25 Measurement and Test Equipment

Correct operation of amateur radio equipment involves measurements to ensure optimum performance, in order to comply with the licence conditions and to avoid interference to other users. This will involve the use of test equipment, as will the repair and maintenance of equipment.

Some professional test gear is very expensive but it is the intent of this chapter to show how some of the cheaper (and perhaps home-built) equipment can be used to good effect.

For further information, equipment and a more detailed discussion of some of the topics the reader should consult [1]. Useful material is also contained in [2] and [3].

One word of caution - whilst the components used are presently available it is a fast changing situation and some components may become obsolete during the life of this book. If a component appears to be unavailable it is always worth putting the part number into a search engine on the Internet.

CURRENT AND VOLTAGE MEASUREMENTS

Most electrical tests rely on the measurement of voltage and current. To this end many types of instrument have been developed, such as meters, oscilloscopes, spectrum analysers etc. These are all examined in this chapter.

The ubiquitous multimeter (**Fig 25.1**) tends to be used for many voltage and current measurements nowadays. The units are either analogue or digital, they are relatively cheap and usually provide resistance measurement as well. Because they are so cheap it is usually not worth making one, except for the experience.

However, when making power supplies, amplifiers etc, it is important to have meters dedicated to a single function, or a group of functions. The following sections deal with this, and how the meters can be adapted to the ranges that need to be used.

One problem with all measuring instruments is how they affect the circuit they are measuring, due to the power they require to provide the input signal.

Fig 24.1: Typical analogue and digital multimeters

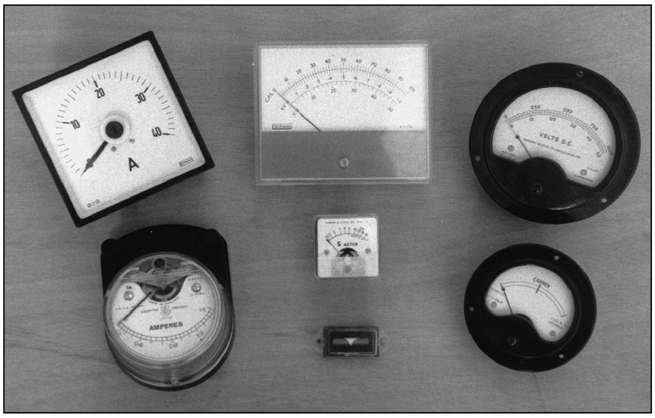


Fig 25.2: Various moving-coil and moving-iron meters

ANALOGUE METERS

These are of electromechanical design and consist (amongst others) of the moving-coil and moving-iron type meters (**Fig 25.2**). The moving coil meter has a linear scale while the moving iron meter is non-linear, the scale being very cramped at the lower end.

The moving-coil meter is the most sensitive and the most accurate, but will respond to DC only, while the moving-iron instrument is AC/DC with a response up to about 60Hz. The modern analogue meter tends to be rectangular, older types usually being round.

The sensitivity of analogue meters is defined by the current that must flow through them in order to provide *full-scale deflection* (FSD). The moving-coil range starts at about 50μA FSD while the moving iron range works from about 100mA FSD.

These analogue meters do draw current from the circuit under test to operate and the coil has resistance (Rm) because it is made from wire. When used for measuring current, **Fig 25.3(a)**, the meter is placed in series with the circuit and so there is a voltage drop - typically 100 to 200mV. When used as a voltmeter, **Fig 25.3(b)**, the current drawn depends on the basic meter movement and, if this takes more than 10% of what is flowing in the circuit, then the circuit conditions are being progressively affected.

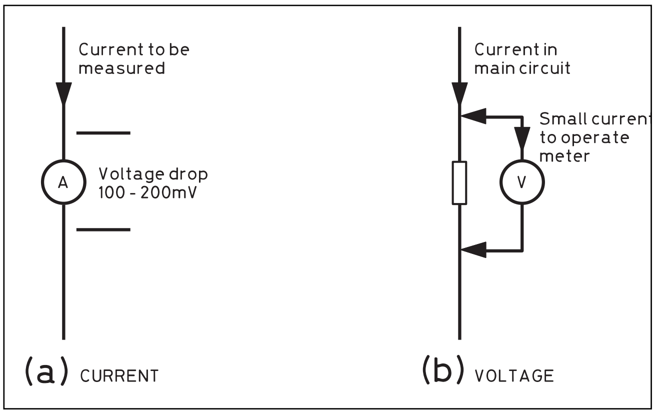


Fig 25.3: The use of meters for measurement

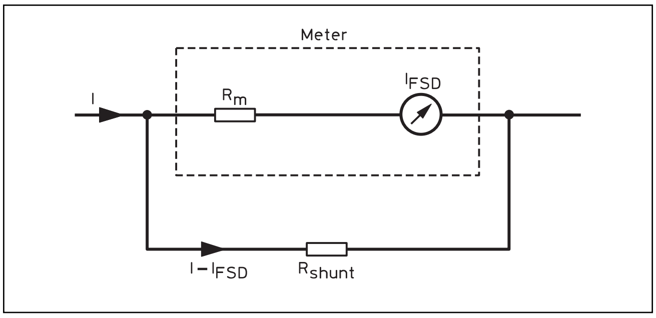


Fig 25.4: Arrangement for current shunts

Extending the Range of Analogue Meters

The meters referred to above come in various fixed arrangements and may not suit the ranges it is desired to measure. It is possible, by the addition of resistors, to extend the range of meters, possibly still using the original scaling.

There is no reason why the scale should not be redrawn by hand or by using transfers. The scale plate can often be removed.

An analogue meter requires current to operate; consider the measurement of current initially. The FSD of the meter cannot be changed so it is necessary to shunt some of the current to be measured around the meter, the typical circuit being shown on **Fig 25.4**.

Here, assuming the maximum current to be measured is I, the shunt resistance is given by:

$$R_{shunt} = \frac{R_m I_{FSD}}{I - I_{FSD}}$$

where I_{FSD} is the current for full-scale deflection of the meter and R_m the resistance of the meter. It is normal to choose I so that only a multiplying factor is required of the scale reading. The power rating of the shunt can be calculated and is:

$(I - I_{FSD})^2 R_{shunt}$.

Example: It is desired to use a 100µA FSD meter to measure a maximum current of 500µA. The resistance of the basic movement is 2000Ω. Substituting these values in the above formula gives:

R_{shunt} = 500Ω with a power rating of 80µW

An alternative way of considering this problem is to consider what the multiplying factor (n) of the scale must be. Using the previous definitions of resistors, the formula for the shunt becomes

$$R_{shunt} = \frac{R_m}{n - 1}$$

Applying this to the above example, then n = 5 and the same value of shunt is found. However, the power rating of the shunt must still be determined.

For use as a voltmeter, the maximum voltage to be read should provide the value of I_{FSD}. The circuit used in this case is shown on **Fig 25.5**. The equation for the resistance of the series resistor R_{mult} is given by

$$R_{mult} = \frac{V}{I_{FSD}} - R_m$$

The power rating for the resistor is given by $I^2_{FSD} R_{mult}$.

Example: A 50µA movement meter with a coil resistance of 3000Ω is required to measure voltages up to 30V. Calculate the multiplier resistor.

R_{mult} = 597kΩ with a power rating of 1.5mW

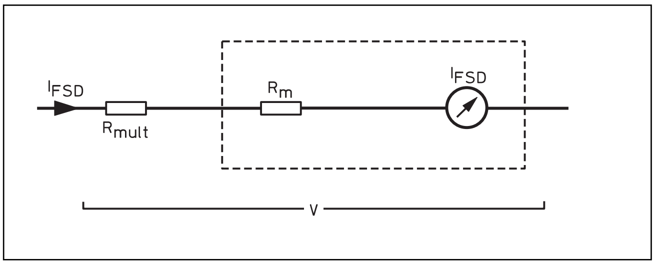


Fig 25. 5: Arrangement for voltage multipliers

These simple calculations show the basis on which the familiar multimeter is based and how they are designed. The switch on the multimeter merely switches in different shunt and multiplier resistors. Remember, these calculations only apply to DC for the moving-coil meter.

Meter Sensitivity

The sensitivity of a voltmeter is usually expressed in ohms/volt. This is merely the reciprocal of the full-scale current sensitivity I_{FSD} of the basic meter. Hence, a 1mA meter used as a voltmeter would be described as 1000Ω/V and a 50µA meter as 20,000Ω/V.

Effect on Circuit Readings

Putting a voltmeter across a resistor may upset the circuit conditions, and the loading effect of a meter has to be considered. For example, putting a meter which requires 50µA across a resistor through which only 100µA flows will disturb the circuit significantly. Putting the same meter across a resistor through which 10mA flows will have little effect. How can one gauge this or guard against it?

Consider a 20,000Ω/V meter. Set on the 10V range this will have a resistance of 10 x 20,000Ω = 200kΩ. It is suggested that any resistance across which this voltmeter is placed should have a maximum value of one-tenth of this, eg 20kΩ. Hence, for any range one can use this rule-of-thumb method. The smaller the percentage, the more accurate will be the reading.

For ammeters the point that must be considered is the voltage drop across the ammeter in relatively low voltage circuits (ie I_{FSD} x R_m). For example, a 0.5V drop across an ammeter is unacceptable in a 12V circuit but it is immaterial in a 100V circuit. One must therefore choose a meter that has as low a coil resistance as possible. This reduces the in-circuit voltage drop and keeps any shunt resistance value as high as possible. If possible, use an ammeter of I_{FSD} equal or just greater than the range required.

For mains circuits of 100V or above the moving-iron meter represents a more viable alternative and tends to be cheaper.

Meter Switching

In order to save cost (and sometimes panel space), it may be worthwhile for a meter to serve several functions. This is more likely to be used in valve circuits for measuring grid and anode voltages and currents. These normally require different ranges for the various parameters being measured. For convenience, two meters would be used - a voltmeter and an ammeter.

In all instances a break-before-make switch should be used. Care should also be exercised in selecting the switch when used in high-voltage circuits.

When measuring current, the resistance of wire and switch contacts may affect the value of low-value shunts.

Voltage measurements are normally made with respect to 0V or earth. This means that one end of the voltmeter is fixed - see

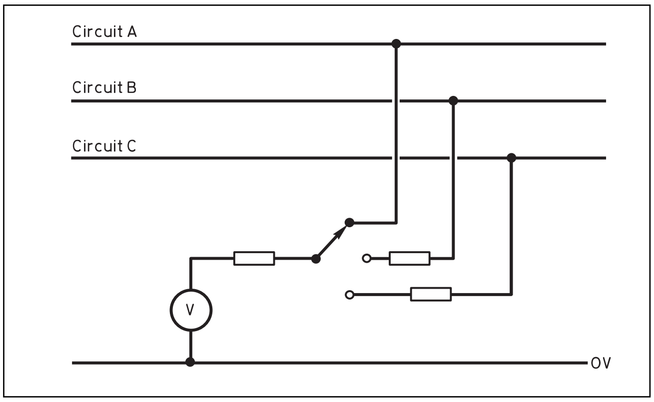


Fig 25.6: Switched voltage measurements

Fig 25.6. Knowing the characteristics of the meter, the various values of series resistance can be calculated. It is suggested that the lowest value is usually wired directly in series with the meter and then the other values chosen such that this value plus the additional one equals the value calculated. Assuming that circuit A in Fig 25.6 has the lowest voltage to be measured, then some current limiting always exists in series with the meter. For current measurements the problem is overcoming contact and wire resistance when low-value shunts are used (ie less than 0.1Ω).

For the purposes of this discussion, a meter is assumed to have 1mA FSD and coil resistance of 100Ω. Fig 25.7 shows how switching could be arranged for the measurement of current on three ranges. Switching/conductor resistance is unlikely to be a problem with circuit A, but it may be a problem on circuit B and certainly will be on circuit C.

One solution is to use a non-switchable meter for any current range which requires a low shunt value, typically less than about 0.5Ω.

A different approach is to consider the meter as measuring volts across a resistor.

The problems of measuring a voltage and the current taken must then be considered as previously discussed. If a 50μA meter was used, then it must be possible to develop a minimum voltage drop of about 150mV; for a 1mA movement it should be about 100mV.

The voltage drop should be equal to or greater than $I_{FSD} \times R_m$. The typical circuit used for this arrangement is given in Fig 25.8.

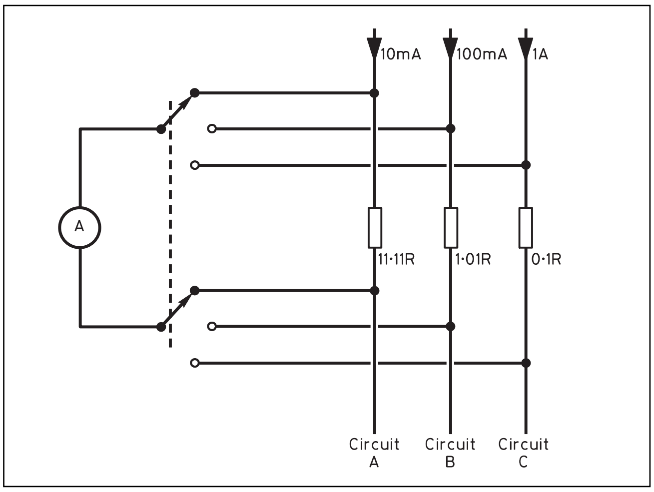


Fig 25.7 Switched current measurements

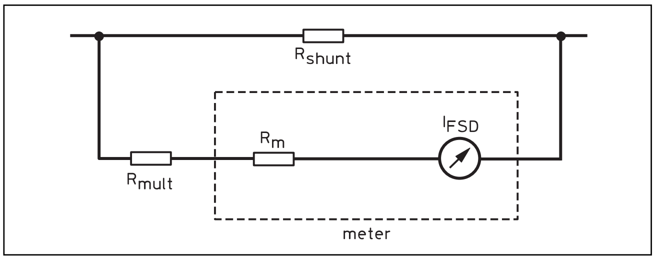


Fig 25.8: Current measurement by volt-drop method

Meter Protection

Meters are relatively expensive and easily damaged if subjected to excessive current. Damage can be prevented simply and cheaply by connecting two silicon diodes in parallel (anode to cathode) across the meter terminals as in Fig 25.9, and this should be regarded as standard practice. No perceptible change of sensitivity or scale shape need occur.

