

T network

$$\text{Attenuation (dB)} = 20\log_{10}\left(\frac{R_0 + R_1}{R_0 - R_1}\right)$$

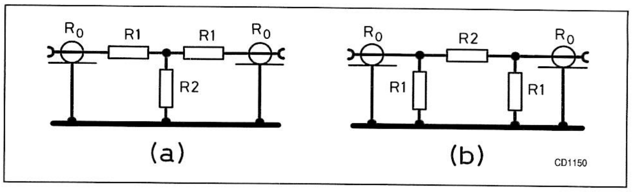


Fig 11.34. (a) T-type attenuator. (b) Pi-type attenuator

when:

$$R_2 = \frac{R_0^2 - R_1^2}{2R_1}$$

Pi network

$$\text{Attenuation (dB)} = 20\log_{10}\left(\frac{R_1 + R_0}{R_1 - R_0}\right)$$

when:

$$R_2 = \frac{2R_0^2 R_1}{R_1^2 - R_0^2}$$

The greatest problem in constructing a good attenuator is the radiation and leakage of signals from within the unit. Because of this, the attenuator should consist of a good RF-tight metal box with high-quality connectors.

Two methods of construction are illustrated in Fig 11.35. A pi-type is shown at (a) and the T-type at (b), but either type could be used in either design. The resistors should be a low-inductance type, the common form of carbon film resistors being particularly suitable. Lead lengths should be as short as possible. For higher-power attenuators at lower frequencies, parallel combinations of 0.5W carbon resistors can be used to increase dissipation. A 10dB attenuator built in this way to handle 10W measures 9.5dB attenuation at 432MHz with low SWR.

Provided that care is taken, these attenuators can be used up to 1–2GHz. The biggest error is likely to arise in the higher-value attenuators, where stray coupling may reduce the attenuation below the expected value. For this reason it is better to use several low-value stages in cascade when a high value of attenuation is required.

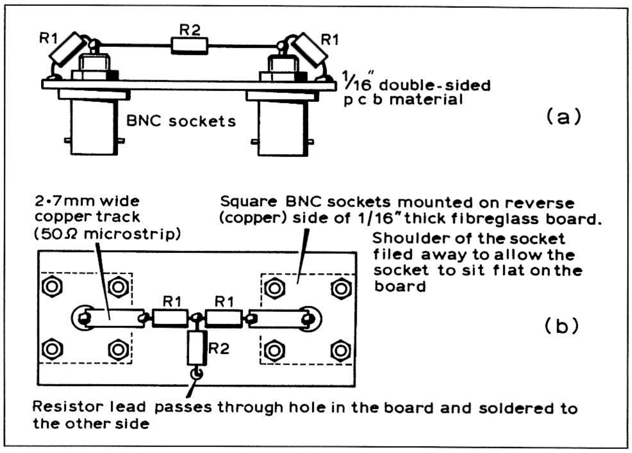


Fig 11.35. Methods of constructing attenuators: (a) below 500MHz, (b) above 500MHz

Two computer programs are given below [3] for assistance as the calculation is somewhat tedious but there is also Table 11.7 for those who do not want to use a computer.

For the pi-attenuator network use the following program:

```
10  PRINT "RESISTIVE PI ATTENUATOR CIRCUIT"
20  DEF FNA(X)=INT(X*100+.5)
30  DEF FNB(X)=LOG(X)/LOG(10)
40  DIM P(12)
50  FOR I=1 TO 12:READ P(I):NEXT
60  DATA 1.2,1.5,1.8,2.2,2.7,3.3
70  DATA 3.9,4.7,5.6,6.8,8.2,10
100 INPUT "CIRCUIT IMPEDANCE (OHMS) ";Z
110 INPUT "ATTENUATION (DB) ";DB
120 V=10^(-DB/20)
130 A=Z*(1+V)/(1-V)
140 B=Z*(1-V*V)/(2*V)
150 PRINT "OPTIMUM VALUES:"
160 PRINT "SIDES: ";A;" OHMS"
170 PRINT "  TOP: ";B;" OHMS"
180 R=Z:GOSUB 500
190 PRINT "NEAREST PREFERRED VALUES: "
200 PRINT " SIDES    TOP    DB      Z"
210 R=A:GOSUB 600:AL=L:AH=H
220 R=B:GOSUB 600:BL=L:BH=H
230 A=AL:B=BL:GOSUB 400
240 A=AL:B=BL:GOSUB 400
250 A=AL:B=BL:GOSUB 400
260 A=AL:B=BL:GOSUB 400
300 INPUT "SIDE RESISTORS ";A
310 IF A<0 GOTO 100
320 INPUT "TOP RESISTOR ";B
330 GOSUB 400
340 GOSUB 500
350 GOTO 300
400 R=A*Z/(A+Z):V=R/(R+B)
410 R=A*(R+B)/(R+A+B)
420 DB=20*FNB(V)
430 PRINT A;TAB(8);B;TAB(16); FNA(DB)/100;
    TAB(24);FNA(R)/100
440 RETURN
500 P1=R/A:P2=P1*V*V:PB=R*(1-V)*(1-V)/B
510 PRINT "POWER:  INPUT ";FNA(P1);" %"
520 PRINT "TOP ";FNA(PB);"%","OUTPUT ";
    FNA(P2);"%"
530 RETURN
600 I=1:M=10^INT(FNB(R)):L=M
610 H=M*P(I)
620 IF H>R THEN RETURN
630 I=I+1:L=H;GOTO 610
```

For T-attenuator design, replace the appropriate lines in the above program by the following:

```
10  PRINT "RESISTIVE T-ATTENUATOR NETWORK"
130 A=Z*(1-V)/(1+V)
140 B=(Z*Z-A*A)/(2*A)
150 PRINT "OPTIMUM VALUES:"
160 PRINT " ARMS: ";A;" OHMS"
170 PRINT " BASE: ";B;" OHMS"
180 R=Z:GOSUB 500
190 PRINT "NEAREST PREFERRED VALUES:"
200 PRINT "  ARMS   BASE    DB      Z"
300 INPUT "ARM RESISTORS ";A
310 IF A<0 GOTO 100
320 INPUT "BASE RESISTOR ";B
400 R=B*(Z+A)/(B+Z+A):R=R+A
500 H=V+V*A/Z:P1=R*(1-H)*(1- H)/A:
    P2=R*V*V*A/(Z*Z):PB=R*H*H/B
```

Table 11.7. Design data for 50Ω T-type and pi-type attenuators

Attenuation (dB)	T-type R1	T-type R2	Pi-type R1	Pi-type R2
1	2.9	433	870	5.8
2	5.7	215	436	11.6
3	8.6	142	292	17.6
4	11.3	105	221	23.9
5	14.0	82	178	30.4
6	16.6	67	150	37.4
7	19.1	56	131	44.8
8	21.5	47.3	116	53
9	23.8	40.6	105	62
10	26.0	35.1	96	71
12	30.0	26.8	84	93
14	33.4	20.8	75	120
16	36.3	16.3	69	154
18	38.8	12.8	64	196
20	40.9	10.0	61	248
25	44.7	5.6	56	443
30	46.9	3.2	53	790

Switched attenuator

A switched attenuator has a number of applications in the amateur station; it may be made either wide or limited range depending on the intended application.

The simple three-section unit illustrated (Fig 11.36) is intended for use between the output of a VHF transceiver having an output in the range 10–20W PEP and a following linear amplifier. With the three stages of 2, 4 and 8dB available, various levels between 0 and 14dB are available.

In an attenuator which is to be used with a power source such as a transceiver, it should be remembered that a significant proportion of the input power will be dissipated in the resistors in the attenuator. The proportion is increased with the level of attenuation. In this case 2dB attenuation will dissipate 36% of the input power, 4dB 52% and 8dB 84%. Therefore adequately rated components are essential if the unit is to be used continuously.

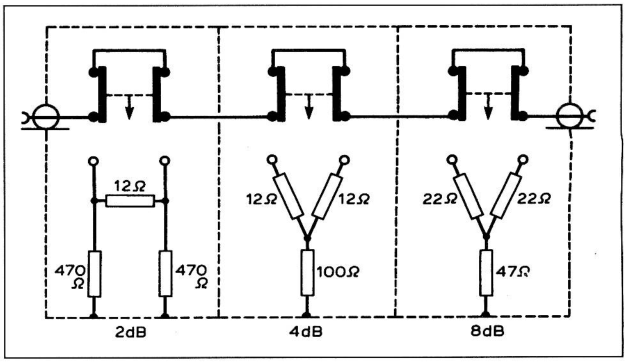


Fig 11.36. Circuit of switched attenuator

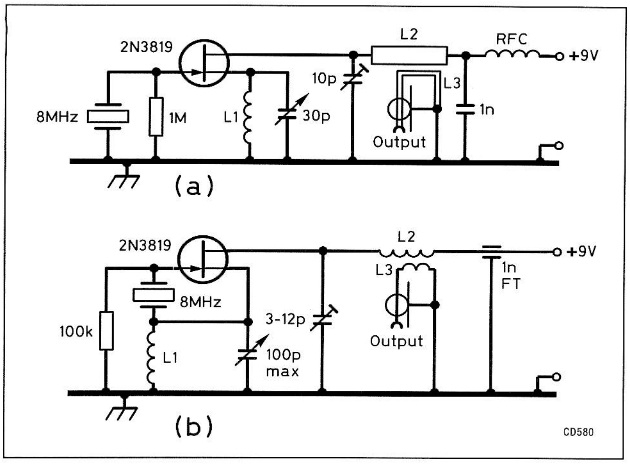


Fig 11.37. Two alternative simple VHF signal sources. Inductor details for (a) are: L1, 20t 8mm dia; L2, 120mm long by 3mm dia; L3, output coupling loop, 16 SWG wire; RFC, 35t 6mm dia. Inductor details for (b) are: L1, 28t 28 SWG wire, 6.5mm dia; L2, 4t 18 SWG wire, 12.7mm dia; L3, output coupling, 1t 18 SWG wire

For a continuous carrier input of 10W, after a 2dB attenuation 6.3W is available to the next stage, after 4dB there is 4W, after 6dB there is 2.5W and after 8dB just 1.6W. The remaining power is dissipated within the attenuator.

Although the switched attenuator may be used, for normal general-purpose operation a single-stage fixed type is more likely to find favour once the degree of attenuation has been established. Such a unit can more readily be built using suitable components and where necessary adequate heat-dissipating construction.

SIGNAL SOURCES

A reliable signal source is a useful adjunct for setting up receivers and converters. This is useful for both the setting of newly built equipment and the repair of equipment – both homebrew and 'black box'. Once the equipment has been aligned on a signal source it is worthwhile trying to tune to a distant beacon or repeater and again trying to optimise reception.

A simple signal source

Fig 11.37 shows two possible crystal oscillators with the appropriate tuned circuit, and both use 8MHz range crystals. The output should be checked with an absorption wavemeter to make sure that the correct harmonic of the crystal frequency has been selected. The possibility of error would be reduced if a higher-frequency crystal were used: this would be particularly desirable if the output circuit were modified to give an output on 432MHz.

In construction it is desirable for the unit to be completely enclosed so that output is only obtained from the output socket – this will largely eliminate unwanted signals and also allow some control of the level. The power supply should be very well decoupled *or* a battery used in the same enclosure.

Dual output signal source

A useful signal source having outputs at 144 and 432MHz can be constructed readily using the familiar Butler crystal oscillator and multiplier circuit with a fifth-overtone crystal of 103MHz – see Fig 11.38.

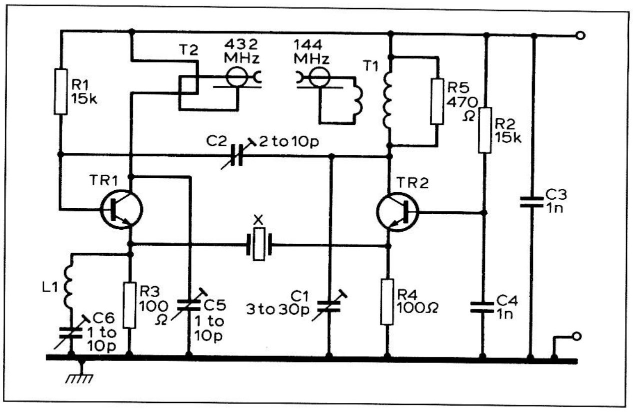


Fig 11.38. Circuit of dual-output signal source

The circuit uses a series-resonant crystal but the two different frequency output tuned circuits are connected to the respective collectors of the transistors in order to obtain the maximum output from the 432.6MHz side. A series resonant circuit (L1/C6) tuned to 288.4MHz is connected in the emitter circuit of TR1. An output power of 30mW is obtainable on 144.2MHz and 10mW on 432.6MHz for an input power of 120mW.

The outputs are inductively coupled and suitably supported and adjusted. The transformer T1 for 144.2MHz output consists of four turns of 0.5mm wire wound with 7mm internal diameter, 9mm long and with a two-turn link output coil. T2, the 432.6MHz output circuit, consists of 43mm lengths of wire formed into suitable loops. The 288MHz idler circuit (L1/C6) is formed of a coil of three turns of 0.5mm wire, 6mm internal diameter, 10mm long and tuned by a 1–10pF trimmer.

The transistors used are BFY90 or equivalent; types such as BSX20 will also be satisfactory, although somewhat lower output will be achieved.

Variable signal source for 144 and 432MHz

The signal sources so far described have been based on crystal oscillators to give high stability. Sometimes a variable frequency oscillator together with variable output level control is more useful. The harmonics of the oscillator are of sufficient strength to give outputs at 144 and 432MHz. Such a circuit is shown in Fig 11.39.

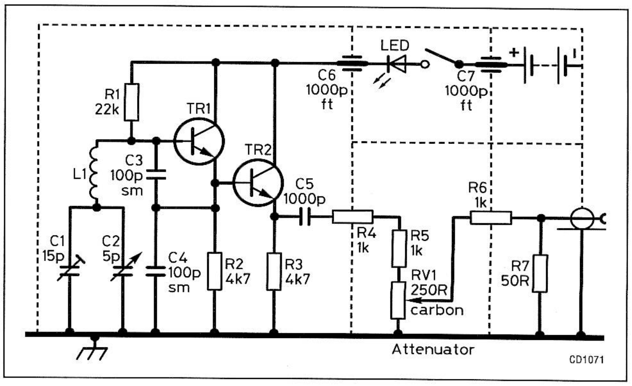


Fig 11.39. Variable signal source for 144 and 432MHz. L1 is 60–70nH

The circuit consists of a VFO (L1/C1/C2) and TR1 followed by a buffer TR2. The output is taken from R5 and goes to a variable level control (RV1). Transistors similar to those *in the previous section* can be used. The whole unit should be mounted in a metal box to minimise stray pick-up. Internal screening should be used as shown around the various parts of the circuit.

The unit may be battery powered but increased stability is likely if the power supply to the circuit is regulated and well decoupled.

ABSORPTION WAVEMETERS

Two important uses of an absorption wavemeter are checking:

1. that the correct harmonic is selected for driving the next stage in circuits such as multipliers; and
2. for the presence of harmonics in the output of a transmitter stage.

An absorption wavemeter provides an unambiguous frequency measurement and it is also a requirement of the UK amateur licence to check regularly the harmonic output.

In the UK, on both 50 and 70MHz the second and third harmonics fall into other user's bands, the second harmonic being in a broadcast band. For the 144MHz band the second harmonic falls in another user's band, and the third harmonic falls in our own 432MHz band. At 432MHz the second harmonic is just above a TV band whilst the third harmonic is 1296MHz – in one of our bands!

The generation of the second harmonic is often too easy, for example, it is often heard that by a little 'tweaking' with the output tuning adjustment a significant increase in output can be obtained – very often the majority of the apparent increase is second harmonic with very little increase in the fundamental.

A good absorption wavemeter should have at least a two-to-one frequency coverage, preferably without switching bands. With care this can readily be attained, although many commercial types do not tune above about 250MHz.

Two designs [4] are described here, one for general purpose use, the other for insertion into the coaxial line. Both types cover a frequency range of 125–350MHz and are sufficiently sensitive to obtain a reasonable indication of the fundamental frequency of a FET dip oscillator having an output of 3–5mW at a distance of 100–125mm. As outputs of appreciably higher levels than this are likely to be involved, the full sensitivity may not always be needed, and a shunt across the meter is provided with a press-to-open switch to give full sensitivity.

General-purpose wavemeter

In this design an edge-mounted meter has been used. There is, however, some advantage in this method insofar as in use it is almost always necessary to 'bend over' to be able to see the meter whereas with it in the end of the box direct observation is possible.

In order to obtain an adequate inductance for coupling to the circuit under test while being small enough to reach the top frequency, it is necessary to make the external loop and the connections to the tuning capacitor of material with low

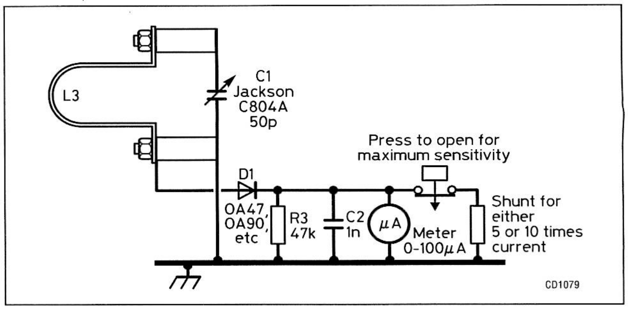


Fig 11.40. Circuit diagram of general-purpose wavemeter

inductance per unit length. The external loop also needs to be mechanically rigid to avoid damage or calibration changes. To avoid restriction of the top frequency the material used for mounting the external loop needs to be of low dielectric constant. The most suitable material for this purpose is PTFE or an equivalent. The circuit diagram of the unit is shown in Fig 11.40.

The construction of the wavemeter is straightforward and is shown in Fig 11.41. The external loop is attached to the internal connections by 4BA (or metric equivalent) brass cheesehead screws and the strip connections to the capacitor are soldered directly to the slots in the heads of the screws. The detector diode (BAT81/83/85 or OA47/90 or equivalent) is connected to one of the screws, thereby providing a tap-down of about 25% of the total inductance and avoiding undue damping of the circuit.

Calibration can readily be carried out by use of a suitable dip oscillator, the actual frequency of which can be verified by a digital frequency meter.

If it is desired to cover 50MHz and 70MHz, then a substantially larger inductance will be needed but this reduces the top frequency. This may take any convenient form – a suitable coil that covers 48–130MHz consists of four turns of 16mm ID of 3.25mm copper close wound with 25mm tails. With this much larger inductance the tapping point of the diode detector will be very much lower down and therefore the full sensitivity of the indicating meter will be required.

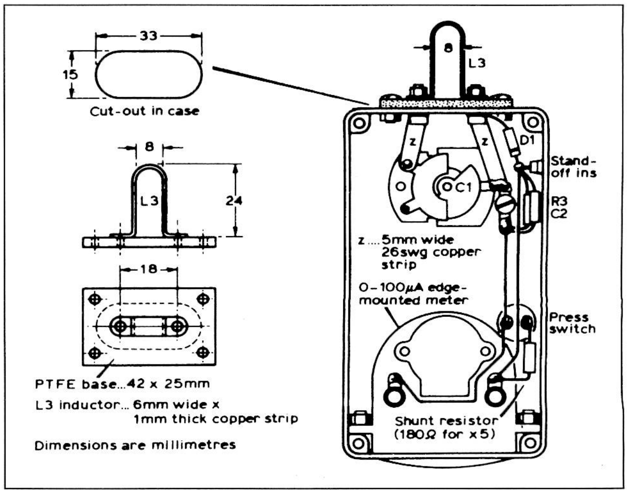


Fig 11.41. Arrangement of components of general-purpose wavemeter

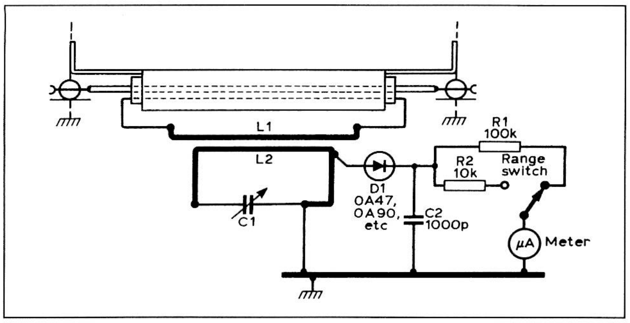


Fig 11.42. Indirect through-line wavemeter

Through-line type wavemeters

This type is intended for connection in the coaxial cable, either between an exciter or the antenna feeder. In the latter case some care is needed to ensure that the unit is not damaged when used with a high-power amplifier, and it should be used with a T connector and a suitable terminating resistor.

The main element of this type of wavemeter is a short section of coaxial line within the unit to which is coupled the tuned circuit.

The dimensions for both types are given for fitting in a die-cast box which is 60mm wide externally (eg Eddystone type 27134P). Adjustments will have to be made to suit other widths of box.

Indirect method

As shown in the circuit diagram (Fig 11.42), the tuned circuit is coupled to the inner of the coaxial line through an intermediate loop. The loop consists of a fine wire inserted in the coaxial line and is connected to a short length of more rigid wire to which the tuned circuit loop is coupled.

In constructing the line within the unit it will be found more convenient to fabricate this from a short piece of semi-air-spaced cable and replace the outer braid with a short length of copper or brass tube of the same diameter as the original braid in order to maintain the correct impedance – see Fig 11.43.

The cable used in the prototype had an insulation diameter of 6.35mm (¼in) which was replaced by a copper tube. If it is required to change the impedance from 50Ω to 75Ω it is only necessary to replace the inner conductor with a thinner wire (the size can be either calculated or obtained from the relevant chart elsewhere in this book).

Fitting the coaxial element into the case (Fig 11.44) requires the ends of the tube to be shaped and bent so that the lugs thus produced can be fixed to the box by the screw holding the connector. BNC sockets have been used in the design shown. Component arrangement is clearly shown in Fig 11.44 and should present no problems – the connection of the diode should be 20mm from the ground end of the tuned loop.

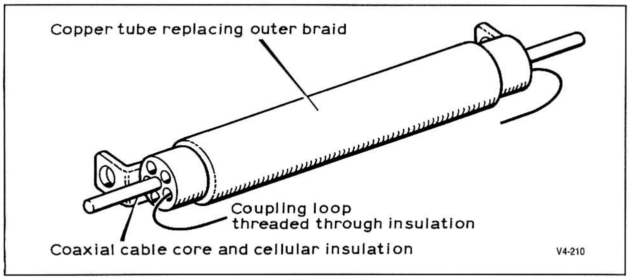


Fig 11.43. Coupling for through-line meter

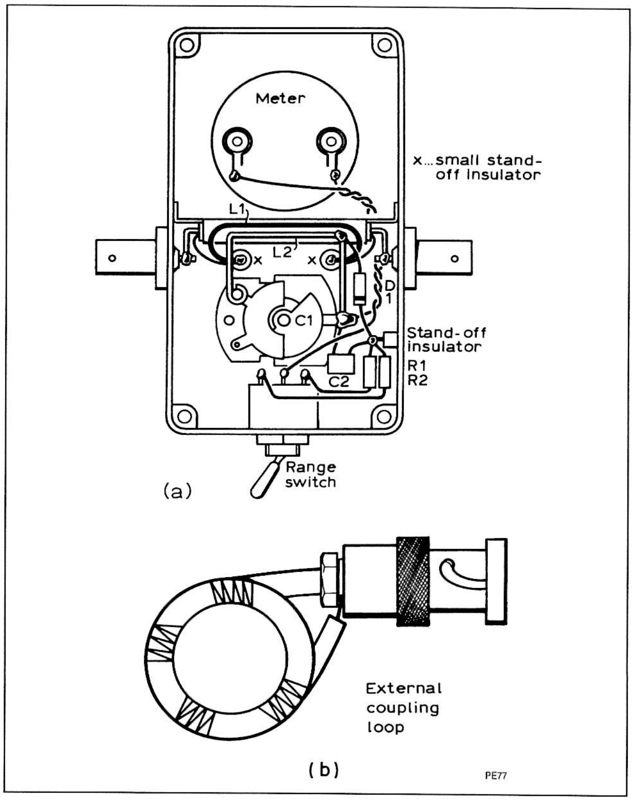


Fig 11.44. (a) Arrangement of components of indirect through-line wavemeter. (b) External coupling loop

Direct method

In this form, as mentioned earlier, the intermediate coupling loop is avoided, and there is direct coupling of the tuned circuit to the inner conductor through a slot in the coaxial outer.

Details of a suitable slotted line are given in Fig 11.45. The tuned circuit inductance consists of a simple U-shaped piece of 18 SWG (1.25mm) enamelled copper wire to which the detector diode is connected at a point 20mm from the earth end. Spacing the loop from the inner conductor is important and should be within the field of the outer of the coaxial line. To assist alignment the slot is arranged to be vertical and this means that the fixing lugs of the tube forming the outer of the line must be at 45° to the slot to enable the line to be fixed by one of the screws holding the coaxial socket.

The general layout for this type of meter is shown in Fig 11.46.

Wide-range UHF cavity wavemeter

Measurement of frequencies above 500MHz becomes difficult using conventional lumped-circuit type wavemeters, so that it is an advantage to use a cavity design. Cavity wavemeters are often constructed to cover relatively narrow bands but for amateur purposes it is desirable to cover several bands,

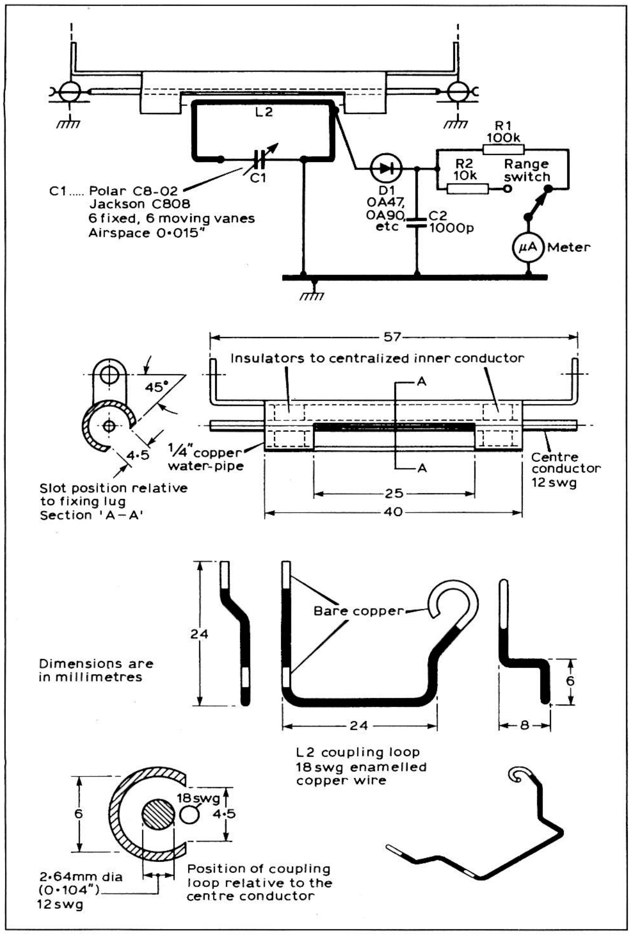


Fig 11.45. Direct through-line wavemeter

if possible, thus providing continuous coverage and allowing their use for second harmonics. The wavemeter described here has been designed to cover from around 400MHz to 2.5GHz

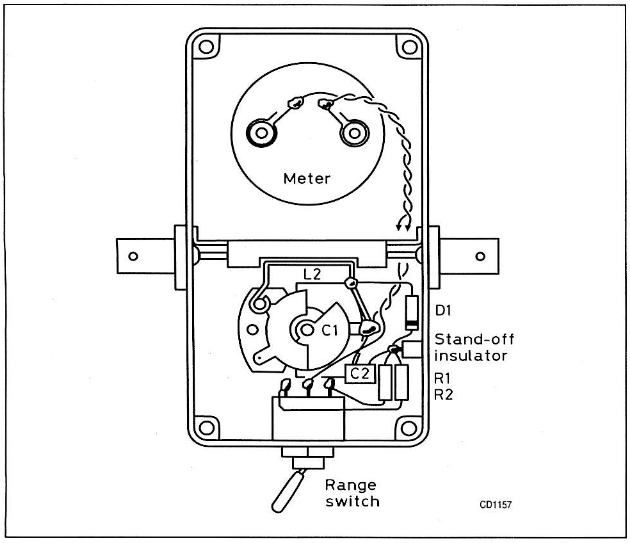


Fig 11.46. General arrangement of through-line wavemeter – direct type

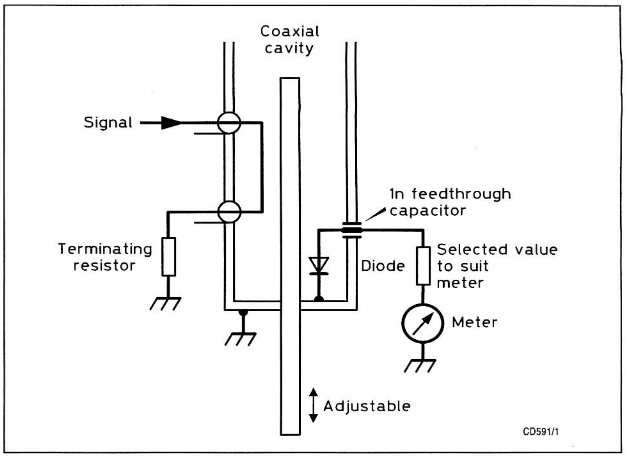


Fig 11.47. Generalised circuit arrangement

using a single cavity with an adjustable inner quarter-wave ($\lambda/4$) element.

Basically, the direct measurement of a $\lambda/4$ element represents the wavelength to which it is resonant. An equivalent circuit of the wavemeter is shown in Fig 11.47. It consists of input coupling, a resonant circuit formed by the $\lambda/4$ line/cavity and output coupling with detection and meter drive (output probe).

When a $\lambda/4$ circuit is energised, a current maximum will occur at the shorted end and a voltage maximum at the open end as shown in Fig 11.48. A current resonance indicator must therefore be coupled to the low-impedance end of the circuit, ie as near as possible to the shorted end. Also, as the input will normally be the output of an oscillator or transmitter, this will also dictate coupling at the low impedance end, ie near the short-circuit.

Both these couplings should be relatively loose so as not to foreshorten the length of the inner conductor by capacitive loading so that the mechanical length of the inner conductor is substantially the electrical length of $\lambda/4$. There will, however, be some apparent shortening of the inner conductor compared with the free-space value due to the stray field from its end to the continuing outer cylinder. This will be most noticeable at the highest frequencies and may be as much as 4mm at 2.5GHz.

The characteristic impedance of the cavity is of no significance in the case of a frequency meter and may be of a value convenient to the materials available. It may be either circular or square in section. The sensitivity will naturally depend to a large extent on the meter used. With a 50μA FSD meter

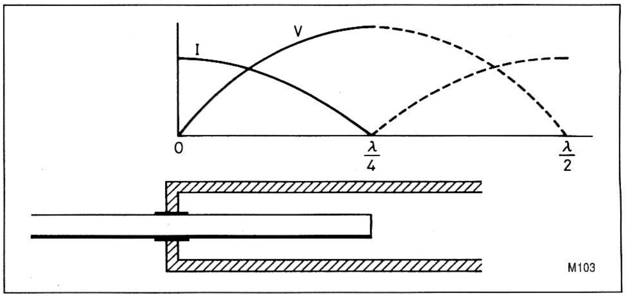


Fig 11.48. Current (*I*) and voltage (*V*) in a $\lambda/4$ cavity

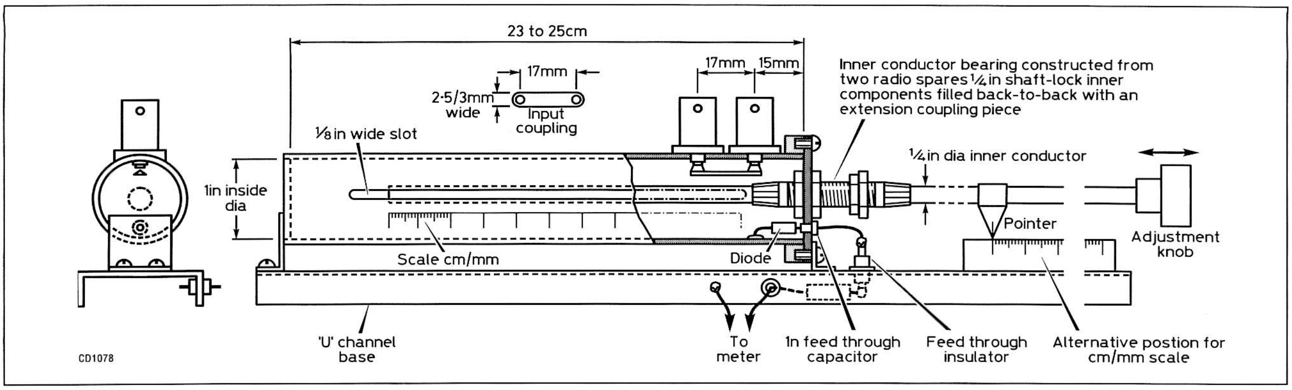


Fig 11.49. Layout of the wide-range cavity wavemeter. Note: 1in = 25.4mm, 1/8in = 3.175mm, 1/4in = 6.35mm

satisfactory indications at levels down to 5mW may be observed.

In the design illustrated in Fig 11.49, the outer consists of a 25mm inside diameter tube with an adjustable length inner conductor of 6.35mm (0.25in) diameter. The outer in this case has a narrow slot running most of the outer tube length so that the position of the inner conductor may be observed directly; the calibration scale can be fixed along the slot (similar to a slotted line). An alternative method of fitting a calibration scale to the extension of the inner conductor outside the cavity is also indicated on the diagram.

Construction

As mentioned, the precise dimensions of the cavity are not critical though, if materials permit, it can be made for 50Ω or 75Ω and may be either round or square in cross-section.

For convenience, the inner conductor should be 6.35mm in diameter and the outer tube 25mm inside diameter. The material may be copper or brass but the latter is more rigid and may be more suitable for the inner conductor. The outer may be for preference copper as it may well be made from a length of water pipe. It is important to provide a reliable sliding contact for the inner conductor. If 6.35mm material is used then two conventional shaft locks should be connected back to back, preferably with an extension tube between them, to provide a long bearing. The input coupling is at the shorted end and consists of a strip drilled and then soldered directly to the connectors.

Using the wavemeter

For frequency measurement, the indicator should be connected to one port of the wavemeter and the other port connected to the output of the source whose frequency is to be measured. In the case of a source that is likely to be sensitive to load impedance it may be preferable to connect the wavemeter via an attenuator or directional coupler since the wavemeter, when tuned to frequencies other than that of the source output, will appear as a variable reactance.

Slide the inner of the wavemeter slowly out whilst watching the indicator for a peak reading. In the case of a wavemeter that is poorly constructed there may be a residual reading whilst still far off the resonant frequency. This is usually due to over-coupling between the input and output circuits. The resonance peak will, in this case, be preceded by a sharp dip in this residual reading.

It is possible with this type of wavemeter to obtain more than one indication due to the wavemeter resonating not only at a quarter-wavelength but also at all odd multiples of a quarter-wave. For example, when measuring the output of a 1296MHz source a reading will also be obtained when the wavemeter is tuned to 432MHz. This characteristic can be used when calibrating the wavemeter since it is possible to obtain several calibration points. These can be plotted on a graph and, by interpolation, the frequencies between can be calibrated. Obviously the more frequencies available when calibrating, the more accurate the overall calibration will be.

DIP OSCILLATORS

A dip oscillator (the valve versions were called *grid dip oscillators*) is an essential tool for the construction of VHF/UHF equipment.

In its simplest form it consists of a stable LC oscillator covering a wide range of frequencies. Depending on the actual range, the inductance will be cut to an appropriate size and will normally be a plug-in type. Some form of indicator, such as a meter, is required to show when the dip oscillator is tuned to the circuit under test.

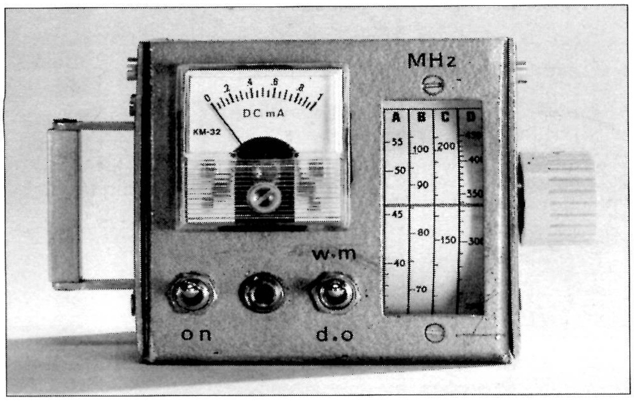
For VHF and UHF dip oscillators, FETs are normally used in a push-pull circuit. With circuits of this type oscillators for use up to 500MHz are practical with careful mechanical layout so that connecting leads between the plug-in coil and the tuning capacitor are short and of as low inductance as possible.