A characteristic of silicon diodes is that they remain very high resistance until the anode is some 400mV above the cathode, at which point they start to conduct and the resistance falls to a low value. Since the voltage drop across the average meter is around 200mV, it follows that a silicon diode connected across the meter will have no effect even when the meter shows full-scale deflection. If, however, the meter is overloaded to twice the FSD and the voltage across the meter rises to 400mV, the diodes will begin to conduct and shunt the meter against further increase of fault current.

Most meters will stand an overload of at least twice the FSD without damage but it is wise to include a series resistor as shown on Fig 25.9 to ensure the protection afforded by parallel diodes without affecting the meter. The series resistance ensures that the voltage drop across the meter/resistor combination is 200mV minimum. Parallel diodes are used because excessive current in either direction can damage the meter.

Example: What series resistor should be added to a 1mA FSD meter with a coil resistance of 100Ω?

At 1mA FSD the voltage drop across the meter is 1mA x 100Ω = 100mV. Thus the drop across the series resistor should also be 100mV, and this requires a resistance of 100Ω. This then means that the meter is protected for currents in excess of 400mV/200Ω = 2mA.

If an additional series resistor is to be included then any shunts to be included to increase the current range should be placed across this combination and the series resistance taken into account when making the calculations.

For most cases, small-signal silicon diodes such as the OA202, 1N914 or 1N4148 are satisfactory - they have the advantage of having an inherently high reverse resistance - ie a low reverse leakage current is required as this shunts the meter circuit. However, it is important that under the worst fault conditions the diode will not fail and go open-circuit, thus affording no protection. An example of this with small-signal diodes would be

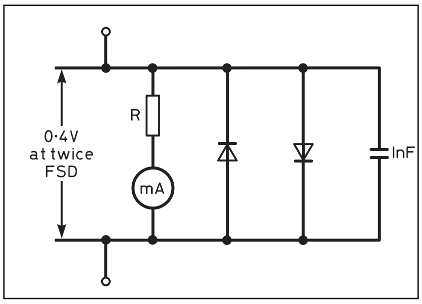


Fig 25.9: Meter protection using diodes

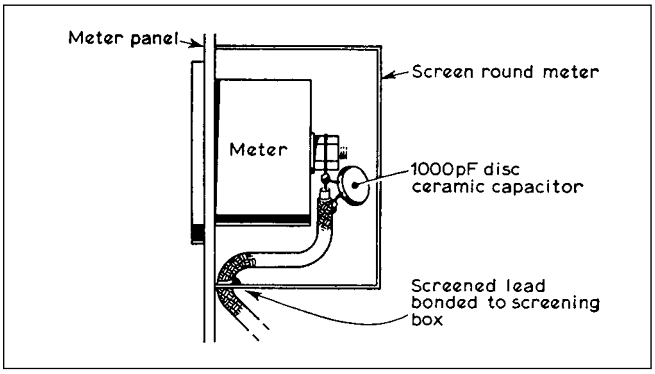


Fig 25.10: Screening and by-passing a meter in a transmitter

in a high-voltage supply where a large current could flow in the event of a short-circuit of the power supply. In these cases a rectifier diode should be used, such as the 1N400X or 1N540X series. The reverse current of these diodes may be a few microamps and, depending on the current to be measured, may have a slight effect on the sensitivity of the meter circuit.

Although diode protection should be applied as routine in order to safeguard instruments, it can cause some unusual effects if measurements are made with an AC signal imposed on a DC signal. This AC component, providing it is symmetrical, should not normally introduce any error but, if the AC is large enough to bring the diodes into conduction at the peak of the cycle, it introduces a dynamic shunt on the meter. This can be partly confusing when back-to-back diodes are used as the meter sensitivity will drop without any offset reading to warn what is happening. These effects are most likely to occur when measuring rectified mains or when RF is present.

Whenever a meter is to be used when RF may be present (this includes even a power supply output voltmeter) it is wise to shunt the meter with a capacitor, typically a 1000pF ceramic type - see **Fig 25.10**. In addition, if strong RF fields are likely to be present, eg in a transmitter, it would also be wise to shield the meter and possibly feed it via screened cable.

AC Measurements

If an alternating current is passed through a moving-coil meter there will normally be no deflection since the meter will indicate the mean value and, in the case of a waveform symmetrical about zero, this is zero. If, however, the AC is rectified so that the meter sees a series of half-sine pulses (full-wave rectification) it will indicate the mean value ($2/\pi$ or 0.637 of the peak value). Commercial instruments using moving-coil instruments for AC sine-wave measurements therefore incorporate a rectifier (see **Fig 25.11** for a typical arrangement) and the scale is adjusted to read RMS values (0.707 of the peak value). They will read incorrectly on any waveform that does not have these relationships. The moral is: *do not use the meter on any waveform other than a sine wave*. This arrangement is normally only used for voltage

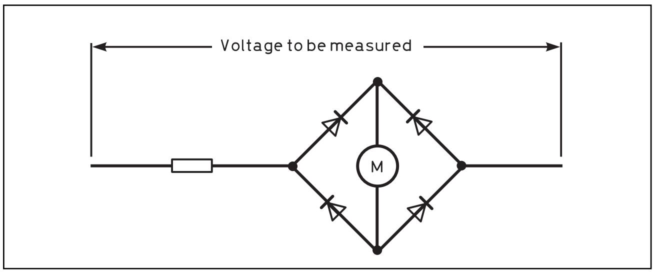


Fig 25.11: Typical arrangements for AC measurements

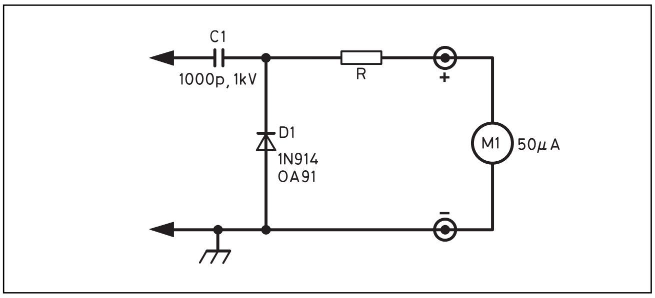


Fig 25.12: RF probe. For R = 270k + 12k, the meter scaling is 0-10V, and full-scale, power in 50 ohms is 2W. For R = 820 + 27k, the meter scaling is 0-30V, and full-scale, power in 50 ohms is 18W

measurements - AC current measurements pose additional problems and are not considered further. The typical frequency range extends to between 10Hz and 20kHz.

Moving-iron instruments, as previously mentioned, do respond to an alternating current and can be used for measurements without rectifiers. This type of meter unfortunately has a square-law characteristic and so the scale tends to be cramped at the lower end. Moving-iron meters normally have a full-scale reading of about 20% more than the normal value to be displayed. They are not used for multimeters.

Other AC measurements can be accomplished by means of electronic voltmeters or oscilloscopes.

ICs do exist (eg AD536, 636, 736, 737, SSM2110) which will provide the RMS of any waveform but their frequency range is limited.

RF Measurements

These probably pose the biggest problem: the circuit under test should not be loaded, capacitance has an increasing effect as frequency rises and the diodes used for rectification must handle the frequencies concerned. The diode characteristics required mean they have a relatively low reverse-voltage rating (1N914 is 100V, OA202 is 150V with slightly poorer RF capabilities) and the forward diode voltage drop. The approach in measuring RF voltages is to rectify as soon as possible and then use DC measuring circuits.

Fig 25.12 shows a typical probe for measuring RF voltages. Capacitor C1 provides DC isolation, D1 rectifies the signal and the resistor is used to convert what is essentially a peak reading to an RMS reading on the meter. For the 50μA meter it is possible to use an individual meter or the most sensitive range on many multimeters. If possible use the precautions for the meter as depicted on Fig 25.10. **Fig 25.13** shows the typical construction of a probe, the exact method being left to the ingenuity of the constructor. A scrap length of 15mm central heating piping may make a good tube. The probe should be useful for frequencies from 50kHz to about 150MHz with an accuracy of about ±10%.

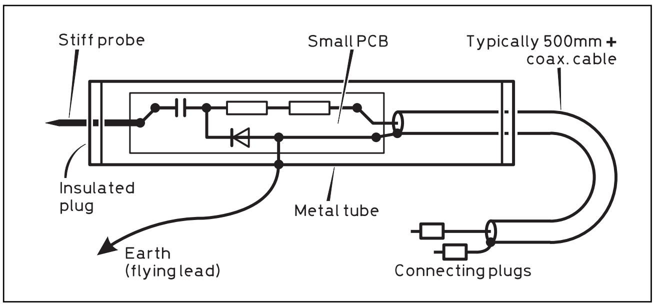


Fig 25.13: Typical construction of an RF probe

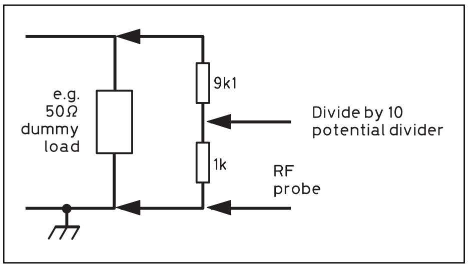


Fig 25.14: Suggested method for higher voltages

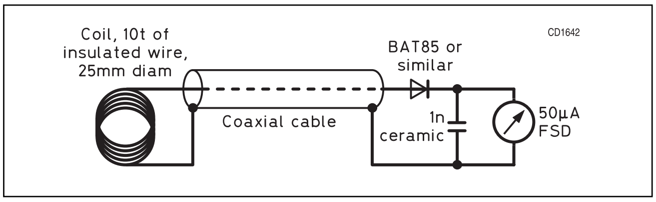


Fig 23.15: Pick-up loop with a diode

Because of reverse-voltage limitations of the diodes, it is necessary to make modifications to take higher voltage readings. **Fig 25.14** shows how a resistive potential divider can be used to effect a ten-fold reduction in voltage to be measured. The resistors should of course be suitable for RF and of adequate power rating. An alternative approach is to use several diodes in series but they will need equalising resistors across them.

An alternative to a probe that makes physical contact with a circuit is the use of a pick-up loop with a diode - see **Fig 25.15**. This cannot give a direct reading of voltage but is capable of indicating the presence of RF energy and may be useful for tuning purposes, ie looking for a maximum or minimum reading. The diode used should have a low forward drop - a germanium or Schottky type would be suitable. To minimise disturbing the RF circuit the pick-up coil should be placed for minimum coupling but give an adequate deflection on the meter.

DIGITAL METERS

The digital meter is fast becoming more common than analogue types and its price is now comparable in most instances. It provides a very accurate meter at reasonable price. Its disadvantages are that the smallest digit can only jump in discrete steps (hence digital) and that it requires a battery.

The digital meter (**Fig 25.16**) works by converting an input analogue voltage to a digital signal that can be used to drive either an LED (light-emitting diode) or LCD (liquid crystal display).

The conversion technique used is either an analogue-to-digital (A-D) converter or the dual ramp technique. A digital meter is often quoted as having, for example, a 3½ digit display. This means that it will display three digits 0 - 9, with the most significant being only a 0 (normally suppressed) or a 1, ie a maximum display showing 1999 as well as + or - signs.

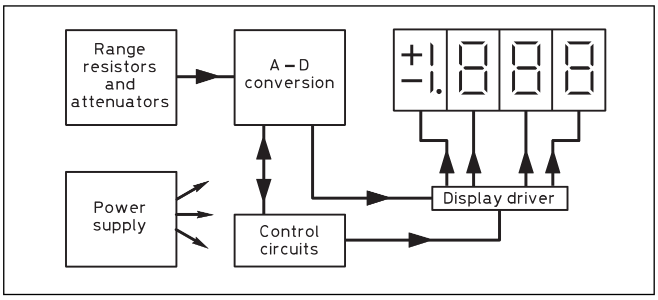


Fig 25.16: Block diagram of a digital meter (3½-digit display)

Fig 25.17: A typical digital panel meter

There are quite a few ICs made by various manufacturers that provide a basic digital voltmeter, external components being required for extending the range, over-voltage protection and displays. These ICs have outputs suitable for driving LEDs, LCDs or provide BCD outputs for further processing.

The digital meter is essentially a DC voltage measuring device (as opposed to the moving-coil meter which is current controlled). Hence all measurements to be made must be converted to a voltage.

The digital form of the multimeter is readily available at reasonable cost and it is not worth the exercise of making one of these meters. The approach here, as for the analogue meter, is to understand the basic principles and how to apply them to specific situations.

The Digital Panel Meter

The best approach for a digital display is to use a panel meter module (which includes the above ICs) and comes with a 3½ or 4½ digit display. These are relatively cheap and provide a good basis for making various types of metering system.

They are normally modules based on LCDs and either plug into a DIL socket or are on a small PCB. They have (typically) a full scale reading for a 199.9mV DC input, work over different supply ranges, from 5V to 14V (depends on model), consume very low current (eg 150-300μA on a 9V supply) and have an input resistance of at least 100MΩ. Because of this high input resistance they present virtually no loading on the circuit under test.

The panel meter itself will provide an accuracy of 0.1% or better but this does not take into account any external signal conditioning circuits such as amplifiers or attenuators. In addition to these parameters, some of the displays will also show units or prefixes such as μ, m, V, A, Ω, Hz etc (referred to as annunciators).

They can be purchased with and without backlighting; a typical meter is shown in **Fig 25.17**. The main design consideration in using these units is to get the parameter to be measured to a DC voltage in the range 0-199.9mV. This can include amplifiers, attenuators and rectifiers.

The following designs are based on the Anders OEM22 module which is readily available. The panel meter consists of a liquid crystal display driven by a 7136 IC which contains an A-D converter and LCD drivers. The unit can be driven from 5V (typically 5mA) providing two links are made on the board or direct from 9V (typically 500μA). It comes with a leaflet containing technical details [4] and the pin designations are marked on the board. The principles explained can, however, be applied to modules available from other manufacturers.

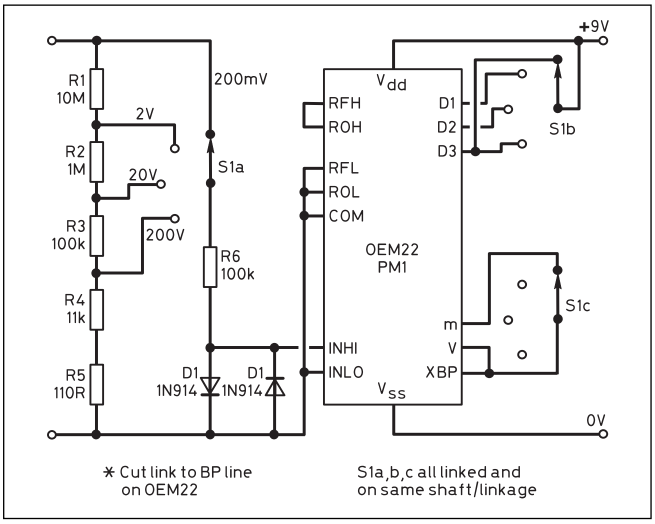


Fig 25.18: A practical digital voltmeter

PM1	Anders Panel Meter type OEM22	R2	1M, 0.5W, 1%
		R3	100k, 0.5W, 1%
SW1	Rotary switch 3p, 4w	R4	11k, 0.5W, 1%
D1,D2	1N914 or similar	R5	110R, 0.5W, 1%
R1	10M, 0.5W, 1%	R6	100k, 0.5W, 5%

Table 25.1: Components for the practical digital voltmeter

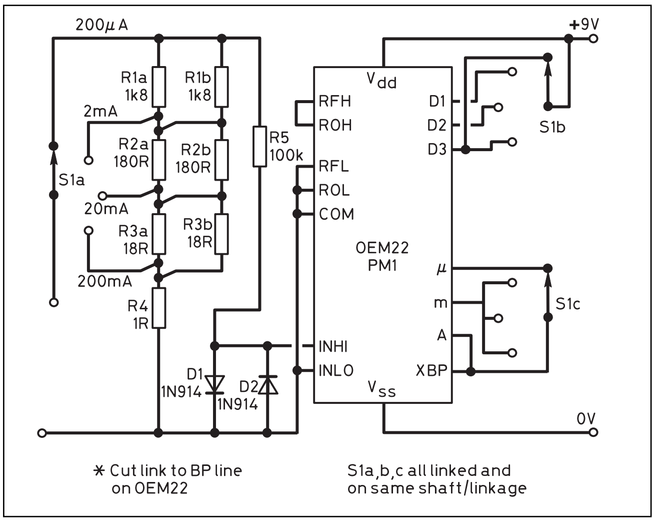


Fig 25.19: A practical digital ammeter

R1a, R1b	1k8, 0.5W, 1%	R5	100k, 0.5W, 5%
R2a, R2b	180R, 0.5W, 1%	PM1	Anders panel meter, type OEM22
R3a, R3b	18R, 0.5W, 1%		
R4	1R, 0.5W, 1%	SW1	Rotary switch, 3p, 4w
D1,D2	1N914 or similar		

Table 25.2: Components for the practical digital ammeter

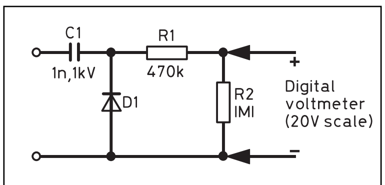


Fig 25.20: RF probe for digital voltmeter

Feature	Analogue Meter	Digital Meter
Operation	Current	Voltage
AC or DC	DC Moving Coil (AC with rectifiers) AC/DC Moving Iron	DC (AC with rectifiers or converters)
Display	Electro-mechanical	Semiconductor
Power supply required	None (taken from circuit under test)	DC supply
Best Sensitivity	50μA FSD typical	199.9mV typical
Circuit Loading	Depends on circuit and sensitivity of meter	Input >10MΩ, may affect high impedance circuit
RF interference	None	Possible due to internal oscillator

Table 25.3: Comparison of Analogue and digital meters

A Practical Digital Voltmeter

Fig 25.18 shows the arrangement for a digital voltmeter for DC voltage ranges of 200mV, 2V, 20V and 200V. The unit requires a 9V DC supply. Components are listed in **Table 25.1**.

Resistors R1 to R5 form a potential divider network with switch S1a selecting the correct input, ie the maximum voltage to the panel meter is to be 199.9mV. Resistor R6 and diodes D1/D2 provide protection for the panel meter should the wrong range (S1a) be inadvertently selected and introduce an error of less than 0.1%. S1b selects the position of the decimal point while S1c selects the annotation to be shown. The link *must* be cut to the BP line on the panel meter. Because the input resistance of the meter module is of the order of 100MΩ it represents negligible loading on the potential divider chain. The overall input resistance of the meter is about 10MΩ.

Sufficient information is provided for the reader to adapt this design to cope with other ranges. For inputs lower than 200mV, then an amplifier is required ahead of the meter input.

A Practical Digital Ammeter

This relies on measuring the voltage drop developed by the current to be measured passing through the measurement resistor, and it must be 200mV for full scale. Hence the circuit of **Fig 25.19** results in a meter measuring 200μA to 200mA in decade ranges.

Resistors R1 to R4 form the load across which the voltage is developed from the current being measured. Resistors R1 to R3 involve resistors in parallel to make up the correct value required. Switch S1a selects the input, S1b selects the decimal point positions and S1c selects the correct annotation for the range being used. The combination R5/D1/D2 provides protection for the panel meter input. The unit can be powered from a PP3 battery or equivalent. **Table 25.2** lists the components used.

RF Measurements

Similar problems arise for the digital meter as explained earlier for the analogue meter. A slight modification is made to the RF probe circuit and this is shown in **Fig 25.20**. This assumes the meter has a scale with 20V full scale and an input impedance in excess of 10MΩ. The resistors provide scaling from peak to RMS for a sine-wave input. The construction should be similar to that shown in Fig 25.13.

Comparison of Analogue and Digital Meters

Table 25.3 assumes that the analogue meter has no electronic circuit associated with it as this may alter its characteristics. It should also be borne in mind that the input to a digital meter might be affected by input amplifiers and attenuators.

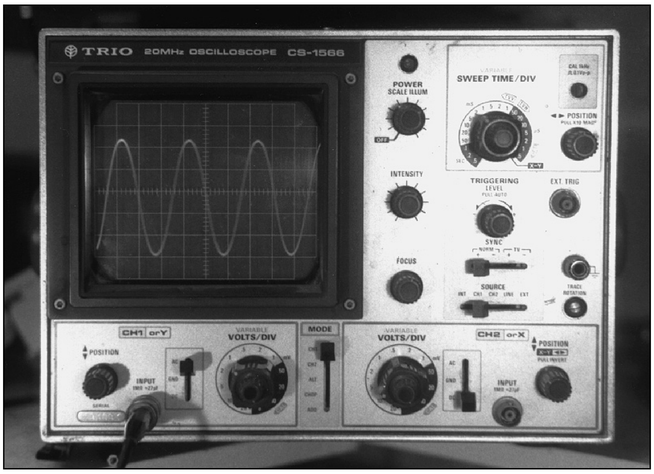


Fig 25.21: A dual trace oscilloscope

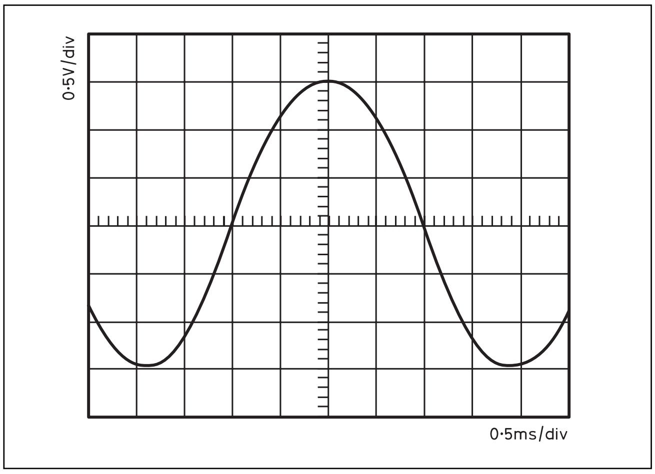


Fig 25.23: A typical displayed waveform

THE OSCILLOSCOPE

The oscilloscope (**Fig 25.21**) is a general-purpose instrument for examining electrical waveforms. It can be used for various sets of measurements depending on how it is has been set up. It is the intention of this section to explain briefly how an oscilloscope works and how it can be used for taking various measurements.

The Basic Oscilloscope

Some oscilloscopes can display a single trace whilst others can display two traces, or even more with adapters. The single-trace oscilloscope has a cathode-ray tube with a single electron gun firing at the phosphor.

The two-trace oscilloscopes fall into two categories - the dual-trace and the dual-beam. In the dual-trace oscilloscope there is a single electron gun but the control of it is split between the two traces to be shown - first one and then the other etc, but using the same timebase. In the dual-beam type there are two electron guns in the same cathode-ray tube which are independent of each other, hence the two beams can use different timebase settings.

Fig 25.22 shows the absolute basics for an oscilloscope. It consists of a display (usually a cathode-ray tube) which shows an electrical waveform. These signals have been processed in some way (eg amplified) for them to be suitable for display. The oscilloscope also contains an oscillator (or timebase) which causes the display beam to traverse the display face in the horizontal plane.

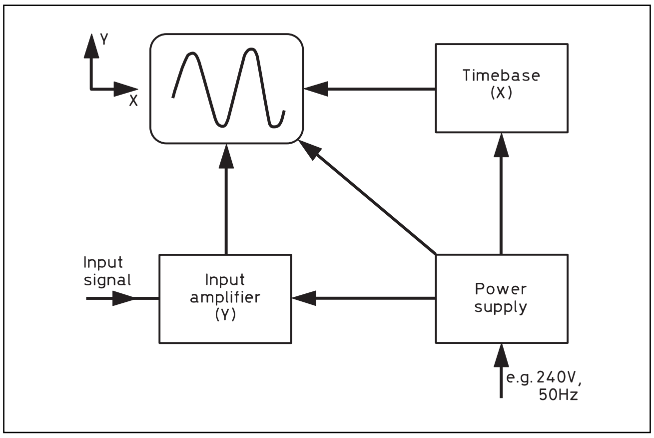


Fig 25.22: Basic block diagram of an oscilloscope

In addition a power supply is required to provide the amplifier and timebase with low-voltage supplies, and the tube with a high voltage (usually in the kilovolt region). The screen is split into two directions, the X (or horizontal) and Y (vertical direction).

Voltage Measurements

Consider the oscilloscope screen display as depicted in **Fig 25.23**. This is obviously a sine wave but what is its voltage? What voltage are we talking about? The easiest voltage measurement to take is the peak-to-peak value. The vertical displacement (Y) is 6 divisions. The setting of the Y-controls must also be taken into account; say these are at 0.5V/div. The peak-to-peak voltage is therefore 6 x 0.5 = 3V. The peak value is half of this, ie 1.5V; the RMS value is 0.7071 times this value, ie 1.06V.

Frequency Measurements

The method for making a frequency measurement is similar to the above, except that now the horizontal (X) axis is used with its setting. One problem with something like a sine wave is estimating a point on a curve and so it is an estimate only. Measure from like point to like point, eg the two negative peaks. The distance between the two negative peaks is 8 units. If the X or timebase setting is 0.5ms/div, then this represents a period of 8 x 0.5 = 4ms. The frequency is the reciprocal of this, ie 250Hz. It should be noted from this exercise that the period of a rectangular waveform is easier to estimate than that of a sine wave.

Equipment Limitations

The Y-amplifiers (plus the tube) limit the frequency response of the oscilloscope. This means that after a certain point the oscilloscope calibration is not valid, but comparative measurements can still be made above this point providing the frequency is not changed.

The capacitance of the oscilloscope and/or its probe may affect the circuit under test if the capacitance in the latter is of the same order as the oscilloscope input (20-40pF), eg in a tuned circuit.

The input voltage on an oscilloscope is normally quoted as x volts DC plus peak AC. Typical of these figures are 400V DC plus the peak AC signal that can be displayed. Exceeding this will damage internal components of the oscilloscope. Although a divide-by-10 probe can be used to extend the voltage range, these have a voltage limit (typically 600V DC) but may have to be derated as frequency rises - check the specification. The divide-by-10 probe will also reduce any loading effects on the circuit.

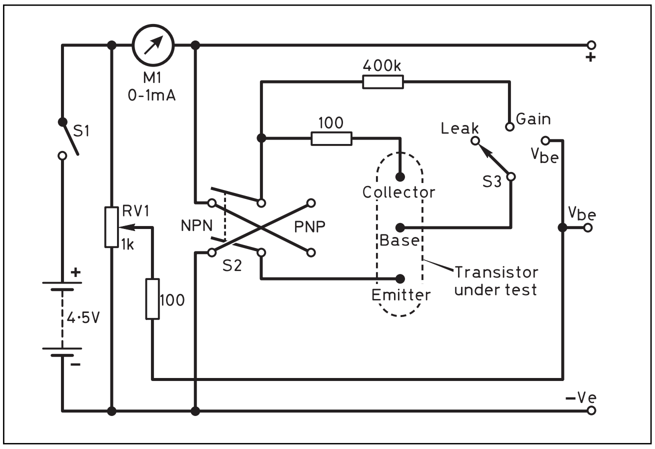


Fig 25.24: Simple transistor and diode tester

For most semiconductor applications the voltage limit never causes a problem, but with high-voltage valve circuits due regard must be paid to the limitations.

COMPONENT MEASUREMENTS

The cost of test equipment has decreased and the complexity of it has increased. It is now possible to purchase relatively cheap instruments to measure resistance, capacitance and inductance as well as testing transistors and diodes.

There are the standard type of analogue LCR meters which normally use a bridge technique for measuring impedance. It is possible to buy these but you may also find ex-commercial units at rallies.

A typical digital LCR tester will cost between £35 and £100 (2005 prices). However, you get what you pay for and the typical resolution for capacitance is 1pF and for inductance is 1µH. The measurement frequency depends on model and varies between about 1kHz and 200kHz. For resistance the minimum resolution is of the order of 1Ω.

The diode and transistor testers again vary in price and similar to the LCR meters. They will certainly test the basic operation of the device but they may not, for example, test the high frequency response.

Below are two circuits that can be built for diode/transistor testing and capacitance measurement.

A DIODE AND TRANSISTOR TESTER

The circuit of **Fig 25.24** shows a simple tester which will identify the polarity and measure the leakage and small-signal gain of transistors plus the forward resistance of diodes.

Testing Transistors

To check the DC current gain h_{FE} (which approximates to the small signal current gain h_{fe} or ß), the transistor is connected to the collector, base and emitter terminals and S2 switched for the transistor type. Moving switch S3 to the GAIN position applies 10mA of base current and meter M1 will show the emitter current. With S3 at the LEAK position, any common-emitter leakage current is shown, which for silicon transistors should be barely perceptible. The difference between the two values of current divided by 10mA gives the approximate value of h_{FE} + 1 which is close to h_{fe} for most practical purposes.

A high value of leakage current probably indicates a short-circuited transistor, while absence of current in the GAIN position indicates either an open-circuited transistor or one of reversed polarity. No damage is done by reverse connection, and PNP and NPN transistors may be identified by finding the polarity which gives normal gain.