Plug-in coils for the higher frequencies are usually made of sufficiently substantial material to be self-supporting. If, however, any support is needed a low-dielectric-constant material such as PTFE should be used. Sockets for the coils should be mounted on similar material and adequate clearance for the sockets from the box should be provided. These precautions assist in the attainment of the highest frequencies.

The indicator may be either a low-reading microammeter or a more robust instrument operated by a simple amplifier. The tuning control should for preference be driven by a slow-motion dial, although a large-diameter dial operated by the thumb has some merit in this type of instrument.

A VHF dip oscillator

This dip oscillator [5] covers the band 29–460MHz in four overlapping ranges with plug-in coils. The ranges are:



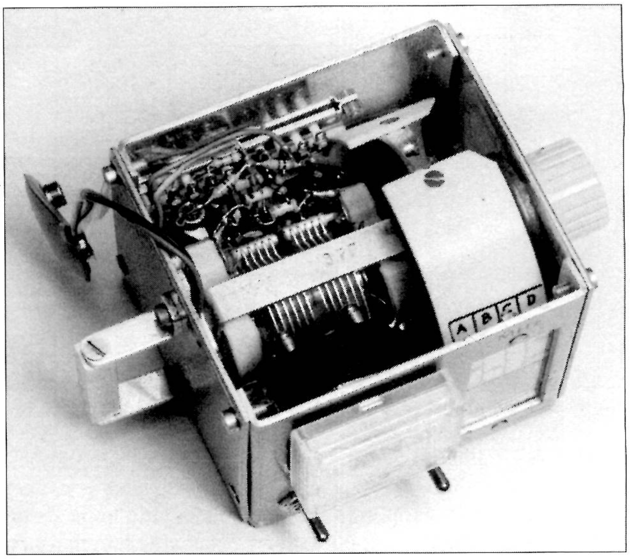


Fig 11.50. VHF dip oscillator: front panel (above) and inside view (right)

Range A	29 to 55MHz
Range B	50 to 109MHz
Range C	97 to 220MHz
Range D	190 to 460MHz

It is shown in the photograph (Fig 11.50).

Circuit description

The circuit is based around a Kalitron oscillator formed by two junction-type FETs TR1 and TR2 – see Fig 11.51. The frequency-determining components are the split-stator capacitor C1 and the plug-in coil L1. The resulting RF signal is then detected by a balanced diode detector D1 and D2, and used to drive meter amplifier TR3. The original design used either 2N5245 or TIS88 for TR1 and TR2 but these are no longer available. These can be substituted by a BF256A or similar N-channel junction FET.

The power can be turned off to convert the instrument to a sensitive absorption wavemeter, or when headphones are plugged into the jack J1 it becomes a modulation monitor.

In this design the existence of spurious dips and 'suck-outs' in the various ranges is very much associated with the quality of the two RF chokes L2 and L3. These are each of 15μH.

If troubles of this kind are experienced, other inductors can be tried. It is very difficult to find components with no strong resonance over the whole of a wide band, but nevertheless the prototype instrument using the inductors with the two series damping resistors R4 and R5 seemed to minimise the problem.

Power supply

This is normally provided by a PP3 battery, the current drain being about 10mA (including the LED). A PP3 replacement mains supply could be substituted.

Construction

Construction can either be on tag strip (Fig 11.52) or using a PCB as given in Appendix 1. The most important points are to keep the leads to the tuning capacitor as short as possible, using copper strip to keep the inductance low. Also keep all other RF leads short, especially any that are associated with the sources of TR1 and TR2.

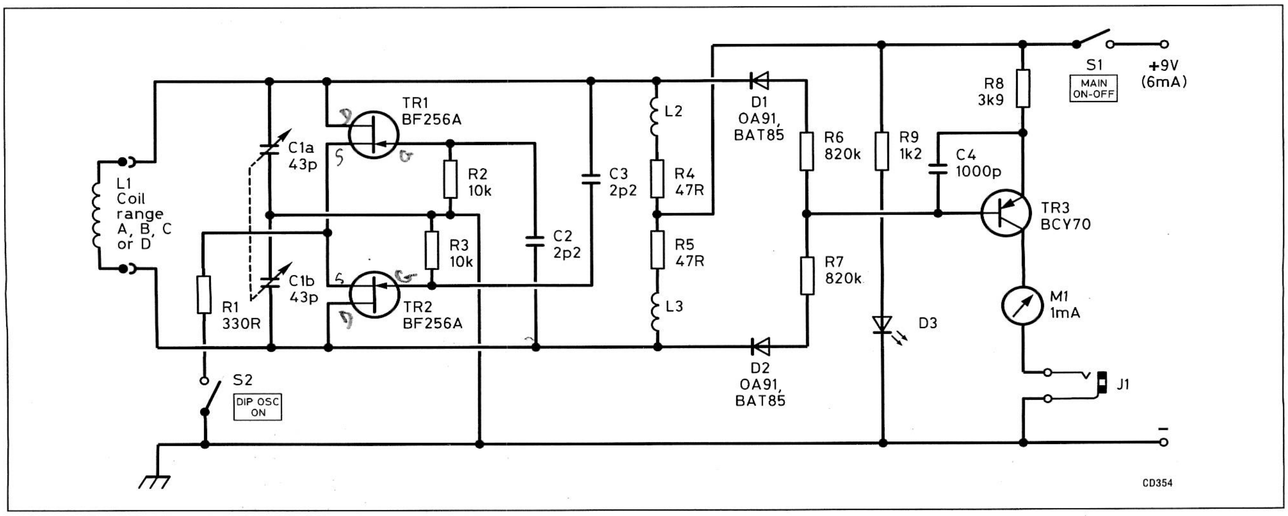


Fig 11.51. Circuit diagram of the VHF FET dip oscillator

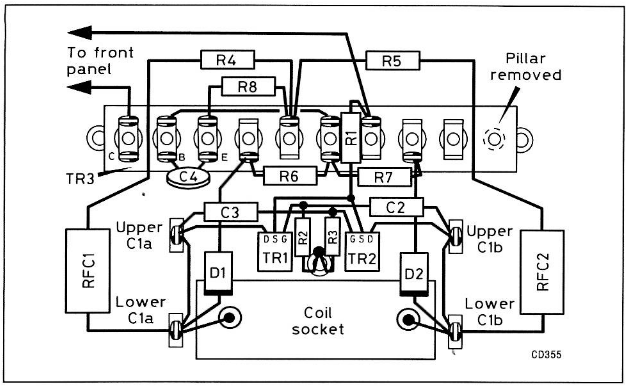
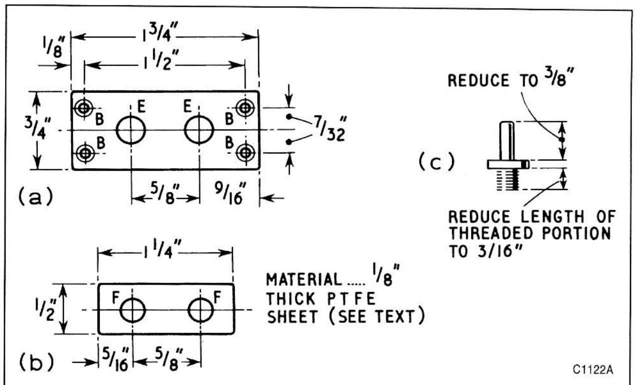


Fig 11.52. Tag board construction



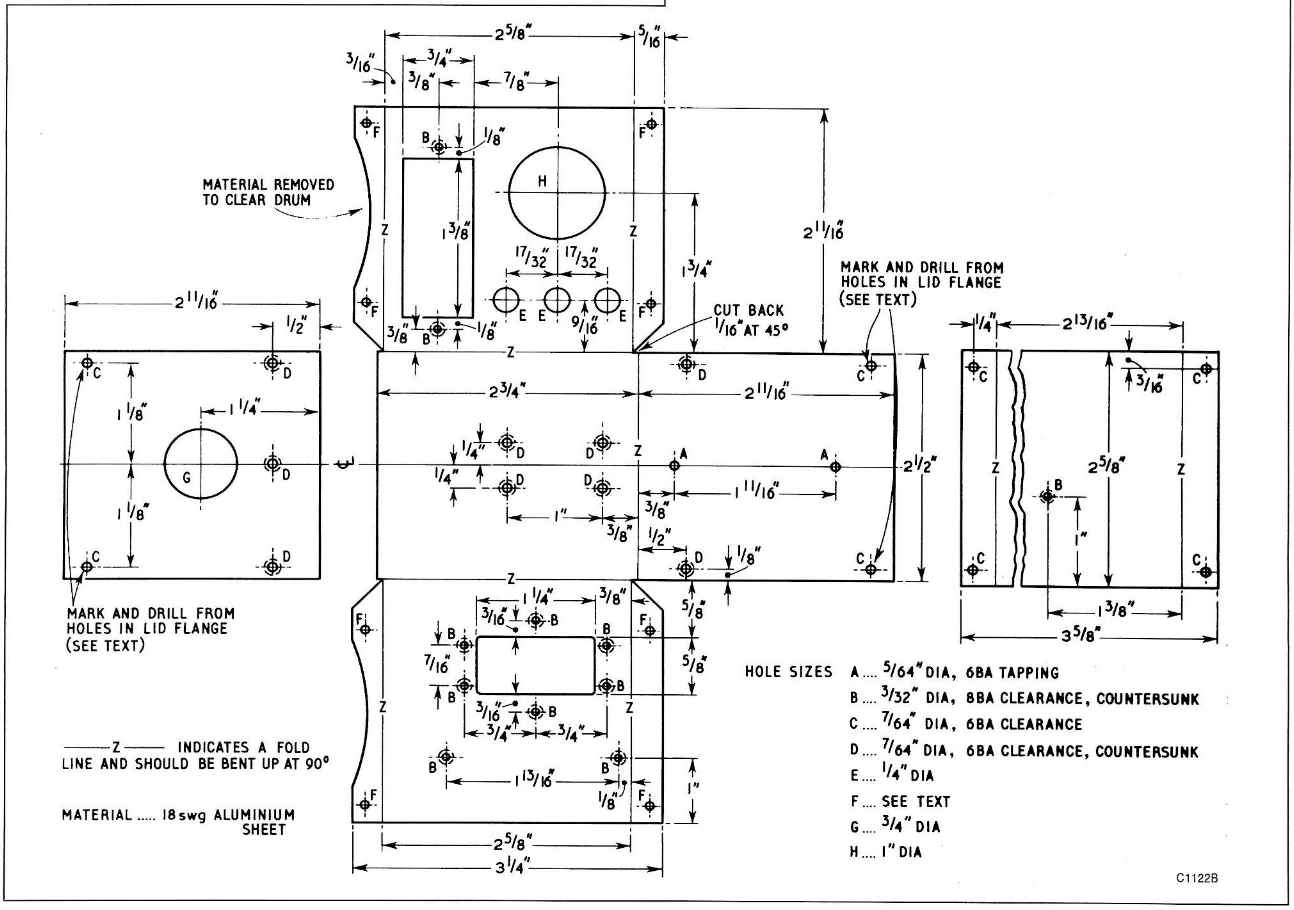


Fig 11.53. Mechanical details for the VHF dip oscillator (1in = 25.4mm)

The original version, as shown in the photographs, was constructed in an aluminium box forming about a 50mm (2in) cube and a design for making this is given in Fig 11.53. Following this construction is, however, a matter of personal choice.

The tuning capacitor C1 with a tuning scale drum is driven by a 6:1 reduction drive. A home-made 22mm (0.875in) wide card or plastic scale is fitted to the drum.

The coil socket and the coil mounting strips are made from 3mm (0.125in) thick PTFE sheet, although polythene or even polystyrene is acceptable (Fig 11.53). Two Belling-Lee 4mm sockets (similar to O-Z pattern) are mounted on the socket strip.

Coil construction

The coil (see Table 11.9) for the lowest frequency range, A, is wound on a short piece of 12.5mm (0.5in) diameter polystyrene rod and then glued in place. Connecting and supporting legs (each 44.5mm or 1.75in long) are made from 14 SWG enamelled copper wire. The next range coil, B, is self supporting and wound directly with 14 SWG wire, also with connection pieces 44.5mm (1.75in) long.

Table 11.8. Component list for the VHF dip oscillator

R1	330R
R2, 3	10k
R4, 5	47R
R6, 7	820k
R9	1k2
C1	43p + 43p variable
C2, 3	2p2 ceramic
C4	1000p ceramic
D1, 2	OA91, BAT85
D3	Low-current red LED
TR1, 2	BF256A or similar – see text
L1	See text
L2, 3	15μH
M1	1mA FSD meter
S1, 2	On/off switch
J1	Jack socket to suit
PP3 battery connector	
Belling-Lee 4mm OZ similar plugs and sockets	

Resistors are 0.25/0.5W, ±5% unless stated otherwise.

Table 11.9. Coil details for the VHF dip oscillator

Range A	29–55MHz	12t 22 SWG enam copper wire on 12.5mm (0.5in) polystyrene rod 25mm (1in) long, 44.5mm (1.75in) legs 14 SWG
Range B	50–109MHz	8t 14 SWG enamelled copper wire wound on 9.5mm (0.375in) drill, 44.5mm (1.75in) legs
Range C	97–220MHz	16mm (0.625in) wide, 73mm (2.875in) long loop of 13 SWG enamelled copper wire
Range D	190–460MHz	8mm (0.3125in) wide, 17.5mm (0.6875in) long 26 SWG (or near) copper or beryllium/copper strip soldered directly across plug ends

Range C has a simple rectangular loop of 13 SWG enamelled copper wire, whilst the highest frequency range D requires the two plug sections to be further shortened (see Fig 11.53) and then a strip of copper foil or beryllium/copper sheet is soldered straight across their ends.

Calibration of dip oscillators

The easiest way to check the calibration of a dip oscillator is to listen for the output on a general-coverage receiver, an amateur receiver or scanner. (*Note:* When the signal is found, especially with scanners which are of very wide bandwidth, check that there is no response at one-half, one-third or one-fifth of the frequency in case the fundamental has not been found.)

This probably allows a good check on the calibration into the VHF range. Additional points can be found by using the second-channel response provided that the IF is known (the second channel response is 2 × IF removed from the normal response).

Another method is to use the resonances of lengths of feeder cables, providing that the velocity factor for the particular cable is known, so that the physical length corresponding to the wanted electrical half- and quarter-waves can be found.

Using the dip oscillator

Although the dip oscillator has a wide range of uses for measurements on both complete equipment and individual components, these all rely on its ability to measure the frequency of a tuned circuit. In use, the coil of the dip oscillator is coupled indirectly to the circuit under test with maximum coupling being obtained with the axis of the oscillator coil at right-angles to the direction of current flow. Coupling should be no greater than that necessary to give a moderate change on the dip oscillator meter. These requirements are shown diagrammatically in Fig 11.54.

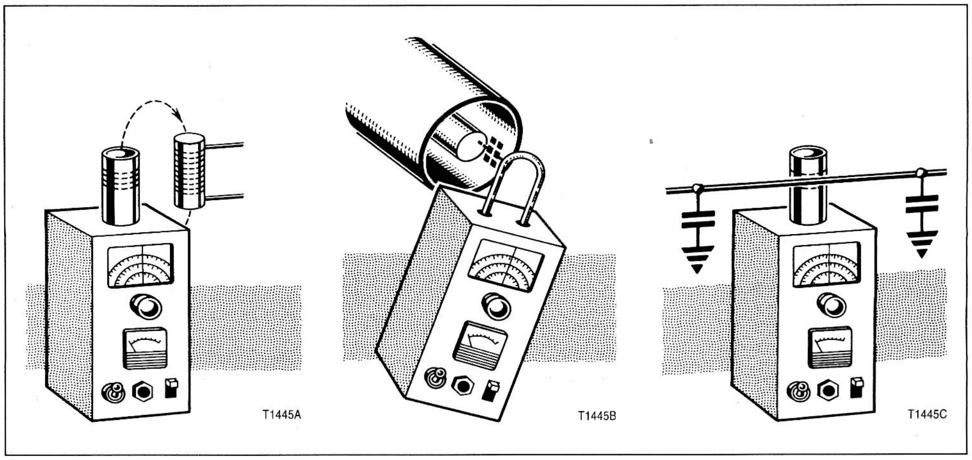


Fig 11.54. Using the dip oscillator

If the tuned circuit being investigated is well shielded magnetically (eg a coaxial line) it may be difficult to use inductive coupling. In such cases it may be possible to use capacitive coupling by placing the open end of the line near to one end of the dip oscillator coil. A completely enclosed cavity is likely to have some form of coupling loop and the dip meter coil can usually be coupled inductively by means of a low-impedance transmission line such as a twisted pair with a coupling loop.

When used as an absorption wavemeter, the oscillator is not energised and the tuned circuit acts as a pick-up loop. This arrangement is useful when looking for harmonic output of a multiplier or transmitter or for spurious oscillations.

Determination of the resonant frequency of a tuned circuit

The resonant frequency of a tuned circuit is found by placing the dip oscillator close to that of the circuit and tuning for resonance. No power should be applied to the circuit under test and the coupling should be as loose as possible consistent with a reasonable dip being produced on the indicating meter. The size of the dip is dependent on the Q of the circuit under test, a circuit having a high Q producing a more pronounced dip than one only having low or moderate Q.

Absorption wavemeter

By switching off the power supply in the dip oscillator and then using in the normal way, the instrument may be used as an absorption wavemeter. In this case power has to be applied to the circuit under test. Resonance is determined by maximum deflection on the meter caused by rectifying the received RF.

The absorption wavemeter can also be used to check for harmonics and spurii.

Capacitance and inductance

Obviously, if an instrument has the ability to measure the frequency of a tuned circuit, it can also be used for the determination of inductance and capacitance providing that one of these components is known or a substitution made. If one component is known then use of the resonance formula is all that is required, ie:

$$f = \frac{1}{2\pi\sqrt{LC}}$$

Otherwise, carry out the following steps. To measure capacitance connect a close-tolerance capacitor C_S (eg 1%) in parallel with a coil (any coil will do providing it will resonate at a frequency within the range of the dip oscillator). This circuit is coupled to the dip oscillator and its frequency f_1 megahertz noted. The unknown capacitor C_X is then connected in place of the known value and the new resonance f_2 megahertz noted. The unknown capacitance is then given by:

$$C_x = \frac{f_1^2 C_s}{f_2^2}$$

A similar substitution can be made for inductance, assuming a known close-tolerance inductor is available. In this case replace C_S and C_X by L_S and L_X respectively.

Signal generator

For receiver tuning, the dip oscillator may be used to provide an unmodulated carrier wave by tuning the oscillator to the desired frequency and placing it close to the antenna terminal. The amplitude of the signal may be controlled by adjusting the distance of the dip oscillator from the antenna terminal.

Checking crystals

The frequency at which a crystal is oscillating can be checked by using the dip oscillator as an absorption wavemeter as described above. It is also possible to check in a similar manner the harmonic on which a crystal is oscillating.

SPECTRUM ANALYSIS

Spectrum analysers are perhaps one of the best ways of examining the performance of filters, transmitters, receivers etc. They allow us to look at the frequency response in a pictorial manner. Fig 11.55 shows a typical display. From this we can see that the frequency is displayed on the horizontal axis whilst the response (or output) of a circuit is shown on the vertical axis.

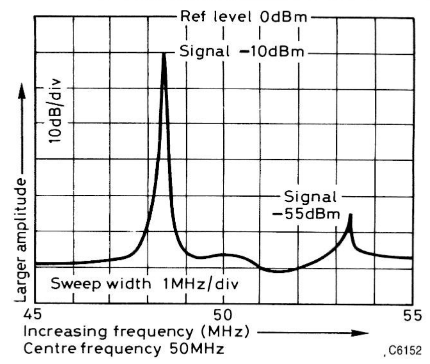


Fig 11.55. A typical screen display of a spectrum analyser

Unfortunately spectrum analysers tend to be quite expensive, although some can be picked up on the second-hand market. An excellent unit to build at home was described in *Radio Communication* [6] but its top frequency range is about 90MHz. This is satisfactory in examining the modulation of 50 and 70MHz signals but we should ideally be able to examine up to about the third harmonic in an item such as a transmitter; if only the modulation pattern or a frequency response of a filter is required we may manage with much less.

An alternative approach to a self contained spectrum analyser is the use of an add-on to an oscilloscope, where the oscilloscope display becomes the spectrum analyser display. These require an oscilloscope with X, Y mode facilities but the bandwidth of the oscilloscope is not important – that is taken care of in the add-on unit. These usually have a lower frequency limit (eg 400kHz), the upper limit being dependent on type, eg 100, 250, 500MHz. They come in formats from a large probe to a stand-alone box and prices vary from about £250 to £1000 (1997 prices).

DIGITAL FREQUENCY COUNTERS

This type of frequency meter utilises integrated circuits to count electronically the number of cycles of an input waveform in a given counting period. Although these digital integrated circuits are themselves complex, the principle of operation of the digital frequency meter (DFM) is quite simple; it consists of five major circuits, *the input wave shaping circuit, clock, gate, counter* and *display.*

The input wave shaping circuit takes the input waveform, amplifies it and converts it to a rectangular waveform of sufficient magnitude to operate the counting circuits.

The clock produces a series of pulses which determine the basic counting period of the DFM. The pulses are typically 10ms, 100ms or 1s long and are derived from a crystal-controlled oscillator (typically 1MHz or 5MHz). The pulses are thus of high accuracy and applied to the gate which can be considered as on on/off switch operated by the clock.

When the clock pulse opens the gate, an input train of pulses from the wave-shaping circuit is sent to the counting circuits. The counting circuits count the number of input pulses for the duration that the gate is open; this count is then frozen and the binary count decoded and used to drive decade displays.

The accuracy of a DFM depends on the accuracy of the internal clock signal. Accuracy can be increased if the clock

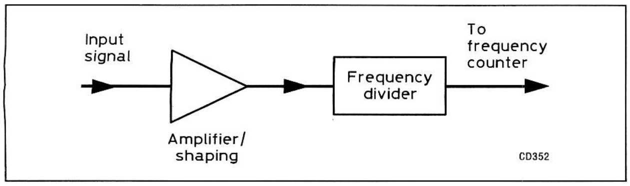


Fig 11.56. Prescaling

oscillator is housed in a crystal oven (thermostatically controlled) and compared with a standard frequency source for calibration. The resolution of a DFM is the smallest digit a display exhibits, usually 1Hz. The resolution will change according to the frequency range in operation and the length of the clock signal.

The DFM can also be made to display the period of an input waveform. In this case one period of the input allows the number of standard pulses to be counted (eg 1μs or 1ms). The number counted in one period is then the time for a period.

Prescalers

To increase the frequency range of a DFM it is common to put components that perform prescaling ahead of the input – Fig 11.56. This consists of an amplifier, wave shaping circuit and high-speed frequency divider (the prescaler). Prescaling will divide the input frequency by a known amount (eg 2, 10, 64 or 100) and the resulting signal is then applied to the basic frequency counter. It should be borne in mind that prescaling reduces the resolution of the counted frequency. For example, a frequency counter without prescaling may measure to 1Hz – if prescaling of 10 is introduced then the same counter will only read to 10Hz.

There are various prescalers on the market and Table 11.10 gives just a representative sample. The devices can have either ECL, TTL or both outputs.

The counter following has prescaling which extends its limit up to 600MHz. For prescalers and amplifiers to extend the frequency range above this the reader is referred to reference [7] or any of the typical data sheets/application notes produced by the various manufacturers. As with all VHF/UHF circuits, great care should be taken with layout – always use an earth plane. Care should be taken in the choice of capacitors – these should be, in order of preference, ceramic chip, silver mica and ceramic disc.

Table 11.10. Typical prescalers available

Device No	Man	F_{max} (MHz)	Division ratio	V_{cc} (V)
CA3179	H	1250	64/256	5
CXA1541M	S	1100	64/65, 128/9	5
MC1690	M	2500		
MC12023	M	225	64	3.2–5.5
MC12073	M	1100	64	5
MC12074	M	1100	256	5
MC12090	M	750	2	5
SP8629	P	200	100	
SP8660	P	150	10	5
SP8680B	P	575	10/11	5
SP8704	P	950	64/65, 128/129	3–5
SP8713	P	1100	64/65 or 72	2.7–5
SP8714	P	2100	32/33, 64/65	2.7–5
SP8715	P	1100	64/65, 128/9	2.7–5
SP8718	P	520	64/65	5
SP8755	P	1200	64	5
SP8799	P	225	10/11	5.2 or 6.8–9.5
SP8830	P	1500	10	5
U666BS	T	1000	256	5

H – Harris, M – Motorola, P – GEC Plessey, S – Sony, T – Telefunken

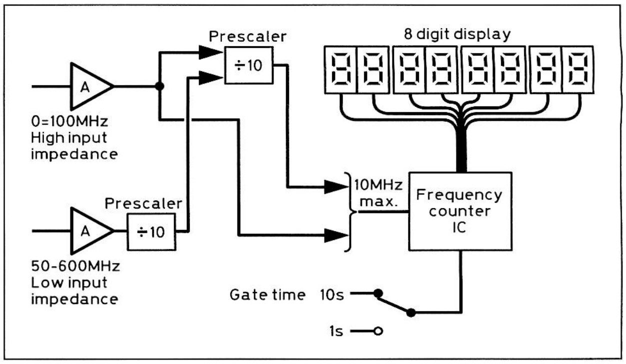


Fig 11.57. Block diagram of frequency counter

A 600MHz digital frequency counter

This frequency counter is capable of operating up to about 600MHz with a sensitivity of 50mV or better. It is based on an eight-digit frequency counter IC which will count to 10MHz. The range is increased by the use of prescalers which divide the incoming frequency down to the 10MHz range. Fig 11.57 shows a block diagram of the system. It offers three input ranges, 0–6MHz (typically 10MHz), 0–60MHz and 0–600MHz. It is possible to build a 0–60MHz counter and then add the additional components later to increase the capability to about 600MHz.

The gate time, ie the count period, can be switched between 1s and 10s. At the 1s position the counter offers a seven-digit display, while at the 10s position the counter provides an eight-digit display. The decimal point is switched to denote the megahertz position, all digits to the left of it representing megahertz. The supply requirement is 5V at approximately 300mA.

References [8], [9], and [10] provided the basic information for the design; for extension in frequency ranges and sensitivity the reader is referred to [7].

Input circuit up to 60MHz

The circuit diagram is shown in Fig 11.58. The input impedance is approximately 1MΩ, with junction FET TR1 acting as a buffer. The signal is then fed to IC3 which contains a group of transistors capable of operation up to 1GHz. This provides amplification and an output capable of driving logic circuits. The output transistor, TR2, is a high-speed PNP switch which translates the ECL output logic level of IC3 to a TTL compatible level.

Input circuit 10–600MHz

This circuit is shown in Fig 11.59 and has an input impedance of approximately 50Ω. IC1 is a monolithic amplifier with 50Ω input and output impedances and operates up to 1GHz with a gain of some 16dB. The resulting output is fed to IC2 which is a high-speed divide-by-10 counter capable of operating up to at least 575MHz (typically more than 600MHz) with a sine-wave input. The output selected is TTL compatible.

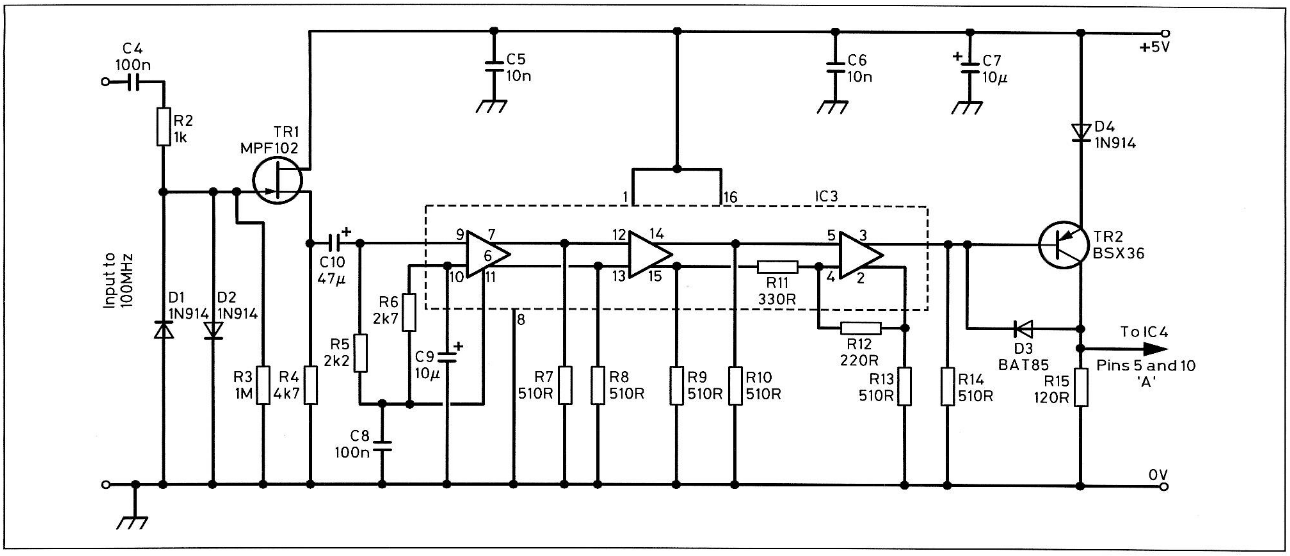


Fig 11.58. Input circuit (up to 60MHz)

The circuit will operate without IC1 but is not as sensitive. Although a Mini-circuits monolithic broad-band amplifier is used it is possible to use those from other manufacturers such as Avantek.

Control, counter and display circuit

Fig 11.60 shows all of the remaining digital side of the frequency counter and displays. The initial stages of the frequency counter consists of the input selection and prescaling circuits.

IC4 and IC5 are both 74F 'FAST' TTL in order to cope with the speed. IC4 is a multiplexer which is controlled by DC signals supplied via S1a and determines the route of signals from the input amplifiers to the counter IC. IC5 is a divide-by-10 circuit which provides prescaling for all signals above 10MHz. IC6 is the heart of the digital frequency meter and provides all the count circuits, decoders and drivers for the eight-digit LED display. Timing is derived from a 10MHz crystal oscillator.

Switches S1b/c determine the position of the decimal point in conjunction with the gate time selection switch (S2) for the various inputs. PB1 is a reset button for the counter.

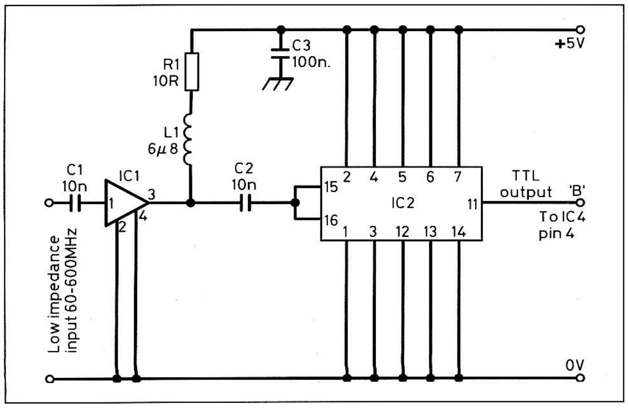


Fig 11.59. Input circuit (10 to 600MHz)

Construction

The components required are listed in Table 11.11. A double-sided PCB and component layout is provided in Appendix 1.

Table 11.11. Component list for 600MHz digital frequency counter

R1	10R
R2, 16, 17	1k
R3	1M
R4	4k7
R5	2k2
R6	2k7
R7–10, 13, 14	510R
R11	330R
R12	220R
R15	120R
R18, 20	10k
R19	22M
R21	100k
R22	3k3
C1, 2, 5, 6	10n ceramic disc
C3, 4, 8, 11, 14, 15	100n ceramic disc
C7, 9	10µ,16V tantalum
C10	47µ, 6V tantalum
C12	47p polystyrene or ceramic
C13	100p polystyrene or ceramic
VC1	65p trimmer
L1	6µ8 RF inductor
D1, 2, 4	1N914, 1N4148
D3	BAT85
TR1	MPF102
TR2	BSX36,2N5771, fast PNP switch
IC1	MAR-1
IC2	SP8680B
IC3	MC10116
IC4	74F153
IC5	74F160
IC6	ICM7216D
XL1	10MHz crystal
LD0–7	Double 7-seg LED (4 off), CC, 0.5in
S1	4p, 3w rotary
S2	DPDT toggle, PCB mounting

Resistors are 0.25/0.5W, ±5% unless stated otherwise.

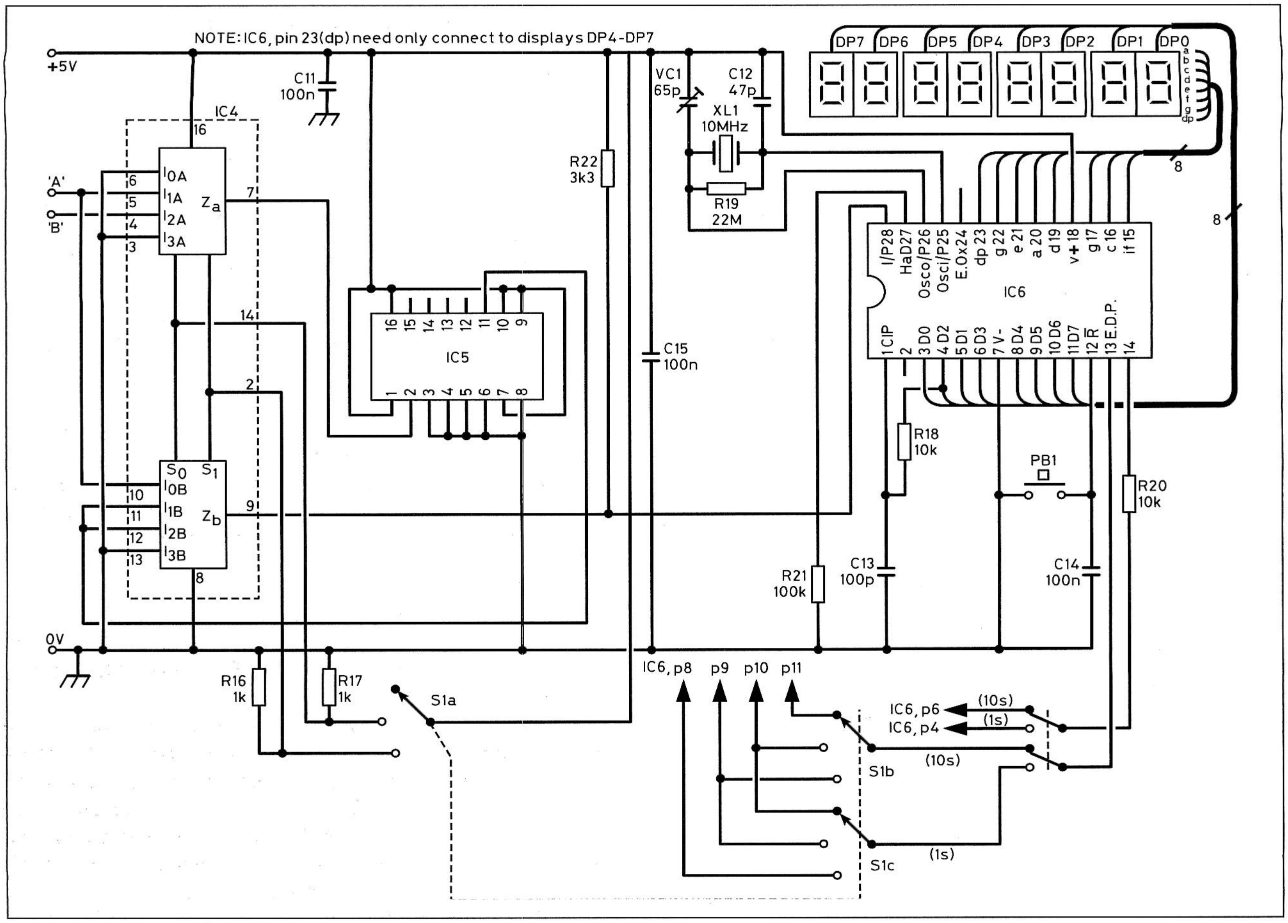


Fig 11.60. Control, counter and display circuit

If you wish to make your own, it is essential that an earth plane is provided on the copper side for the two input circuits on Fig 11.58 and Fig 11.59. It is wise to use good layout practice for the digital circuits, especially where they can run up to 100MHz. Adequate decoupling is also essential.

Do not use IC sockets for IC1, IC2, IC3, IC4 and IC5. The dual LED displays should be mounted on single-row sockets. The PCB is double sided – through connections are made mainly via component leads and this includes IC sockets. The sockets may therefore have to stand off the board slightly in order to solder 'on the top side'. It is suggested that turned pin sockets are used. There are three through connections to be made with wire links. Where necessary, clearance must be made around holes on the earth plane where connection is not made to the earth plane – use a sharp drill or cutter for this.

If a counter is only required up to 10MHz then the output of the circuit of Fig 11.58 can be fed straight to pin 28 of IC6, omitting IC4 and IC5 and the range switch S2. If an input is required only up to 60MHz then omit the components associated with the 600MHz input amplifier and alter the stop on S1 for only two positions.

Calibration

Using a highly accurate frequency counter which is calibrated against a standard, feed the same input signal into both counters and adjust VC1 for the same reading.

HF/UHF DEVIATION METER

The HF/UHF deviation meter described herein by G3BIK [11] is simple to construct, of relatively low cost and based on readily available components. It is housed in a small RF-shielded plastic enclosure. It can be used with FM signals over a wide frequency range from about 3MHz to 450MHz. The power supply requirement is 15–20V DC at about 50mA – see later.

Fig 11.61 shows the block diagram of the instrument, the interconnections and signal routing, Fig 11.62 gives the circuit diagram, and the components are specified in Table 11.12. A PCB pattern and component layout is given in Appendix 1.

Fundamentally the circuit is a type of superheterodyne receiver with an FM demodulator output which is displayed on a moving-coil meter showing kilohertz deviation. The instrument can also be used to monitor the demodulated audio signal on a small internal loudspeaker, and provision has been made for external access to the pulsed RF harmonics of the VFO as a simple source of test signals, extending up to VHF, which could be useful for calibration or general receiver testing.

Circuit operation

A restricted-range signal f_2 (1.5MHz) is generated by the Colpitts oscillator (VFO) formed by TR1 and associated components. The variable-capacitance diode in conjunction with

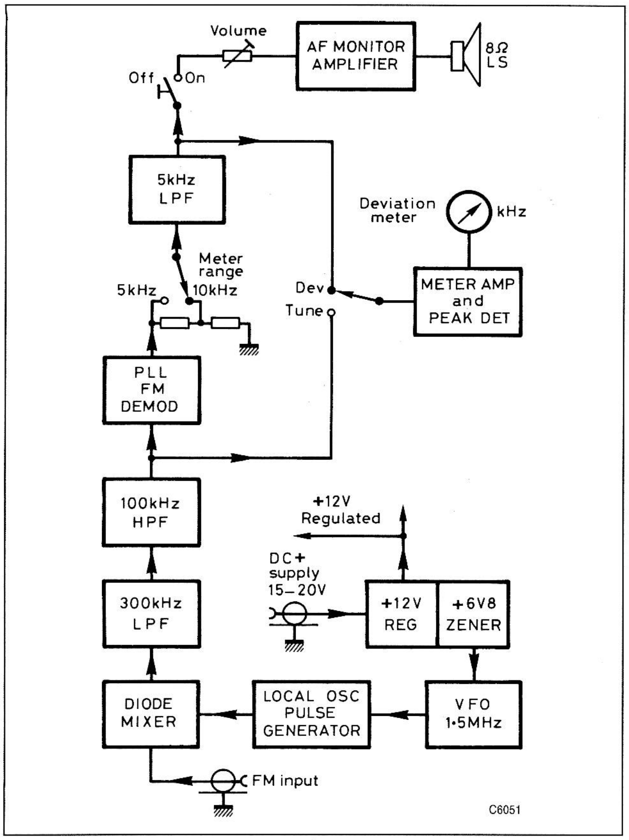


Fig 11.61. Block diagram of the deviation meter

RV1 allows variation of this frequency by about ±100kHz. The sine-wave output from this oscillator is buffered by TR2 and then amplitude limited by TR3–5 and D3. It is then differentiated by TR6 and TR7 which results in a train of sharp pulses at the VFO repetition rate.

The pulses are applied to a fast-switching diode (D4). The action of the diode causes current pulses through the primary of pulse transformer T1. These current pulses are rich in harmonics and extend well up into the VHF region. *Note*: the usable upper frequency is to a large extent determined by the switching diode. A microwave step-recovery diode would probably extend the frequency range but these devices are not cheap. Transformer T1 is wound in bifilar manner, it being important to get the phasing correct.

The two phase-related outputs from the transformer are applied to the diode bridge mixer formed by D5 and D6 and here mixed with the FM signal (f_1) being measured. The mixing products are developed across R41 and the signal $f_1 - f_2$ (harmonic) filtered by the action of the various constituents of IC1 (range 100–300kHz). This signal is then demodulated by the phase-locked loop of IC2. The output at pin 7 of the PLL is a true replica of the input modulating signal and this is attenuated by R57/RV3 for the two meter ranges. The output is selected by S1 and then amplified by a low-pass active filter IC3a. With S2 in the TUNE position, the IF signal at about 150kHz is passed to the meter amplifier IC3b/c. Diode D7 with capacitor C58 acts as a peak detector and the resulting DC drives the meter. With S2 set to DEV, then the demodulated signal is applied to the meter circuit.

An AF monitor is provided (IC4). The drive to this is via a push switch (S3) so that the user is deterred from using the monitor during measurement of deviation.

Power supply

A single supply philosophy is adopted for the deviation meter via an on-board 12V regulator, thus ensuring stability of deviation calibration and flexibility in the choice of external voltage supply.

All of the discrete transistors are fed from the full +12V regulated rail except the VFO which is further stabilised by a 6.8V zener diode. Split-voltage supplies for IC1, IC2 and IC3 are provided by resistive dividers R50/51, R52/55 and R71/72 and well decoupled.

Adequate decoupling of the DC supply connections is of the utmost importance in this circuit because of the pulsed nature of the local oscillator and mixer signals – hence the provision of low-value decoupling capacitors to filter out the higher frequencies from the supply rails and larger capacitors (microfarads) to deal with the lower frequencies.

Construction

The prototype circuit was constructed on SRBP copper stripboard of 0.1in hole spacing and dual parallel planes were allocated on the stripboard for each of the several DC supply and return rails.

Component layout is not critical provided that, as with any HF circuit, due attention is given to the shortness of component leads and interconnecting wires, and to mechanical rigidity. Sub-miniature 50Ω coaxial cable is employed for the interconnection between signal input socket and mixer, and miniature screened cable is used between all other panel-mounted components and the circuit board.

The RF screened enclosure is electrically bonded to the negative supply terminal, shown on the circuit diagram as the 0V rail, and the decoupling capacitors associated with the integrated circuits should be connected directly between the supply pins of the appropriate IC.

The panel-mounted miniature volume control was found to be not really necessary in practice, so this could be replaced by a presettable potentiometer mounted internally on the circuit board.

As the upper operational frequency limit of the instrument is largely determined by the performance of the local oscillator harmonic-generator diode D4, pulse transformer T1 and the D5/D6 diode mixer circuit, particular care should be taken to ensure shortness of component leads and symmetry of component layout for this section of the circuit.

Setting up

Before any attempt is made to use or calibrate the instrument the free-running frequency of the phase-locked loop VCO must be set to 150kHz.

Connect a frequency counter to either pin 4 or 5 of IC2 and adjust RV2 until the frequency is virtually 150kHz.

Calibration

As for the calibration of most measuring instruments the crunch comes with the requirement for a signal source of known accuracy. The ideal source for this particular application would be an RF signal generator with a frequency range of 2–450MHz, and a frequency modulation facility with

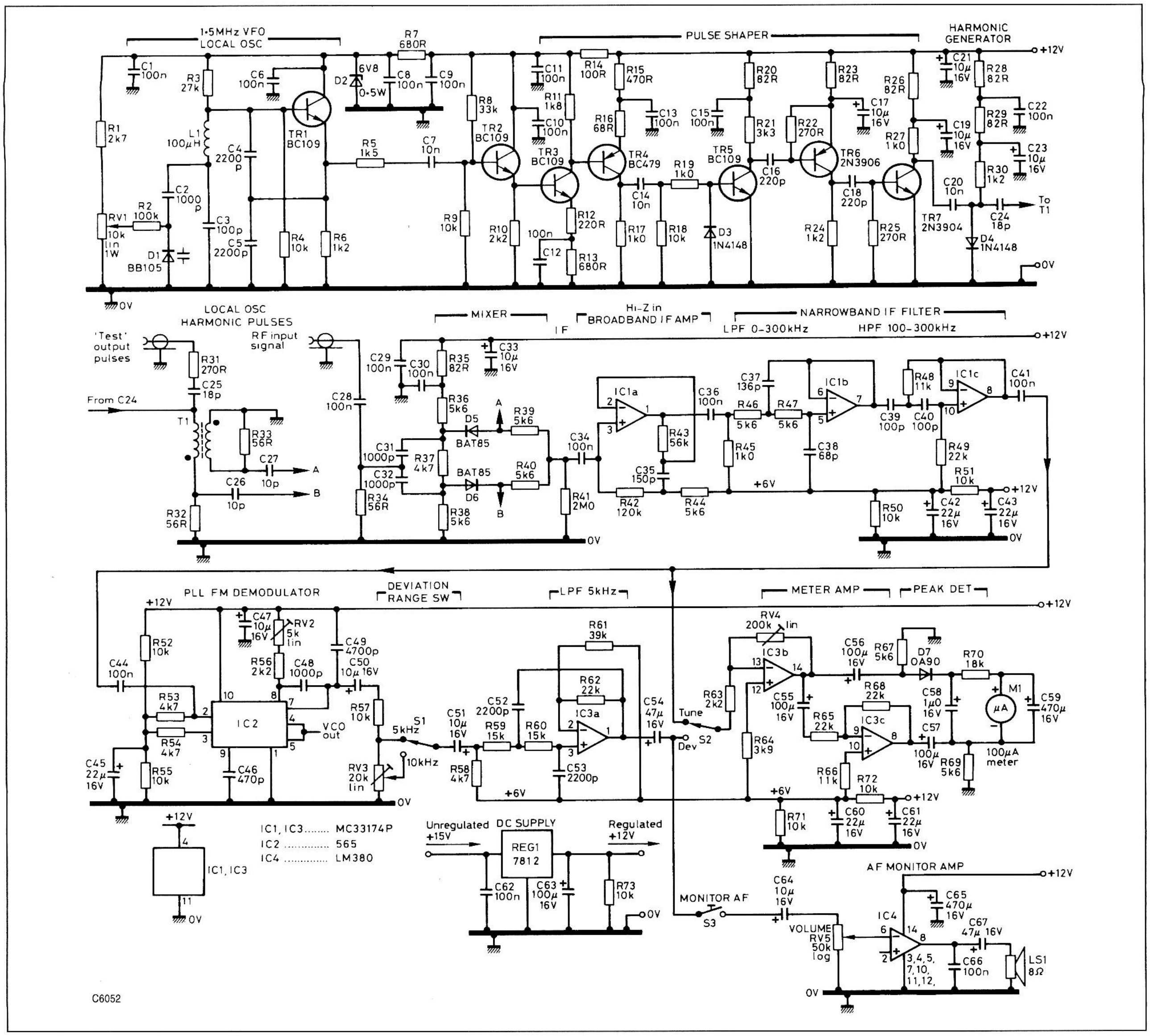


Fig 11.62. Circuit diagram of the deviation meter

calibrated deviation from zero to 30kHz at an audio frequency of 1kHz. Output attenuation down to 5μV PD across a 50Ω load is also required.

With the deviation meter switch turned ON, set switch S2 (TUNE/DEV) to TUNE. Connect the signal generator to the input of the deviation meter, and set the frequency of the generator to 50MHz with an output of 50μV PD and zero deviation. Rotate the VFO knob slowly until a maximum deflection is obtained on the meter. If the meter deflection is greater than full-scale adjust the signal generator attenuator to bring it back on to scale.

A number of maxima of differing amplitudes will be observed as the VFO knob is rotated, the choice of which is not critical. It is sufficient to opt for one which appears to be dominant, then carefully tune for the peak. Switch from TUNE to DEV (S2) and the meter reading should fall to zero (no deviation on input signal).

Set the deviation switch of the meter (S1) to the 5kHz position. Frequency modulate the signal generator with a 1kHz signal at 5kHz deviation. The resultant meter deflection can now be adjusted with preset resistor RV4 until the meter reads full-scale. Now switch S1 to the 10kHz setting and the meter will take up a lesser deflection. Adjust RV3 until a mid-scale reading is obtained. Assuming that the meter is scaled 0–10, then with S1 set for 10kHz, the scale reading is read as kilohertz deviation.

Change the signal generator to 10kHz deviation and note the meter reading. It should read full-scale – if not adjust RV3 slightly. Reduce the deviation in steps of 1kHz on the signal generator to confirm the true linearity of the indicating meter circuit.

While the signal generator is still connected it is prudent to determine the upper frequency limit of the meter. Turn off the deviation on the signal generator and set S2 to TUNE.

Table 11.12. Components list for the HF/UHF deviation meter

R1	2k7	R61	39k	C58	1µ electrolytic
R2	100k	R64	3k9	C59, 65	470µ electrolytic
R3	27k	R70	18k	L1	100µH
R4, 9, 18,		R71–73	10k, 0.5W	T1	Pulse transformer,
57	10k	RV1	10k lin,1W cermet		10mm ferrite ring, 4t
R5	1k5	RV2	5k lin, min cermet		bifilar ECW
R6, 24, 30	1k2	RV3	20k lin, min cermet	D1	BB105, BB405B, BB809
R7	680R, ½W	RV4	200k lin, min cermet		varicap
R8	33k	RV5	50k log carbon pot with	D2	6V8, 400mW
R10, 56, 63	2k2		S3	D3, 4	1N4148
R11	1k8	C1, 6, 8–13,		D5, 6	BAT85
R12	220R	15, 22,		D7	OA90
R13	680R	28–30, 34,		TR1-3, 5	BC109, 2N3904 or
R14	100R	36, 41, 44,			similar
R15	470R	62, 66	100n ceramic	TR4	BC479, 2N3906 or
R16	68R	C2, 48	1n ceramic		similar
R17,19, 27,		C3	100p SM/polystyrene	TR6	2N3906
45	1k	C4, 5, 52,		TR7	2N3904
R20, 23, 26,		53	2n2 polystyrene	IC1, 3	MC33174P
28, 29, 35	82R	C7,14, 20	10n ceramic	IC2	NE565
R21	3k3	C16, 18	220 ceramic	IC4	LM380
R22, 25, 31	270R	C17, 19, 21,		REG1	12V, eg 7812
R32, 33, 34	56R	23, 33, 47,		LS1	8Ω miniature
R36, 38, 39,		50, 51, 64	10µ electrolytic	M1	Meter, 100µA, scaled 0–10
46,47	5k6	C24, 25	18p SM/ceramic		
R37, 53, 54,		C26, 27	10p SM/ceramic	S1,2	SPDT min toggle
58	4k7	C31, 32	100p ceramic	S3	Push-button, NO (part of RV5)
R40, 44, 67,		C35	150p SM/ceramic		
69	5k6	C37	136p (2 × 68p)	Enclosure: ABS, RF shielded, 190 × 110 × 60mm	
R41	2M (2 × 1M)	C38	68p SM/ceramic		
R42	120k	C39, 40	100p SM	Socket (DC): 3.5mm chassis mounting	
R43	56k	C42, 43, 45,			
R48, 66	11k	60, 61	22µ electrolytic	Socket (RF): 50Ω BNC panel mounting	
R49, 62, 65,		C46	470p SM/ceramic		
68	22k	C49	4n7 ceramic	Socket (pulse): 50Ω BNC panel mounting	
R50–52, 55	10k, 0.5W	C54, 67	47µ electrolytic		
R59, 60	15k	C55–57, 63	100µ electrolytic		

Resistors 0.25W, ±5% metal film unless stated otherwise.

Progressively increase the carrier frequency in sensible increments, retuning at each step, until it is no longer possible to obtain a suitable maximum on the meter. If required, the frequency tuning range of the VFO may be conveniently measured at the test pulses output socket.

The instrument is now ready for use.

Use

Apply the RF signal to be measured to the input socket via a coaxial cable or by plugging into the socket an elementary pick-up antenna of some 100–200mm length. In either case the instrument will behave correctly with an input as low as 5µV but a higher signal is preferred, eg 50µV.

With no modulation of the RF carrier and S2 set to TUNE, rotate RV1 for maximum deflection on the meter. Switch S2 to DEV and read the meter, altering S1 if necessary.

NOISE MEASUREMENTS

Noise can be defined as any unwanted disturbance that is superimposed on a wanted signal. It will interfere with the information contained within the wanted signal and in the limit will prevent it being decoded.

In TV reception it can result in a grainy picture with possible white and black spots, and in the limit loss of picture. In radio reception it can produce crackling or hissing which in the limit will mask the speech or music. In data transmission it will affect the reliability with which the data is decoded, a poor signal with a high noise level providing a much higher error rate.

Where does noise come from? It can be generated externally to any equipment and can be classified as *man-made* (ie from electric motors, ignition systems etc) or *natural* (such as that from lightning discharges or stellar sources). In addition to these, there is what is known as *thermal noise*. At frequencies up to around 21MHz, the external noise is generally greater than any noise generated in a receiver. Above this frequency and up into microwaves the receiver-generated noise is generally dominant, especially at 144MHz and above.

Thermal noise is generated within components by the random agitation of atoms and results in a random voltage. At absolute zero this voltage is zero, increasing as temperature rises. It consists of frequencies that start virtually at DC and rise well into the gigahertz region. Some components generate more noise than others and, if the semiconductor literature is scanned, devices classified as 'low noise' will be noticed.

If a group of components in the front-end of a receiver generate noise in the microvolt region then this can well mask received signals of the same order. It is therefore important that receiver front-ends are well designed and set up in order to minimise the effect of noise. The following sections deal with equipment for this and how to perform these measurements.

Noise figures

In any amplifier, noise is added to the signal so that the signal-noise ratio at the output of the amplifier is worse than at the input even though the signal has been amplified. This is especially important in receiver RF amplifiers which deal with low-level signals. The ratio:

$$\frac{\text{Signal / noise in}}{\text{Signal / noise out}}$$

is defined as the *noise factor* but, as this ratio can have a wide range of values, it is convenient to express it in decibels (dB). The above ratio then becomes:

$$\text{Noise figure} = 10\log_{10}\frac{\text{s / n in}}{\text{s / n out}} \text{ decibels}$$

In a receiver, there are a number of cascaded stages which will each contribute noise but the effect of the noise contribution of each successive stage is reduced by the power gain of the preceding stage. Thus if F_1, F_2 and F_3 are the respective noise figures of each successive stage and P_1, P_2 and P_3 are the stage gains, then the overall receiver noise factor F will be given by:

$$F = F_1 + \frac{F_2 - 1}{P_1} + \frac{F_2 - 1}{P_1 \times P_2}$$

In most cases only the first and second terms are significant.

Noise performance is measured by noting the noise output of the receiver when its input terminals are terminated with the value of source resistance for which it is designed and then adding a known amount of noise at the input such that the value of output noise is doubled. It is then obvious that the added noise is equal to the noise generated by the receiver, although two assumptions are made for this to be true:

1. All of the known output from the noise source is, in fact, coupled into the receiver; and
2. the receiver output doubles when the effective input is doubled (ie the receiver is linear over this range of inputs).

The first point will be met provided that:

(a) none of the noise is shunted;
(b) transit time effects are negligible; and
(c) the output of the noise source is coupled into the receiver by a very short length of low-loss cable of the correct characteristic impedance.

The linearity of the receiver can be established by providing two identical noise sources and shunting the output of the receiver by a 3dB attenuator when the second source is switched on, but for amateur use this is scarcely worthwhile.

Noise figure measurements [12]

If the radio amateur can get hold of a calibrated noise source then it is possible to obtain accurate noise figures for a piece of equipment. Fig 11.63 shows a typical arrangement of equipment. The noise source should be matched to the receiver and the impedance across which the audio output power is measured must be known.

The first reading is taken with the noise generator turned off. The receiver audio gain is adjusted for a convenient noise reading in decibels as observed on the audio power meter.

The noise generator is next turned on and its output is increased until a convenient power ratio, expressed by N_2/N_1, is observed. The ratio N_2/N_1 is referred to as the *Y factor*, and

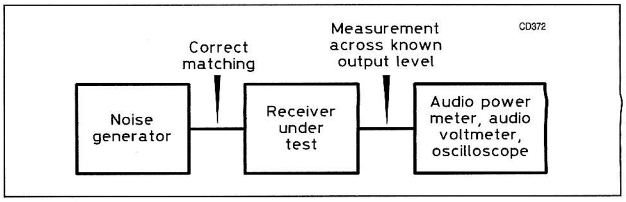


Fig 11.63. Arrangement for performing noise measurements

this noise figure measurement is commonly called the *Y-factor method.* From the Y factor and output power of the noise generator, the noise figure can be calculated, ie:

$$\text{NF} = \text{ENR} - 10\log_{10}(Y - 1)$$

where NF is the noise figure in decibels, ENR is the excess noise ratio of the noise generator in decibels and Y is the output noise ratio N_2/N_1. The excess noise ratio of the generator is:

$$\text{ENR} = 10\log_{10}(P_2/P_1 - 1)$$

where P_2 is the noise power of the generator and P_1 is the noise power from a resistor at 290K.

Most manufacturers of amateur communication receivers rate the noise characteristics with respect to signal input, and a common expression is:

$$\frac{\text{Signal + Noise}}{\text{Noise}}$$

or the *signal-to-noise ratio.* Usually the sensitivity is given as the number of microvolts for a signal-to-noise ratio of 10dB.

Typically, the noise figure for a good receiver operating below 30MHz is about 5 to 10dB. Lower noise figures can be obtained but they are of no real value due to the external noise arriving from the antenna. It is important to remember also that optimum noise figure in an RF amplifier does not always coincide with maximum stage gain, especially at VHF and higher. This is why actual noise measurements must be used to peak for best noise performance.

Noise sources

In the past various devices have been used as noise generators such as saturated thermionic diodes or the argon-filled fluorescent discharge tube. However, these devices have now become obsolete and semiconductor devices must be considered.

One noise source has been the reverse-biased germanium diode. Unfortunately the law relating noise with diode current varies for each diode and cases have been observed where the noise actually decreases for an increase of current over a limited range. Although individual diodes whose characteristics have been determined can be useful to optimise the noise performance of a receiver, they cannot be regarded as a measuring instrument. Fig 11.64 shows a typical circuit.

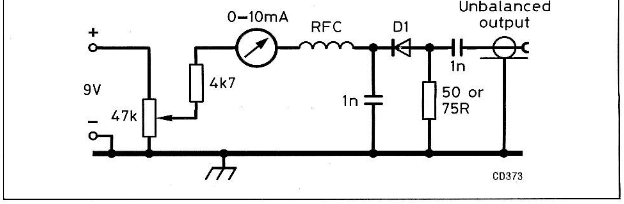


Fig 11.64. Noise generator using reverse-biased diode

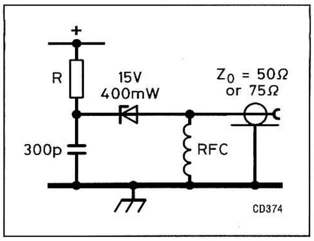


Fig 11.65. Zener diode reference noise source. Note that R should be selected for 5mA diode current from a supply of somewhat more than 15V: the components may conveniently be built into a coaxial plug

Silicon diodes in the avalanche region often produce wide-band noise. Zener diodes (especially those exhibiting avalanche breakdown) also provide a simple source of consistent noise over a wide frequency range. The noise output is not completely dependent on the current and they are not suitable as adjustable noise sources. A simple reference noise source may be made using the circuit of Fig 11.65, it being possible to mount the components all within a coaxial plug.

A zener/transistor noise source

This design is taken from reference [13] and the circuit diagram shown in Fig 11.66. The major noise source is zener diode D1 which is used to bias the first amplifier TR1. As there is no bypassing of the zener diode at the base of the transistor and the current in the diode is small, the excessive noise currents in the diode will flow through the base of the transistor. The resulting amplified output is applied to a second stage of gain, TR2. The second amplifier has a 51Ω resistor (R3) in the collector in order to provide a controlled output impedance.

The noise output of this circuit has been measured on a spectrum analyser. The detailed distribution of noise with frequency will not be presented since it will vary considerably with zener diode and transistor characteristics. Generally noise in the HF region was quite robust, reaching levels of 80dB higher than the noise output from a room-temperature resistor. The noise output is still 20dB above a 290K resistor at 432MHz.

The constructor should not attempt to estimate noise figure with a device as crude as this. It may be used, however, as a source for tuning receivers or amplifiers. If one were to build a free-running multivibrator using a 555 timer, with a total period of one to two seconds, it could be used to automatically turn the generator on and off. The system could then be used in conjunction with a step attenuator to adjust a VHF preamplifier for low noise figure. The output detector would be the constructor's ears, although refined circuitry could be built for this purpose.

Table 11.13. Components list for the zener/transistor noise source

R1, 2, 5	330R
R3	51R
R4	100R
C1–4	10n, ceramic
D1, 2	6V2, 400mW zener, eg 1N753A
TR1, 2	2N5179

Resistors are 0.25/0.5W, ±5% unless stated otherwise.

A gated noise source

The circuit [12] described in this section may be used to construct a simple low-cost device to optimise a converter/receiver for best noise figure. The simplicity of this system makes effective alignment possible without a lot of test equipment.

The circuit diagram is given in Fig 11.67. TR1 and TR2 form an astable multivibrator operating at about 700Hz. The value of C1 is chosen to be greater than C2 so that the duty cycle is deliberately not 50%. The output from TR2 is capacitively coupled via C3 to the base of TR3, which acts as the current source for D1 via R7 and R8.

The diode generates broad-band noise which is output to a socket via R9. R7, C4 and C5 form a low-pass filter to prevent high-order harmonics of the switching pulses from appearing at the output. It should be noted that an absolute value of noise figure is not obtainable with this unit.

The circuit uses readily available components and may be easily duplicated. The lead placement in and around the diode should follow good VHF practice with short leads and direct placement. The influence of stray RF signals entering the device under test through the generator may be minimised by shielding the components shown. The unit can be housed in a metal box or one made from PCB scraps.

For best match, this source should be connected directly to the input of the equipment under test; therefore the unit should be equipped with a male connector. This matching becomes a greater consideration as the frequency of interest increases.

The gated noise source does not require a special detector, or any detector other than one's ear. By turning the noise

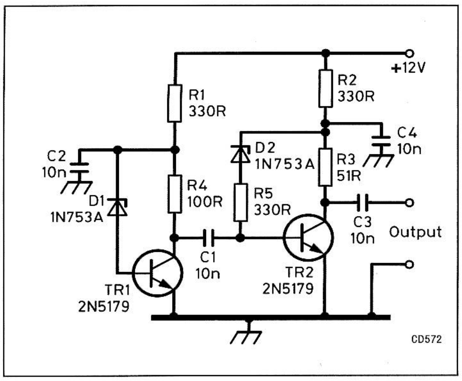


Fig 11.66. Circuit diagram of a noise generator (ARRL)

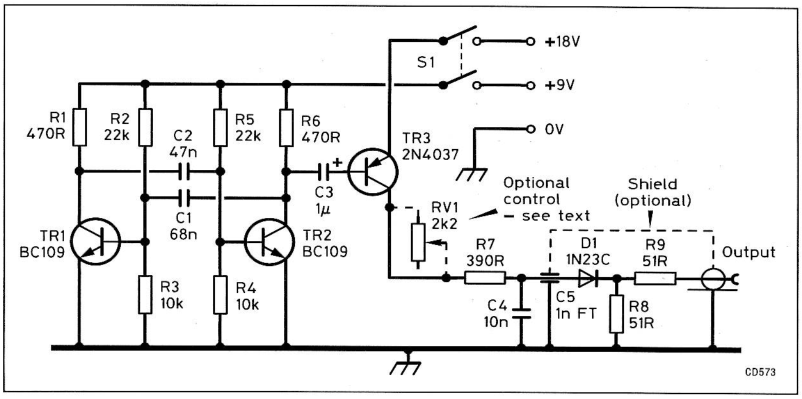


Fig 11.67. Circuit diagram of gated noise source (ARRL)

Table 11.14. Components list for the gated noise source

R1, 6	470R
R2, 5	22k
R3, 4	10k
R7	390R
R8, 9	51R
RV1	2k2 pot (see text)
C1	68n
C2	47n
C3	1μ, 50V
C4	10n
C5	1n feedthrough
D1	1N23C or equivalent
TR1, 2	BC109 or equivalent
TR3	2N4037 or equivalent
S1	DPDT switch

Resistors are 0.25/0.5W, ±5% unless stated otherwise.

source on and off at an audio rate, the ratio of noise contributed by the system to noise of the system plus excess noise appears as an audio note. The louder the note, the greater the differential in levels. Hence, the greater the influence of the excess noise, the better the noise figure.

If greater precision is desired than that obtained by subjectively listening to the signal, an oscilloscope may be used. Connect the vertical (Y) input to any point in the audio system of the receiver, eg loudspeaker terminals. Adjust the oscilloscope for a display of several multiples of the train of rectangular pulses. Proceed by adjusting the device being tested for greatest vertical deflection.

In some cases the available noise generated by this unit may be too great. The output may be reduced by inserting attenuators between the generator output and the equipment under test. Alternatively a potentiometer (RV1) may be added as indicated in Fig 11.67. The use of an attenuator is preferred because it reduces the apparent output VSWR of the generator by increasing the return loss. If a control is used it must be returned to its minimum insertion loss position when starting a test or no signal may be heard.

Some contemporary receivers and transceivers cannot be operated in the AM mode and consequently the noise source seems unusable. The detection of noise is the process by which the noise source operates; therefore it will not work through an FM detector, nor will it work through a product detector since one of the terms of the detection (the noise) is not coherent.

The oscilloscope jack on many receivers is loosely coupled to the IF amplifier preceding the detector. A wide-band oscilloscope connected to this point will show the train of pulses and eliminate the need for aural detection. The alignment of the later IF stages of a system should have the least impact on the noise performance and maximum signal response will always occur at the same setting. Therefore the simple detector of Fig 11.68 will generally work for aural AM detection. Connect point A to the last IF amplifier anode, collector or drain. Connect point B to the audio amplifier at or near the volume control and ground point C. With this arrangement the normal output detector is turned down with the volume control and the temporary detector provide AM detection.

The gated noise source has been used for literally hundreds of applications and has proved to be a powerful yet simple addition to the test bench. While no guarantee of duplication

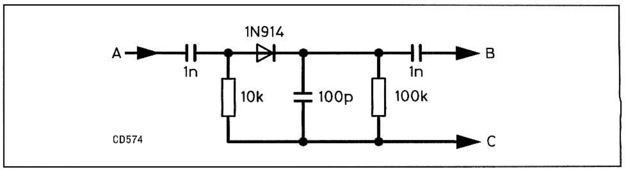


Fig 11.68. A simple detector that can be used when aligning SSB and FM receivers *(ARRL Radio Amateur's Handbook)*

may be made, these units develop approximately 18dB of excess noise in the region 50–300MHz. This unit was originally described by Hartsen in *QST* January 1977.

MEASUREMENTS ON MODERN VHF/UHF FRONT-ENDS

This section is taken from an article by G3SEK in *Radio Communication* [14] and expands on the principles of the previous sections. Power gains and losses are the basic currency of front-end system design. It is important to be able to measure them reasonably accurately.

The gain or loss of a device is the power level at its output relative to the signal level fed into the input – strictly speaking it is the insertion gains and losses that are being measured. To measure gain and loss a signal source is required together with an instrument for measuring relative power and calibrated in decibels. Although the absolute power level in watts need not be known, measurements must be made at the right sort of power levels. To make valid measurements on receiver front ends, the power levels must be well below the gain compression point [15], ie at levels of tens to hundreds of microwatts. At this low level accurate power measurements require a special instrument – see later.

RF amplifier gain

Set up the measuring system as shown in Fig 11.69 but with a coaxial adaptor instead of the amplifier. The two 20dB attenuator pads are used to establish 50Ω source and load impedances for the amplifier under test. Adjust the meter so that it reads 0dB, then remove the adapter and insert the amplifier. The meter reading should increase and the amount which it rises is the amplifier gain in decibels. If a long length of 50Ω coaxial cable had been inserted instead the meter would have fallen below the 0dB setting – this would then represent the cable loss in decibels. As well as simple losses and gains, this technique can be used for practically any other measurement that involves relative RF power levels, from HF to microwaves. For instance one can measure filter responses, crosstalk in diplexers and coaxial relays, antenna gains and radiation patterns and the VSWR of anything that has got a VSWR!

Stage-by-stage measurements of gains and losses is an extremely powerful technique for checking out a newly-built RF system such as a receiver front-end. If the system has been designed correctly [14] one knows already what to expect. This is one of the big rewards for spending time on the design before picking up a soldering iron: if there are problems they will hopefully be spotted. It's always useful to design a system with 50Ω interconnections so that each stage can be tested separately. This does not mean that each stage has to be built in separate boxes with plug and socket connections. Even a single-board layout can be designed to include interstage

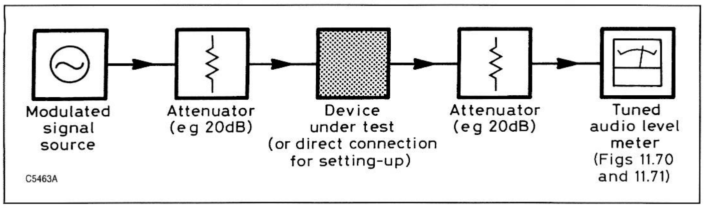


Fig 11.69. Typical set-up for gain or loss measurement

connections of 50Ω stripline with in-line coupling capacitors. By removing a capacitor, a coaxial test lead can be connected to the output of one stage or the input of the next. Thus, an entire system can be stepped through stage by stage, checking that all the gains and losses agree with design values.

Attenuators

One way to calibrate relative power measurements is by comparison with fixed resistive attenuators. These should be of good design for use in a 50Ω system (see 'Attenuator' section earlier). They should include a switched set giving any combination of 1, 2, 3, 4, 10 and 10dB) to allow 1 to 30dB to be selected in 1dB steps), plus fixed attenuators such as 3, 6, 10 and 20dB. It is also useful to own or have access to a few attenuators of known high quality to check the others against.

Almost all the measurements will be made on 50Ω systems, feeding the device from a 50Ω source impedance and terminating it in a 50Ω load. This can be ensured by placing the device under test between two 50Ω attenuators of 20dB or so. Any power reflected from a mismatch at the far side of an attenuator will itself be attenuated on the way back, hence the VSWR seen by the device under test will not vary very much from 1:1. For example, the VSWR looking into a 20dB attenuator theoretically cannot exceed 1.02, no matter what is connected at the other end. (In practice the VSWR would probably be rather higher owing to errors in the attenuator itself.) If all the gain and loss measurements are made with impedance-stabilising attenuators at the input and output of the device under test, the results should be well defined. If the practice is not followed then confusing and misleading results may ensue. For example, even the so-called standard attenuators will not perform as designed unless they themselves are in a 50Ω system.

Tuned AF level meters

As mentioned earlier, gain and loss measurements on receiver front-ends need to be made at power levels of tens to hundreds of microwatts. Conventional power meters using diode detectors (as mentioned elsewhere in this chapter) and DC voltmeters either are not sensitive enough or tend to suffer from noise and drift. A better solution is to use a test signal which is amplitude modulated by a steady audio tone and to measure the relative level of the tone instead of the carrier. The signal is detected by a diode in the usual way, but instead of trying to dredge the DC component out of the hum and noise, the audio tone signal is amplified. AC-coupled amplifiers neatly side-step the DC drift problem, and hum and noise are reduced by a sharp audio filter tuned to the tone frequency. The level of the tone is finally measured on a meter calibrated directly in decibels, those being the units of relative power level.