With S3 in the V_{be} position, the base-emitter voltage is controlled by RV1 which should be near the negative end for NPN and near the positive end for PNP. V_{be} may be measured by a voltmeter connected between the terminal marked 'V_{be}' and either the positive or negative rail depending on the polarity of the device. This test position may be used for FETs but only positive or zero bias is possible.

Testing Diodes

The forward voltage drop across a diode may be measured by connecting it across the terminals marked '+' and 'V_{be}' with a voltmeter in parallel. The forward current is set by RV1.

Diodes may be matched for forward resistance and, by reversing the diode, the reverse leakage can be seen (which for silicon diodes should be barely perceptible). The value of forward voltage drop can be used to differentiate between germanium and silicon diodes.

The unit can also be used to check the polarity of LEDs as the maximum reverse voltage of 4.5V is hardly likely to damage the device (note: the reverse voltage applied to an LED should not exceed 5V). For this test RV1 should be set to about mid-position.

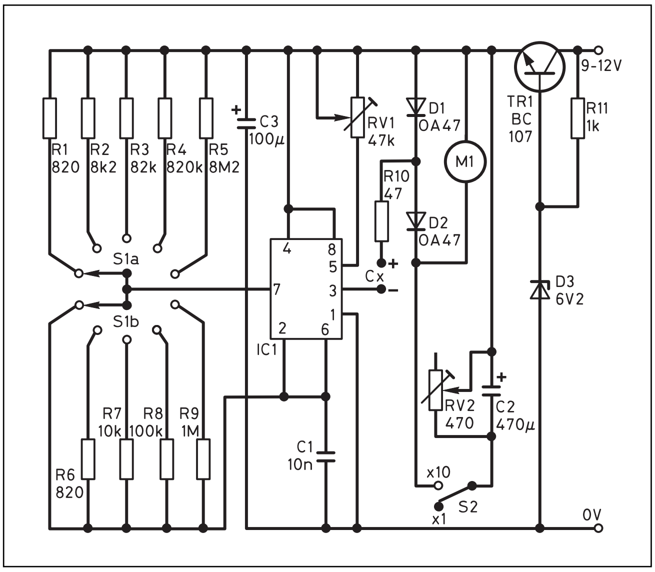


Fig 25.25: Circuit of the linear-scale capacitance meter

A LINEAR-SCALE CAPACITANCE METER

This instrument is based on the familiar 555 timer and the circuit is shown in **Fig 25.25**. It has five basic ranges with a x10 multiplier. This gives the equivalent of six ranges of full scale values 100pF, 1nF, 10nF, 100nF, 1µF and 10µF.

The meter works by charging the unknown capacitor Cx to a fixed voltage and then discharging it into a meter circuit. The average current is proportional to the capacitance and hence a direct reading on the meter. If measuring small electrolytic capacitors please observe the polarity. The unit requires a low current 9V DC supply.

Construction

A components list is given in **Table 25.4**. The layout of the components is not critical (**Fig 25.26**). A PCB pattern (**Fig 25.27**) is given in Appendix B. The builder can either make a box or, as is more usual, purchase one of the cheaper plastic types.

R1, R6	820R	RV2	Single turn trimmer, 470R, 0.5W
R2	8k2	C1	Polystyrene, 10nF, ±1%
R3	82k	C2	Electrolytic, 470μF, 16V
R4	820k	D1,D2	OA47 or BAT85
R5	8M2	D3	6V2, 400mW zener
R7	10k	TR1	BC107 or similar NPN
R8	100k	IC1	555 Timer
R9	1M	M1	Moving Coil, 50μA FSD
R10	47R	S1	2p, 6way, Rotary, PCB mounting
R11	1k	S2	PCB Mount SPCO switch
RV1	Single turn trimmer, 47k, 0.5W		

Resistors are metal film type, MRS25, 1% unless specified otherwise.

Table 25.4: Components for the linear-scale capacitance meter

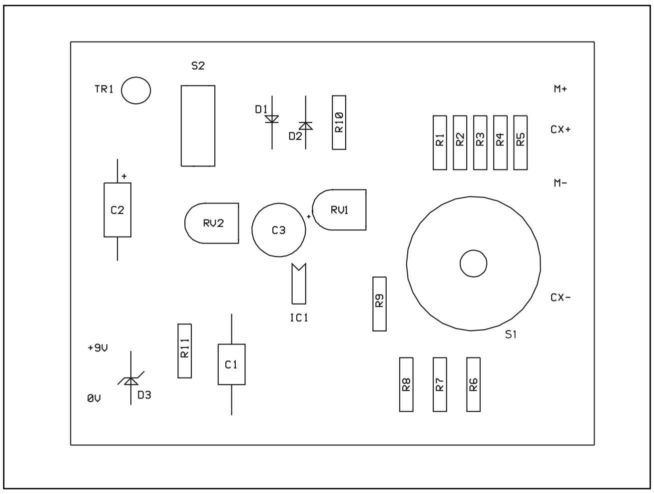


Fig 25.26: Component layout for the linear-scale capacitance meter (not to scale). The PCB layout can be found in Appendix B

Calibration

Calibration may be carried out on any range; if possible obtain 100pF, 1nF and 10nF capacitors with ±1% tolerance. With the range switch set to position 2 and multiplier switch S2 in the x1 position, connect the 1nF capacitor. Adjust RV1 for full-scale deflection. Switch to the x10 position of S2 and adjust RV2 for a meter reading of 0.1. Use the other capacitors to check the other ranges. Calibration is now complete.

Warning: If a large-value capacitor is to be measured, the meter will be overloaded.

IMPEDANCE MEASUREMENTS

Impedance measurement for the radio amateur probably means antenna or feed point impedance. The following two items of home-built equipment allow impedance of various circuits to be analysed. The first one merely provides the impedance value. It does not give the resistive and reactance components separately, the second one will provide this in series form. A previous design in [1] (page 70) will give the equivalent parallel components of an unknown impedance.

A good design was produced by G3BIK and is described in *RadCom*, Dec 1999 with full constructional details; this has an internal VFO and does not use the noise generator/communications receiver as described here.

Proprietary equipment can be purchased [2] which is designed to cover the amateur bands up to 470MHz and would be useful for other frequencies as well. However they are likely to set you back about £350. The alternative is to look for second-hand commercial gear or build your own.

Further information can be obtained from more specialised books on amateur antennas and experimentation.

An RF Impedance Bridge

The need for an instrument which will measure impedance is felt at some time or other by every experimenting amateur. The instrument normally used is the full RF bridge, but commercial RF bridges are elaborate and expensive. On the other hand it is possible to build a simple RF bridge which, provided the limitations are appreciated, can be inexpensive and a most useful adjunct in the amateur workshop. In fact it is essential if experiments with antennas are undertaken.

The instrument described here will measure impedances from 0 to 400Ω at frequencies up to 30MHz. It does not measure reactance or indicate if the impedance is capacitive or inductive. A good indication of the reactance present can be obtained from the fact that any reactance will mean a higher minimum meter reading.

Circuit description

There are many possible circuits, some using potentiometers as the variable arm and others variable capacitors, but a typical circuit is shown on **Fig 25.28**. The capacitors have to be differential in action, mounted in such a way that as the capacitance of one decreases the capacitance of the other increases. The capacitors should be of the type which have a spindle protruding at either end so that they can be connected together by a shaft coupler. To avoid hand-capacitance effects, the control knob on the outside of the instrument should be connected to the nearest capacitor by a short length of plastic coupling rod. These capacitors form two arms of the bridge, the third arm being the 100Ω non-inductive resistor and the fourth arm the impedance to be measured. Balance of the bridge is indicated by a minimum reading on the meter M1.

Construction

Construction is straightforward, but keep all leads as short as possible. The unit should be built into a metal box and screening provided as shown in Fig 25.28.

Signal source and calibration

The instrument can be calibrated by placing across the load terminals various non-reactive resistors (ie not wire-wound) of known value. The calibration should preferably be made at a low

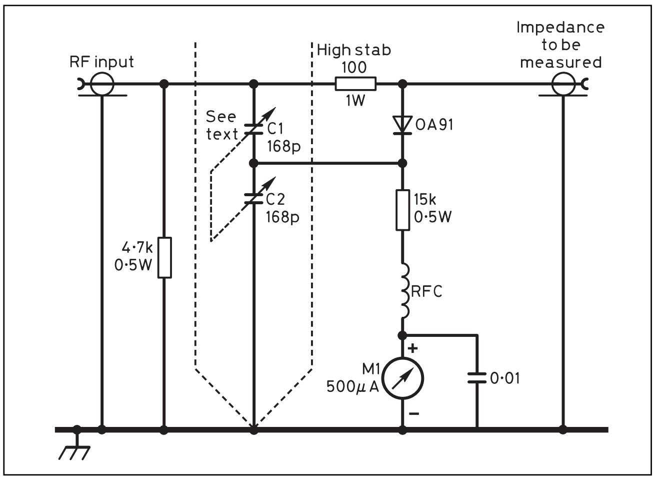


Fig 25.28: Simple RF bridge. Note that a BAT85 diode may be used instead of the OA91

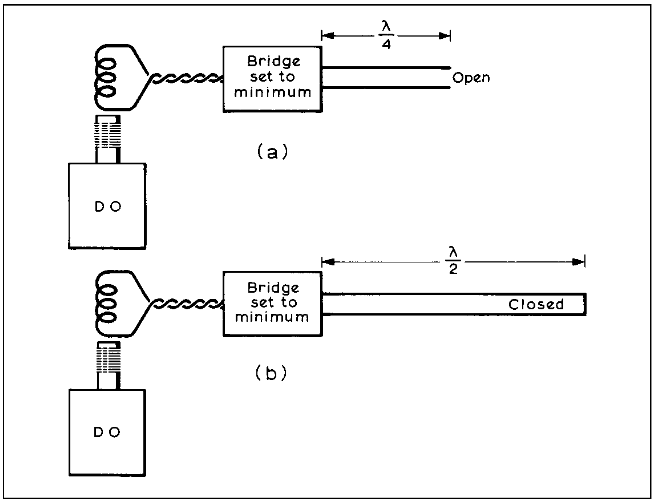


Fig 25.29: Use of the RF bridge with a dip oscillator

frequency where stray capacitance effects are at a minimum, but calibration holds good throughout the frequency range. In using the instrument, it should be remembered that an exact null will only be obtained on the meter when the instrument has a purely resistive load. When reactance is present, however, it becomes obvious from the behaviour of the meter; adjusting the control knob will give a minimum reading but a complete null cannot be obtained.

The RF input to drive the bridge can be obtained from a dip oscillator, signal generator or low-power transmitter capable of giving up to about 1W of signal power. The signal source can be coupled to the bridge by a short length of coaxial cable directly or via a link coil of about four turns as shown on Fig 25.29.

If using the dip oscillator, care should be exercised in order to not over-couple with it as it may pull the frequency or, in the worst circumstances, stop oscillating. As the coupling is increased it will be seen that the meter reading of the bridge increases up to a certain point, after which further increase in coupling causes the meter reading to fall. A little less coupling than that which gives the maximum bridge meter reading is the best to use. The bridge can be used to find antenna impedance and also used for many other purposes, eg to find the input impedance of a receiver on a particular frequency.

Some practical uses

One useful application of this type of simple bridge is to find the frequency at which a length of transmission line is a quarter- or half-wavelength long electrically. If it is desired to find the frequency at which the transmission line is a quarter-wavelength, the line is connected to the bridge and the far end of it is left open-circuit. The bridge control is set to zero ohms. The dip oscillator is then adjusted until the lowest frequency is found at which the bridge shows a sharp null. This is the frequency at which the piece of transmission line is one quarter-wavelength. Odd multiples of this frequency can be checked in the same manner. In a similar way the frequency at which a piece of transmission line is a half-wavelength can also be found but in this case the remote end should be a short-circuit.

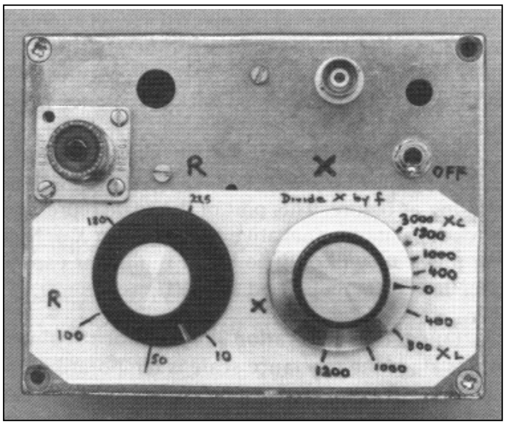
The bridge can also be used to check the characteristic impedance of a transmission line. This is often a worthwhile exercise, since appearances can be misleading. The procedure is as follows.

1. Find the frequency at which the length of transmission line under test is a quarter-wavelength long. Once this has been found, leave the oscillator set to this frequency.
2. Select a carbon resistor of approximately the same value as the probable characteristic impedance of the transmission line. Replace the transmission line by this resistor and measure the value of this resistor at the preset frequency. (Note: this will not necessarily be identical with its DC value).
3. Disconnect the resistor and reconnect the transmission line. Connect the resistor across the remote end of the transmission line.
4. Measure the impedance now presented by the transmission line at the preset frequency. The characteristic impedance (Z_0) is then given by:

$$Z_0 = \sqrt{Z_s \times Z_r}$$

where Z_s is the impedance presented by the line plus load and Z_r is the resistor value.

Fig 25.30: The G3ZOM noise bridge



An RF Noise Bridge

The noise bridge uses the null method. Wide-band RF noise is used as a source, and a receiver is used as frequency-selective null detector. Noise bridges do not have a reputation for accuracy but they are small and convenient to use. The accuracy and the depth of the null depends mostly on the layout of the bridge network and the care taken in balancing out the bridge.

The description of a noise bridge that follows is by G3ZOM. The front panel is illustrated in Fig 25.30, the circuit diagram is shown in Fig 25.31 and parts list in Table 24.5. Fig 25.32 shows the layout of the bridge components. It can be used to measure impedance in terms of series resistance and reactance, within the frequency limits 1 to 30MHz. The useful range is approximately 0 to 200Ω. The reactance range is dependent on frequency and the capacitance swing of the variable capacitor used in the variable arm of the bridge. As a rough guide, using the suggested 250pF variable:

R1	220R	ZD1	5V6 Zener, 400mW
R2	100k	C1, C2	10n
R3	1k2	C3,C4	10n
R4	68k	C5	82p
R5	680R	C6	10µ, 16V electrolytic
RV1	220R pot, carbon	VC1	250p variable
TR1,TR2	BC182, 2N3903 or similar	S1	SPCO switch
T1	Dust iron core FT50-43, FT50-5 or similar. See Fig 25.31 for winding details		

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

Table 25.5: RF Noise bridge components list

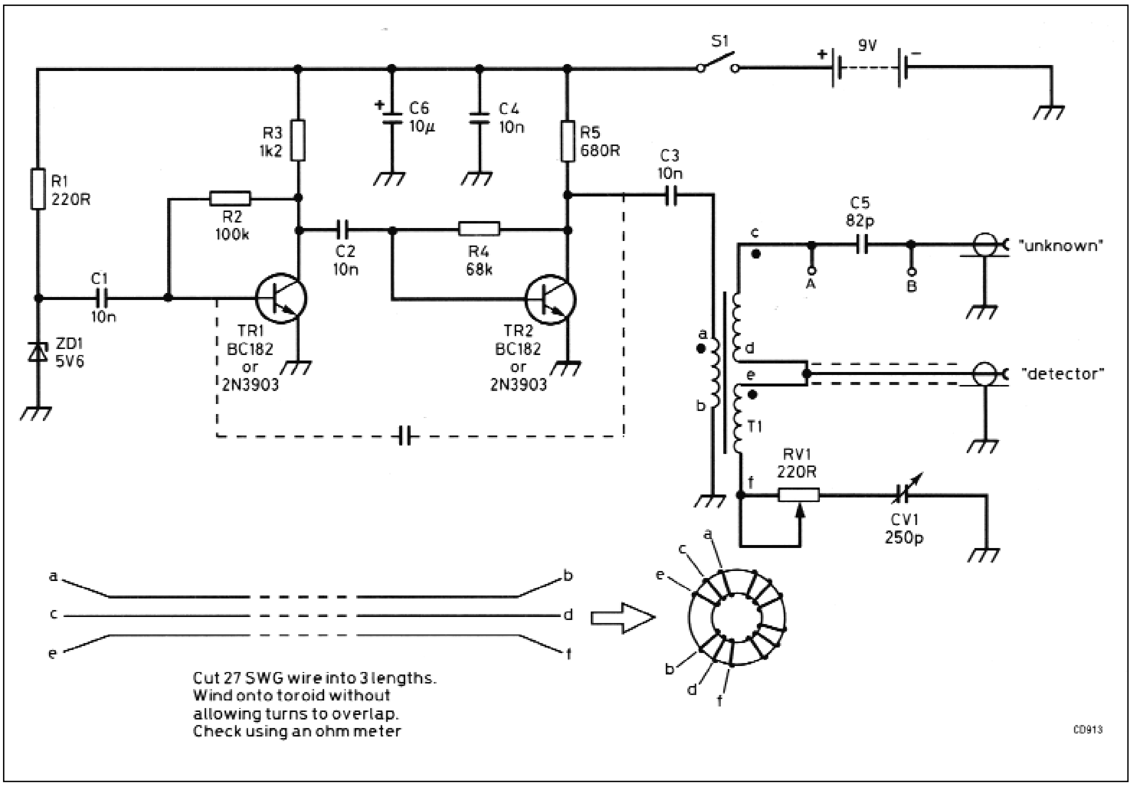


Fig 25.31: Circuit of the G3ZOM noise bridge

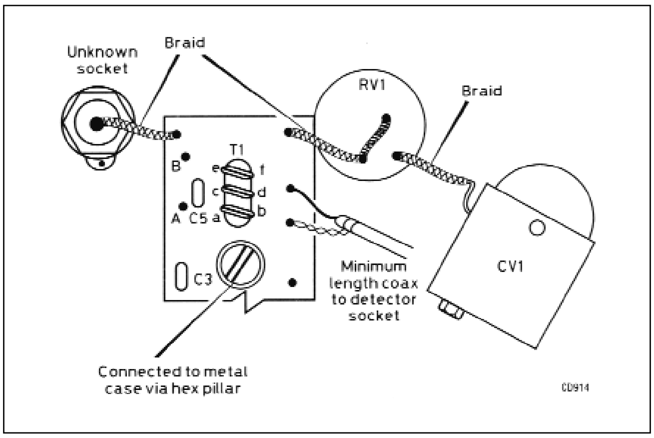


Fig 25.32: Layout of bridge components

At 1MHz: 5000Ω capacitive to 1200Ω inductive
At 30MHz: I70Ω capacitive to 40Ω inductive

Balancing the bridge

The bridge has to be balanced to obtain a reasonable calibration over the intended range. Connect a suitable receiver to the detector socket and a non-reactive (carbon or metal film) across the UNKNOWN socket. The resistor leads must be kept as short as possible to reduce the unwanted reactance to a minimum (important at the high-frequency end of the range). Set RV1 to maximum resistance and CV1 to maximum capacitance (fully meshed). Tune the receiver to around 14MHz and switch the noise bridge on. A loud 'rushing' noise should be heard from the receiver, and the S-meter (if fitted) should show a good signal strength.

By listening to the noise level, and observing the S-meter (if fitted), adjust CV1 to obtain a decrease in volume (a null). Then adjust RV1 for a deeper null. Repeat these two adjustments until the deepest null has been reached. Temporarily mark the null positions of RV1 200Ω and CV1 to zero.

Set the receiver within the 1 to 2MHz range and repeat the nulling procedure. This time the null will be much sharper so careful adjustment is needed. The positions of RV1 and CV1 should be the same or close to those obtained previously. If not, the wiring around the bridge components is probably too long. Short wiring lengths are essential.

Repeat the procedure again with the receiver set to around 30MHz. This time the null will he much wider. The position of RV1 should again be close to that obtained previously, but it will probably be found that the position of CV1 is somewhat different to before. If this is the case, the situation can be remedied by adding a small-value balancing capacitor between pin A and chassis in Fig 25.31.

Both the value and the position of this balancing capacitor will need to be determined by trial and error. Try, say, 10pF to pin A and repeat the nulling procedure at 2 and 30MHz. If the situation is worse than before, try a 10pF capacitor between the chassis and RV1 (where it connects to T1). One or the other position will result in an improvement which is worth the effort to obtain reliable measurements. Even greater accuracy can be obtained by adding compensating inductance to the bridge but this has not been found necessary to date.

The instrument shown in Fig 25.31 has a better than 50dB null at 10MHz.

Calibration

Calibration can now be carried out. Tune the receiver to around 14MHz.

Resistance scale

Connect suitable resistors, one at a time across the UNKNOWN socket, nulling the bridge and marking the resistance values of the test resistors on the RV1 scale. CV1 should remain at its zero position. The resistance scale should be fairly linear, allowing simple interpolation of unmarked values.

Reactance scale

Connect a 51Ω resistor across the UNKNOWN socket, using short leads. Null the bridge and mark the position CV1 as zero '0'.

Leave the 51Ω resistor in place and connect a selection of capacitors across C5 (use terminals A and B), nulling the bridge each time and marking the reactance scale with the capacitance value. This part of the scale represents series inductance (positive reactance or X_L).

With the 51Ω resistor still in place, connect a series of capacitors across CV1, again nulling and marking as before. This part of the scale represents series capacitance (negative reactance, or X_C). Note that the scale will only be linear if a linear capacitance law variable is used for CV1.

Reactance scale calibration - capacitance or ohms?

At this stage, the reactance scale is temporarily calibrated in capacitance. Most published designs leave the reactance scale calibrated this way and use either a graph or a formula to make the conversion to the required reactance value in positive or negative ohms. You can use either method of calibration, using the conversion graph of **Fig 25.33** or the formula -

$$X = \frac{10^6}{2\pi f} \times \frac{S}{C_5(S + C_5)}$$

where X is the reactance in ohms, f is the frequency in MHz, S is the scale reading (+ or -) and C5 = 82 (the value of C5 in pF).

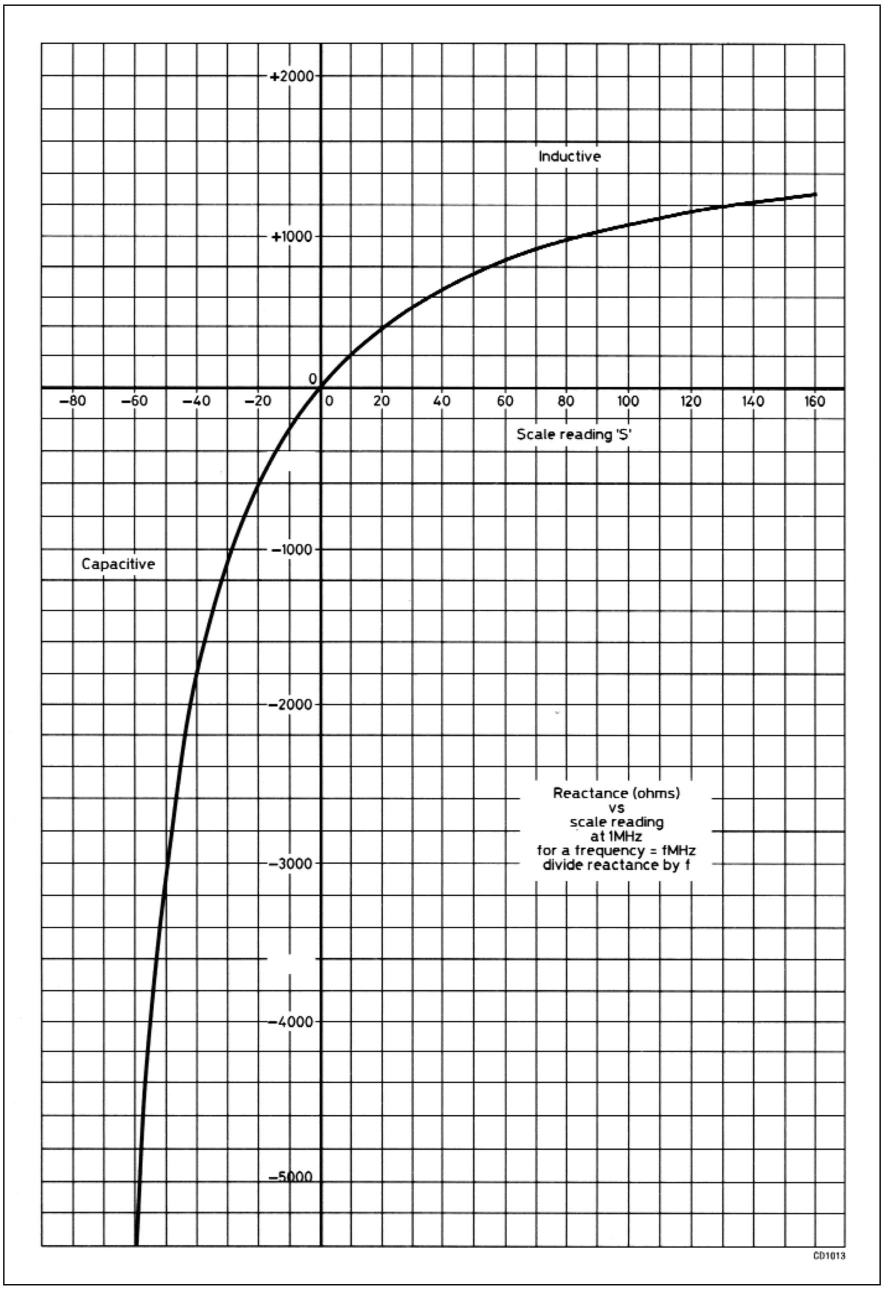


Fig 25.33: Calibration graph for converting capacitance value s to reactance (+ or - j)

Using the noise bridge

This bridge, in common with all other impedance measurement bridges, measures the impedance presented to the UNKNOWN socket. This may not be the same as the antenna impedance because of the impedance transformation effect of the coaxial cable connecting the antenna to the bridge.

RV1 and CV1 are then varied alternately to obtain the best null. The equivalent series resistance is obtained directly from the RV1 scale.

Impedance at the UNKNOWN socket is measured by connecting the noise bridge as shown in **Fig 25.34**. The receiver or transceiver is tuned to the measurement frequency and the R and X controls adjusted for minimum noise. These controls interact and the sharpest dip must be found by trial and error. Antenna impedance measurements can be accomplished in one of two ways:

1. At the transmitter end of the feeder, **Fig 25.34(a)**. By using a multiple of λ/2 at the frequency of measurement, the antenna feed impedance is reflected back to the transmitter end of the feeder. The disadvantage of this method is that the antenna matching network (eg gamma match) is at the antenna, remote from the impedance measurement, making the method rather cumbersome.
2. At the antenna end of the feeder, **Fig 25.34(b)**. The adjustment, and the measurement of the

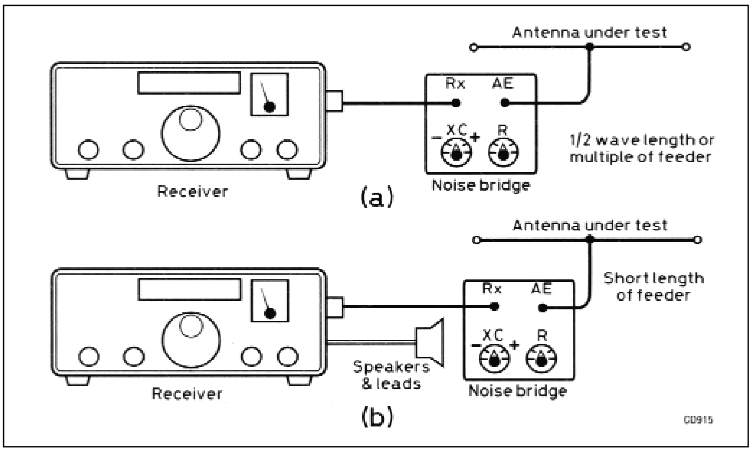


Fig 25.34: Noise bridge and receiver connections for antenna impedance measurements