Test gear for the modulated signal technique is very easy to build. The only special instrument is the meter which measures the level of the audio modulating tone. Fig 11.70 shows a partial circuit diagram [16]. It is simply a tuned AF amplifier with adjustable gain, followed by an active rectifier which drives a meter. Gain is adjustable by both a switched attenuator and a continuously variable control. The switched attenuation is in 10dB steps; the total of 60dB is split into two separate 30dB attenuators to maintain a good signal/noise ratio without overdriving any stage. The amplifiers should be a modern low-noise type such as ZN459CP, OP-27G, NE5534N, TL071/2 or similar. Construction should follow standard hi-fi preamplifier practice. The first 20dB (×10 voltage) gain block is a low-noise audio preamplifier and requires careful shielding and grounding around the input. The two 40dB gain blocks (×100 voltage) are much less critical. Any frequency around 1kHz will do for the two tuned circuits as long as they are both tuned to the same frequency. The two inductors L1 and L2 should be about 80–100mH and are resonated by capacitors C4 and C6 which should each be about 220nF. Capacitors C3 and C5 are chosen to match the low output impedances of the gain blocks to the high-impedance tuned circuits; the values shown should be satisfactory, though some adjustment maybe necessary. Transformer T1 should be a typical interstage type with a step-up ratio of about 1:5. The active rectifier will typically be an op-amp with diodes in the feedback circuit in order to minimise the effect of the diode forward voltage drop. The resistors should be metal film types.

Fig 11.71 shows an alternative instrument [17] which uses 88mH inductors for selectivity. (Note: 88mH inductors used

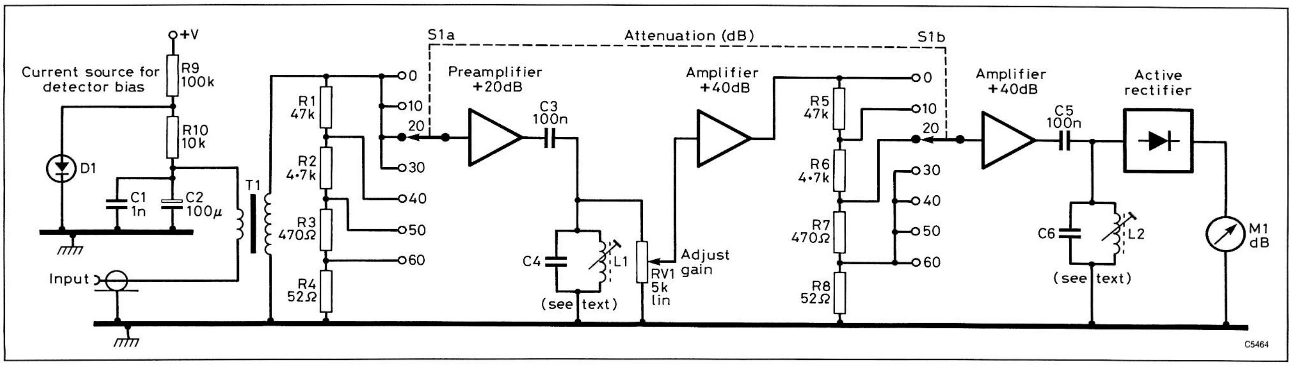


Fig 11.70. Partial circuit of a tuned audio level meter for modulated-signal measurements

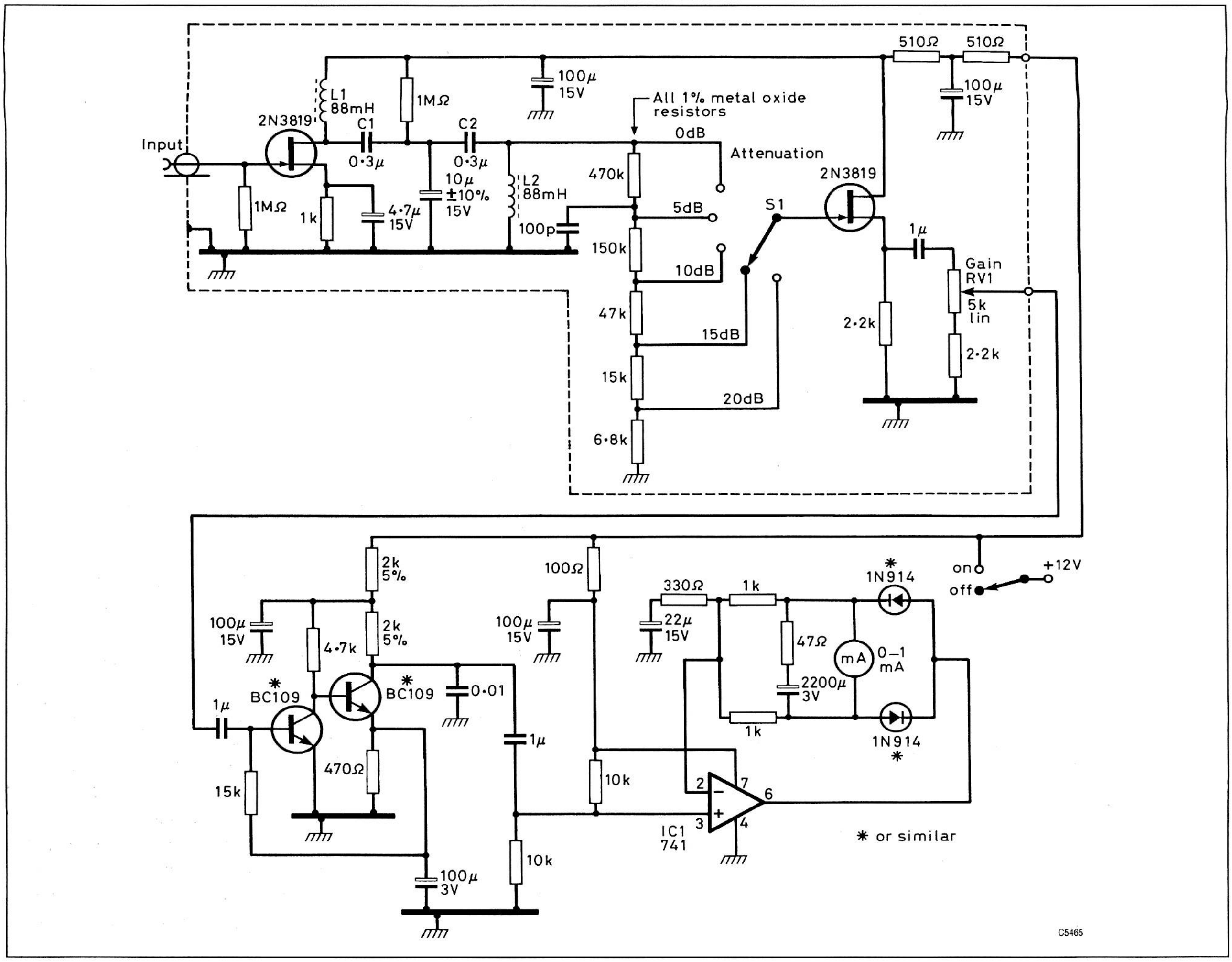


Fig 11.71. A simplified tuned audio level meter

to be readily available but it may be necessary to substitute the more easily available 82mH type or alternatively to wind them).

To get the best from either instrument, a large and accurate analogue meter is required with some existing form of linear calibration. The instrument should preferably be battery powered in order to avoid direct pick-up of the modulating signal via ground loops instead of via the device under test.

Detectors and VSWR bridges

A variety of diode detectors can be used for the modulated-signal technique. The most convenient are ordinary Schottky diodes fed with a few microamps of DC forward bias provided by the level meter of Fig 11.71. At or below the milliwatt RF level, these diodes are in their square-law region, which means that the rectified audio level is accurately proportional to the RF power *(not the voltage)*. If the meter scale already has an accurate linear calibration, a decibel scale can be added using nothing more than a calculator. Detectors usually need to provide a 50Ω RF termination, and this can be made with a Schottky diode in a BNC tee adapter – see Fig 11.72. The RF impedance is not exactly 50Ω but this can be preceded by an attenuator to 'flatten' the VSWR.

Since the modulated-signal technique basically measures power ratios, it can also measure VSWRs. Although VSWR is defined as a ratio of impedances, it is more often measured as a ratio of the 'forward' and 'reverse' power levels, as detected by directional sensors. An ordinary VSWR meter contains these directional couplers but typically requires at least

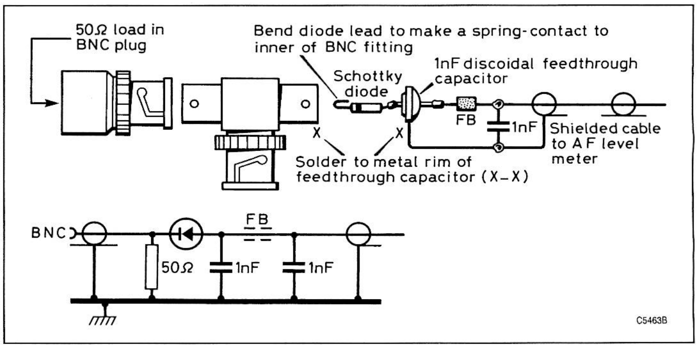


Fig 11.72. Exploded view of a detector and 50Ω load using a BNC T-fitting

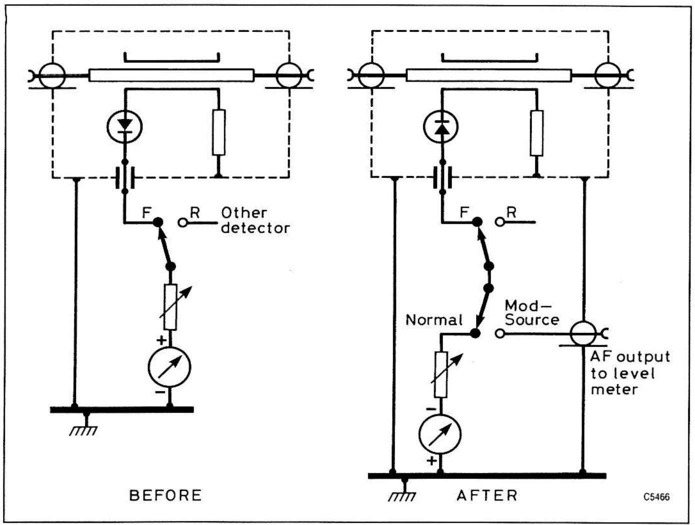


Fig 11.73. Adapting the ordinary VSWR bridge for modulated-signal measurements. The two detector diodes in the bridge are reversed, as are the meter connections

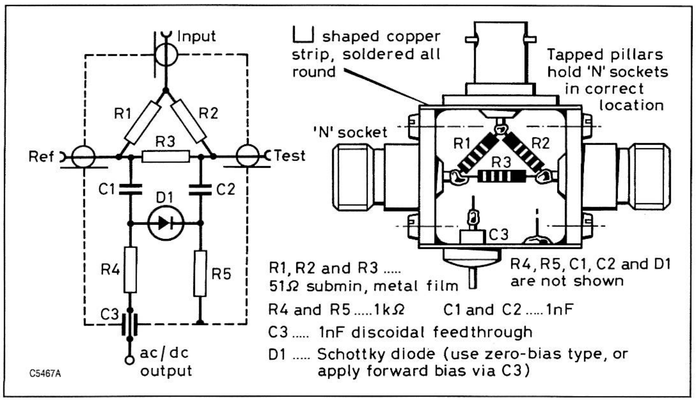


Fig 11.74. Circuit and sketch of a return loss bridge

1W of RF to give a good indication on the DC meter display. Simply reversing the diode detectors allows you to use the modulated-signal technique to measure the VSWR of delicate circuits at very low power levels – Fig 11.73. For example, it is possible to measure the input or output VSWRs of preamplifiers, mixers and other allegedly 50Ω devices (prepare for nasty surprises!).

Other kinds of VSWR sensors can also be used. Fig 11.74 shows a home-made VHF/UHF return loss bridge. Return loss is an alternative way of expressing VSWR and can be read directly in decibels from the meter scale [18]. W7ZOI has recently described many uses for an HF/VHF test set based on an HF-type return loss bridge and unmodulated RF sources [19]. By using the modulated-signal technique and the bridge of Fig 11.74, the same measurements can be extended to UHF. For microwaves you can use a slotted coaxial line or waveguide to observe and measure standing waves more directly [20].

Modulated signal sources

A wide variety of modulated signal sources can be used with the same detectors and level meter. For example, on 432MHz a crystal-controlled source, modified from a converter local oscillator strip, which provides a few milliwatts of RF can be used. An RSGB Microwave Committee board [21] would serve equally well if throttled back to the same power level. Since any signal source will always need a 50Ω output attenuator to establish the correct source impedance, a suitable pad can be permanently built-in. Modulation can be very simple; almost any waveform will do, so long as the frequency is adjustable to match the tuned circuits in the level meter. Typically a 432MHz source can be modulated by an AF oscillator using a 555 timer IC, which chops the RF output by supplying square-wave forward bias to the base of the final transistor – a bit brutal but effective! Whatever kind of modulated source is used, it should be very well screened, with supply leads decoupled to avoid stray pick-up into the detector. The RF output should also be spectrally pure, because the device under test may have significant – and misleading – responses to spurious frequencies from the signal source.

Modulated-signal sources can be made for all the VHF/UHF amateur bands of interest following the above guidelines. The modulated-signal technique can also be used outside the amateur bands, and is equally useful at HF. For example, a suitable general-coverage AM signal generator could be used with its internal modulation adjusted to the peak audio response of the meter. This would allow measurements of gains and losses in the IF stages of a front-end, and to manually sweep the frequency response of filters. If sweeping over a wide frequency range or measuring the gain or loss of a frequency-translating device (eg a mixer, transverter or complete front-end) you are also relying on the detector having a flat frequency response. It would be wise to check the detector first, using a good signal generator.

Accuracy

The modulated-signal technique can measure gains and losses with excellent accuracy. From the 0dB reference at full scale on the meter, the first 1dB of loss is spread over 21% of the meter scale, so changes of less than 0.1dB can be resolved. Changes of 10 or 20dB are taken care of by the range switch, while larger changes require external RF attenuators. Since calibration of the instrument relies only on resistor values and the linearity of the 0–10 meter scale, you can use it to cross-check the calibration of attenuators. Take care to make all measurements at power levels within the square-law region of the detector diode. This can be checked by increasing the applied power by a known amount and observing that the meter reading increases by exactly the same amount. All the calibrations – the meter scale, the 10dB range switches and all fixed and variable RF attenuators – are capable of being consistent within a small fraction of a decibel at frequencies up to at least 432MHz.

Noise measurements

Measurements related to noise are another stock-in-trade of front-end performance testing. In the test methods described the test signal itself is a known amount of broad-band RF noise, and you are interested in measuring the relative difference in noise levels when the test signal is applied. Noise

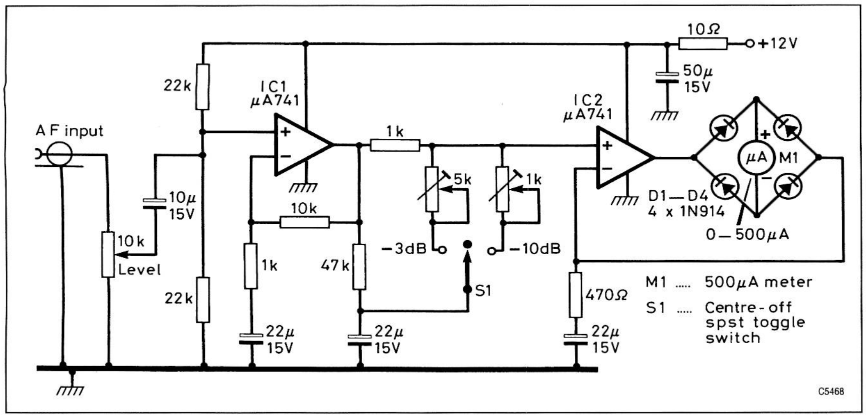


Fig 11.75. A peak-responding audio voltmeter for use in measuring (signal + noise)/noise ratio. For further details see [13]

levels are most conveniently measured at the audio output of the receiver. When this is done, you are relying on the audio output being strictly proportional to the noise or signal level, and hence the receiver's AGC must be switched off for all the tests to be described. Care must also be taken to avoid limiting or distortion in any RF or AF stage. Always repeat the measurements, especially where random noise is involved. Repeat the measurements under the same conditions to get some idea of the random errors. Then alter the RF and AF gain settings and repeat the measurements. If the results change, find out why.

A suitable instrument for relative AF noise power measurements is a rectifying meter as shown in Fig 11.75, which can be pre-calibrated in decibels by calculation [13]. This is essentially a peak-detecting instrument, so be careful that noise peaks are not being significantly clipped by overloading of any stage in the receiver. More complex RMS-detecting audio noise level meters are less prone to this error, but with care the simple meter of Fig 11.75 is quite adequate. With such a meter, (signal + noise)/noise ratios can be measured accurately as required in any of the following techniques for sensitivity measurement.

Sensitivity measurements

VHF/UHF receivers are extremely sensitive, as mentioned in [15]. A good 144MHz front-end would have a noise figure of about 2dB and would be able to detect signals as weak as 20 nanovolts. At those levels, accurate sensitivity measurements using an attenuated signal generator are extremely difficult. The signal-generator technique is still satisfactory for HF receivers and relics from the older generation of 'deaf' VHF/UHF receivers; but accurate measurements on low-noise systems require an extremely well-screened signal generator and meticulous attention to both practical and theoretical details. For modern VHF/UHF front-ends, better techniques for sensitivity measurement are those that stay as close as possible to the fundamental concept of noise temperature [15]. Best of all measure the noise temperature itself.

Noise temperature measurement techniques are based on supplying a known amount of excess noise to the RF input of the receiver and measuring the change in noise output. If the receiver noise temperature to be measured is T_{RX} and the noise source has a noise temperature T_{ON} when switched on and T_{OFF} when switched off, the system noise temperatures are:

$$T_{SYS} = T_{RX} + T_{ON} \text{ – noise source on}$$
$$T_{SYS} = T_{RX} + T_{OFF} \text{ – noise source off}$$

The noise power from the receiver is proportional to the system noise temperature, so the on/off power ratio Y is given by:

$$Y = (T_{RX} + T_{ON})/(T_{RX} + T_{OFF})$$

If T_{ON} and T_{OFF} are known and Y is measured, T_{RX} can be calculated from:

$$T_{RX} = (T_{ON} - YT_{OFF})/(Y - 1)$$

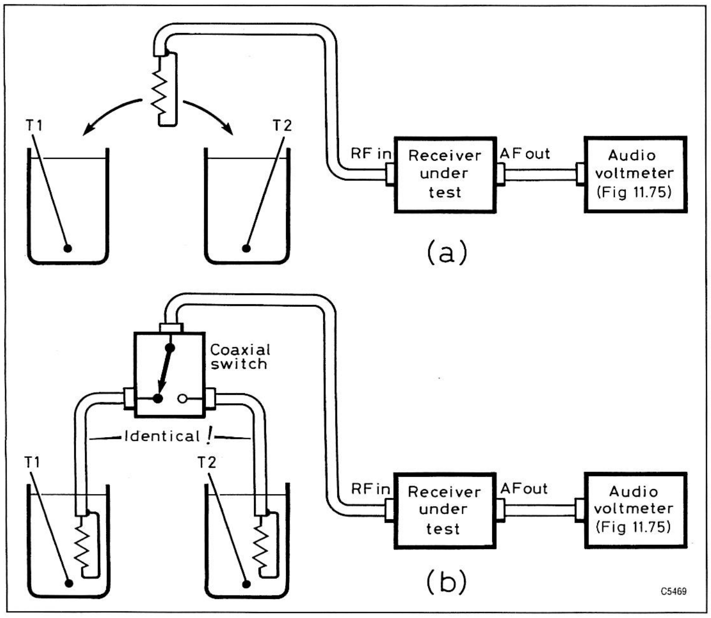


Fig 11.76. The hot/cold method for measuring noise temperature. The basic method (a) involves transferring one resistor between two temperature baths. The two-resistor method (b) is more convenient but requires even more care in setting up

The hot/cold method

The main problem is to generate RF noise at two different known noise temperatures T_{ON} and T_{OFF}. The simplest way is to use the thermal noise from a 50Ω resistor at two known physical temperatures – Fig 11.76(a). This is called the *hot/cold method* and is the closest you can get to a fundamental measurement of noise temperature. For the best accuracy you need a large difference between the two temperatures. Resistors do not like being 'roasted', so the hot end of the range cannot go too high. The cold end can be extended down to 77K if liquid nitrogen can be obtained from your friendly local laboratory (in rural areas ask the AI man!), or to about 196K in crushed solid CO_2.

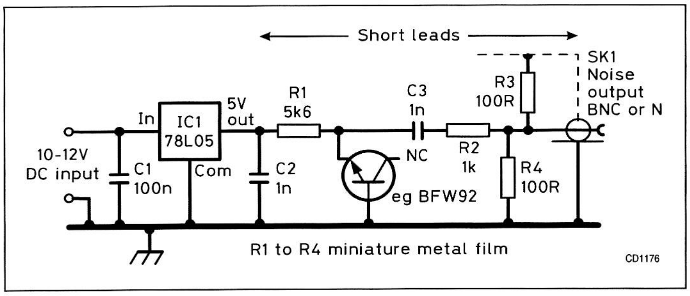


Fig 11.77. A diode noise generator with low VSWR in on and off states. The circuit from C2 to the output socket SK1 is built with the shortest possible leads on a copper groundplane (PCB) soldered directly to the back of SK1. R1 may be altered to maximise noise output

Note: the use of liquid nitrogen and solid CO_2 can be dangerous and unless it is known how to handle these potentially hazardous materials properly, it is safer to use a shorter temperature baseline, eg between melting ice and hot water. With care and a good laboratory thermometer, it is still possible to obtain quite accurate results.

Even the fundamental and basically simple hot/cold method has pitfalls for the unwary. To present an accurate 50Ω termination at both temperatures, the resistor must have good RF characteristics and a low temperature coefficient. Corrections need to be made for losses and thermal noise from connecting cables. There may be errors owing to drift in system gain while the resistor is being warmed up or cooled down, and measurements can become very tedious and slow. It is far quicker to compare noise powers from two 50Ω resistors at different temperatures – Fig 11.76(b), but then the two resistors and all their associated connections need to be identical in all respects (except temperature) so the test system itself requires careful preliminary checking before the results can be trusted.

Non-thermal noise sources

Another way of generating noise is to use a non-thermal source such as the noise generated in certain kinds of diode. The traditional method was to use a thermionic diode or a silicon microwave detector diode in forward conduction.

Nowadays a better and cheaper method for amateurs is to use a diode in reverse breakdown, eg a zener or the base-emitter junction of a silicon transistor. With careful control of the diode current, the long-term noise characteristics can be quite repeatable.

It is absolutely vital to make sure that the noise generator represents an accurate 50Ω source, and the impedance must not change when the generator is switched on or off – many published noise generator circuits ignore this precaution. The noise temperatures of many modern preamplifiers (especially GaAsFET types) are extremely sensitive to source impedance, and measurements can be totally falsified by changes in the on/off impedance of the noise source.

A suitable diode noise generator is shown in Fig 11.77. The diode is a transistor base-emitter junction operated in reverse breakdown, the resistor in series with the regulated 5V supply having been selected to maximise the noise output. The on/off impedance change is stabilised by over 30dB of RF attenuation; the RF attenuator arrangement in Fig 11.77 was optimised at 432MHz with the aid of a home-made return loss bridge (Fig 11.74) and its performance was subsequently verified using commercial equipment.

Regardless of whether the DC supply is on or off, the noise generator presents a good 50Ω termination up to 432MHz, and is still passable at 1.3GHz. Although its noise temperature is not calibrated, the generator can be used in comparative measurements and for checking system performance. It can also be used with the G4COM noise figure comparator [22] (see p11.41) which is highly recommended for aligning front-ends to give the best possible noise figure.

The problem with all semiconductor noise generators is that they cannot be used for absolute measurements until their noise temperatures have been calibrated against some other standard, preferably a hot resistor. Even so, their convenience and repeatability has made semiconductor noise sources the norm in commercial noise generators for VHF, UHF and at least the lower microwave region.

To summarise, relative measurements of noise temperature with uncalibrated noise sources are quite straightforward, given a certain amount of care and basic understanding. But there are no simple methods for absolute measurements. The hot/cold method is probably the most promising for use at home, though even this needs a great deal of care.

Strong-signal (dynamic range) measurements

As explained in [15], strong off-frequency signals can have several possible effects on receivers. The following tests evaluate the three effects that are of most concern to amateur VHF/UHF operators because of their potential to interfere with weak signals. The three effects are:

- *Third-order intermodulation* – mixing between at least two strong in-band signals, to give products which are also in-band and can interfere with the wanted signal.
- *Reciprocal mixing* – raising of the receiver's apparent noise level, due to the noise sidebands of the receiver's own local oscillator.
- *Gain compression* – when a strong off-frequency signal causes one of the stages in the receiver to limit, suppressing all other signals including the wanted one.

Different levels of strong signals are needed to provoke each of these effects. These levels can each be expressed in decibels above the receiver's noise floor, giving the dynamic range for the overload effect in question [15]. Remember, there is no single definition of dynamic range: a receiver has a separate dynamic range for each overload effect, and thus needs a separate test. Even slight overload could interfere severely with a barely copyable weak signal. When testing receivers meant for amateur DX working, the so-called 'spurious-free dynamic ranges' are measured, which relate to the situation where the particular overload effect just begins to be noticeable.

Strong-signal sources

Strong-signal tests require signal sources capable of delivering CW carriers at known power levels of several milliwatts, with low levels of spurious signals including noise sidebands. For gain compression measurements on modern front-ends

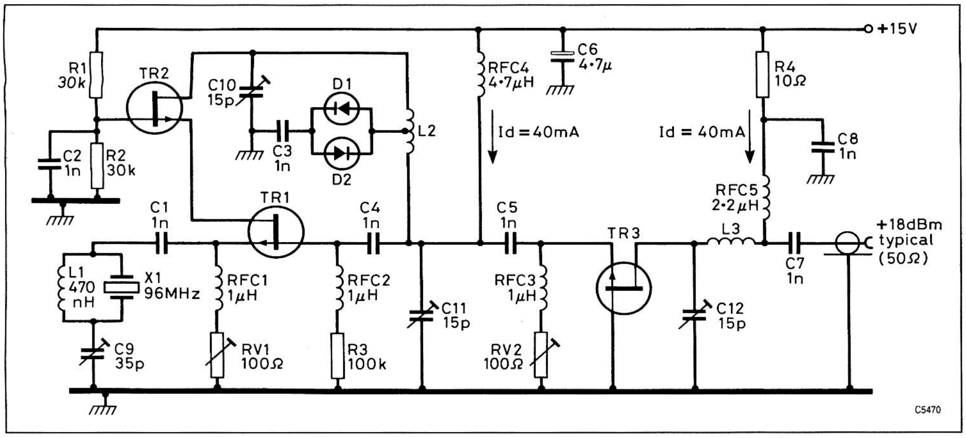


Fig 11.78. A low-noise crystal oscillator for 96MHz. The circuit is adaptable to 72MHz (×2 = 144MHz) or 108MHz (×4 = 432MHz). For details see [23]. D1, D2: Schottky diodes; TR1–3: P8000 power FETs (or U310 at I_d = 20mA); X1: 96MHz; L1: to resonate at 96MHz with X1; L2: 6t, 1mm dia wire on 6mm dia former, centre-tapped; L3: as L2 but no tap

the level of the single-tone test signal needs to be variable up to about 0dBm (1mW). A power level of about –30dBm is required to test for reciprocal mixing – and many present-day commercial VHF/UHF receivers are in deep trouble with test signals of much lower power!

Home-built signal sources for VHF and above need to be crystal controlled, and designs for low-noise crystal oscillators are available. Fig 11.78 is a typical example of a low-noise oscillator [23] for the frequencies of interest. The oscillator is based on a FET cascode amplifier with positive feedback through the pi-network C10, L2 and C11. The overtone crystal bypasses the source of TR1 to ground at its series-resonant frequency, producing sufficient loop gain to permit crystal-controlled oscillation. The active devices in most oscillator circuits not only provide gain but also limit the amplitude of the oscillation, so the devices have to operate in a non-linear mode. This can lead to increased levels of noise sidebands. In the circuit of Fig 11.78 the amplifying and limiting functions are separated; D1 and D2 provide the limiting, allowing TR1 and TR2 to run in linear Class A for lower noise. Other low-noise features include the use of power FETs at high bias currents, and the generally high signal levels maintained around the oscillator loop. Note the RF chokes RFC1 and RFC3 in series with RV1 and RV2, the source bias resistors of TR1 and TR3. Without these chokes, RV1 and RV2 would become sources of thermal noise because their resistances are quite well matched to the circuit impedances prevailing at these points.

A low-noise oscillator deserves low-noise amplifiers and frequency multipliers. Amplifier TR3 is a power FET running at high current in Class A. Higher-level amplifiers can also use power FETs or bipolar VHF/UHF power transistors, which should be operated in Class A and considerably below their normal RF output ratings. Strong signal sources for 50MHz and 70MHz can use overtone oscillators directly on those frequencies, while sources for the higher bands require frequency multipliers. Multipliers have to be non-linear or else they would not work, but the operating conditions of ordinary transistor multipliers are very vague indeed, making it hard to design for low-noise performance as well as reasonable efficiency. Avoid high-order multiplication in a single stage; frequency double if you can, treble if you must, but nothing higher. Thus a 144MHz source requires an oscillator on 72MHz followed by a doubler (overall, this will probably be easier, cheaper and better than going for an oscillator on 144MHz direct), and 432MHz requires an 108MHz oscillator and two doublers. Following this logic, a 1.296GHz source requires an 81MHz oscillator and four doublers but it may be worthwhile trying a solution with one trebler in it.

A frequency-doubling technique worth exploring is the balanced diode circuit of Fig 11.79. If this circuit does look similar to a power supply, that is because it is a full-wave rectifier. (Remember how a 50Hz mains input produces a 100Hz ripple on the smoothed but unregulated DC?). In Fig 11.79 an input RF signal of frequency *f* is applied through a suitable centre-tapped transformer T1. In this application one is after the ripple at 2*f* and so smoothing is minimal. The output therefore is mostly at 2*f* because the balanced arrangement tends to cancel out the neighbouring frequencies, *f* and 3*f*. A single tuned circuit is enough to give a very pure output at 2*f*. The diodes should be fast Schottky switching types for VHF/UHF, and for low noise and good doubling efficiency they need to be driven hard so they cannot dither in partial conduction. Diode multipliers are passive, so they need a Class A amplifier at the output frequency to make up the loss in the diode circuit. Comparing a diode multiplier chain with a conventional chain using transistor multipliers, about the same number of transistors are required but also the diodes and their transformers. But diode multipliers do not require double-tuned circuits and they give a far better chance of obtaining low noise and good spectral purity without special setting up.

Intermodulation measurements

Intermodulation requires at least two strong signals, so a test for intermodulation performance requires two separate signals fed into the receiver input. For accurate results, the only source of intermodulation between these two signals must be the actual receiver under test. This can be tricky! Somehow the two signals must be brought together without letting them

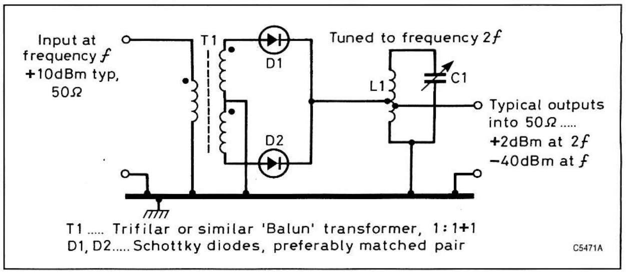


Fig 11.79. Balanced-diode frequency multiplier for VHF/UHF [13]

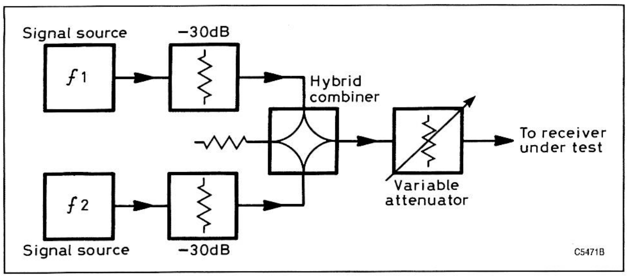


Fig 11.80. Intermodulation test set-up. Note the precautions to isolate the two signal sources from each other

intermodulate. Intermodulation could occur in any non-linear components outside the receiver – most likely in the output amplifiers of the signal generators themselves, where signal levels are the highest. So the two signal generators need to be isolated from each other by in-line attenuators plus a special hybrid combiner to add the two signals while keeping the sources isolated. Another attenuator after the combiner varies the level of the two-tone test signal. The basic set-up is shown in Fig 11.80 and more details are given in the references [24, 25].

For measurements on modern VHF/UHF front-ends, the level of the two-tone test signal needs to be variable up to about −40dBm, with incidental intermodulation products suppressed by more than 60dB. (It is conventional to quote the levels of intermodulation products against the power level of *one* of the two equal test signals; rather than the sum). Since the output of each signal generator has to pass through a high-value attenuator plus another variable attenuator after the combiner, each generator must provide an initial output of at least +10dBm. This means making *two* high-level low-noise signal sources following the guidelines outline earlier. The direct RF signal paths are not the only source of intermodulation between the two generators. One RF source can modulate the other via leakage along the *outside* of coaxial cables, or by RF getting on to the DC supply lines. Each signal source must be very carefully screened and all DC supplies must be fanatically filtered and shielded; typical precautions include boxes within boxes, each with soldered-up lids, double filtering of DC supplies through screened compartments and permanently soldered connections of solid-wall coaxial cable for the high-level signals [24, 25].

Having assembled the whole test set-up, try to get it tested on a spectrum analyser to check for unwanted intermodulation and to get an absolute calibration of the RF levels. You *can* calibrate it, but why not let it be done on professional equipment?

Conclusion

With the home-built test equipment described in this section, you can make all the necessary tests and measurements on modern VHF/UHF receiver front-ends. Simple relative measurements may be all that is needed to optimise front-end performance and to check that it stays that way. Beyond that, absolute measurements *can* be made with all the accuracy needed in amateur radio. It will take longer than the person using professional gear but with care and understanding good results can be obtained.

What really matters is getting the best possible on-the-air performance from your receiver. Any measurement, no matter how simple, is better than guesswork and wishful thinking!

RECEIVER ALIGNMENT AID

By far the most common method of aligning amateur low-noise receivers relies upon listening to a weak signal from a distant station, such as a beacon, and adjusting the matching components of the receiver input stage for maximum signal-to-noise ratio [7].

Signals from distant sources are notoriously unreliable, varying rapidly in strength over a range of many decibels. This makes it necessary to repeatedly check the strength of the beacon to ensure that an improvement in signal-to-noise ratio has been achieved.

A locally generated signal which can be adjusted in level down to barely detectable would appear to be ideal, since it would not suffer from the vagaries of propagation. In practice it can be very difficult to attenuate the test signal to the required level because of the amount of screening needed.

A second, and not often considered, problem with this approach is matching between the source and the receiver. A well-attenuated signal generator output will provide a good 50Ω match, whereas the antenna may not provide the same degree of matching. The result can be less than optimum.

A better approach to aligning low-noise receivers is to use a noise generator in place of the signal generator [22]. With this technique, broad-band noise is injected into the receiver input. The noise source is turned on and off and the receiver matching adjusted until the ratio of *noise on* to *noise off* at the receiver output is at a maximum. It can be very difficult to judge aurally when the ratio is maximum, so that some form of visual indicator becomes desirable.

Such an instrument is known as *an automatic noise figure meter* when it indicates directly the true noise figure of the item under test. It is, however, necessary to use a source with an accurately known noise output in order to make an accurate measurement. If the noise output of the source is not accurately known the instrument can still be used to adjust the receiver for best signal-to-noise performance, although the actual noise figure will not be known.

The instrument described in the following sections can be used to adjust receivers operating at any frequency for optimum sensitivity. It provides a continuous readout of the difference between the audio output of a receiver with no RF input and the output when a wide-band noise generator is connected to the receiver's antenna socket. The meter indicates the ratio between the outputs under these two conditions.

By design the meter reading is not affected by changes in audio level over a wide range of volume settings. The circuit has a logarithmic response so that the meter scale can be linearly calibrated in signal-to-noise ratio in decibels. Unless the absolute level of noise output from the noise source is known, the scale cannot be marked in noise figure.

The unit uses a reverse-biased diode as a noise source.

Circuit description

The circuit diagram is shown in Fig 11.81. Audio input from the loudspeaker socket of the receiver under test is connected to a small speaker (LS1) at the instrument input. This speaker provides a means of monitoring the receiver output which would otherwise be inaudible due to the muting action of most

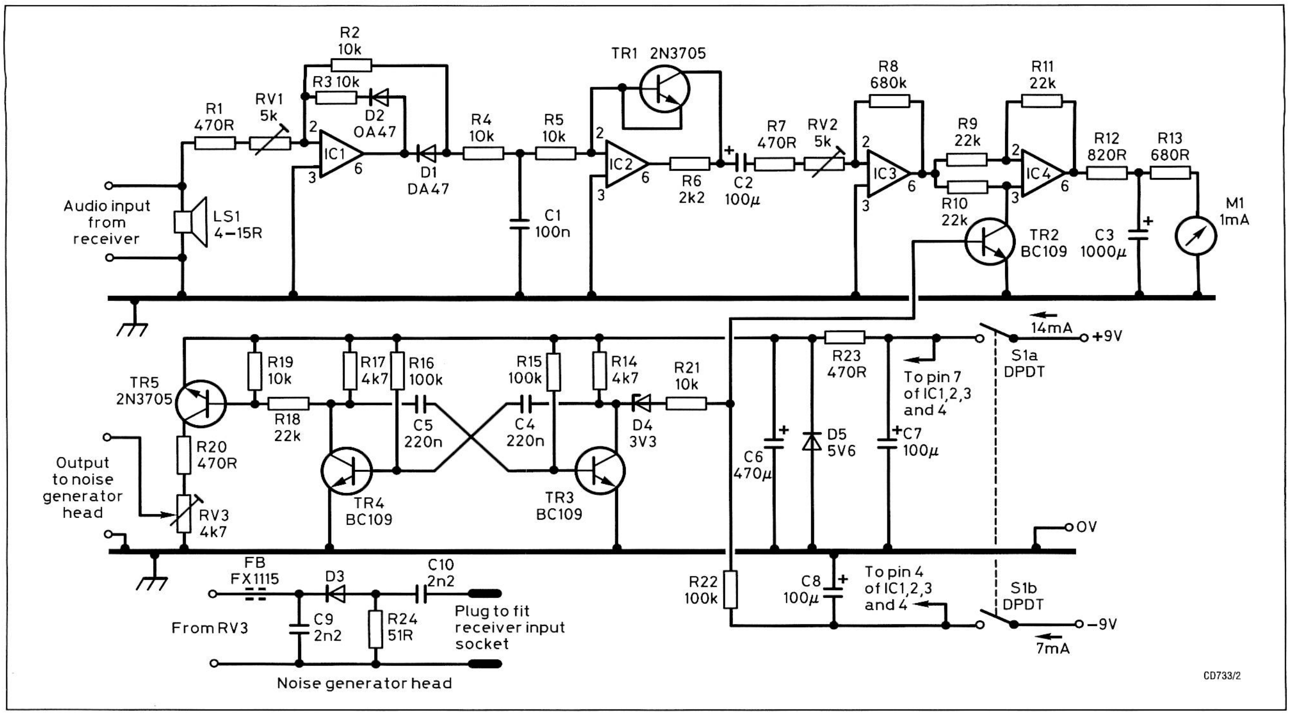


Fig 11.81. Circuit diagram of the alignment aid

loudspeaker external connection sockets. The AC across the speaker is rectified by the precision rectifier formed around IC1. This arrangement effectively overcomes the forward voltage diode drop of a rectifier and thus very-low-level AC signals can be accommodated. The voltage gain of this circuit is given by:

$$\frac{R_2}{R_1 + RV_1}$$

while D2 and R3 prevent the operational amplifier saturating on the negative half-cycles of the input. R4/C1 act as a low-pass filter to the input of IC2.

IC2 is formed into a logarithmic amplifier by the use of TR1 in the feedback loop. *Note:* the voltage across the base/emitter junction of a transistor with its base connected to its collector is proportional to the logarithm of the current through the transistor.

Because the receiver is fed with two signals then IC1 and hence IC2 are also fed with them. The difference (in the millivolt range) between the output voltages under these two conditions is a function of the ratio between the two input voltages and this ratio is independent of the average input level. Provided that the various stages of the receiver and the circuit around IC1 are working within their linear range, the AC output from the circuit formed around IC2 at the pulse frequency used will be dependent only on the overall signal-to-noise ratio. As the output from IC2 is only a fraction of a volt peak to peak it is amplified by the following stage formed around IC3 which has a voltage gain given by:

$$\frac{R_8}{R_7 + RV_2}$$

The output of IC3 is fed to a unity gain phase-sensitive detector (PSD) based on IC4. The reference signal is fed via TR2 from the pulse generator (or multivibrator) TR3 and TR4. A PSD is ideally suited to applications such as this, where an indication is required of the magnitude of an AC signal which has a known frequency and phase but a high accompanying noise level. In this application the PSD gives a usable output when the signal is accompanied by so much noise that it is undetectable by ear.

IC4, which has a relatively low output impedance, can drive a 1mA meter. Full-scale deflection of the meter in the prototype was set at approximately 10dB signal to noise with the scale reading linearly in decibels. R12 and R13 limit the current through the meter with C3 providing smoothing of the detected signal, otherwise the meter would show an erratic response due to the nature of the noise inputs.

The pulse generator is formed from a conventional astable multivibrator (TR3 and TR4) operating at about 30Hz, the output of this being fed to amplifier TR5 which is used to pulse the noise source.

The noise generator uses a reverse-biased diode mounted in a separate enclosure with matching and decoupling components. An ideal arrangement would be to mount this within a coaxial plug. The diode D3 used in the prototype was a CV364 microwave mixer, but alternatives are 1N21, 1N23, 1N25 and 1N32. A possible, but not tried, alternative is a BAT31 silicon avalanche device which is intended as a noise source from 10Hz to 18GHz.

Construction

Construction of the receiver alignment aid is not critical and audio techniques can be used with the exception of the noise head which must be built using VHF techniques if it is to operate reliably at the highest frequencies. The circuit requires

Table 11.15. Components list for the receiver alignment aid

R1, 7	470R
R2–5, 19, 21	10k
R6	2k2
R8	680k
R9–11, 18	22k
R12	820R
R13	680R
R14, 17	4k7
R15, 16, 22	100k
R20, 23	470R
R24	51R or 75R
RV1, 2	5k skel preset, 0.1W
RV3	4k7 lin carbon pot
C1	100n polystyrene
C2	100μ, 6V3 tantalum
C3	1000μ, 10V electrolytic
C4, 5	220n polystyrene
C6	470μ, 16V electrolytic
C7, 8	100μ, 16V electrolytic
C9, 10	2n2 ceramic disc
D1, 2	OA47, OA79, OA90, BAT85
D3	See text
D4	3V3, 400mW zener
D5	5V6, 400mW zener
TR1, 5	2N3705, 2N3703, 2N4126
TR2–4	BC109, 2N2926
IC1–4	741, 8-pin
FB	FX1115 or equivalent
LS1	4–15Ω min speaker
S1	DPDT switch
M1	1mA FSD meter

Resistors are 0.25W/0.5W, 5% unless specified otherwise.

a symmetrical ±9V DC supply at about 20mA. A PCB and component layout are given in Appendix 1.

Alignment

The unit requires little alignment and no test equipment is needed. Plug the noise head into a receiver and gradually increase the diode current until an audible 'purring' sound is heard in the receiver loudspeaker. Connect the audio output of the receiver to the input of the unit. The 'purring' should now transfer to the unit's loudspeaker and the meter should show a fairly steady reading which can be varied by adjusting the noise diode current (using RV3). Set RV1 so that the meter reading is constant over a wide range of receiver volume settings. Set RV2 to give a full-scale deflection of the meter at maximum diode current on the highest frequency band of interest. The unit is now ready for use.

Operation

Connect the unit and noise head to the receiver under test and adjust RV3 for about half-scale deflection on the meter. Any adjustment to the receiver that results in an improved signal gain with no change in the noise figure, or a reduced noise figure with no change in signal gain, or both simultaneously, will result in an increased meter reading. By noting the reading of the meter before and after any circuit adjustments, improvements in performance can readily be seen.

Although the unit is not especially sensitive to small temperature changes, it is best to switch the unit on at least 10 minutes before use and to ensure that the ambient temperature is reasonably constant.

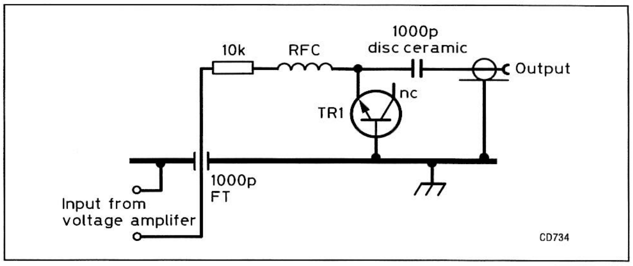


Fig 11.82. HF-to-UHF noise head. TR1: 2N2369, BFY90 etc – see text; RFC: 3t of the 10kΩ resistor lead, 2mm inside dia

Additional notes on use

Use of the receiver alignment aid assumes reasonable linearity of the receiver, therefore care must be taken when aligning FM receivers to ensure that the receiver does not limit with the noise source on. With most receivers this will mean that the level of noise injected must be as small as possible, consistent with still exceeding the FM threshold. With AM/SSB receivers the noise blanker and AGC must be disabled if meaningful results are to be obtained.

Care must be taken if the alignment aid is to be used for initial alignment of a converter or receiver. Noise output from the unit is constant over a wide range and it is therefore possible to inadvertently align on a spurious or image frequency, especially if the receiver has a low intermediate frequency. A signal generator or similar should therefore be used for initial alignment to avoid the problem.

Some receivers have been encountered that have a small DC voltage appearing at the loudspeaker socket. When connected to the alignment aid this voltage can bias IC1 beyond its linear range, thus resulting in false readings on the meter. Connecting an electrolytic capacitor of about 47μF in series with the input overcomes this problem. The negative terminal of this capacitor should be connected to the junction of R1 and the monitor loudspeaker.

Additional improvements

Considerable development work has been carried out to the receiver alignment aid since it was first published and this has resulted in several very useful improvements [26].

The original noise head was designed primarily for VHF operation. An alternative design that can be used throughout the HF range and up to at least 1.3GHz is shown in Fig 11.82. Useful output may still be available at 2.3GHz when a suitable transistor is used for TR1. It is best to select a transistor with a high f_T for TR1. It may be necessary to try several transistors before one with enough output is found.

Better phase detector performance is achieved at low levels with an FET (eg 2N3819) in place of the bipolar transistor TR2. Sometimes difficulties have been encountered with the meter reading not being independent of audio drive level. This can be just a matter of incorrect use or it can arise when the comparator is used in conjunction with receivers possessing an odd audio frequency response. This can be cured by replacing the components R4, C1 and R5 by the circuit shown in Fig 11.83.

Fluctuating meter readings can also be a problem at times. Changing the meter to one of 50μA FSD and altering the time constant of the meter circuit can noticeably improve matters.

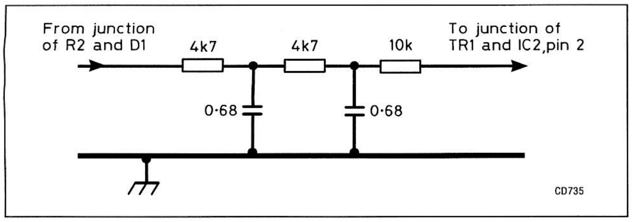


Fig 11.83. An improved interstage coupling network between IC1 and IC2

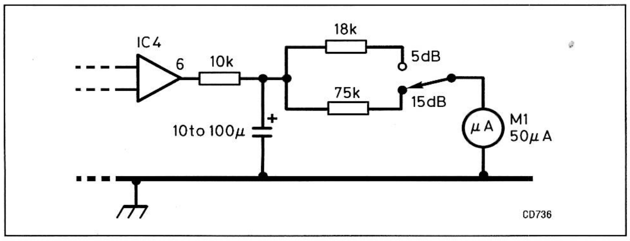


Fig 11.84. Modified meter circuitry

Fig 11.84 incorporates these modifications as well as including a switch to give different full scale readings of signal-to-noise ratio.

STATION MONITOR FOR FM TRANSMISSIONS

The unit described here [27] is designed to enable an amateur with some constructional experience to build a simple FM monitor, capable of monitoring FM transmissions from his or her station from HF to UHF. The use of a sampling mixer allows a very wide frequency coverage with a simple low frequency oscillator.

Principle of operation

The sampling mixer essentially consists of a switch which is opened and closed by short pulses derived from the local oscillator signal (Fig 11.85). When the switch is closed, the input signal is fed to the hold capacitor so that, when it opens again, the capacitor holds a charge proportional to the value of the input signal during the sampling interval. Operation is shown in somewhat idealised form in Fig 11.86. The low-frequency (LF) signal which appears on the hold capacitor is filtered and forms the intermediate frequency (IF).

There are several assumptions implicit in Fig 11.86.

1. The sampling pulse is short compared with the period of the input signal.
2. The switch resistance is small so as to allow the input signal to charge C to the correct value during the short period

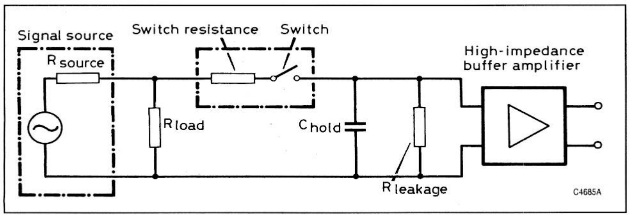


Fig 11.85. Simple model of a sampling gate

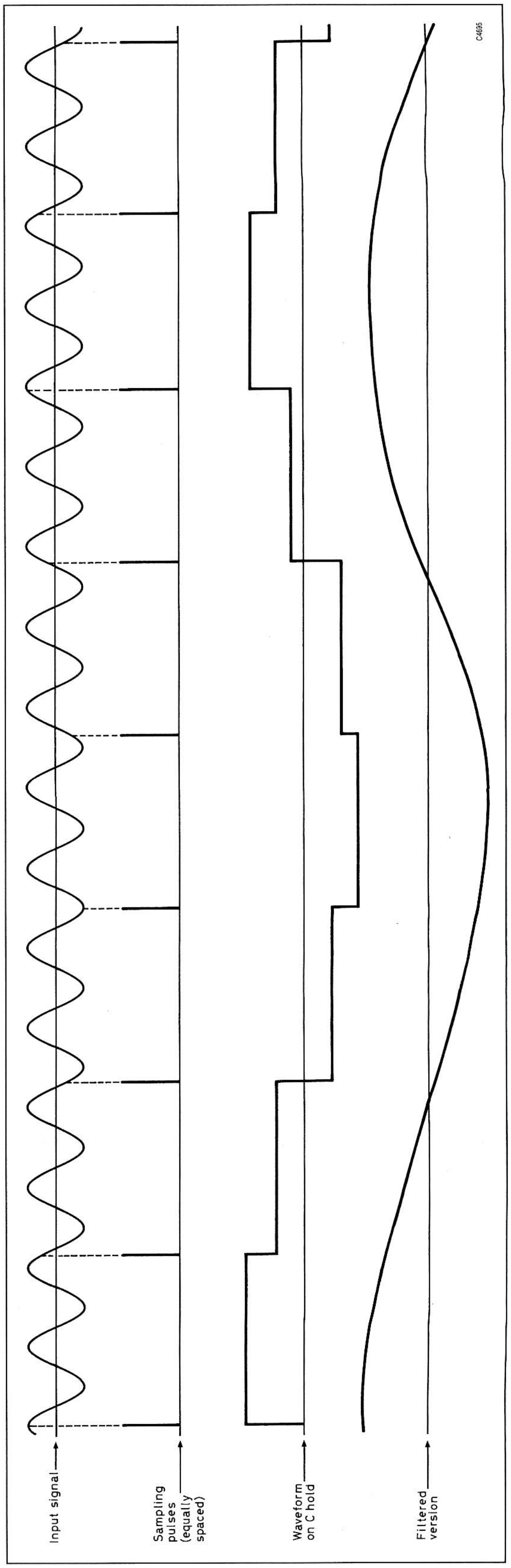


Fig 11.86. Idealised sampling action

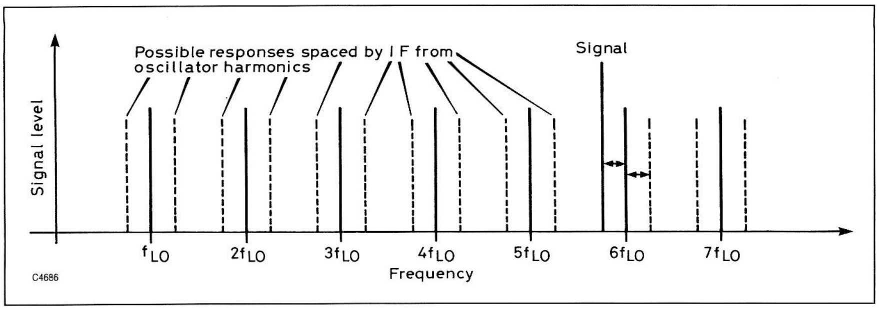


Fig 11.87. Spectrum of sampling gate showing multiple responses

when the switch is closed. Finite resistance of the switch means that the hold capacitor is not charged to the peak of the input signal and manifests itself as loss in the mixer which increases as the intermediate frequency is raised.

3. The leakage resistance must be sufficiently large that the capacitor voltage does not fall appreciably inbetween samples, causing mixer loss.

The sampling mixer can also be considered in the frequency domain (Fig 11.87). The local oscillator (LO) pulse, if infinitely short in duration, would have a spectrum extending out to infinity, a so-called harmonic 'comb', with a 'tooth' spacing equal to the LO frequency. Practical pulses of finite duration have a comb which is not flat but falls at high frequencies. The wider the pulses, the lower the roll-off frequency.

Also depicted in Fig 11.87 is the fact that each LO harmonic can produce an IF with any signal spaced an IF away on either side. The sampling mixer thus has a multitude of responses and is not therefore recommended as a receiver front-end. For transmitter monitoring, there will (it is hoped) be only a single frequency present and this is then of no concern. The sampling mixer described enables transmitter monitoring of frequencies beyond the 432MHz band.

Monitor block diagram

The block diagram is shown in Fig 11.88. The sampling mixer, consisting of a diode gate fed by an LO pulse generator, is followed by a high-impedance FET buffer amplifier. The IF signal is selected by a low-pass filter and then amplitude limited. The FM signal is rendered intelligible by a pulse-counting discriminator. This form of discriminator is used because it is wide band and removes the need for excessively tight tolerance on the LO frequency. For example, a LO operating in the region of 4MHz *will* produce an IF with a 432MHz signal on around its hundredth harmonic. A narrow-tuned discriminator would require very careful setting of the LO.

Circuit description

Economy was a major design consideration and the unit uses relatively cheap components throughout. The circuits are shown in Fig 11.89 and Fig 11.90. A diode gate (D1 to D4) acts as the switch, with C4 the 'hold' capacitor. The diodes should be either germanium or of the Schottky type. A FET amplifier (TR1) buffers the sampling gate. IC1, an ECL triple line receiver, forms a somewhat unorthodox pulse generator. The first two sections square up the oscillator signal, and the resulting square wave feeds the third section (with a delay to one input caused by the insertion of a short length of coaxial cable, eg 300mm). The square-wave edge at pin 7 causes the output at pin 14 to go *high*. After the very short cable delay, the delayed edge returns pin 14 to the *low* state, giving a very narrow pulse. An anti-phase signal is available at pin 15. The two outputs are very convenient for driving the sampling gate and the symmetrical drive to some extent balances out the LO and reduces breakthrough into the IF.

The LO is not shown. Frequencies from 4 to 100MHz have been used, with levels above about 20mV being suitable. Hence, the LO may be based on a large number of designs appropriate to the frequency chosen. A crystal source is recommended.

The IF at TR1 is low-pass filtered to remove any LO signal and applied to a limiter, IC2. The limited IF signal is then fed to a pulse-counting discriminator based on the design by G3JGO [28].

The value of C20 shown on the circuit allows operation with an IF up to 500kHz. The sampling gate, however, is capable of operating with an IF of more than 1MHz. If this extended range is required, the capacitor value should be reduced in proportion to the maximum IF; eg if 1MHz maximum IF is used, C20 should be reduced to 90pF.

The penalty paid for this is that the discriminator then produces a smaller output for a given deviation and more audio gain must be used to restore the level. IC5 provides audio gain with the gain set to 48. This is given by (R30 + R31)/R30 and should be adjusted if C20 is changed.

With component values as in the circuit, the audio output will be approximately 1V peak for 1kHz deviation. A signal with ±2.5kHz deviation will therefore give an audio output of 5V peak to peak. The op-amp (IC5) will clip outputs some way under 12V peak to peak which (fortunately) limits the amount of noise when the input signal is disconnected.

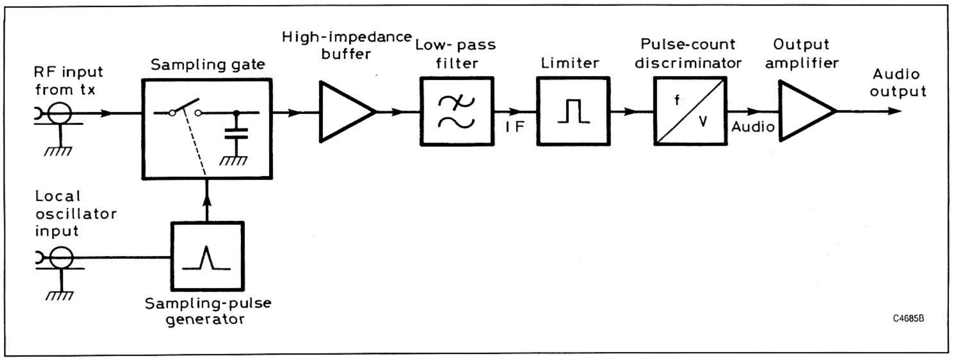


Fig 11.88. Block diagram

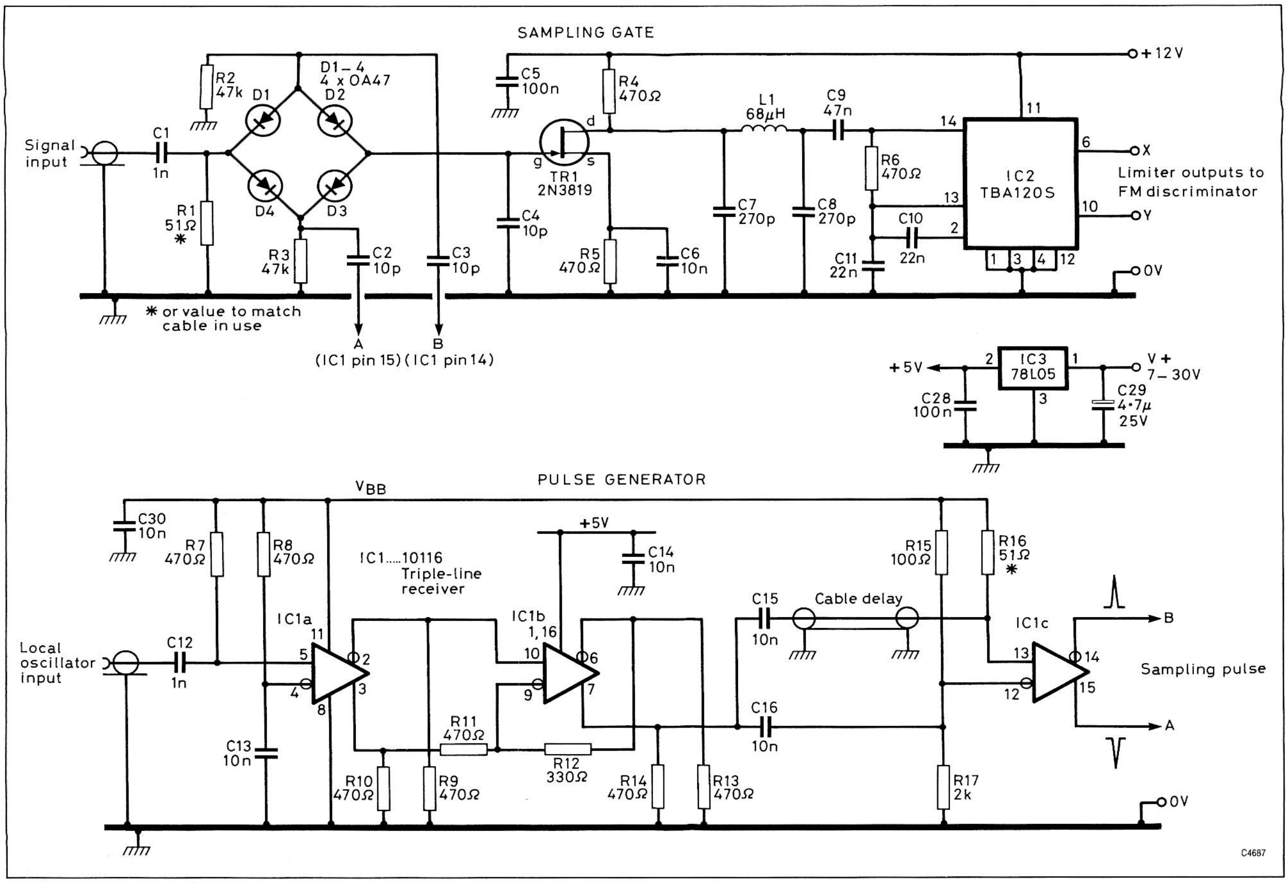


Fig 11.89. Circuits of the sampling gate and pulse generator

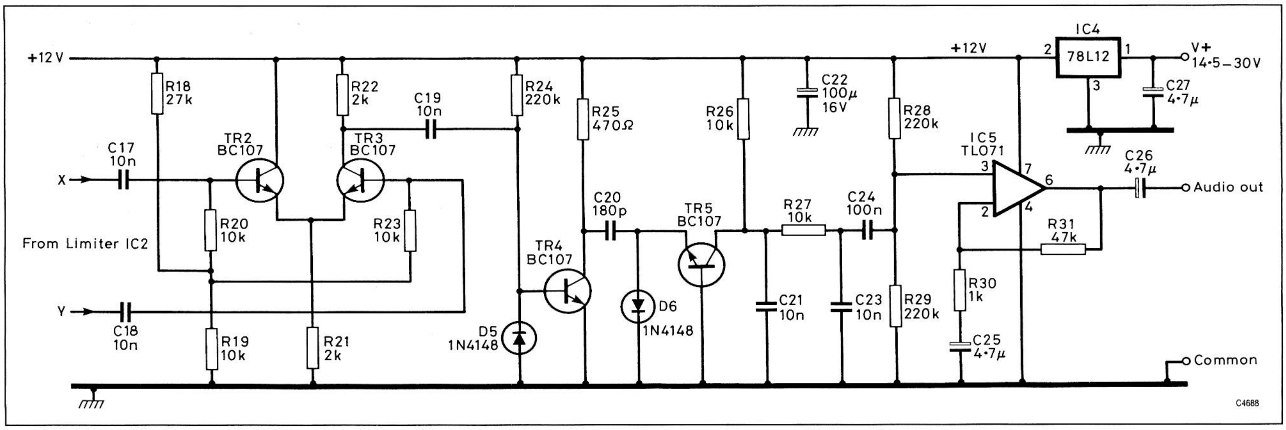


Fig 11.90. Pulse-count discriminator circuit

Construction

Several prototypes have been built. The quickest, surest method of construction of the high-frequency circuitry of the sampling gate and pulse generator is to build them on an earth plain of plain copper-clad glassfibre board, with components soldered together using minimum lead lengths, and decoupling capacitors and other earthed parts soldered directly to the board. IC1 should be mounted upside down and connections made direct to the pins – unfortunately this leads to an ugly-looking circuit. The braid at each end of the delay cable should be soldered direct to the copper (see Fig 11.91). The input and LO leads can be soldered similarly. If 75Ω cable is used,

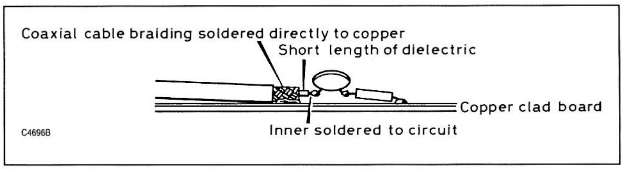


Fig 11.91. Soldering coaxial cable to earth plane

Table 11.16. Component list for sampling monitor

R1, 16	51R
R2, 3, 31	47k
R5–11, 13, 14, 25	470R
R12	330R
R4, 15	100R
R17, 21, 22	2k
R18	27k
R19, 20, 23, 26, 27	10k
R24, 28, 29	220k
R30	1k
C1, 12	1n ceramic disc
C2, 3, 4	10p ceramic
C5, 24, 28	100n ceramic disc
C7, 8	270p ceramic
C9	4n7 ceramic disc
C6, 13–21, 23, 30	10n ceramic disc
C10, 11	2n2 ceramic disc
C20	180p polystyrene
C22	100μ, 16V electrolytic
C25–27, 29	4μ7, 25V electrolytic
L1	68μ RF choke
D1–4	OA47 or similar – see text
D5, 6	1N4148
TR1	2N3819
TR2–5	BC107, 2N2369 or similar
IC1	10116
IC2	TBA120S
IC3	78L05
IC4	78L12
IC5	TL071

Resistors are 0.25/0.5W, ±5% unless stated otherwise.

R16 should strictly be increased to 75Ω or a close value to provide a match, but this is not critical. No advantage is gained by shortening the cable.

Providing good RF layout is used for the high-frequency circuits, performance of the unit should be satisfactory. Following the FET buffer stage, construction becomes non-critical and any of the large range of constructional techniques available may be used.

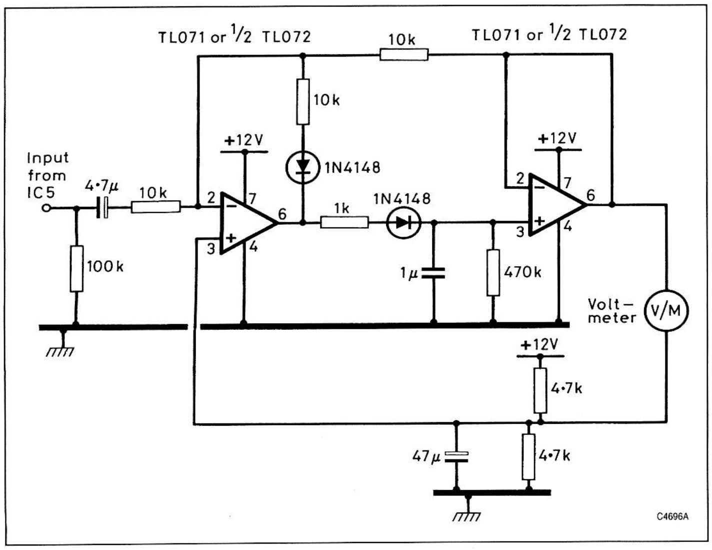


Fig 11.92. Suggested circuit for peak detector

Choosing the local oscillator frequency

Using a single LO frequency, the availability of a range of several hundred kilohertz of IF allows a similar range of frequencies within an amateur band to be used. For example, suppose that the discriminator capacitor is set for a maximum IF of 500kHz. An attempt should be made to find a local oscillator frequency which gives an IF from about 100 to 500kHz and thus a 400kHz band coverage. Of course, the image response will give another 400kHz on the other side of the local oscillator harmonic if this is useful. The relation between the various frequencies is:

$$\text{Input frequency} = N.f_{LO} \pm f_{IF}$$

where N is the harmonic number in use. Taking a realistic case:

$$f_{LO} = 12.0333\text{MHz (crystal)}$$
$$36f_{LO} = 433.1988\text{MHz}$$

Thus FM simplex channels 433.375 to 433.500MHz can be covered with an IF of 176.2 to 301.2kHz and discriminator values shown would be suitable. The 12th harmonic of the same oscillator frequency is 144.3996MHz. An IF up to 1MHz would then give coverage up to 145.400MHz. A 8.0222MHz input gives the same results with $N = 54$ and 18 respectively.

As the IF is lowered below 100kHz, more IF signal appears in the audio output and the application determines whether this is tolerable.

Operating the unit

Having chosen the LO frequency, apply this signal to the unit at a level exceeding 20mV EMF. If a fast oscilloscope is available, the pulses should just be seen at pins 14 and 15 of IC1. Arrange for a signal of 30mV to 300mV RMS (PD, −13 to +3dBm) from the transmitter. Excessive drive will cause conduction of the gate diodes and degrade performance. The limiter should now provide square waves from the IF and audio should be available at the output. The audio can be used in a number of ways:

1. The audio can simply be used for listening to the modulation.
2. An oscilloscope can be used to check on the peak deviation.
3. A peak detector can be used (suggested circuit in Fig 11.92) to give readings proportional to peak deviation. If the recommendations in the text are followed, the 1kHz/V of peak deviation will be maintained.

Conclusion

It is hoped that the unit described will be taken as the basis for further experimentation and used by amateurs to ensure that the quality of modulation and efficient spectrum use are kept to a high standard.

TWO-STUB TRANSFORMER

For impedance matching, a two-stub transformer (tuner) provides a satisfactory method. It is effectively a coaxial or VHF pi-coupler.

The general arrangement is shown in Fig 11.93 and

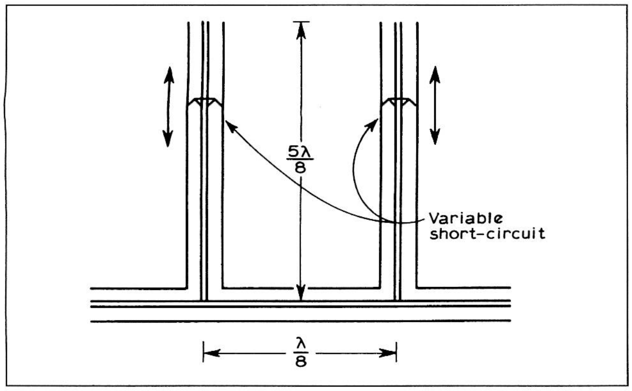


Fig 11.93. General arrangement of the two-stub transformer

this is made using 3.2mm diameter centre conductor and 12.7mm outer. The transformer will have a useful range of $Z_0/2$ to $2Z_0$.

If the section of line between the two stubs is made adjustable and the stubs themselves are long enough for the lowest frequency required, then a unit for use on more than one band can be made.

COAXIAL SWITCH

Although manually operated coaxial switches are available, these are generally fitted with SO239/PL259 UHF-type connectors which are a variable-impedance type. Many stations use N-, BNC- or TNC-type connectors which have superior performance and are of a constant impedance.

Of course, adapters may be added to a commercial unit with UHF-type connectors but this is both expensive and cumbersome. It is relatively simple to construct a unit in a small die-cast box using a two-way single wafer switch or a slider switch. In such a unit the preferred connector can be used to fit into the existing system.

The unit illustrated (Fig 11.94) is fitted with N-type connectors. The two slide connectors are positioned as close to the end of the box as possible and are soldered directly to the switch terminations. The centre connector, mounted on the box end, is connected by a short piece of copper strip (lower inductance than wire unless rod with 'turned down' ends is used).

An additional connector has been included at the other end of the box, to which a coupling loop is connected so that a DFM or other monitor may be attached without having to 'break in' to the main circuit. The coupling loop is tuned by a small piston capacitor.

A switch of this type is useful for switching two different antennas or from antenna to dummy load. The performance of the prototype, with all ports matched to 50Ω and the switch set to connect ports 1 and 2, is shown in Table 11.17.

Table 11.17. Performance of prototype coaxial switch

Frequency (MHz)	Insertion loss ports 1 to 2 (dB)	Isolation ports 2 to 3 (dB)	VSWR
70	<0.1	44	1.15
144	0.2	38	1.3
432	0.85	26	2.1

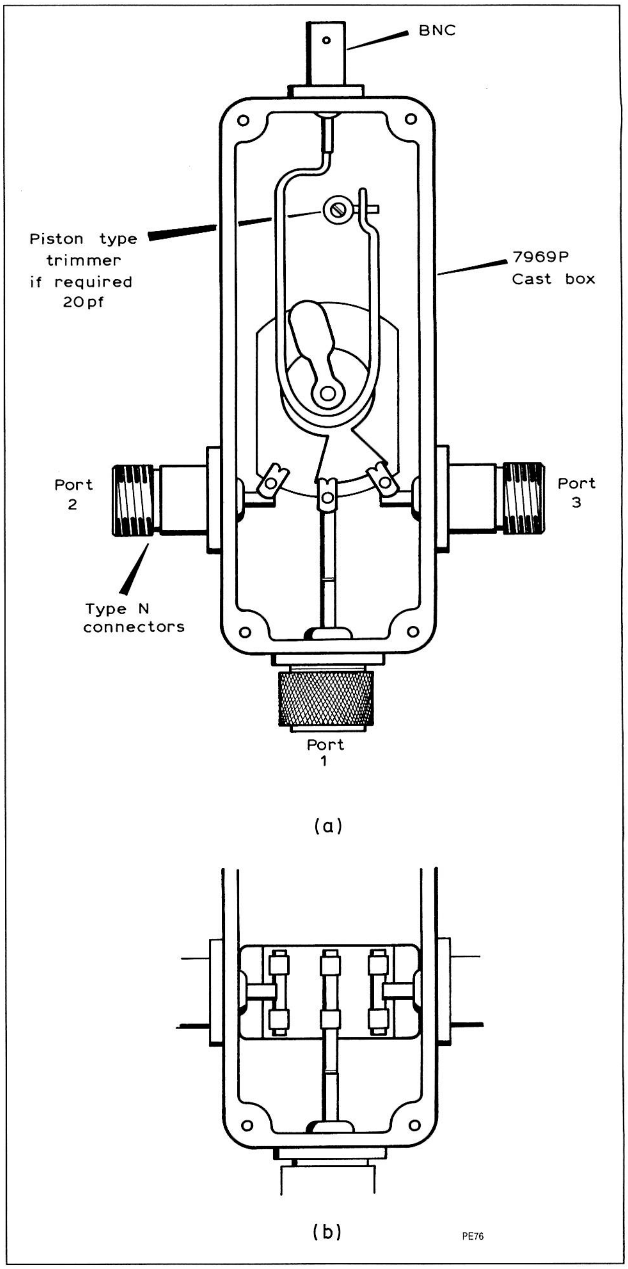


Fig 11.94. Coaxial switch using (a) rotary switch or alternatively (b) a slide switch

REFERENCES

[1] *Test Equipment for the Radio Amateur*, 3rd edn, Clive Smith, G4FZH, RSGB, 1995.