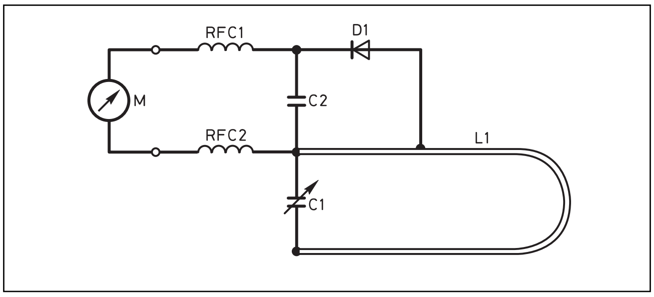


Fig 25.35: A simple absorption wavemeter for 65-230MHz

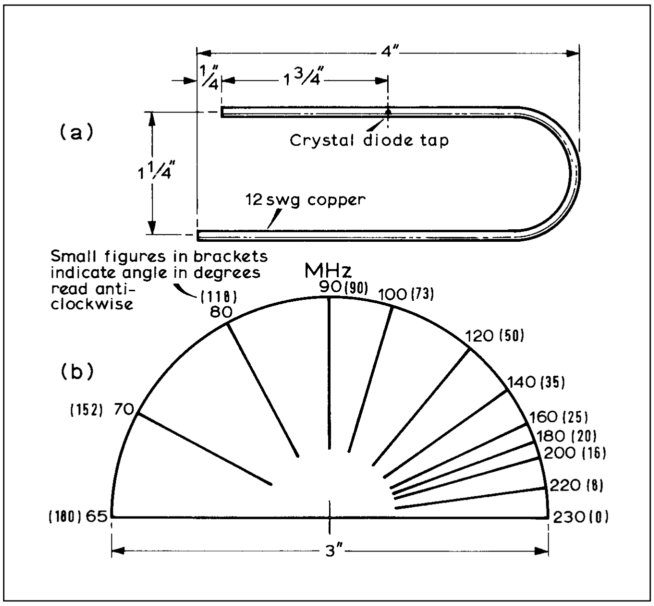


Fig 25.36: Constructional details of simple absorption wavemeter: (a) inductor; (b) dial plate. ¼in = 6.3mm, 1¼in = 31.8mm, 1¾in = 44.5mm, 3in = 76.2mm, 4in = 101.6mm

RFC1,2	80t of 40SWG ECW wound on 10k, 0.5W resistor	M1	1mA FSD or better
		C1	4-50p, Jackson C804 or equivalent
L1	See Fig 25.36(a)	C2	470p ceramic
D1	OA91, BAT85 or similar		

Table 25.6: Simple absorption wavemeter components list

results of the adjustment, is far more convenient. However, the method is limited to situations where there is access to the antenna in situ. A further disadvantage is that the noise null detector, the receiver, also has to be close at hand, which may be rather inconvenient 20m up a mast or on the roof of a house. The problem can be overcome by leaving the receiver in the shack. A small speaker or a pair of headphones can be connected to the output of the receiver via another feeder or a couple of wires from the rotator cable. The feeder length is immaterial. *Make sure that the receiver/headphone arrangement is earthed to prevent an electric shock hazard.* Alternatively, a pair of low-cost PMR446 handheld transceivers may be used to relay the audio.

A SIMPLE ABSORPTION WAVEMETER FOR 65-230MHz

The absorption wavemeter shown in Fig 25.35 is an easily built unit covering 65-230MHz. For a lower-range unit the dip oscillator described in the next section can be used.

Construction is straightforward and all the components, apart from the meter, are mounted on a Perspex plate of thickness 3 or 4mm and measuring 190 x 75mm. Details of the tuned circuit are shown on Fig 25.36(a) and should be closely followed. The layout of the other components is not critical provided they are kept away from the inductor. A components list is given in Table 25.6.

For accurate calibration, a signal generator should be used but, provided the inductance loop is carefully constructed and the knob and scale are non-metallic, the dial markings can be determined from Fig 25.36(b).

In operation the unit should be loosely coupled to the circuit under test and the capacitor tuned until the meter indicates resonance (a maximum). For low-power oscillators etc a more sensitive meter should be used (eg 50µA or 100µA).

The wavemeter can also be used as a field strength indicator when making adjustments to VHF antennas. A single-turn coil should be loosely coupled to the wavemeter loop and connected via a low-impedance feeder to a dipole directed towards the antenna under test.

DIP OSCILLATORS

It is possible to buy these units and radio rallies would be a good starting point. Alternatively one can build them - see later and reference [1] and [5].

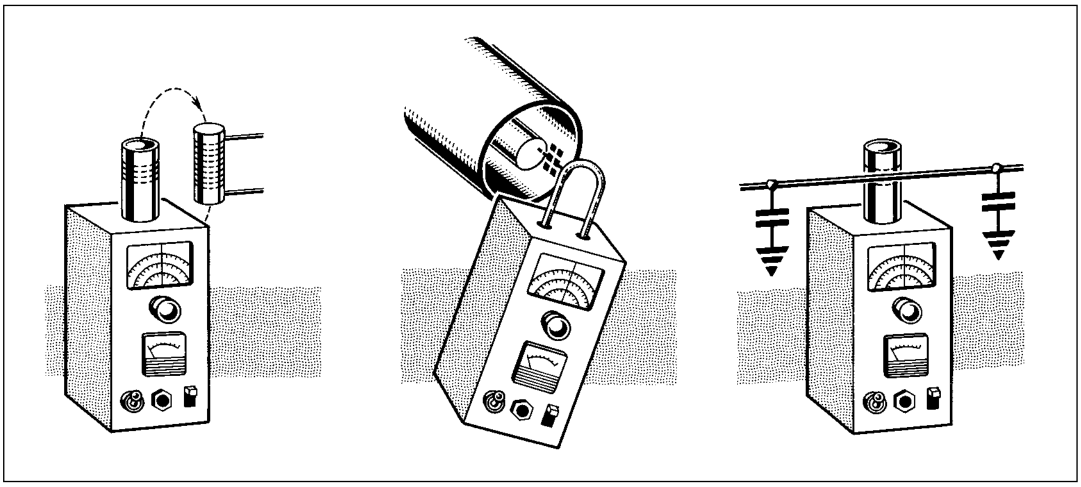


Fig 25.37: Using a dip oscillator

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 are shown diagrammatically in **Fig 25.37**.

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.

To determine unknown capacitor						
Known L µH	1.0	2.0	4.7	6.8	10	22
C pF is	25330÷F²	11513÷F²	5389÷F²	3725÷F²	2533÷F²	1151÷F²
To determine unknown inductor						
Known C pF	10	22	33	47	68	100
L µH is	2533÷F²	1151÷F²	768÷F²	539÷F²	373÷F²	253÷F²

Table 25.7: Determination of L and C using known values

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 a 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 can be 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.

Measurement of L and C

The following is by G3BIK [6] and describes a method of measuring L (inductance) and C (capacitance).

The dip oscillator provides a quick and easy means of checking (to a degree of accuracy acceptable for experimental purposes) the inductance value of coils in the microhenry (µH) range and capacitors in the picofarad (pF) range, such as are commonly used in radio circuits. This can be very useful, for example, when constructing an ATU, a crystal set, a short-wave receiver, a VFO or a band-pass filter for a direct conversion receiver.

For this purpose the following are kept with the dip oscillator:

- two fixed value RF coils of known inductance - 4.7µH and 10µH
- one capacitor each of 47pF and100pF

The choice is yours and you can decide to keep several of each to be selected from **Table 25.7**. A personal choice of coil-type is the moulded RF choke (Maplin) or RF inductor (Mainline or RS). These are axial-leaded, ferrite based, encapsulated, easy to handle, and readily available at low cost in a range of fixed-value microhenries. The capacitors are 5% tolerance polystyrene, also axial-leaded.

To determine or verify the value of either an RF coil or a capacitor, simply connect the unknown component in parallel with the appropriate known component to form a parallel LC tuned circuit, ie an unknown L in parallel with a known C (or vice versa), then use the dip oscillator to determine the resonant frequency of the parallel LC circuit. The value of the unknown component can then be obtained easily to an acceptable approximation, by using the relevant formula from **Table 25.7** and a pocket calculator. The formulas were derived from the accepted formula for the resonant frequency of a parallel tuned circuit-

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

(f in Hz, L in Henries and C in Farads)

NOTE: in **Table 25.7**, F is the frequency in MHz as given by the dip oscillator.

Example 1: An unknown capacitor in parallel with an known 10µH inductor, produces a dip at 6.1MHz, hence F = 6.1. From **Table 25.7**, the value of the unknown capacitor is given by:

C pF = 2533 ÷ F²

= 2533 ÷ 6.1 ÷ 6.1

= 68pF

Example 2: An unknown coil in parallel with an known 47pF capacitor, produces a dip at 12.8MHz, hence F = 12.8. From **Table 25.7**, the value of the unknown inductance is given by:

L µH= 539 ÷ F²

= 539 ÷ 12.8 ÷ 12.8

= 3.3µH

Bear in mind that because the accuracy of results relies upon the frequency as derived from the dip oscillator, it would be sensible to keep the coupling between the LC circuit and the dip oscillator as loose as possible, consistent with an observable dip. This minimizes pulling of the dip oscillator frequency. Also, rather than relying upon the frequency calibration of the dip oscillator itself, it might be useful to monitor the frequency on an HF receiver or a digital frequency meter.

A final point worth considering is that each fixed-value inductor of the type mentioned might have its own self-resonant frequency, but these would typically lie above the HF range so should not be a problem. For example, the self-resonance of the selected 10µH inductor is about 50MHz and that of 4.7µH is about 70MHz. You could quickly and simply find out the self-resonant frequency of an inductor, by taping it to each of the dip oscillator coils in turn and tuning across the full frequency span.

It is best to make L and C measurements at frequencies much lower than the self-resonant frequency of your chosen test-inductor, but perhaps better be safe than sorry and stick with the lower µH values if your interest lies between 1.8 and 30MHz.

Tone Modulation

The following is also by G3BIK [6].

Sometimes it is also useful to be able to hear an audio tone when using the GDO as an RF signal source in association with a radio receiver.

If your dip oscillator does not have tone modulation, you might like to construct the simple add-on 1kHz audio oscillator circuit shown in **Fig 25.38**. It uses a unijunction transistor, the frequency of oscillation being given approximately by 1/(R1 x C1).

The 1kHz tone output connects via C2 to the positive supply line of the dip oscillator, which it modulates. R2 acts as the modulator load and its value helps to determine the level of modulation. This

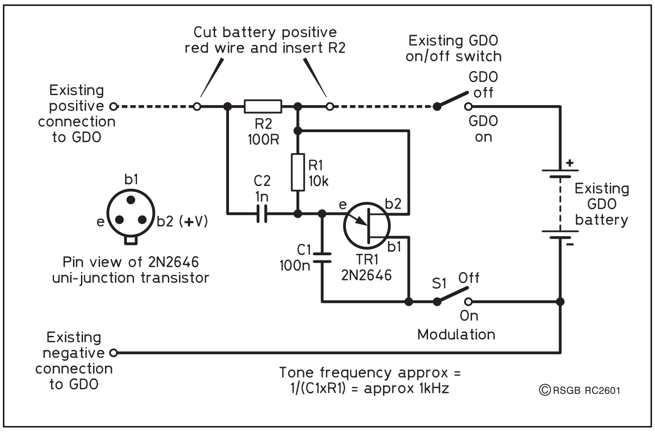


Fig 25.38: Add-on audio 1kHz oscillator for a dip oscillator (sometimes referred to as a GDO)

produces a simple but effective tone modulation of the dip oscillator's RF signal, which can be heard on an AM or an FM receiver.

The method of construction is really left to the builder but it can be fitted on a small piece of strip board without having to cut copper tracks. The finished board might conveniently mount onto one of the dip oscillator meter terminals, provided care is taken to isolate the copper-tracks from the terminal.

Use as an Absorption Wavemeter

A dip oscillator may be used as an absorption wavemeter by switching off the oscillator's power supply and then using in the normal way. In this case, power has to be applied to the circuit under test. Resonance is detected by a deflection on the meter due to rectified RF received. It should be noted that an absorption wavemeter will respond to a harmonic if the wavemeter is tuned to its frequency.

A SIMPLE DIP OSCILLATOR [2]

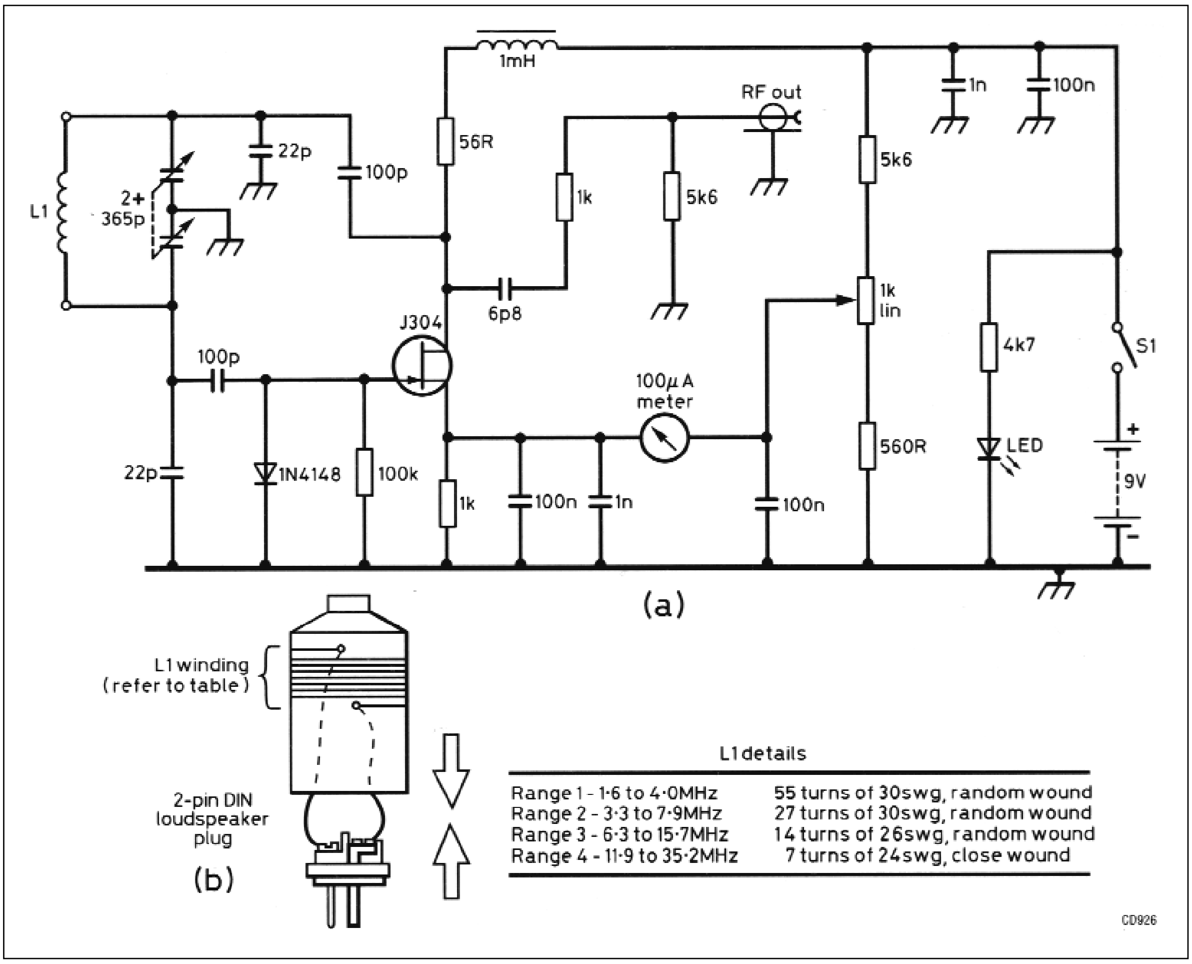
The best of the simple solid-state dip oscillators so far discovered were two very similar designs published in the 'Technical Correspondence' column of *QST* [7, 8]. The circuit of **Fig 25.39** is a variation of these two designs, plus improvements incorporated by G3ZOM.

This circuit does not measure gate current directly; instead it measures the total current through the FET. This is large compared with that flowing in a base or gate of a solid-state oscillator. However, the variation of current through resonance is only a small part of the total current through the FET. The dip is enhanced by offsetting the meter reading using a potentiometer in a bleeder network. This is set so that the meter reads about 75% FSD) when the instrument is not coupled to a load.

This design does not perform very well on the VHF bands with the circuit values shown and the dip tends to reverse if the coupling is too tight. This suggests that reducing the 100pF capacitors in the tuned circuit to a smaller value would improve the performance at VHF. This instrument met all the criteria of a good dip oscillator on all the HF and lower VHF bands.

The simple dip oscillator is easy to construct and, provided the necessary components are available, can be constructed in an evening. It is possible to extend the range up to 60MHz plus by designing another plug in coil L1.

This is not a complete description of how to make the instrument, but rather a few notes to emphasize the important aspects of construction.



L1 details	
Range 1 - 1·6 to 4·0MHz	55 turns of 30swg, random wound
Range 2 - 3·3 to 7·9MHz	27 turns of 30swg, random wound
Range 3 - 6·3 to 15·7MHz	14 turns of 26swg, random wound
Range 4 - 11·9 to 35·2MHz	7 turns of 24swg, close wound

Fig 25.39: A simple dip oscillator

The most important part of a dip oscillator is the tuning capacitor and frequency read-out dial. Sometimes a whole assembly can be obtained from an old transistor radio. The coil socket should be to-as close to the tuning capacitor as possible so that the coil leads can be kept short. The rest of the circuit can be wired around these main components. Choose a coil plug and socket arrangement that is practical. All the circuits so far discussed use two-pin coil plugs and sockets. This means that simple arrangements using crystal holders or phono plugs sockets can be used. The arrangement shown in Fig 25.39 uses two-pin DIN loudspeaker plugs for the coils.

CALIBRATION OF DIP OSCILLATORS

A frequency counter is the most convenient instrument for calibrating the dial although it is possible to check the calibration by listening for the output on a general-coverage receiver, an amateur receiver or scanner. 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 x 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-waves and quarter-waves can be found.

SIGNAL SOURCES

Signal sources of controlled frequency and amplitude are necessary for setting up both transmitters and receivers. Ideally, for receiver adjustment, it is desirable to have an rf source covering from a few hundred kilohertz up to the highest frequency used at the station. The amplitude should be known from a fraction of a microvolt up to tens of millivolts. In a good signal generator both frequency and amplitude are accurately known but such instruments are costly and certainly difficult to make and calibrate in an amateur workshop.

Fortunately many good instruments appear on the surplus market although the frequency calibration is sometimes not too accurate. This is not too important as the amateur almost always has means of checking frequency. Therefore, in selecting an instrument, the quality of the attenuator and the effectiveness of the screening are all-important. At very low levels, a poorly screened oscillator will emit sufficient to by-pass the attenuator and prevent low microvolt output levels being attained.

For less onerous requirements, simple oscillators can be constructed for tuning over a limited range and several examples follow. The dip oscillator is a simple form of signal source but suffers from the defect that the frequency is easily pulled with changes in coupling and it has no attenuator. However, a dip oscillator placed remotely from the receiver under test is often useful. It should be borne in mind that the output may have significant harmonic power and the possibility of interference with domestic receivers, including televisions, should be considered. An audio frequency generator is useful for testing audio amplifiers and for checking the performance of transmitters. The design given in this chapter provides a sinewave output and a frequency range well in excess of the audio range.

LOW-FREQUENCY SINEWAVE OSCILLATOR FOR 10Hz-100kHz

The circuit diagram for this oscillator is shown on Fig 25.40. It is based on a Wien bridge oscillator formed around IC1a and buffered by IC1b. The main frequency-determining components are R1/R2 and RV2 with capacitors C1 to C8. In the configuration shown, stable oscillation can occur only if the loop gain remains at unity at the oscillation frequency. The circuit achieves this control by using the positive temperature coefficient of a small lamp to regulate the gain as the oscillator varies its output. Potentiometer RV3 forms the output level control, with R4 giving a defined output resistance of approximately 600Ω and C11 providing DC isolation. Capacitors C9 and C10 provide power supply line decoupling.

The approximate ranges provided are:

Range 1 10Hz-100Hz
Range 2 100Hz-1kHz
Range 3 1kHz-10kHz
Range 4 10kHz-100kHz

The exact range is dependent on the tolerance of the components used and ambient temperature variations.

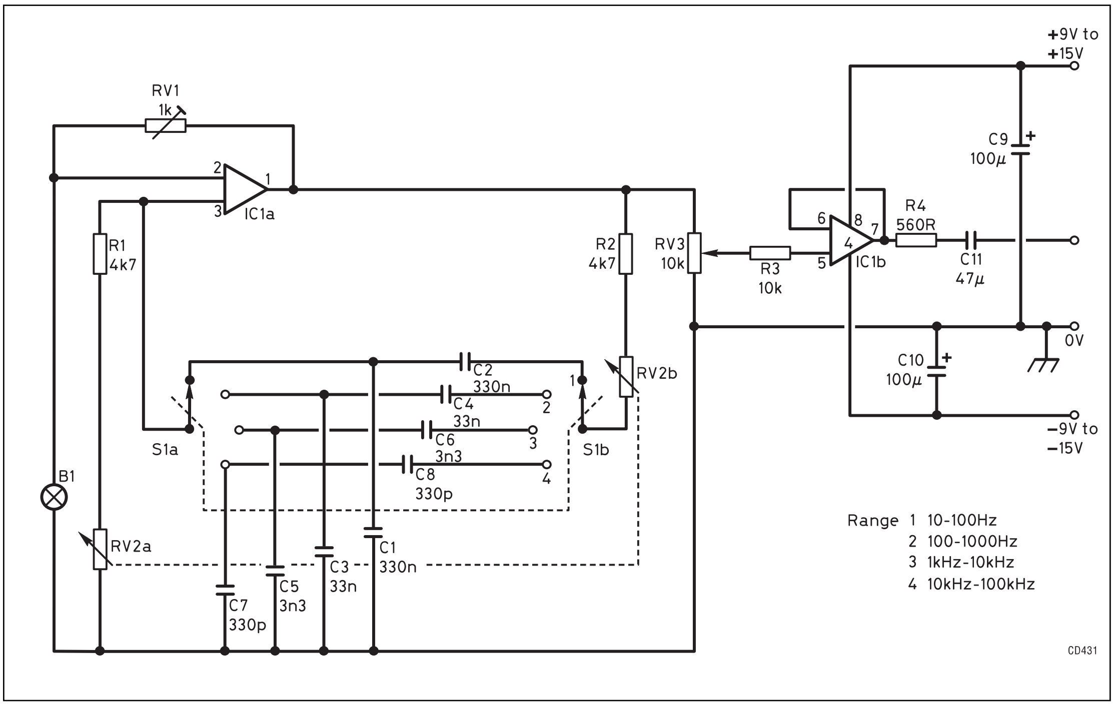


Fig 25.40: Circuit of the low frequency oscillator

R1,R2	4k7	C1,C2	330n
R3	10k	C3,C4	33n
R4	560R	C5,C6	3n3
RV1	1k trimmer	C7,C8	330p
RV2	47k dual gang pot.	C9,C10	100µ, 25V
RV3	10k lin. pot.	C11	47µ, bipolar
B1	28V, 40mA bulb	IC1	LM358

Table 25.8: Low Frequency Oscillator, 10Hz to 100kHz

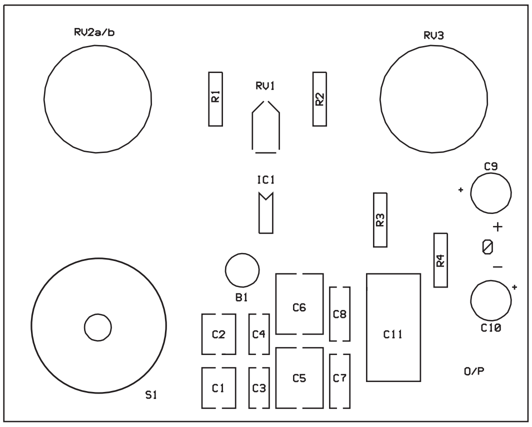


Fig 25.42: Component layout of the low frequency oscillator (not to scale)

The circuit requires a symmetrical plus and minus supply between 9 and 15V.

Construction

A components list is given in Table 25.8. The layout of the circuit is not critical but a PCB pattern (Fig 25.41) is given in Appendix B and a component layout in Fig 25.42. If some ranges, or the output level control, are not required then the layout can be tailored accordingly. The feedback resistor RV1 should be adjusted so that the output on all ranges is just below the clipping level.

Testing

No frequency calibration is required of this circuit but it would be wise to check with a frequency counter that the ranges are as suggested. An oscilloscope is required for setting up the adjustment of RV1.

A CRYSTAL-BASED FREQUENCY MARKER

The purpose of this unit is to produce a 'comb' of output frequencies which are all based on a crystal. The unit described here gives outputs at harmonics of 1MHz, 100kHz, 25kHz, 12.5kHz and 10kHz with an additional output of a sine wave at 1kHz which may be useful as an accurate modulation signal. The sine wave output has an output resistance of approximately 600Ω and maximum amplitude of approximately 2.5V peak to peak.

Circuit Description

The circuit diagram is shown in Fig 25.43. The signal is derived from a 1MHz crystal-controlled oscillator formed by XL1 and IC1 plus various components. Capacitor VC1 allows a slight varia-

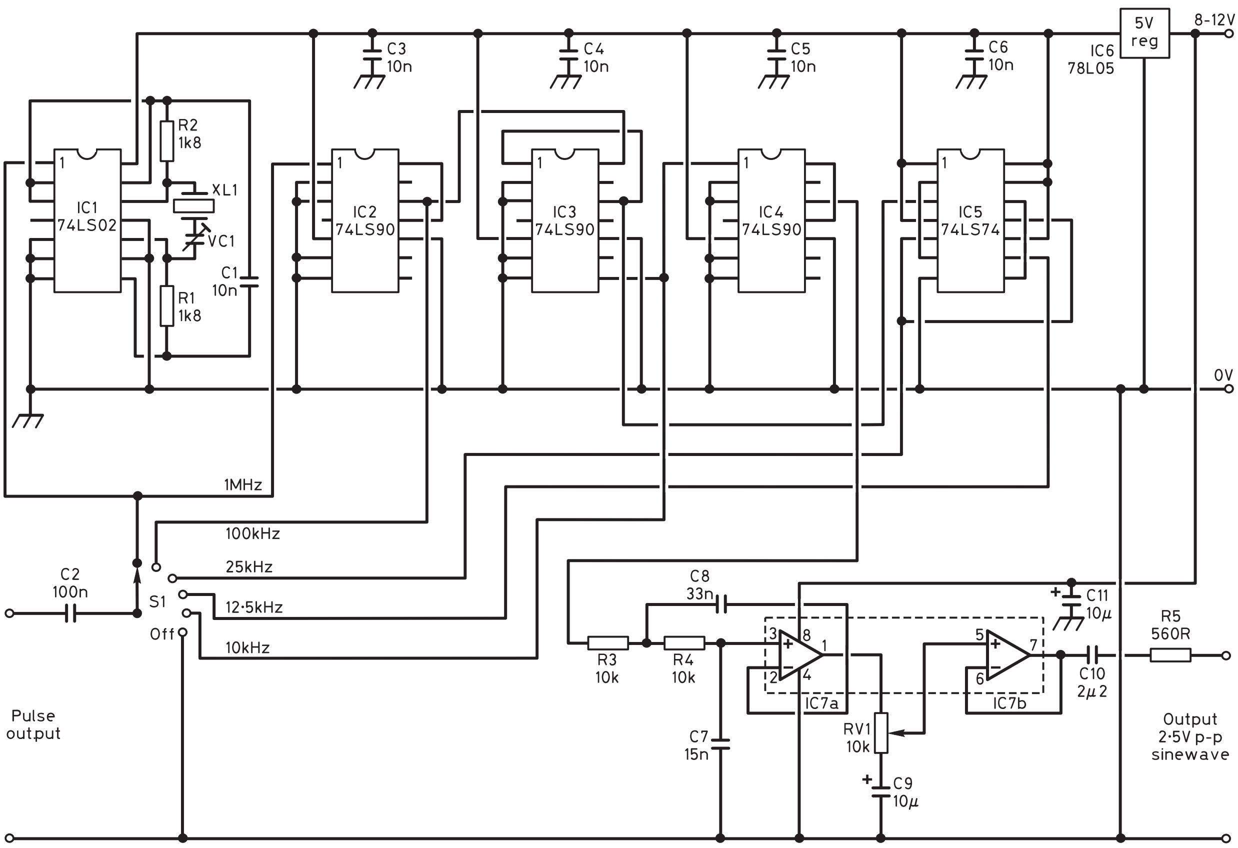


Fig 25.43: Circuit of the crystal-based frequency marker

R1,R2	1k8	C10	2µ2
R3,R4	10k	IC1	74LS02
R5	560R	IC2,IC3,IC4	74LS90
VR1	10k lin pot.	IC5	74LS74
C1	10n, ceramic	IC6	78L05
C2	100n, ceramic	IC7	LM358
C3,C4	10n, ceramic	SW1	2p, 6w rotary switch
C5,C6	10n, ceramic	XL1	1MHz crystal, HC6U
C7	15n	VC1	30p trimmer
C8	33n	IC Sockets	14p, 5 off; 8p, 1 off
C9,C11	10µ, 25V tant bead		

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

Table 25.9: Crystal-based frequency marker components list

tion of the crystal frequency for calibration as described later. This 1MHz signal is divided by 10 by IC2 to give a 100kHz signal. This signal is then passed to IC3 which has a 50kHz output and also a 10kHz output. The 50kHz output is divided by dual flip-flop IC5 to give a 25kHz and 12.5kHz output. The 10kHz signal from IC3 is divided by 10 by IC4 to give a 1kHz square-wave output.