[2] *Amateur Radio Licence (A) or (B): Terms, Provisions and Limitations Booklet BR68*, Radiocommunications Agency.

[3] *Amateur Radio Software*, J Morris, GM4ANB, RSGB, 1985.

[4] 'Absorption Wavemeters for 144MHz', G R Jessop, G6JP, *Radio Communication* September 1982.

[5] 'A VHF dip oscillator', A L Mynett, *Radio Communication* September 1970.

[6] 'Simple Spectrum Analyser', R Blackwell, G4PMK, *Radio Communication* November 1989.

[7] *Microwave Handbook*, Vol 2, Ed M W Dixon, G3PFR, RSGB, 1991.

[8] 'A frequency counter for a 144MHz transmitter', N B Pritchard, *Radio Communication* May 1979.

[9] *Data sheet for ICM7216A/B/D*, Harris Semiconductors.

[10] *Data Sheet for 8680B*, Plessey Semiconductors.

[11] 'HF/UHF deviation meter', E Chicken, G3BIK, *Radio Communication* May 1989.

[12] *The ARRL Handbook for the Radio Amateur*, ARRL.

[13] *Solid State Design for the Radio Amateur*, Wes Hayward, W7ZOI, and Doug DeMaw, W1FB, ARRL.

[14] 'Measurements on modern VHF/UHF front ends', I F White, G3SEK, *Radio Communication* October and November 1986.

[15] 'Modern VHF/UHF front-end design, I F White, G3SEK, *Radio Communication* April-July 1985.

[16] From an article by W6SVS in a pre-1970 edition of *Ham Radio*.

[17] 'Antenna gain measurements', Fred Brown, W6HPH, *QST* December 1982.

[18] 'More about return loss bridges', Ian White, G3SEK, *The Lunar Letter* May 1983.

[19] 'Beyond the dipper', Wes Hayward, W7ZOI, *QST* May 1986.

[20] 'How to use an SWR indicator', Bob Stein, W6NBI, *Ham Radio* January 1977.

[21] *Microwave Newsletter Technical Collection,* RSGB.

[22] 'An alignment aid for VHF receivers', J R Compton, G4COM, *Radio Communication* June 1976.

[23] 'An extremely low-noise 96MHz crystal oscillator for UHF/SHF applications', B Neubig, DK1AG, *VHF Communications* Winter 1981.

[24] 'A two-tone test generator for the two-metre band', M Ploetz, DL7YC, *VHF/UHF Technik* (Berliner DUBUS Gruppe 1978). *UHF-Unterlage*, Teil 3, Ed K Weiner, DJ9HO, contains designs for 144MHz and 432MHz (in German).

[25] 'Intermodulation measurement techniques in mixers and amplifiers', P E Chadwick, Application Note AN1009, Plessey Semiconductors Ltd.

[26] 'Microwaves', C Suckling, G3WDG, *Radio Communication* October 1979 and March 1980.

[27] 'A sampling FM monitor', I Braithwaite, G4COL, *Radio Communication* October 1983.

[28] 'A pulse count discriminator unit', B Priestley, G3JGO, *Radio Communication* Sept 1971.

12 General data

Capacitance

The capacitance of a parallel-plate capacitor is:

$$C = \frac{0.224\,KA}{d} \text{ picofarads}$$

where K is the dielectric constant (air = 1.0), A is the area of dielectric (sq in), and d is the thickness of dielectric (in).

If A is expressed in centimetres squared and d in centimetres, then:

$$C = \frac{0.0885\,KA}{d} \text{ picofarads}$$

For multi-plate capacitors, multiply by the number of dielectric thicknesses.

The capacitance of a coaxial cylinder is:

$$C = \frac{0.242\,K}{\log_{10}(D/d)} \text{ picofarads per centimetre length}$$

where D is the inside diameter of the outer and d is the outside diameter of the inner.

Capacitors in series or parallel

The effective capacitance of a number of capacitors in *series* is:

$$C = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3} + \text{etc}}$$

The effective capacitance of a number of capacitors in *parallel* is:

$$C = C_1 + C_2 + C_3 + \text{etc}$$

Characteristic impedance

The characteristic impedance Z_0 of a feeder or transmission line depends on its cross-sectional dimensions.

(i) *Open-wire line*:

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

where D is the centre-to-centre spacing of wires (in) and d is the wire diameter (in).

(ii) *Coaxial line*:

$$Z_0 = \frac{138}{\sqrt{K}} \log_{10}\frac{d_o}{d_i} \text{ ohms}$$

where K is the dielectric constant of insulation between the conductors (eg 2.3 for polythene, 1.0 for air), d_o is the inside diameter of the outer conductor and d_i is the diameter of the inner conductor.

Decibel

The *decibel* is the unit commonly used for expressing the relationship between two power levels (or between two voltages or two currents). A decibel (dB) is one-tenth of a *bel* (B). The number of decibels N representing the ratio of two power levels P_1 and P_2 is 10 times the common logarithm of the power ratio, thus:

$$\text{The } \textit{ratio}\ N = 10 \log_{10}\frac{P_2}{P_1} \text{ decibels}$$

If it is required to express *voltage* (or *current*) ratios in this way, they must relate to identical impedance values, ie the two different voltages must appear across equal impedances (or the two different currents must flow through equal impedances). Under such conditions the *power* ratio is proportional to the square of the *voltage* (or the *current*) ratio, and hence:

$$N = 20 \log_{10}\frac{V_2}{V_1} \text{ decibels}$$

$$N = 20 \log_{10}\frac{I_2}{I_1} \text{ decibels}$$

Dynamic resistance

In a parallel-tuned circuit at resonance the dynamic resistance is:

$$R_D = \frac{L}{Cr} = Q\omega L = \frac{Q}{\omega C} \text{ ohms}$$

where L is the inductance (henrys), C is the capacitance (farads), r is the effective series resistance (ohms), Q is the Q-value of the coil and $\omega = 2\pi \times$ frequency (hertz).

Frequency – wavelength – velocity

The velocity of propagation of a wave is:

$$v = f\lambda \text{ centimetres per second}$$

where f is the frequency (hertz) and λ is the wavelength (centimetres).

For electromagnetic waves in free space the velocity of propagation v is approximately 3×10^8m/s and, if f is expressed in kilohertz and λ in metres:

$$f = \frac{300{,}000}{\lambda} \text{ kilohertz}$$

$$\lambda = \frac{300{,}000}{f} \text{ metres}$$

$$\text{Free space } \frac{\lambda}{2} = \frac{492}{\text{MHz}} \text{ feet}$$

$$\text{Free space } \frac{\lambda}{4} = \frac{246}{\text{MHz}} \text{ feet}$$

Note that the true value of v is 2.99776×10^8m/s.

Impedance

The impedance of a circuit comprising inductance, capacitance and resistance in series is:

$$Z = \sqrt{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2}$$

where R is the resistance (ohms), L the inductance (henrys), C the capacitance (farads) and $\omega = 2\pi \times$ frequency (hertz).

Inductors in series or parallel

The total effective value of a number of inductors connected in *series* (assuming no mutual coupling) is given by:

$$L = L_1 + L_2 + L_3 + \text{etc}$$

If they are connected in *parallel*, the total effective value is:

$$L = \frac{1}{\frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3} + \text{etc}}$$

When there is mutual coupling M, the total effective value of two inductors connected in series is:

$$L = L_1 + L_2 + 2M \text{ (windings aiding)}$$

$$\text{or } L = L_1 + L_2 - 2M \text{ (windings opposing)}$$

Ohm's Law

For a unidirectional current of constant magnitude flowing in a metallic conductor:

$$I = \frac{E}{R} \qquad E = IR \qquad R = \frac{E}{I}$$

where I is the current (amperes), E is the voltage (volts) and R is the resistance (ohms).

Power

In a DC circuit, the power developed is given by:

$$W = EI = \frac{E^2}{R} = I^2R \text{ watts}$$

where E is the voltage (volts), I is the current (amperes) and R is the resistance (ohms).

Q

The Q-value of an inductance is given by:

$$Q = \frac{\omega L}{R}$$

where L is the inductance (henrys), R is the effective resistance (ohms) and $\omega = 2\pi \times$ frequency (hertz).

Reactance

The reactance of an inductance and a capacitance respectively is given by:

$$X_L = \omega L \text{ ohms}$$

$$X_C = \frac{1}{\omega C} \text{ ohms}$$

where L is the inductance in henrys, C is the capacitance in farads and $\omega = 2\pi \times$ frequency (hertz).

The total reactance of an inductance and a capacitance in series is $X_L - X_C$.

Resistors in series or parallel

The effective value of several resistors connected in *series* is:

$$R = R_1 + R_2 + R_3 + \text{etc}$$

When several resistors are connected in *parallel* the effective total resistance is:

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \text{etc}}$$

Resonance

The resonant frequency of a tuned circuit is given by:

$$f = \frac{1}{2\pi\sqrt{LC}} \text{ hertz}$$

where L is the inductance (henrys) and C is the capacitance (farads).

If L is in microhenrys (μH) and C is picofarads (pF), this formula becomes:

$$f = \frac{10^3}{2\pi\sqrt{LC}} \text{ megahertz}$$

The basic formula can be rearranged thus:

$$L = \frac{1}{4\pi^2 f^2 C} \text{ henrys}$$

$$C = \frac{1}{4\pi^2 f^2 L} \text{ farads}$$

Since $2\pi f$ is commonly represented by ω, these expressions can be written as:

$$L = \frac{1}{\omega^2 C} \text{ henrys}$$

$$C = \frac{1}{\omega^2 L} \text{ farads}$$

See Fig 12.1.

Time constant

For a combination of inductance and resistance in series the time constant (ie the time required for the current to reach $1/\varepsilon$ or 63% of its final value) is given by:

$$t = \frac{L}{R} \text{ seconds}$$

where L is the inductance (henrys) and R the resistance (ohms).

For a combination of capacitance and resistance in series, the time constant (ie the time required for the voltage across the capacitance to reach $1/\varepsilon$ or 63% of its final value) is given by:

$$t = CR \text{ seconds}$$

where C is the capacitance (farads) and R is the resistance (ohms).

Transformer ratios

The ratio of a transformer refers to the ratio of the number of turns in one winding to the number of turns in the other winding. To avoid confusion it is always desirable to state in which sense the ratio is being expressed, eg the 'primary-to-secondary' ratio n_p/n_s. The turns ratio is related to the impedance ratio thus:

$$\frac{n_p}{n_s} = \sqrt{\frac{Z_p}{Z_s}}$$

where n_p is the number of primary turns, n_s is the number of secondary turns, Z_p the impedance of the primary circuit (ohms) and Z_s the impedance of the secondary circuit (ohms).

COIL WINDING

Most inductors for tuning in the HF bands are single-layer coils and they are designed as follows. Multilayer coils will not be dealt with here.

The inductance of a single-layer coil is given by:

$$L\,(\mu H) = \frac{D^2 \times T^2}{457.2 \times D + 1016 \times L}$$

where D is the diameter of the coil (millimetres), T the number of turns and L the length (millimetres). Alternatively:

$$L\,(\mu H) = \frac{R^2 \times T^2}{9 \times R + 10 \times L}$$

where R is the radius of the coil (inches), T is the number of turns and L is the length (inches).

Note that when a ferrite or iron dust core is used, the inductance will be increased by up to twice the value without the core. The choice of which to use depends on frequency. Generally, ferrite cores are used at the lower HF bands and iron dust cores at the higher. At VHF, the iron dust cores are usually coloured purple. Cores need to be moveable for tuning but fixed thereafter and this can be done with a variety of fixatives. A strip of flexible polyurethane foam will do.

Table 12.1. Wire data

Diameter	Approx SWG	Turns/cm	Turns/in
1.5	16–17	6.6	16.8
1.25	18	7.9	20.7
1.0	19	9.9	25
0.8	21	12.3	31
0.71	22	13.9	35
0.56	24	17.5	45
0.50	25	19.6	50
0.40	27	24.4	62
0.315	30	30.8	78
0.25	33	38.5	97
0.224	34–35	42.7	108
0.20	35–36	47.6	121

Note: SWG is Imperial standard wire gauge. The diameters listed are those which appear to be most popular; ie they are listed in distributor's catalogues. The 'turns/cm' and 'turns/in' are for enamelled wire.

Designing inductors with ferrite pot cores

This is a simple matter of taking the *factor* given by the makers and multiplying it by the square of the number of turns. For example, an RM6-S pot core in 3H1 grade ferrite has a 'factor' of 1900 nanohenrys for one turn. Therefore 100 turns will give an inductance of:

$$100^2 \times 1900\text{nH} = 10000 \times 1900\text{nH} = 19\text{mH}$$

There are a large number of different grades of ferrite; for example, the same pot as above is also available in grade 3E4 with a 'factor' of 3300. Manufacturers' literature should be consulted to find these 'factors'.

Table 12.2. Coaxial cables

Type	Nominal impedance (Ω)	Outside diameter (mm)	Velocity factor	Capacitance (pF/m)	Maximum RF voltage (kV)	Attenuation per 10m of cable 10MHz (dB)	100MHz (dB)	1000MHz (dB)
CT100	75	6.65	0.84	56	—	0.2	0.6	2.1
H100 (1)	50	—	—	—	—	—	0.44	1.35
LDF4-50A (2)	50	16.0	0.88	75.8	8.0	0.07	0.224	0.77
LDF5-50A (3)	50	28.0	0.89	75	8.0	0.037	0.121	0.43
RA519 (1)	50	10.3	0.80	84	5.0	0.10	0.35	1.25
RG58BU	50	4.95	0.66	100	3.5	0.5	1.7	5.6
RG58CU	50*	4.95	0.66	100	2.5	0.5	1.7	5.6
RG59BU	75	6.15	0.66	68	3.5	0.5	1.5	4.6
Min RG59	75	3.7	0.84	51	—	0.4	1.2	3.9
RG62AU	95	6.15	0.84	44	—	0.3	0.9	2.9
Min RG62	95	3.8	0.84	41	—	0.4	1.4	4.5
RG174AU	50*	2.8	0.66	101	2.1	0.3	0.9	2.9
RG178PE	50*	1.83	0.85	99	—	1.5	4.8	16
RG179PE	75*	2.54	0.85	69	—	1.2	4.0	13
RG213/URM67 (1)	50	10.3	0.66	100	6.5	0.2	0.7	2.7
RG402U	50	3.58†	—	—	—	—	—	—
RG405U	50	2.20†	—	—	—	—	—	—
UR43	50	5.0	0.66	100	2.6	0.4	1.3	4.5
UR67	50*	10.3	0.66	100	6.5	0.2	0.68	2.5
UR70	75*	5.8	0.66	67	1.8	0.5	1.5	5.2
UR76	50*	5.0	0.66	100	2.6	0.5	1.6	5.3
UR95	50	2.3	0.66	100	1.3	0.9	2.7	6.9
UR202	75*	5.1	0.84	56	—	0.4	1.1	4.2
UR203	75	7.25	0.84	56	—	0.2	0.8	2.7
WF103 (1)	50	10.3	0.85	78	5.0	0.09	0.32	1.30

* Indicates cable with flexible core. † Indicates cable with solid drawn outer, ie rigid.

(1) Minimum bending radius is 60mm. Obtainable in the UK from W H Westlake Electronics, West Park, Clawton, Holdsworthy, Devon, EH22 6QN.
(2) Minimum bend radius is 125mm. Obtainable in the UK from Andrew Ltd, Lochgelly, Fife, KY5 9HG. This needs special coaxial fittings.
(3) Minimum bend radius is 250mm. Obtainable from Andrew Ltd. This needs special coaxial fittings.

There are many further types of coaxial cable but these are the most popular ones; ie those listed in distributors' catalogues. The characteristics of others are listed in earlier editions of the *VHF/UHF Manual*.

Table 12.3. Colour coding for glass fuses

Colour	Rating (mA)	Colour	Rating (A)
Green/yellow	10	Green	0.75
Red/turquoise	15	Blue	1.0
Eau-de-Nil	25	Light blue	1.5
Salmon pink	50	Purple	2.0
Black	60	Yellow and purple	2.5
Grey	100	White	3.0
Red	150	Black and white	5.0
Brown	250	Orange	10.0
Yellow	500		

Note that this coding does not apply to the ceramic-bodied fuse commonly found in 13A plugs etc.

Table 12.4. Useful twist drill sizes

Screw size	2	3	4	5	6
Clearance drill	2.10	3.10	4.10	5.10	6.10
Tapping drill	1.55	2.65	3.50	4.50	5.20

All sizes are in millimetres.

Twist drill no	¼in 0.250 in	1 0.228 in	9 0.196 in	17 0.173 in	24 0.152 in	32 0.116 in	43 0.089 in	50 0.070 in
Clearance for woodscrew no	14	12	10	8	6	4	2	0
Clearance for BA	0	1	2	3	4	6	8	10
Tapping size for BA	—	—	0	1	2	4	6	8

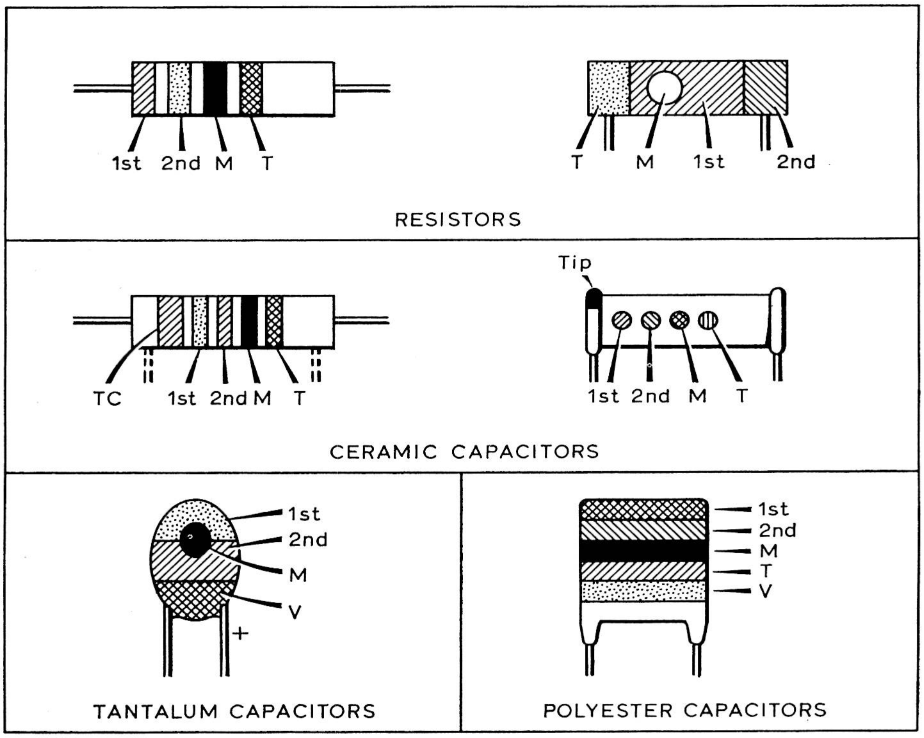


Table 12.5. Component colour codes

Colour	Significant figure (1st, 2nd)*	Decimal multiplier (M)	Tolerance (T) (per cent)	Temp coeff (TC) (parts/10^6/°C)	Voltage (V) (tantalum cap)	Voltage (V) (polyester cap)
Black	0	1	±20	0	10	—
Brown	1	10	±1	−30	—	100
Red	2	100	±2	−80	—	250
Orange	3	1000	±3	−150	—	—
Yellow	4	10,000	+100, −0	−220	6.3	400
Green	5	100,000	±5	−330	16	—
Blue	6	1,000,000	±6	−470	20	—
Violet	7	10,000,000	—	−750	—	—
Grey	8	100,000,000	—	+30	25	—
White	9	1,000,000,000	±10	+100 to −750	3	—
Gold	—	0.1	±5	—	—	—
Silver	—	0.01	±10	—	—	—
Pink†	—	—	—	—	35	—
No colour	—	—	±20	—	—	—

Units used are ohms for resistors, picofarads for ceramic and polyester capacitors, and microfarads for tantalum capacitors.

* Some close-tolerance resistors have *three* significant figures followed by a multiplier. These are usually metal-film types and have a tolerance of ±1% or less.

† A pink fourth ring on a resistor indicates 'high stability'.

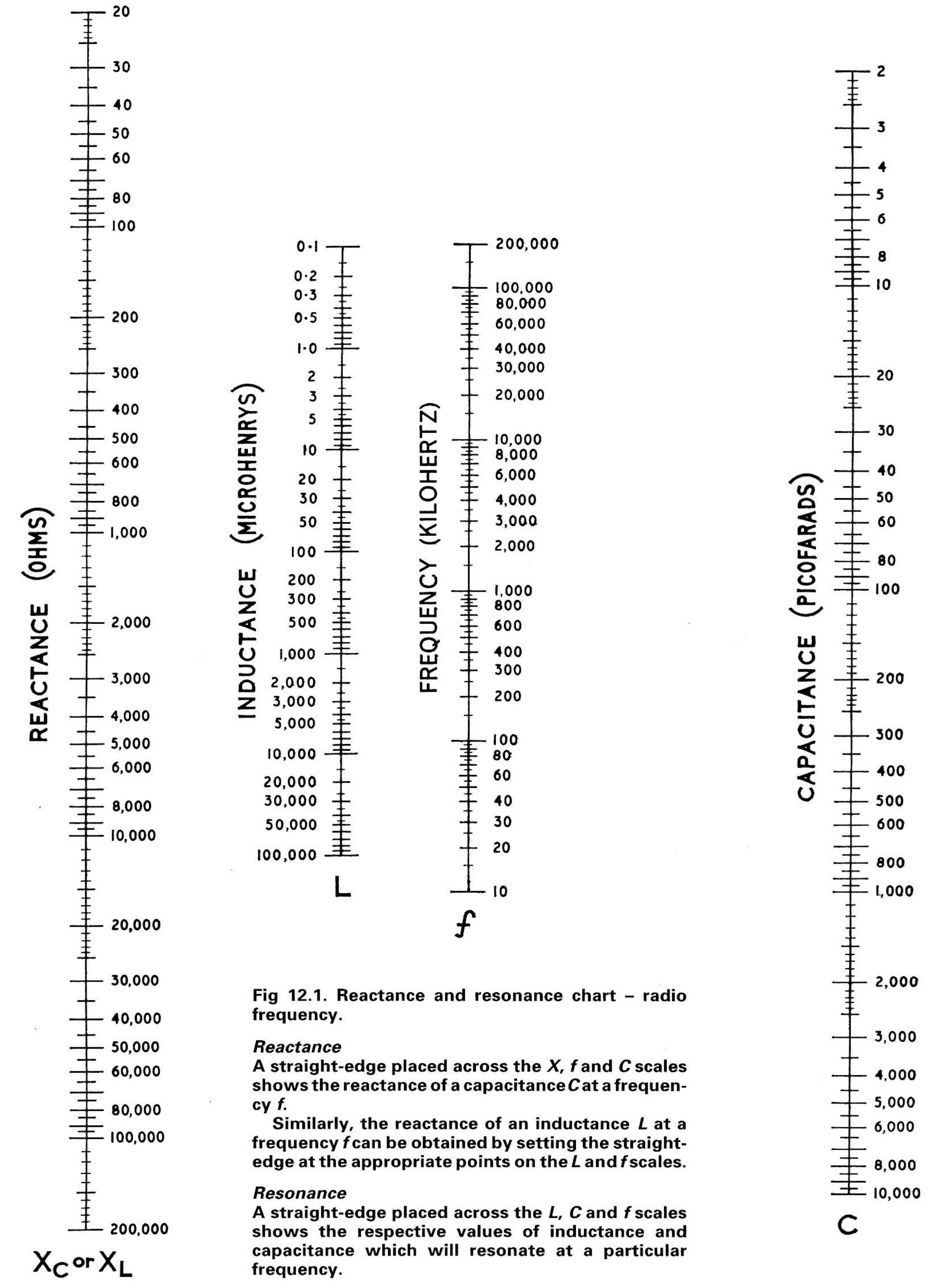


Fig 12.1. Reactance and resonance chart – radio frequency.

Reactance

A straight-edge placed across the *X*, *f* and *C* scales shows the reactance of a capacitance *C* at a frequency *f*.

Similarly, the reactance of an inductance *L* at a frequency *f* can be obtained by setting the straight-edge at the appropriate points on the *L* and *f* scales.

Resonance

A straight-edge placed across the *L*, *C* and *f* scales shows the respective values of inductance and capacitance which will resonate at a particular frequency.

1 PCB layouts

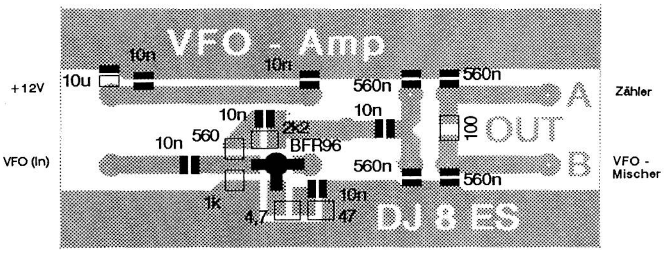


Fig 4.54. Component layout with the surface-mount parts on the track side of the PCB

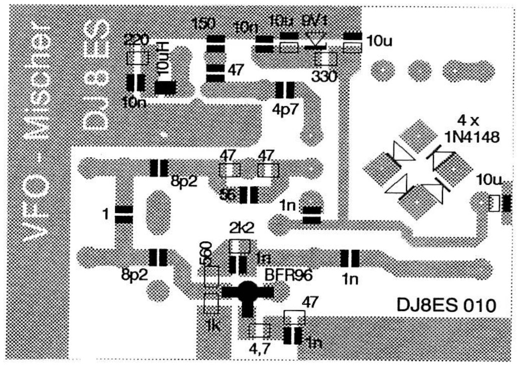


Fig 4.59. Layout of the mixer PCB (ground-plane side)

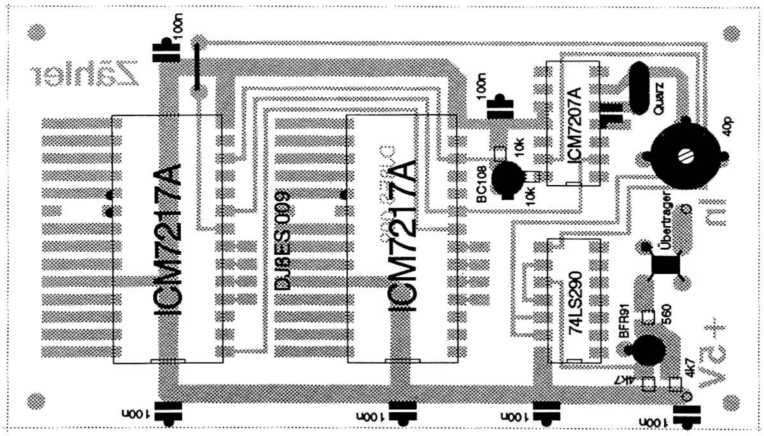


Fig 4.57. Layout of the counter module. Surface-mount components are placed on the track side of the PCB

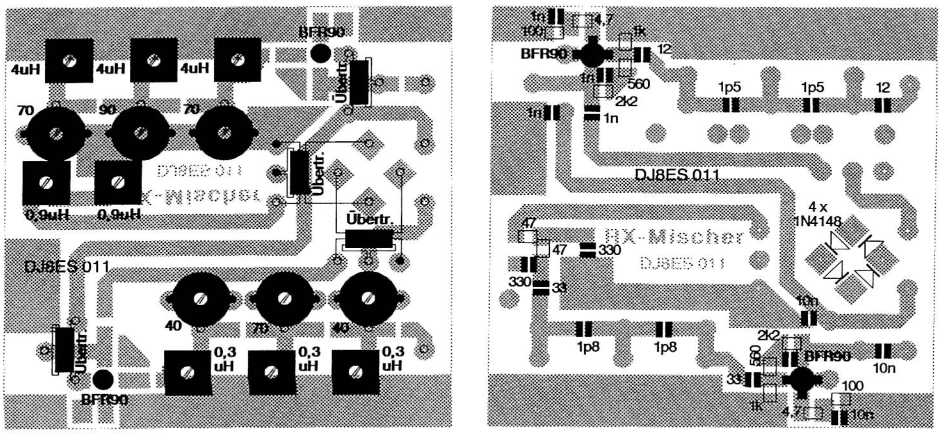


Fig 4.61. Layout of the receive mixer: component side and track side

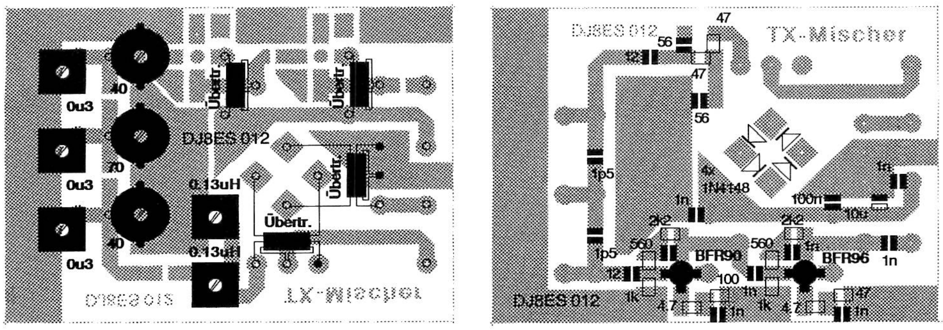


Fig 4.63. Layout of the transmit mixer. Left: component side, right: SMD component layout on track side

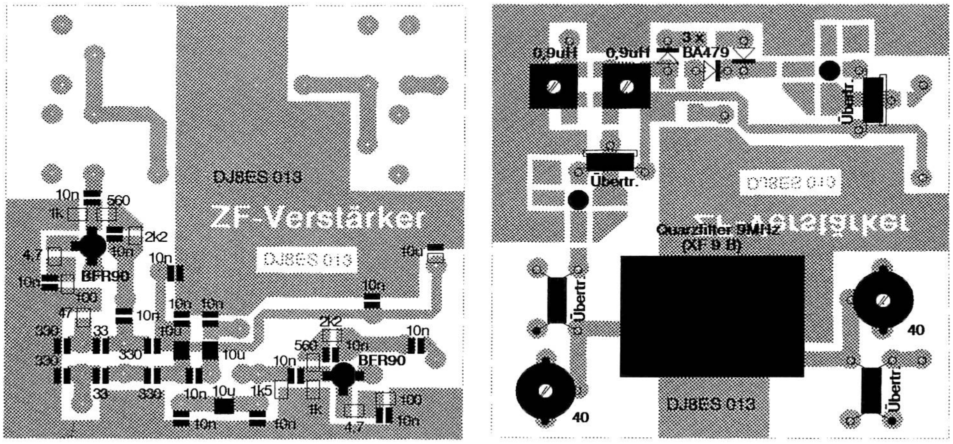


Fig 4.65. Layout of the IF amplifier. Left: component side; right: track side with SMDs

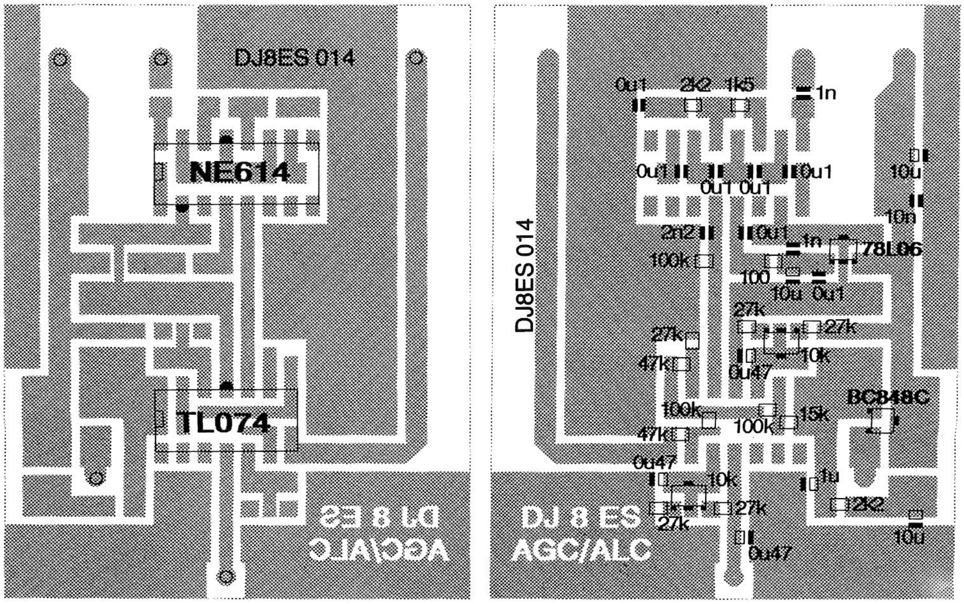


Fig 4.67. Layout of the AGC/ALC module. Left: component side; right: track side

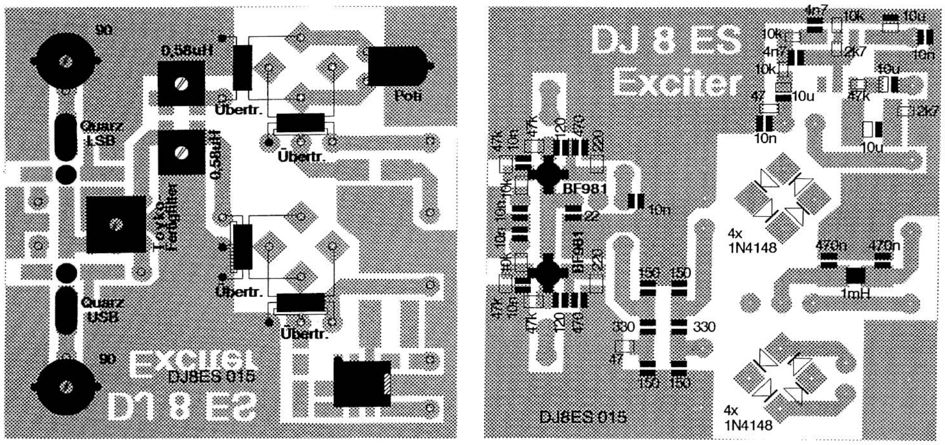


Fig 4.69. Layout of the SSB/CW exciter. Left: the component side with only a few components on the earth plane; right: the track side

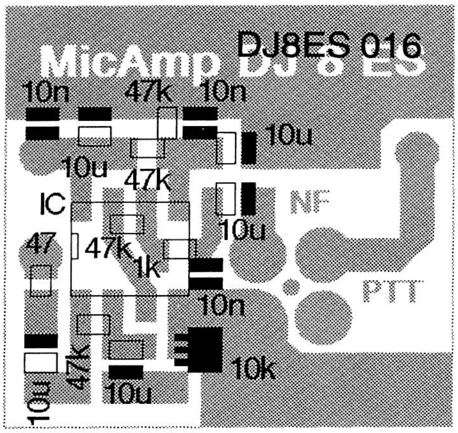


Fig 4.71. The layout of the microphone amplifier (components side)

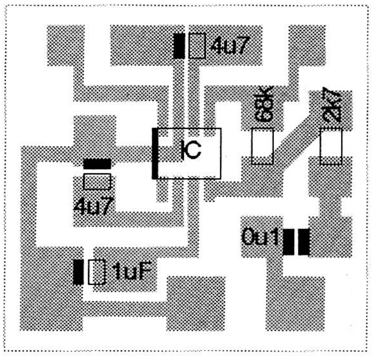


Fig 4.73. Layout of the AF amplifier

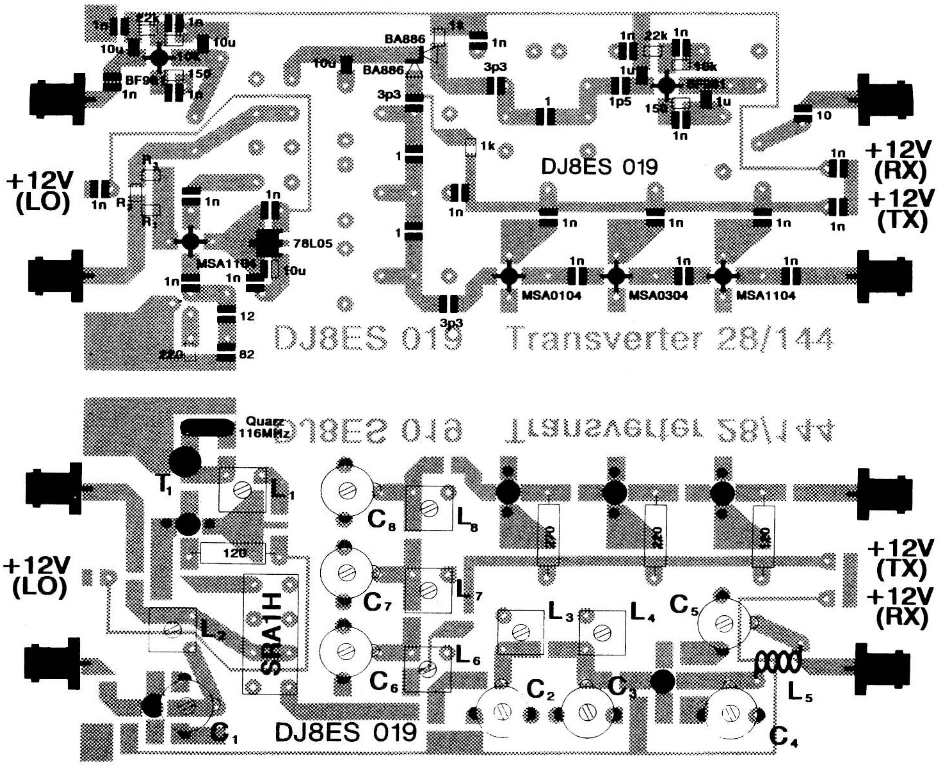


Fig 4.78. Transverter component layout. Top: track side with semiconductors and SMD components; bottom: components side

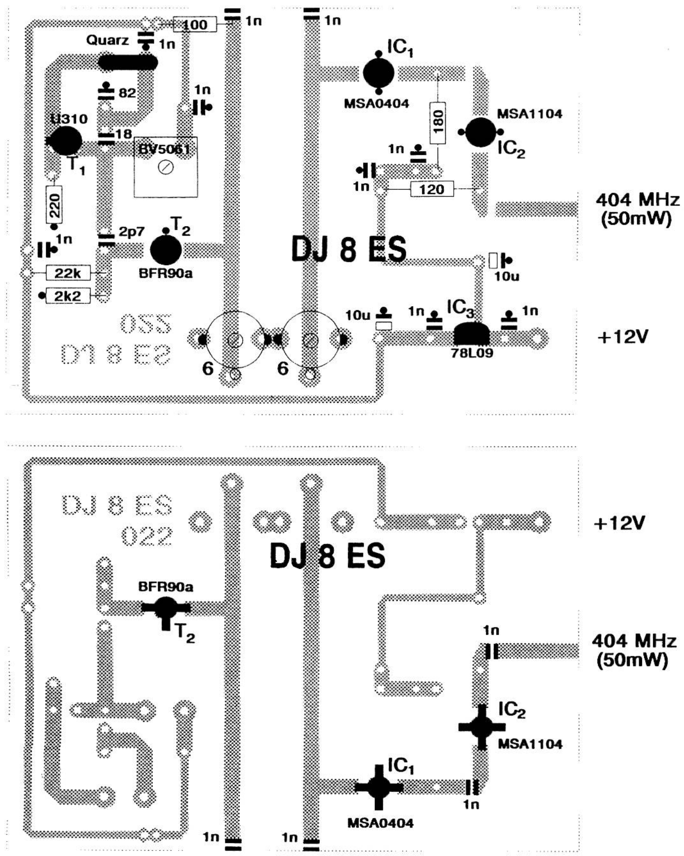


Fig 4.81. Layout of frequency synthesiser. Top: component side; bottom: track side

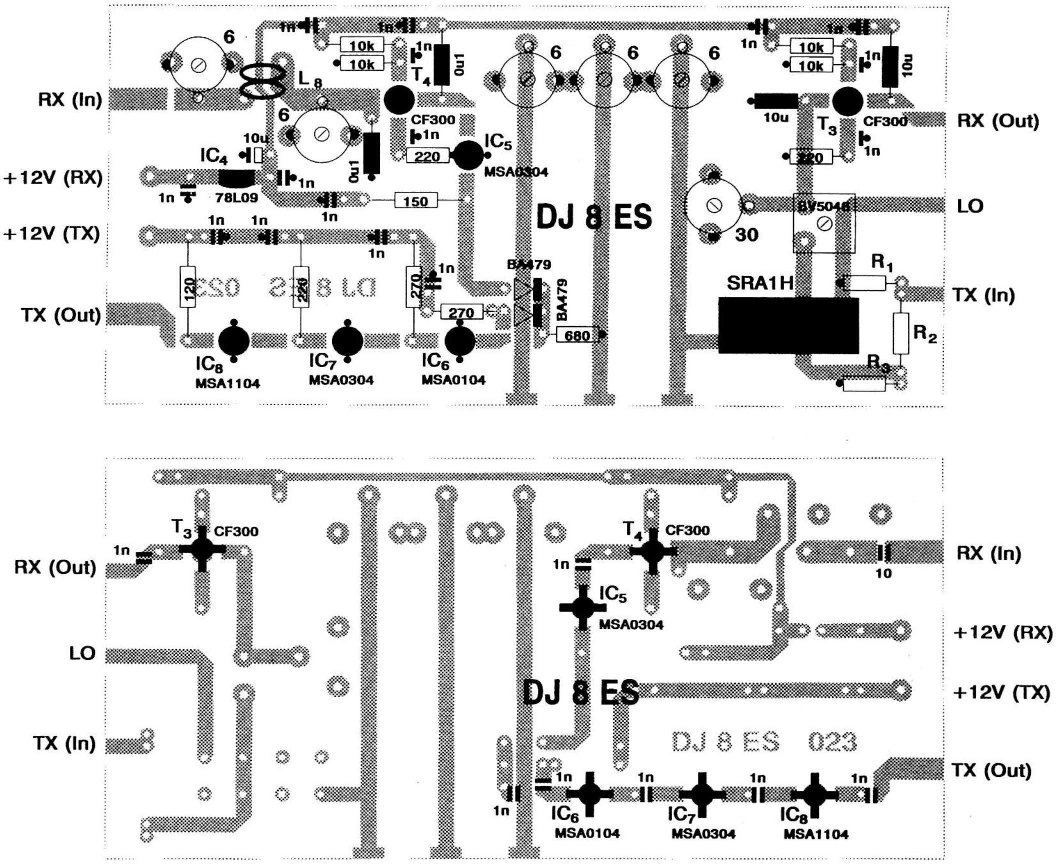


Fig 4.82. Layout of main board of transverter. Top: component side; bottom: track side with semiconductors and coupling capacitors

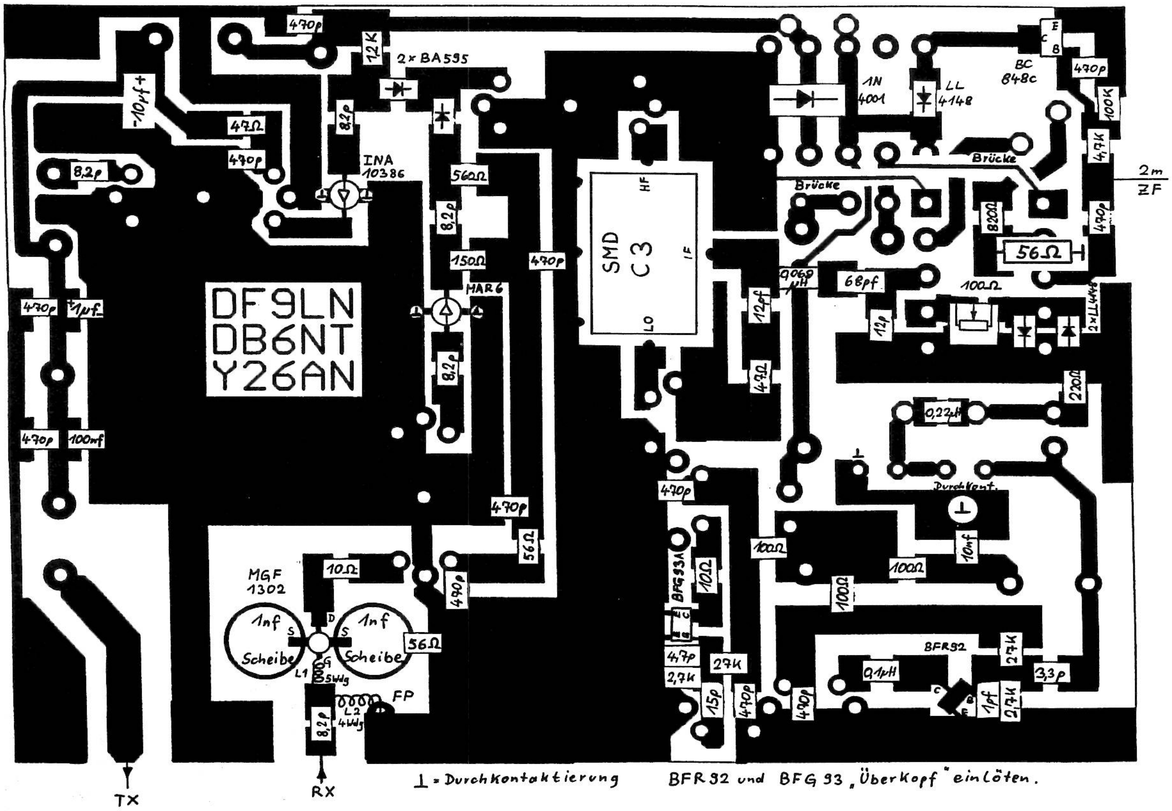


Fig 4.84. PCB track-side layout *(DUBUS Technik IV)*

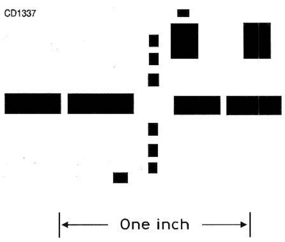


Fig 4.136. Etching pattern for the driver amplifier. The board should be 1/16in G-10 or FR-4 double-sided glass-epoxy circuit board. One side is unetched to act as a ground plane *(QEX)*

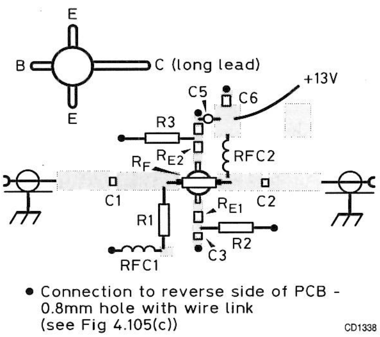


Fig 4.137. Layout diagram for the driver amplifier. Six 0.8mm holes are drilled to allow connecting ground pads with copper foil to the ground plane *(QEX)*

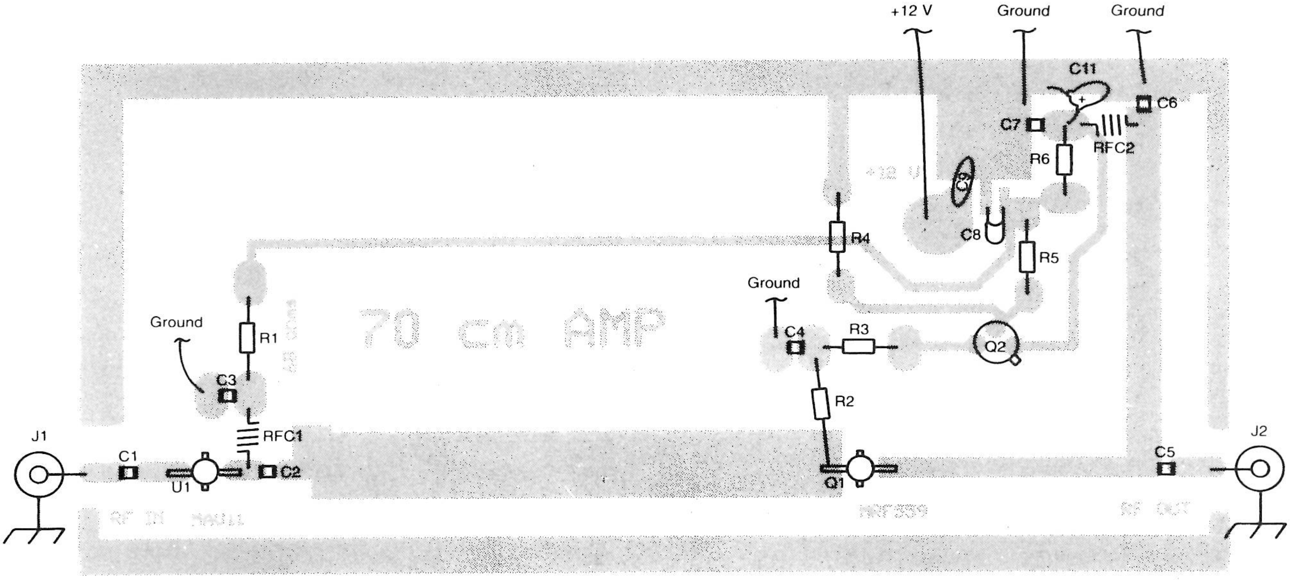


Fig 4.139. Component layout diagram for the amplifier *(QEX)*

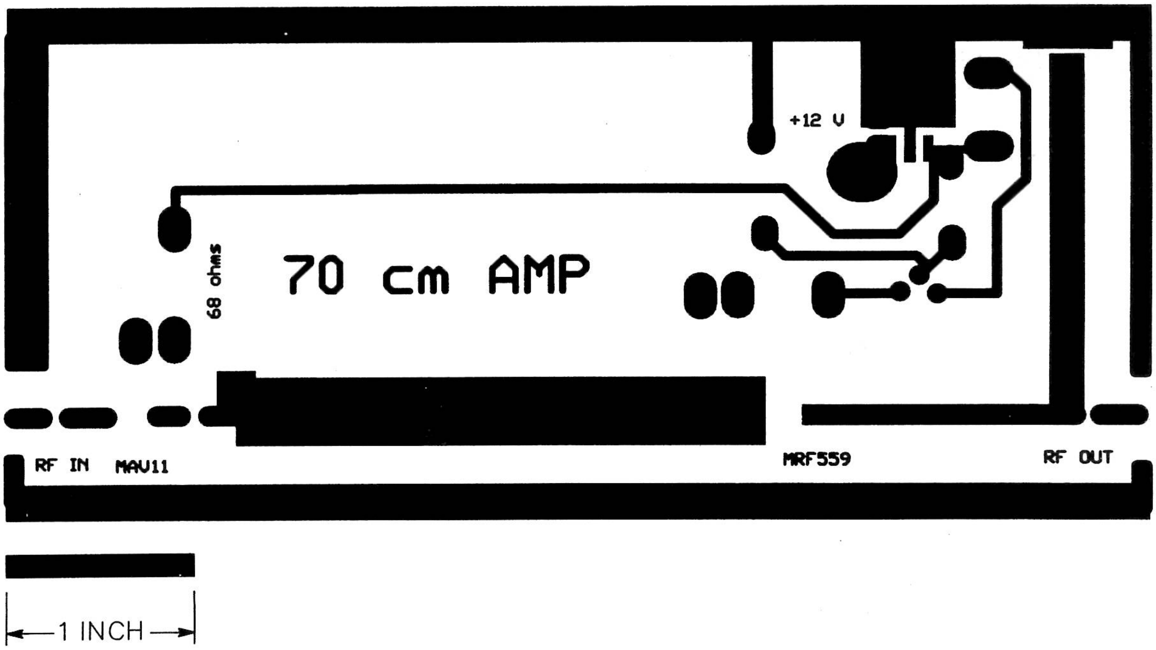


Fig 4.140. Circuit board layout of the amplifier. This shows the component side of the PCB. The reverse side is copper ground plane with links to the component side ground connections with wires/pins (copper foil for U1 and Q1) *(QEX)*

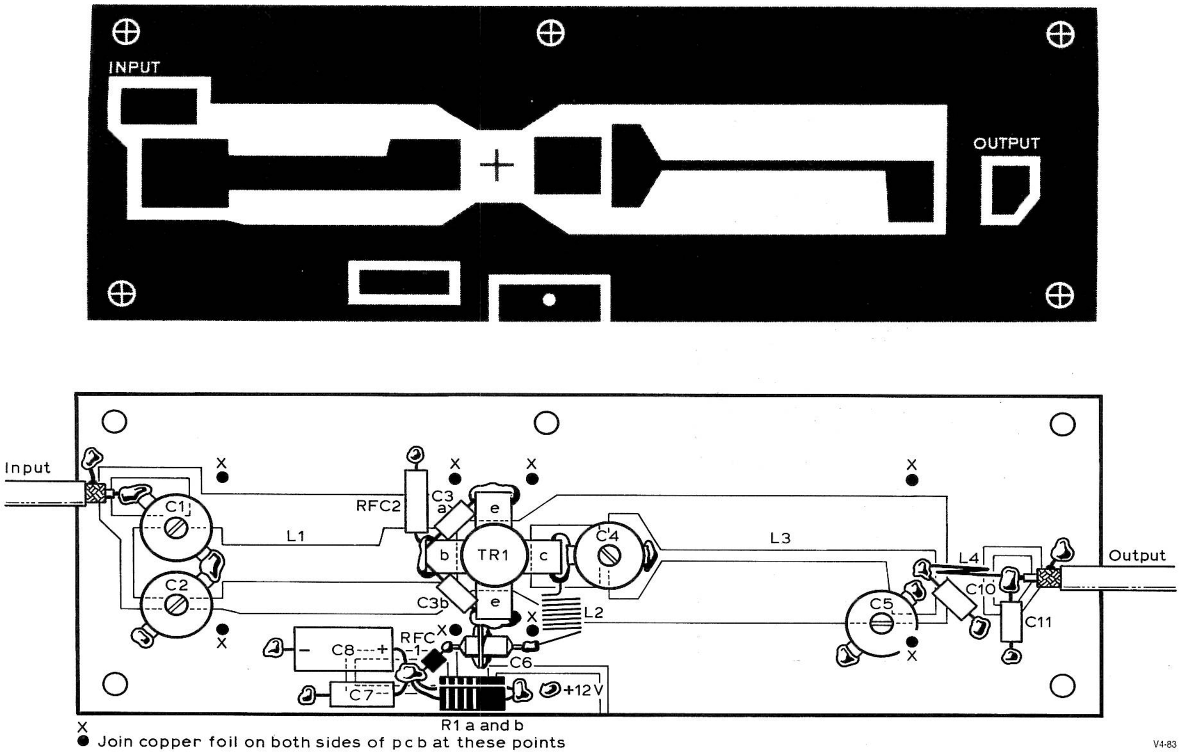


Fig 4.155. Transmitter PA board, actual size, and component layout. This board is double-sided (see text)

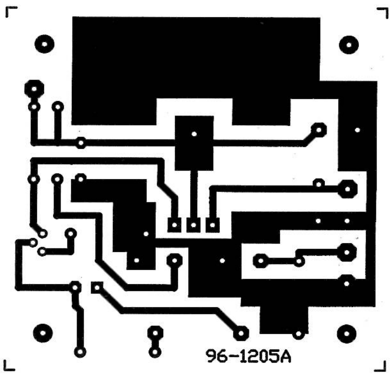


Fig 8.15. PCB for 70cm ATV modulator

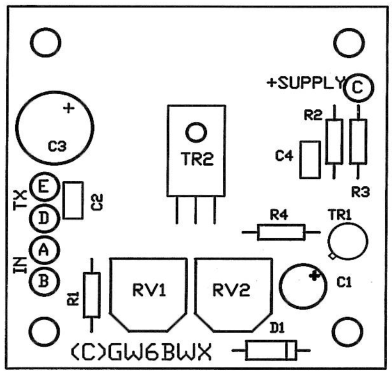


Fig 8.16. Component layout for 70cm ATV modulator

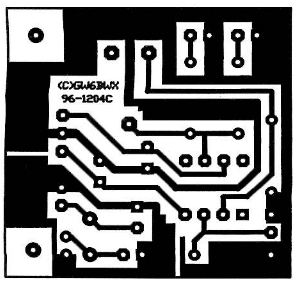


Fig 8.17. PCB for video amplifier

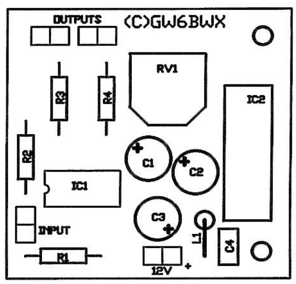


Fig 8.18. Component layout for video amplifier

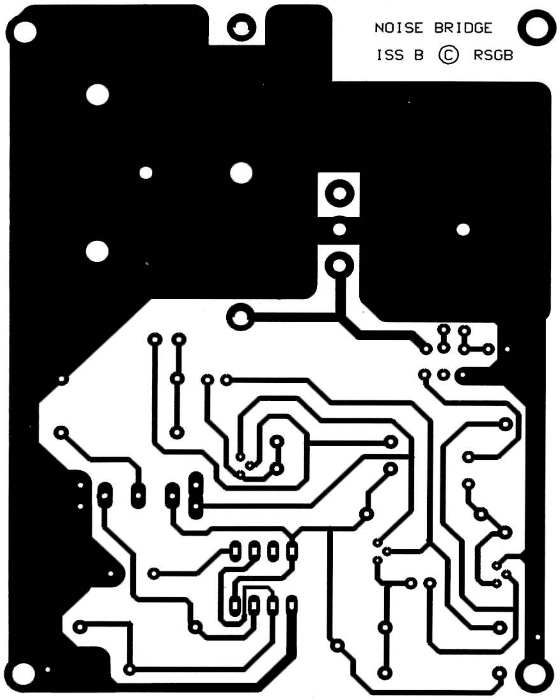


Fig 11.95. Modulated noise bridge PCB

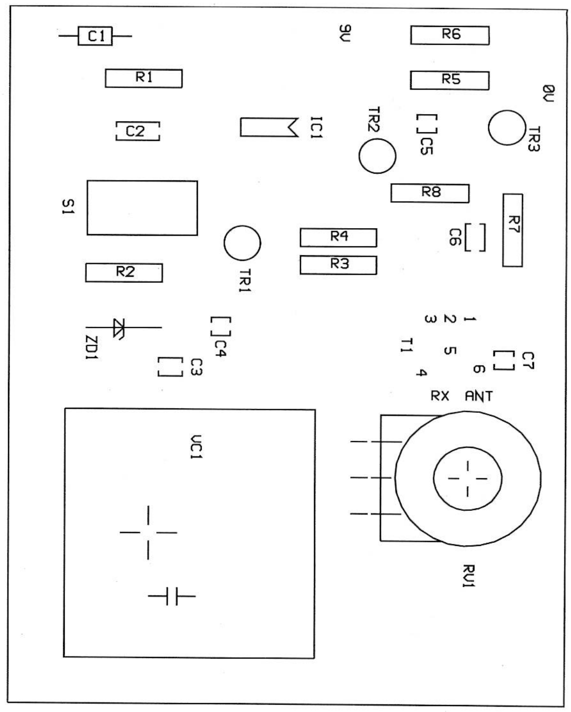


Fig 11.96. Modulated RF noise bridge layout

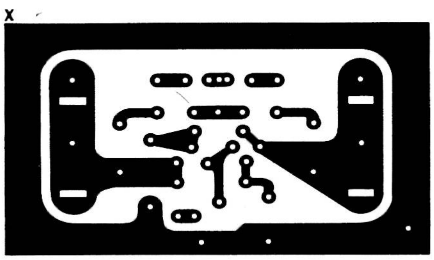


Fig 11.97. VHF dip oscillator PCB

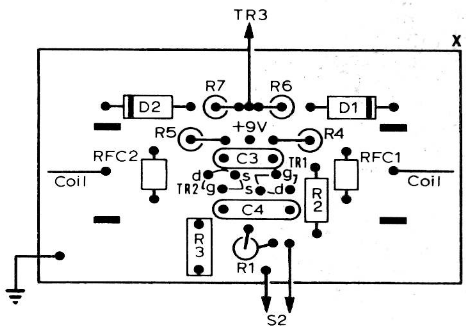


Fig 11.98. VHF dip oscillator layout

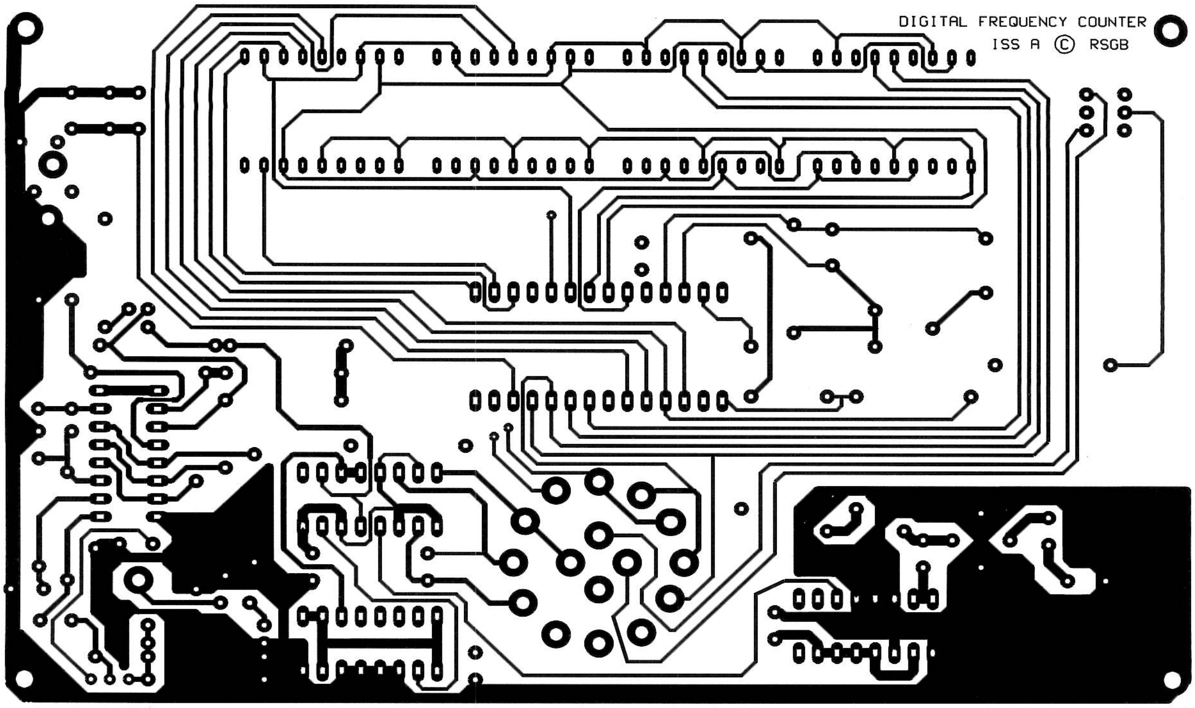


Fig 11.99. 600MHz frequency counter PCB (bottom layer)

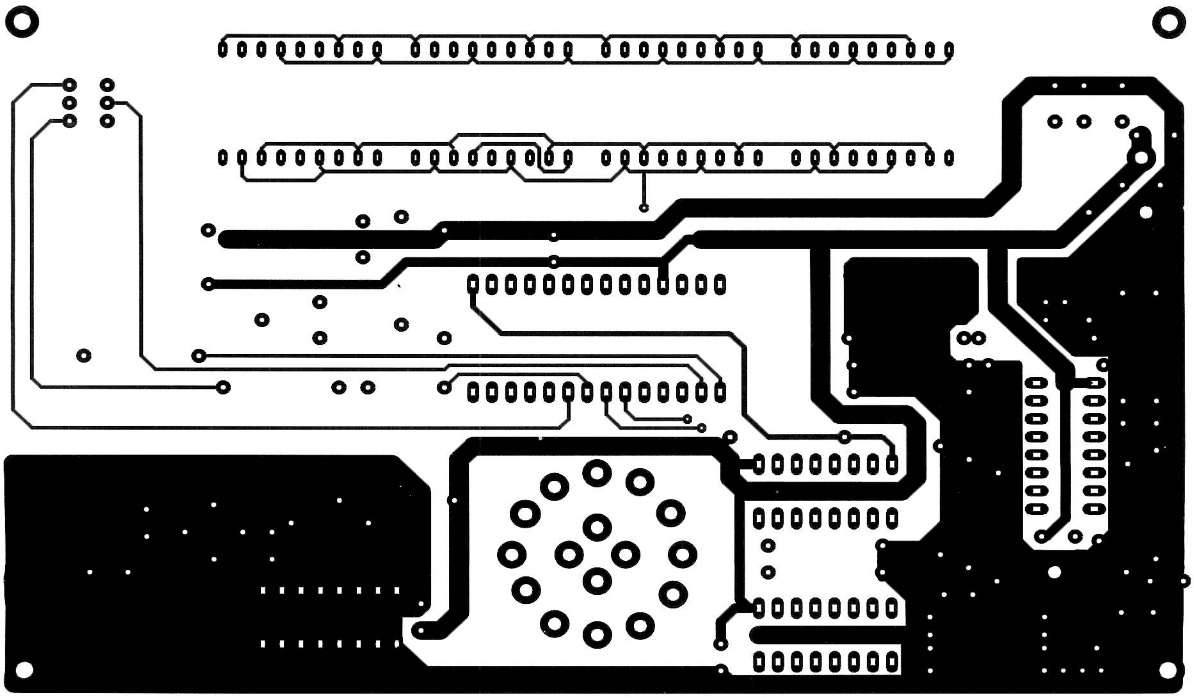


Fig 11.100. 600MHz frequency counter PCB (upper layer)

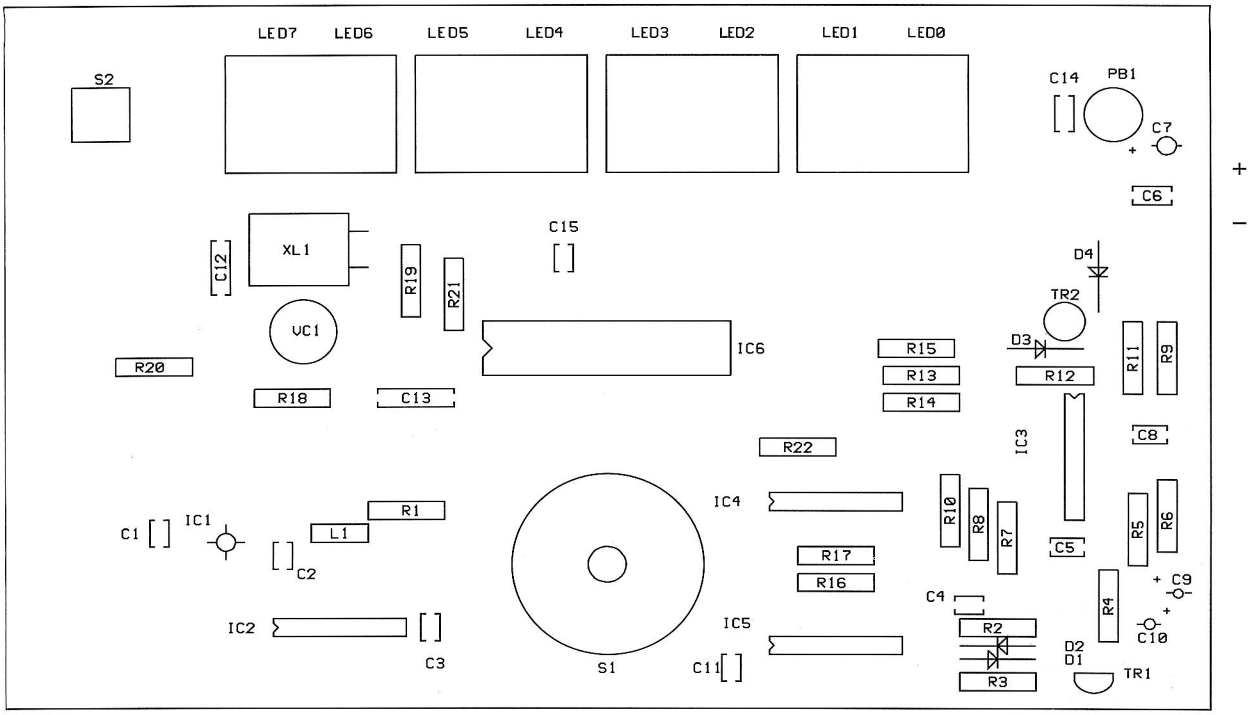


Fig 11.101. 600MHz frequency counter layout

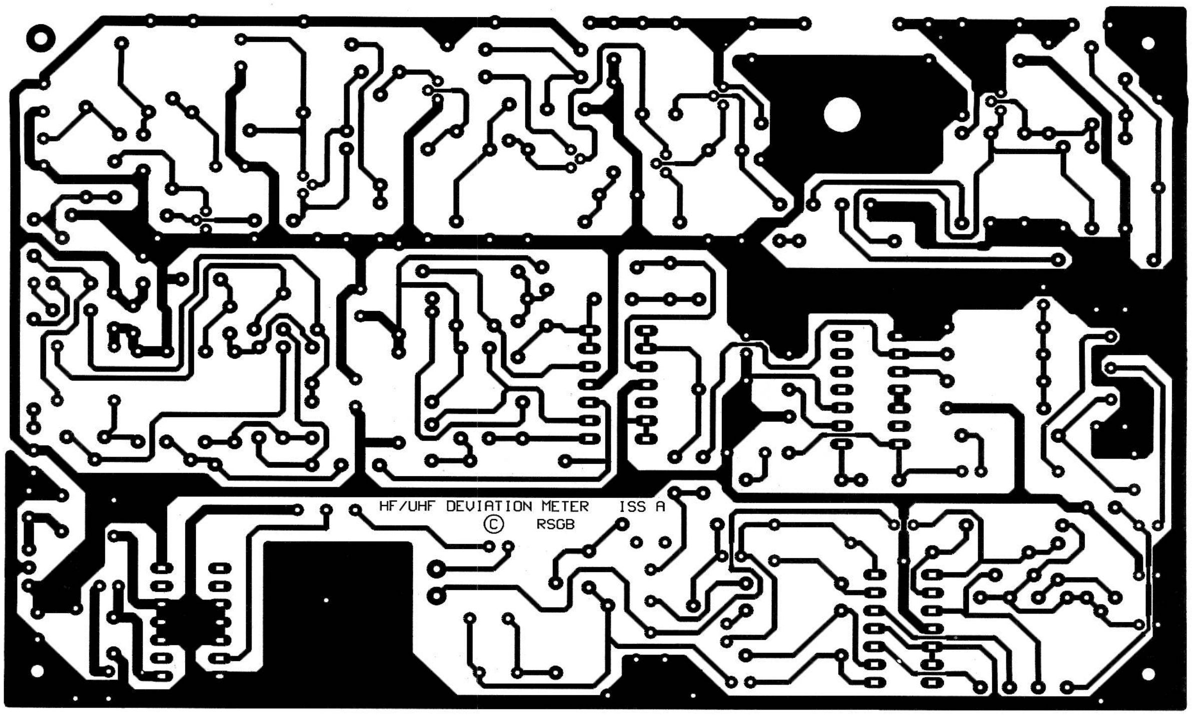


Fig 11.102. HF/UHF deviation meter PCB

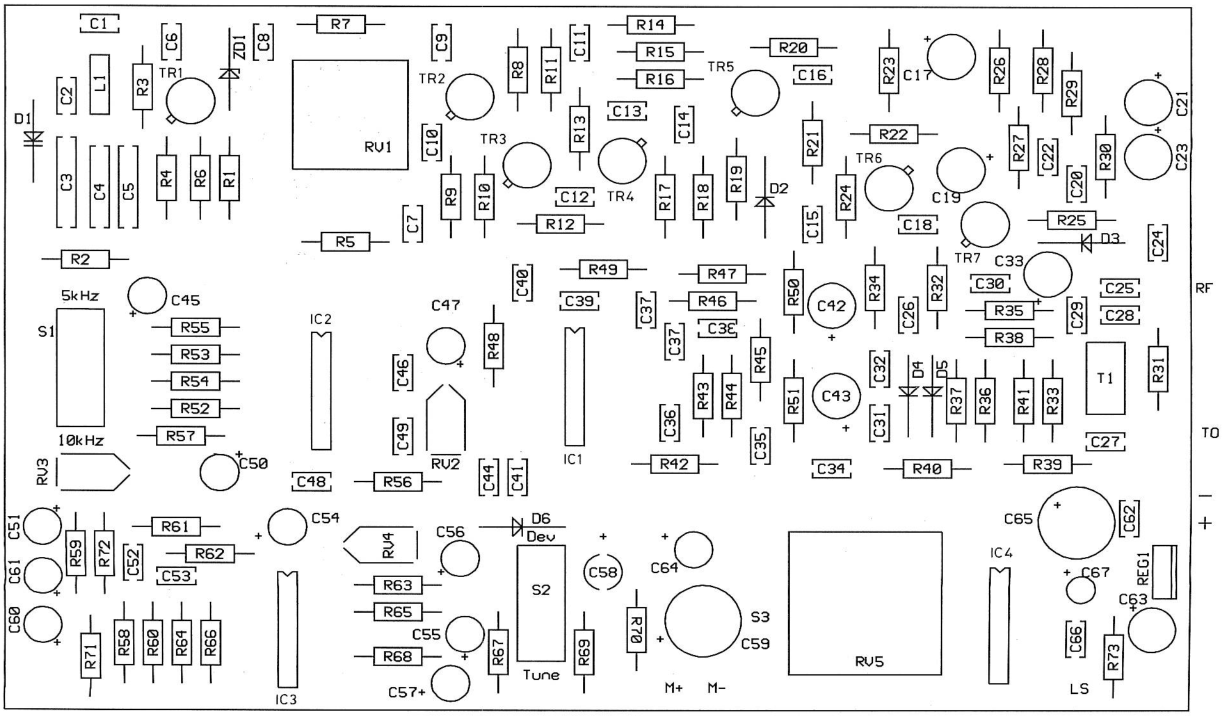


Fig 11.103. HF/UHF devaition meter layout

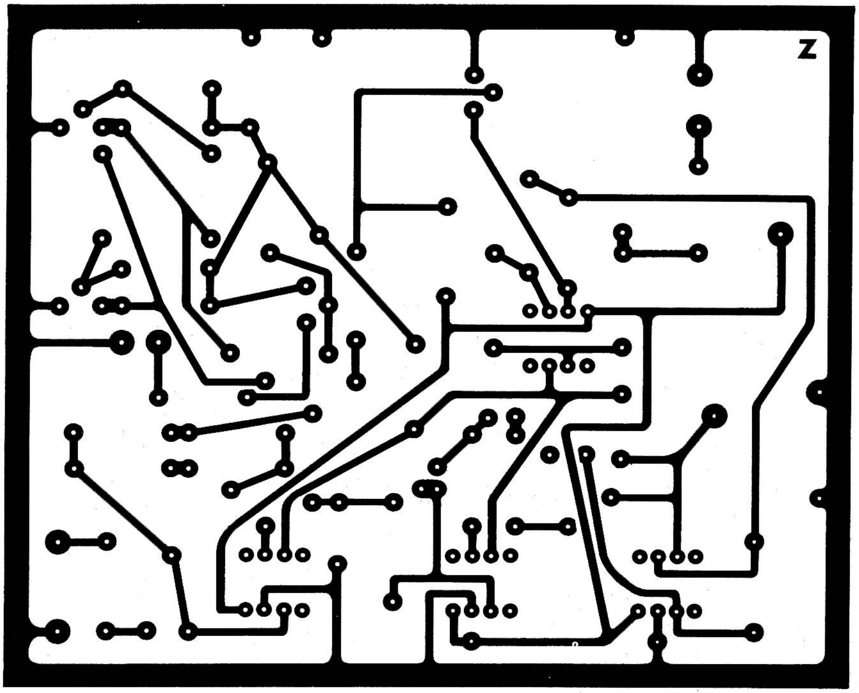


Fig 11.104. Receiver alignment aid PCB

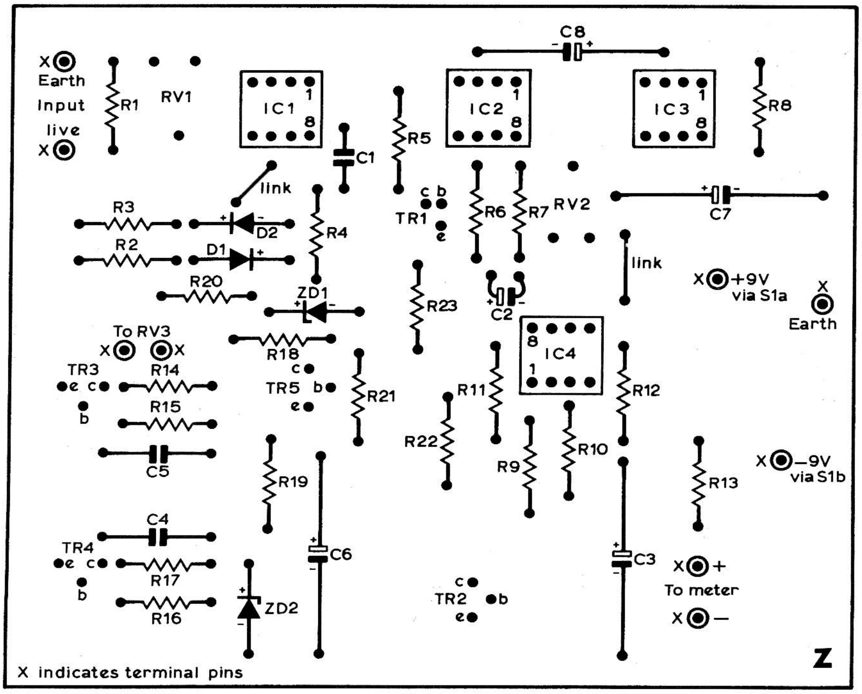


Fig 11.105. Receiver alignment aid layout

2 Fitting coaxial connectors

FITTING coaxial connectors to cable is something we all have to do, and like most things there are probably more wrong ways of fitting connectors than right ones. The methods I'm going to describe are not necessarily the right ones, but they work. Although specific styles of connector and cable are mentioned, the methods are applicable to many others.

CABLES AND CONNECTORS

The main secret of success is using the right cable with the right connector. If you're buying connectors, it is important to be able to recognise good and bad types, and know what cables the good ones are for. Using the wrong connector and cable combination is sure to lead to disaster. Any information you can get, such as old catalogues, is likely to prove useful, especially if you can get the cable cutting dimensions and equivalents lists. Two further sources of dimensions and techniques are [1] and [2]. Some excellent general advice on cable and connector selection is contained in [3].

Cables commonly are of one of two families, the American 'RG' (RadioGuide MIL specification) types and the English 'UR' (UniRadio) series. For details of the most-popular types, see Chapter 12. URM67 is equivalent to RG213, is 10.5mm in diameter and is the most common cable used with type N and PL259 connectors. URM43 (5mm OD) is one usually used with BNC connectors, although these also fit RG58 cable since both have similar dimensions. If there is any doubt about the quality of the cable, have a look at the braid. It should cover the inner completely. If it doesn't it is unlikely to be worth buying.

Having obtained your cable, the easy bit is over. Now to select the connector. The three most popular connector types are the UHF, BNC and N ranges. I'll cover these in some detail, and mention a few others later. If you can, buy connectors from a reputable manufacturer. Some names that spring to mind are RS Components, Greenpar, Suhner, Radiall, Transradio, Kings and Amphenol, among many. There are some good surplus bargains about at rallies.

It cannot be too widely known that the iniquitous UHF connector is no good much beyond 200MHz, because the impedance through the plug-socket junction is not 50Ω. The suitability of N and BNC connectors for use at UHF and beyond is due to them maintaining the system impedance (50Ω) through the connector. PL259 plugs, like the RG8 cable they were intended for, have a lot of nasty imitations. Beware of any that don't have PTFE insulation. They might be OK, but many cheap types are lossy and badly made. The plating should be good quality (silver solders best, although some proprietary plated finishes are just about as good), and there should be two or more solder holes in the body for soldering to the braid. There should be two small tangs on the outer mating edge of the plug which locate in the serrated ring of the socket and stop the body rotating. If you are going to use small-diameter cable with these plugs, get the correct reducer. Often two types are available, one being for 75Ω cable. The 50Ω type is often called 'UG175'. Using the wrong one is certain disaster. Incidentally, buy your reducers at the same time, as some manufacturers use different reducer threads.

With BNC, TNC (like the BNC but threaded) N and C (like N but bayonet) types, life can be more complicated. All these connectors are available in 50 and 75Ω versions. Be sure that you get the right one! To help those of you who are hunting for bargains at rallies, Table A2.1 shows some common manufacturers' designations. All of these connectors have evolved over the years, and consequently you will meet a number of different types. The variations are mostly to do with the cable clamping and centre pin securing method. The original cable clamp is usually called 'unimproved MIL', the later modification the 'improved' and the best for most uses is the 'pressure-sleeve' type. If you are buying new then for normal use go for the pressure-sleeve type. It is much easier to fit. If you are fortunate enough to have some of the double-braided PTFE dielectric cable such as RG142, you may find it easier to use the older clamp types, although the pressure-sleeve type will fit properly with care.

All original clamp types use a free centre pin that is held in place by its solder joint onto the inner conductor. Captive

Table A2.1. Some common connectors and equivalents

Type	Pin	Clamp	Fits cable	Part numbers: MIL No	RS Components	Greenpar
BNC types						
Plug	C	P	URM43	UG88D/U	455-624	GE35070C10
Plug	F	I	URM43	UG88C/U	—	GE35018-10
Plug	F	O	URM43	UG83	—	GE35001-10
Angle plug	C	P	URM43	—	455-646	GE35002C10
Line skt	C	P	URM43	UG89C/U	455-652	GE35060C10
N types						
Plug	C	P	URM43	UG536B/U	455 949	GE15055C10
Plug	C	P	URM67	—	455-753	GE15015C1
Plug	F	I	URM67	UG21E/U	—	—
Angle plug	C	P	URM67	UG594/U	455-393	GE15003C1
Line skt	C	P	URM67	UG23D/U	455-775	GE15022C1

Pin types are: C, captive, and F, free. Clamp types are: P, pressure sleeve, I, improved and O, original.

The Greenpar part numbering system

Greenpar connectors are numbered systematically in a way that should enable you to quickly identify connectors suitable for your use, and to check through those rally ' bargains'. The part number is 'GE' followed by a five-digit number, a letter, another number and lastly some more letters. The first digit is the connector series (N, BNC etc) which is already apparent from looking at the connector. The second digit is vital – it is '5' for 50Ω connectors, and '7' for 70 or 75Ω types. The next three numbers are the connector style. The letter refers to the cable clamp method – it is 'C' for pressure sleeve types, 'A' for modified MIL clamp with captive contact, 'D' for crimp types and '-' for MIL clamps with non-captive pins. The next group of numbers is the cable series. Useful ones are '1' for URM67 and RG213, '4' for RG214 *and* URM67 and RG213, '10' for URM43, URM76, RG58 and 142, and '22' for URM95 and RG174. There are many others for less common cables. The final group of one or more letters refers to the panel mounting holes and optional finish (if any). So a connector numbered GE3507C22 is a BNC plug suitable for RG174 or URM95 50Ω cable with a pressure sleeve clamp.

contact types have a two-part centre insulator between which fits the shoulder on the centre pin. Improved MIL clamp types may have either free or captive contacts. Pressure-sleeve types have a captive centre pin. As an aid to identification, Fig A2.1 shows these types. Pressure-clamp captive-pin types are easy to spot; they have a ferrule or 'top hat' that assists in terminating the braid, a two-piece insulator and a centre pin with a shoulder. Unimproved clamp types have a washer, a plain gasket, a cone-ended braid clamp and a single insulator, often fixing inside the body. Improved types have a washer, a thin ring gasket with a V-groove and usually a conical braid clamp with more of a shoulder. There are variations, so if you can get the catalogue description it helps!

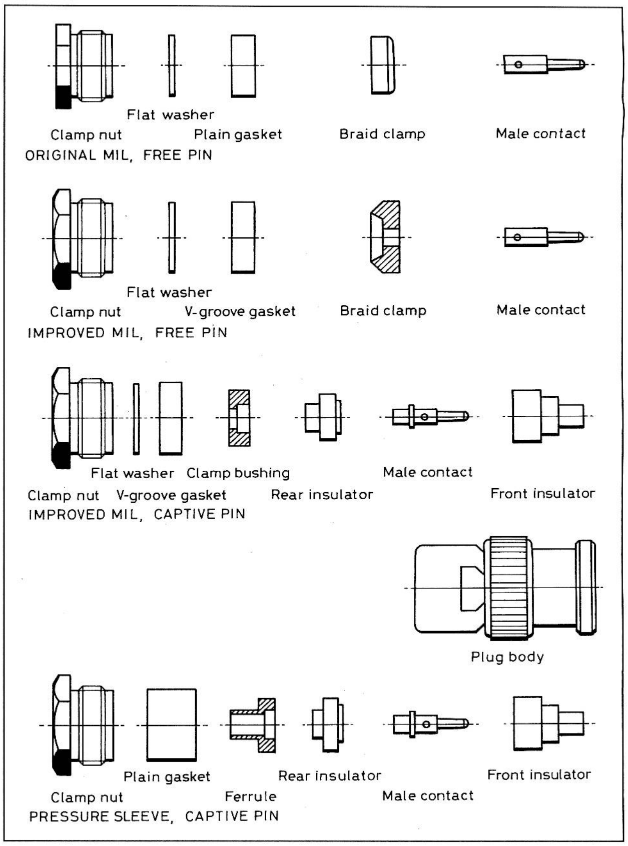


Fig A2.1.Types of BNC/N cable clamps

TOOLS FOR THE JOB

To tackle this successfully, you really need a few special tools: while they may not be absolutely essential, they certainly help. First and foremost is a good soldering iron. If you never intend to use a PL259, then a small instrument type iron is sufficient. If you use PL259s, or intend to use some of the 'dirty tricks' described later, something with a lot more heat output is required. Ideally a thermostatically-controlled iron is best; as with most tools, a little extra spent repays itself handsomely in the future.

A *sharp* knife is another must. A Stanley-type is essential for larger cables, provided that the blade is sharp. For smaller cables, you can use a craft knife or a very sharp penknife. I use a scalpel. A word or two of warning is in order, however. Scalpels excel at the job they were designed for – cutting *flesh*. Make sure it isn't yours! Use sharp blades, cut away from you, and keep the object you're cutting on the bench, *not in your hand*. Although sharp, the steel blades are brittle and will shatter if you apply excessive force or bend them. Dispose of used blades in a box or plastic jar. Model shops have a good range of craft knives which will also do an excellent job.

A pair of sharp small scissors is needed for cutting braids, and a blunt darning needle (mount it in a handle made from a piece of wood dowelling) is useful for unweaving the braid: so too is a scriber. You will find a small vice a great help as well. For BNC, TNC and N type connectors, some spanners are essential to tighten the gland nuts. The BNC/TNC spanners should be thin $^{7}/_{16}$in AF. Those for type N need to be $^{11}/_{16} \times {}^{5}/_{8}$in AF. BNC spanners are sold in pairs by RS Components (available via Electromail in the UK) and are $^{7}/_{16} \times \frac{1}{2}$in AF – the other end suitable for BNC line sockets. A junior hacksaw is needed to cut larger cables such as URM67. Finally, if you intend to put heatshrink sleeves over the ends of plugs for outdoor use, some form of heat gun helps, although the shaft of a soldering iron may work. (You probably have a heat gun already, thinly disguised as a hot-air paint stripper).

PREPARING CABLES

Fitting a plug requires you to remove various bits of outer sheath, braid and inner dielectric. The important knack to acquire is that of removing one at a time, without damaging what lies underneath. To remove the outer sheath, use a sharp knife or scalpel. Place the knife across the cable and rotate the cable while applying gentle pressure. The object of doing this is to score right round the cable sheath. Now score a line from the ring you just made up to the cable end. If you have cut it just enough, it should be possible to peel away the outer sheath leaving braid intact underneath. If this is not something you've tried before, practise on a piece of cable first. For some connectors, it is important that this edge of the sheath is a smooth edge at right-angles to the cable, so it really is worth getting right.

Braid removal usually just requires a bit of combing out and a pair of scissors. Removal of the inner dielectric is most difficult with large-diameter cables with laid multi-strand inner conductors like URM67. Again, it is important that the end is a clean, smooth cut at right-angles to the cable. This is best achieved by removing the bulk of the dielectric first, if necessary in several stages, and finally trimming the dielectric to length. There is a limit to how much dielectric you can remove at one go: 1–2cm is about as much as can be attempted with the larger sizes without damaging the lay of the inner. For the larger cables, it is worthwhile to pare down the bulk of the unwanted material before trying to pull the remainder off the inner. If you can, fit one plug on short cables before you cut the cable to length (or off the reel if you are so lucky). This will help to prevent the inner sliding about when you are stripping the inner dielectric.

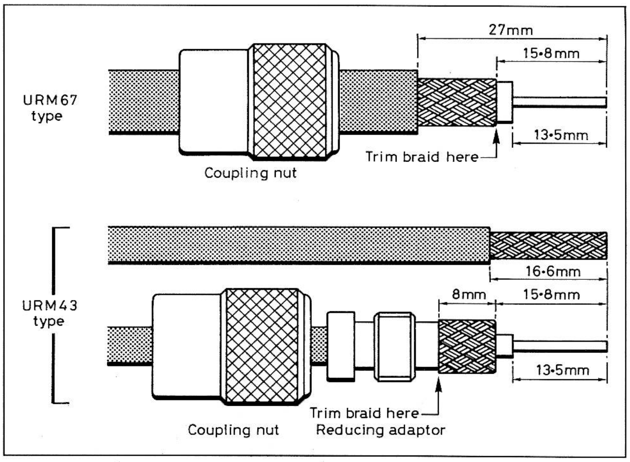


Fig A2.2. PL259 plug assembly

FITTING PL259 PLUGS

Without reducer, URM67-type cable

First make a clean end. For this large cable, the only satisfactory way I have found is to use a junior hacksaw. Chopping with cutters or a knife just spoils the whole thing. Having got a clean end, refer to Fig A2.2 for the stripping dimensions. First remove the sheath braid and dielectric, revealing the length of inner conductor required. Do this by cutting right through the sheath and braid, scoring the dielectric, then removing the dielectric afterwards. Next carefully remove the sheath back to the dimension indicated, *without disturbing the braid.* Examine the braid: it should be shiny and smooth. If you have disturbed it or it looks tarnished, start again a little further down. Now the tricky bit. With a hot iron, tin the braid carefully. The idea is to do it with as little solder as possible. Lightly tin the inner conductor also at this stage. Take a breather while the cable cools.

Now slide the coupling piece onto the cable (threaded end towards the free end). Examine the plug body. If it isn't silver plated, and you think it might not solder easily, apply a file around and through the solder holes. Now screw the body onto the cable, hard. When you've finished, the sheath should have gone into the threaded end of the connector, the inner should be poking out through the hollow pin, and the end of the exposed dielectric should be hard up against the inside shoulder of the plug. Look at the braid through the solder holes. It should not have broken up into a mass of strands; that's why it was tinned. If it has, it's best to start again.

If all is well, lightly clamp the cable in the vice, then apply the iron to the solder holes. Heat it up and then apply solder. It should flow into the holes: if it stays there as a sullen blob, the body isn't hot enough. Now leave it undisturbed to cool before soldering the inner by heating the pin and feeding solder down the inner. Finally, when it's all cool, cut any excess protruding inner conductor and file flush with the pin, then screw down the coupling ring. Merely as a confidence check, of course, test for continuity on both inner and outer from one end of the cable to the other, and check that the inner isn't shorted to the braid.

With reducer, URM43 type cable

First, slide the outer coupler and the reducer on to the cable. Next, referring to Fig A2.2, remove the outer sheath without nicking the braid. Now, using a blunt needle, gently unweave the braid a bit at a time until it is all straight and sticking out like a ruff around the cable. Remove the inner dielectric without nicking the inner conductor, so as to leave the specified amount of dielectric. Tin the inner conductor. Bring up the reducer until the end of it is flush with the end of the outer sheath. Fold the braid back so it lies evenly over the shank of the reducer, then cut off the excess braid with scissors so that it is not in danger of getting trapped in the threads. Smooth it down once more, then offer up the plug body and, *while holding the reducer and cable still,* screw on the plug body until it is fully home. The only really good way of doing this is with two pairs of pliers. Now hold the assembly in the vice and ready the soldering iron. There has been a spirited discussion from time to time about the advisability of soldering the braid through the holes: the best information that I have is that you should. If you don't, the cable will sooner or later fail. So, *with a big iron,* solder the braid through the holes. See the section above for advice. Finally, solder and trim the inner conductor and test the assembly as described earlier.

FITTING BNC AND TYPE N PLUGS

These are 'constant impedance' connectors: that is, when correctly made up, the system impedance of 50Ω is maintained right through the connector. It is vital that the cable fits the connector correctly, therefore check that each part fits the cable properly after you prepare it. Refer to Fig A2.3 for BNC dimensions and Fig A2.4 for N types.

Original or unmodified clamp types

Slide the nut, washer and gasket onto the cable in that order. With the sharp knife, score through the outer sheath by holding the knife and rotating the cable, without nicking the braid. Run the knife along the cable from the score to the end, then peel off the outer sheath. Using a blunt needle, for example, start to unweave the braid enough to enable the correct length of dielectric to be removed. Now slip the braid clamp on, pushing it firmly down to the end of the outer sheath. Finish unweaving the braid, comb it smooth, then trim it with scissors so that it just comes back to the end of the conical section of the clamp. Be sure that the braid wires aren't twisted. Now fit the inner pin and make sure that the open end of the pin will fit up against the dielectric. Take the pin off and lightly tin the exposed inner conductor. Re-fit the pin and solder it in

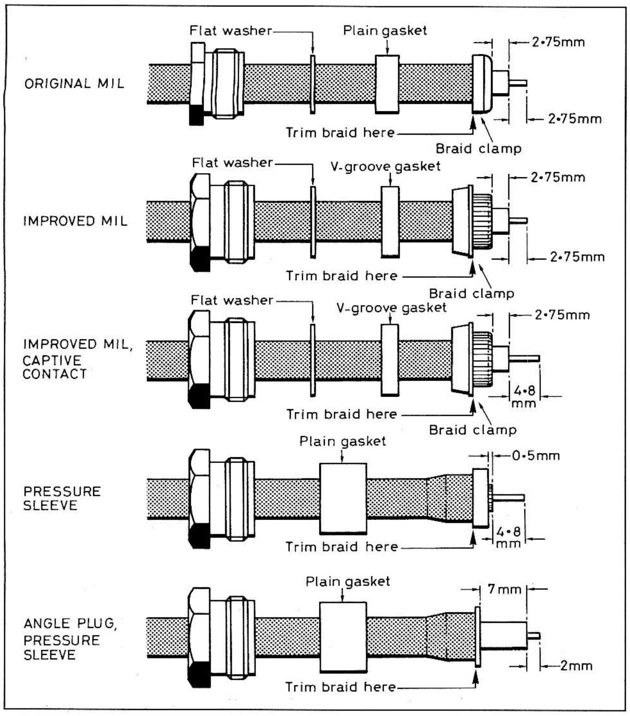


Fig A2.3. BNC dimensions, plugs and line sockets

place by placing the soldering iron bit (tinned but with the solder wiped off) on the side of the pin opposite the solder hole. Feed a small quantity of solder (22 SWG or so works best) into the hole. Allow to cool and examine. If you've been careful enough, the dielectric should not have melted. Usually it does, and swells up, so with the sharp knife trim it back to size. This is essential as otherwise the plug will not assemble properly. Remove any excess solder from around the pin with a fine file. Now push the gasket and washer up against the clamp nut, check the braid dressing on the clamp, then push the assembly into the plug body. Gently firm home the gasket with a small screwdriver or rod and then start the clamp nut by hand. Tighten the clamp nut by a spanner, using a second spanner to hold the plug body still; it *must not rotate.* Finally, check the completed job with the shack ohmmeter.

Modified or improved clamp types

In general, this is similar to the technique for unmodified clamp types described above. There are some important differences, however.

The gasket has a V-shaped groove in it, which must face the cable clamp. The clamp has a corresponding V-shaped profile on one side: the other side may be conical or straight sided, depending on the manufacturer. If the clamp end has straight sides, then the braid is fanned out and cut to the edge of the clamp only, not pushed down the sides. Some types have a small PTFE insulator which is fitted before the pin is put on (common on plugs for the small RG174 cable).

You now appreciate why having the assembly instructions for your particular plug is a good idea! Still, by using these instructions as a guide, it shouldn't be too difficult to get it right, even if it does not fit the first time.

One important point – if the plug has been assembled correctly and tightened up properly, the clamp will have

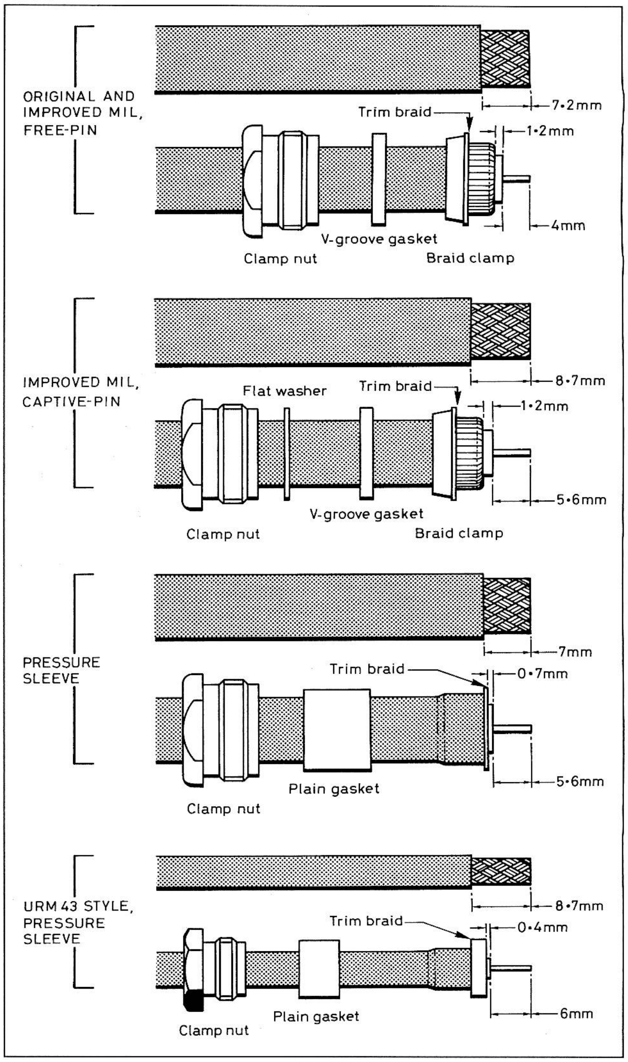


Fig A2.4. N-type dimensions, plugs, angle plugs and line sockets

(intentionally) cut the gasket. It is then rather difficult to re-use it as the gasket, being thin, will not stand a second attempt. The thicker gasket types will often allow careful re-use.

Captive-contact types

These have a small shoulder on the pin and a rear insulator which fits between the pin and the cable. Most types use a thick gasket and a ferrule, although some use a V-grooved braid clamp and thin gasket. I shall describe the ferrule type, as these are the most commonly available and the easiest to fit.

First, slip the nut and gasket on to the cable. Refer to Fig A2.3 or Fig A2.4 for cutting dimensions, then strip off the correct amount of outer sheath by rotating the cable, producing a neat scored circle. Score back to the end of the cable and peel off the unwanted sheath. Comb out the braid, and with it fanned out evenly around the cable, slide the ferrule (small end first) on to the dielectric-covered inner conductor. Push it home so that the narrow portion of the ferrule slides *under* the braid, and the end of the outer sheath rests against the ferrule shoulder. Trim the braid with scissors to the edge of the ferrule. Slide up the gasket so that it rests gently against the ferrule shoulder, which will prevent the braid from being disturbed. Using the sharp knife, trim the dielectric back to the indicated dimension without nicking the inner conductor. Fit the rear insulator, which will have a recess on one side to accommodate the protruding dielectric. Incidentally, if you don't have the size for your particular plug, trim the dielectric until it fits but don't overdo it! Now trim the exposed inner conductor to length and check by fitting the pin, whose shoulder should rest on the rear insulator unless the inner has been cut too long. Tin the inner lightly, then fit the pin and solder it by applying the iron tip (cleaned of excess solder) to the side of the pin opposite from the solder hole and feed a small amount of solder into the hole. Allow to cool and remove excess solder with a fine file. Now fit the front insulator (usually separate from the body) and push the whole assembly into the body. Push down the gasket gently into the plug body with a small rod or screwdriver. Start the nut by hand, then tighten fully with one spanner, using the other to prevent the body from rotating. Check with the ohmmeter, then start on the other end, remembering to put the nut and gasket on first!

VARIATIONS

Angle plugs generally follow a similar pattern to the straight types, except that connection to the inner is via a slotted pin, accessed via a removable cap screw. Tighten the connector nut before soldering the inner. Line sockets are fitted in the same way as plugs.

DIRTY TRICKS

Most of these were originally described in 'The Golden Treasury of Connector Abuse' in the December 1985 issue of the VOWHARS newsletter. The author gratefully acknowledges the contributions by a number of connector abusers, who wish to remain anonymous, and also including G3SEK.

We would like to use new connectors every time, but often a pressure-sleeve type can be re-used if the gasket is not too deformed. Get all the solder you can out of the pin and then carefully ream out the rest with a small drill, held in a pin chuck. The sizes to use are 1mm for URM43-style pins, and 2.6mm for URM67 ones.

Tarnished silver-plated connectors can be made to shine by dipping the metal parts in Goddards 'Silver Dip' silver cleaner. Rinse carefully afterwards, then bake in a slow oven.

BNC connectors for URM67 cable can be rather hard to find. A standard captive-contact BNC plug can be fitted to URM67 in the following way. First discard the nut, gasket and ferrule, and prepare the rear insulator by removing the ridge from it with a sharp knife. Now prepare the cable by cutting with a knife, right through the jacket, braid and insulator about 5mm back from the end. Cut sufficiently deep so that you notch the inner conductor strands, and remove the remains. Carefully bend the six individual outer strands of the inner so they break off flush with the end of the dielectric, leaving one straight inner strand. Now remove sufficient outer jacket (about 2cm) such that when the body is pushed on the cable, some braid is still visible. Tin the braid and inner conductor lightly, then fit the rear insulator, pin and front insulator and push home the assembly into the plug body. With the big iron, heat the plug body and feed solder down the joint with the braid. After it has cooled, put some heatshrink adhesive-lined sleeving over the plug and cable join to protect it. Testing of this trick with a TDR has shown it to be almost as good as the real plug, and certainly better than an adapter. This assembly will happily stand 100W of 1296MHz.

An N plug can be *carefully* pushed on to a BNC socket; OK for quick test equipment lash-ups, but don't do it too often or too hard as you will eventually damage the socket. In a similar vein, the pin of a PL259 is about the same diameter as a 4mm wander plug; after all, what is a PL259 but a screened wander plug?

To make a PL259 to BNC adapter, solder a length of copper wire to the back of a BNC single-hole socket. Drop it (without the nut) on to the top of a PL259 so that the wire pokes through the pin of the plug. With a big iron or a careful blowtorch, solder the body of the socket and plug together. After it has cooled, solder the inner wire to the pin. Not exactly a precision job, but good enough for a PL259!

Finally, to waterproof a connector-cable joint and to provide added strength where flexing of the cable will occur, heatshrink a piece of adhesive-lined heatshrink sleeving over the plug body and cable. For N connectors, a 19mm diameter variety (that shrinks to a minimum of 6mm, such as RS399-748) can be slid on to the cable and connector after assembly.

CONCLUSIONS

With a little practice, care and patience, I hope that these notes make the fitting of connectors a little less of a chancy business. Practice on some short leads (there is no such thing as too many spare short coaxial leads in any shack) and remember that the best time to put new connectors on the feeder isn't two minutes before the start of the contest!

REFERENCES

[1] *The Radio Amateur's Handbook* (any edition), ARRL.

[2] *Microwave Measurements and Techniques,* T S Laverghetta, Artech House, 1976. (Good advice on cable, connectors and how to fit them. Much useful practical advice on many subjects, not just for microwavers!)

[3] *The Buyer's Guide to Amateur Radio,* Angus McKenzie, MBE, G3OSS, RSGB, 1986.

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A magazine which covers a wide range of interests and which features the best and latest amateur radio news. The Society's journal has acquired a worldwide reputation for its content. With 100 pages, many in colour, '*RadCom*' strives to maintain its reputation as the best available and is circulated free of charge to members throughout the world.

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One of the most vital features of the work of the RSGB is the ongoing liaison with the UK Licensing Authority – presently the Radiocommunications Agency (RA) of the Department of Trade and Industry. Setting and maintaining the proper framework in which amateur radio can thrive and develop is essential to the well-being of amateur radio. For example, the Novice Licence was introduced by the RA after long discussions with the RSGB. The Society also spares no effort in defence of amateur radio's most precious assets – the amateur bands.

Beacons and Repeaters

The RSGB supports financially all beacons which are looked after by the appropriate committee of the Society, ie 1.8–30MHz by the HF Committee, 30–1000MHz (1GHz) by the VHF Committee and frequencies above 1GHz by the Microwave Committee. For repeaters, the Society's Repeater Management Committee has played a major role. Society books such as the *RSGB Yearbook* give further details, and computer-based lists giving operational status can be obtained from HQ.

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A wide range of operating awards is available via the responsible officers: their names can be found in the front pages of *Radio Communication* and in the *RSGB Yearbook*. Details of these awards can be found in the latter and also the *Amateur Radio Operating Manual* published by the Society.

Contests (HF/VHF/Microwave)

The Society has two contest committees which carry out all work associated with the running of contests. The HF Contests Committee deals with contests below 30MHz, whilst events on frequencies above 30MHz are dealt with by the VHF Contests Committee.

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The Society has responsibility for Morse testing of radio amateurs in the UK. If you wish to take a Morse test, write direct to RSGB HQ (Morse tests) for an application form.

Slow Morse

Many volunteers all over the country give up their time to send slow Morse over the air to those who are preparing for the 5 and 12 words per minute Morse tests. The Society also produces Morse practice tapes.

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The Society publishes a range of books for the radio amateur and imports many others. Members are entitled to a discount on all books purchased from the Society – this discount can offset the cost of membership.

Propagation

The Society's Propagation Studies Committee is highly respected – both within the amateur community and professionally – for its work. Predictions are given in the weekly GB2RS news bulletins and the Society's monthly magazine *Radio Communication*.

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Although the role of the Society's Technical and Publications Advisory Committee is largely to vet material intended for publication, its members and HQ staff are always willing to help with any technical matters.

Breakthrough in domestic entertainment equipment can be a difficult problem to solve as well as having licensing implications. The Society's EMC Committee is able to offer practical assistance in many cases. The Society also publishes a special book to assist you. Additional advice can be obtained from the EMC Committee Chairman via RSGB HQ.

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There is a special booklet and expert help available to members seeking assistance with planning matters.

GB2RS

A special radio news bulletin transmitted each week and aimed especially at the UK radio amateur and short wave listener. The script is prepared each week by the Society's HQ staff. The transmission schedule appears in the *RSGB Yearbook*. The GB2RS bulletin is also available via the packet radio network, the RSGB website and a premium-rate telephone number.

RSGB Exhibitions and Mobile Rallies

The Society's Exhibition and Rally Committee organises an annual exhibition and an annual mobile rally. Full details and a rally calendar can be found in *Radio Communication*.

RSGB Conventions

The Society's diary in *Radio Communication* contains details of all special conventions which are open to all radio amateurs. The Society holds several major conventions each year.

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A number of leading national radio societies have volunteers who monitor the amateur bands as a service to the amateur community. Their task is to spot licence infringements and defective transmissions, and report them in a friendly way to the originating station.

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This helps to protect the exclusive amateur bands by monitoring for stations not authorised to use them.

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RSGB

EDITED BY DICK BIDDULPH, G8DPS

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