The 1kHz square wave is then filtered by an active low-pass filter formed by IC7a. The variable-amplitude sine-wave output is then buffered by IC7b. R5 forms the output resistance of the buffer.

Construction

A components list is given in **Table 25.9**. The layout for this circuit is not critical but the completed circuit should be housed in a metal box to prevent unwanted radiations. The output should be via a coaxial socket to a small antenna when in use. It requires a power supply of 8 to 12V DC at about 50mA. If the voltage regulator IC6 is omitted the circuit can be fed straight from a 5V supply but ensure there is a supply to the 1kHz filter IC7. A PCB pattern (**Fig 25.44**) is given in Appendix B and component placement details in **Fig 25.45**.

Calibration

The frequency of the 1MHz crystal oscillator can be adjusted by a small amount by VC1. The output from the oscillator or a harmonic should be checked against an accurate frequency source.

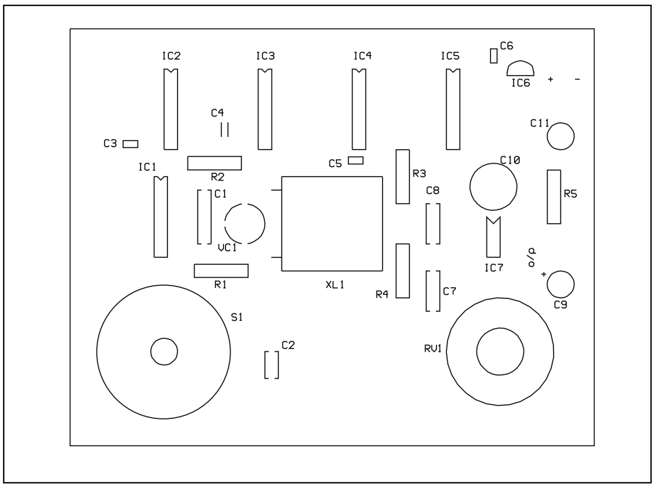


Fig 25.45: Component placement details for the crystal-based frequency marker (not to scale)

AN HF SIGNAL SOURCE

This project uses a Maxim MAX038 IC which is available from several sources. This IC is a high-frequency precision function generator and can produce triangular, sawtooth, square and pulse waveforms as well as a sine wave. In this application it is configured for a sine-wave output of 2V p-p from the IC before any load or filtering. Further details can be obtained from the data sheet [9]. To make this into a useful HF source it should be followed by an attenuator - see end of this chapter.

Fig 25.46 shows the circuit diagram for this signal source which should provide an output frequency from about 2MHz to 20MHz. Frequency of oscillation fo is determined by R1, R2 and C3. The circuit can also be adapted so that the output frequency can be controlled using a voltage applied to pin 8 - this is further explained in the data sheet. A component list is given in **Table 25.10**.

Care should be taken with the circuit layout - use a double-sided circuit board with the top side as an earth plane. C3 should be chosen for low temperature coefficient - NPO ceramics or silvered mica types should be satisfactory. It should be placed so that its connection to the ground plane is close to pin 6 of IC1; C1, C2 and C4 should be placed as close to the IC as possible.

Keep *all* capacitor leads as short as possible in order to minimise series inductance. The variable resistor R2 should be a multi-turn cermet type potentiometer or a single-turn type with slow-motion drive. The MAX038 is followed by a 25MHz low-pass filter. The inductors for this filter can be bought or made. To make L1/L2; for each use a 10kΩ, 0.25W resistor and wind on it 18 turns of 30SWG (0.315mm dia).

The circuit must be supplied from a well-regulated ±5V supply with adequate decoupling at RF. The complete circuit should be placed in a well-screened metal box and the capacitance to earth held constant by fixing the circuit board on suitable spacers.

A suitable double-sided board for experimentation is shown in Appendix B (**Fig 25.47**) and **Fig 25.48** is the component overlay;

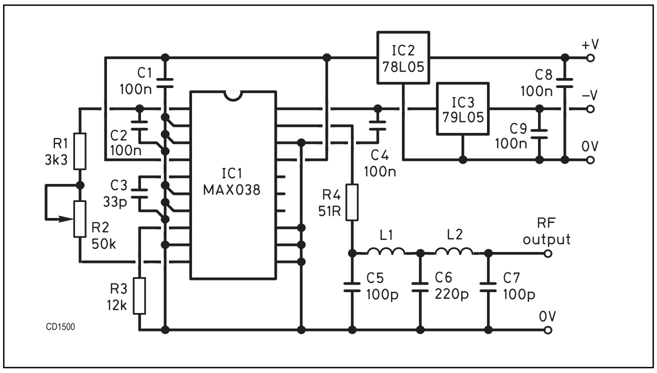


Fig 25.46: Circuit of the HF signal source

R1	3k3	C5,C7	100p low K, ±5%
R2	50k multi-turn (see text)	C6	220 low K, ±5%
		C8,C9	100n ceramic ±10%
R3	12k		
R4	51R	L1, L2	390nH (see text)
C1,C2,C4	100n ceramic ±10%	IC1	MAX038
C3	33p NPO ceramic orl silver mica, ±2% or better	IC2	78L05
		C3	79L05

Table 25.10. HF signal source component list

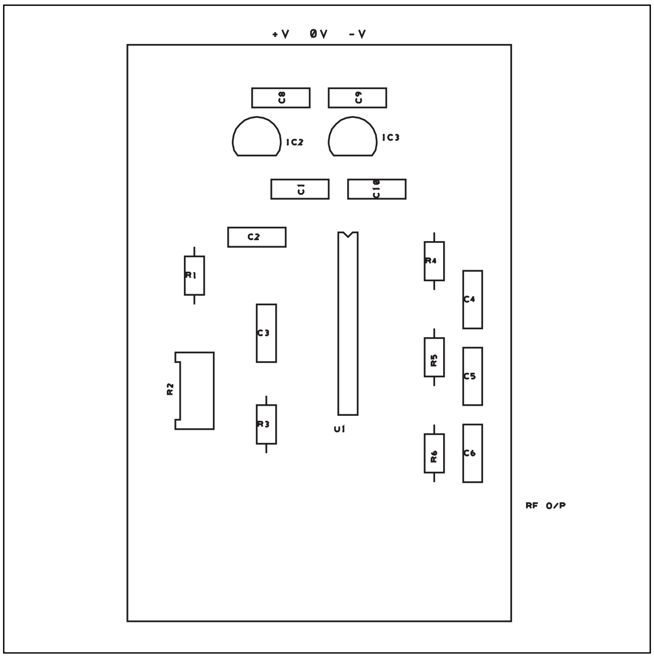


Fig 25.48: Component placement for the HF signal source not to scale)

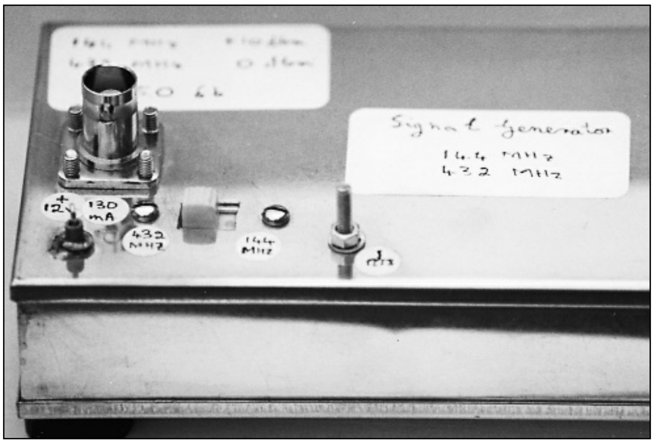


Fig 25.49: The combined 2m and 70cm signal source showing the BNC output socket, power supply terminals and frequency selector circuit

this includes the regulated power supply. *Do not* use an IC socket for the MAX038 as this will affect capacitance to earth.

A COMBINED 2M AND 70CM SIGNAL SOURCE

This signal source was developed by John Brown, G3DVV [10]. The project (**Fig 25.49**) uses one of the low-cost crystal oscillator modules that are available - the one that is really useful is the module for 48MHz. To a VHF/UHF enthusiast 48 x 3 = 144MHz and 48 x 9 = 432MHz, ie 2m and 70cm. These modules have a lot going for them. Their cost is less than £10 - compare that with purchasing a separate crystal, semiconductors, capacitors and resistors, and then putting them all together on a PCB. What's more, the modules work first time, which makes a pleasant change.

NOTE: The original device for TR2 (MSA0304) appears to be obsolete. If you can find one then all well and good. An alternative is one of the MAR monolithic amplifiers from Mini-Circuits. Several of the devices match the specification but the bias resistor R8 must be changed - see *Biasing MMC Amplifiers* at www.minicircuits.com/application.html.

R1	3k8
R2	1k
R3	68R
R4, 5, 6, 7	1k SMD type 1206
R8	See text. For MSA0304 use 270R, 0.5W
R9, R11	292R (270R + 22R) SMD type 1206
R10	18R
C1	4µ7 tantalum
C2	100n
C3, C4	1n SMD type 0805
C5 - C9	100p type 0805
RFC1, RFC2	1mH
L1	44nH (3T of 0.56mm ecw, 65mm long, inside 47.5mm, pulled out until there is 10mm between the ends
L2	7.5nH (loop of 0.56mm ecw 15mm long, ends 10mm apart
IC1	48MHz crystal oscillator module
TR1	BFR96
TR2	MMIC. See text. Original used MSA0304
VR1	78L05
D1 - D4	1N914 or any suitable switching diode
F1	Toko CBT3
F2	Toko 7HW
S1	2-way single pole switch
Double sided glass-fibre PCB type FR4, 16mm thick	
Tin-plate box, 74 x 148 x 30 mm	
Track pins	
All resistors are 0.25W, 5% tolerance unless specified otherwise	

Table 25.11. Component list for combined 2m and 70cm signal source

Preview

The component list is given in **Table 25.11**. The circuit, shown in **Fig 25.50**, starts with a TTL crystal oscillator module. The 2.5V rectangular-wave output it provides is fed into a BFR96 amplifier/multiplier, the output of which is selected by diode switches between the filters. The output from the filters, 144 or 432MHz, is then amplified by a standard MMIC (monolithic microwave integrated circuit); sometimes called a modamp. In spite of the fearsome name, MMICs are easy to handle and will work over a wide range of frequencies without fuss or the need for tuned circuits. After amplification the output is passed through a 3dB attenuator, connecting finally to a BNC socket.

The power supply is external. VR1 provides the necessary +5V for the crystal oscillator module. As the unit consumes a total of 130mA, an internal battery is not used. The crystal oscillator module takes 80mA and the MMIC will vary according to device used.

The Circuit in Detail

The output from the crystal oscillator module (IC1) is rich in harmonics. This brew is fed into a UHF-type transistor (TR1), working as a frequency multiplier/amplifier. A wide range of harmonics is produced at the collector, up to and including the 32nd. The output from the collector is diode switched between two filters. The filters are standard types by TOKO.

The switch selects the filter to be used by connecting a positive voltage to one pair of diodes or the other, causing them to conduct and thus complete one of the circuits.

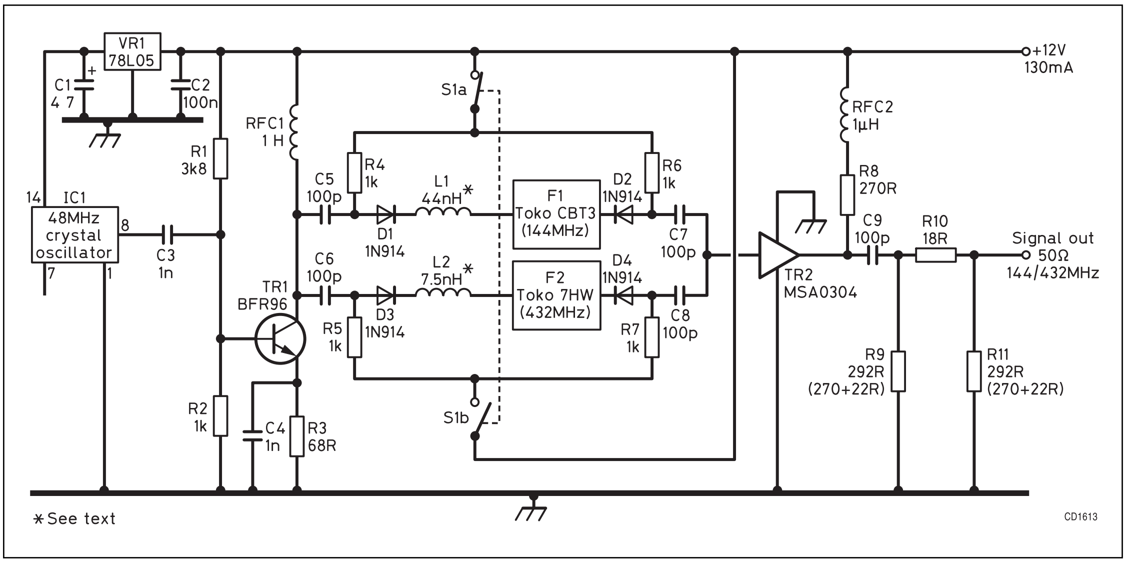


Fig 25.50: Circuit of the combined 2m and 70cm signal source. Steering diodes are used to switch between filters F1 and F2

The two coils (LI and L2), one for each filter, give an approximate match from the collector of the BFR96 into the 50Ω input of the filters. Their value is not critical. During development, a number of diodes were tried as switches, and the 1N914 was found to be as good as any.

The output from the filters, both at 50Ω, is fed into the 50Ω input of the MMIC. After being amplified by the MMIC, the output is still 50Ω, as is the 3dB attenuator and the BNC socket. The attenuator acts as a buffer against open-circuits and short-circuits.

Construction

The PCB pattern is given in **Fig 25.51** (in Appendix B) and the layout in **Fig 25.52(a)**. It is important to use the right type of printed circuit board, as the tracks form a transmission line with the ground plane, thus cutting down radiation and providing the correct matching. The PCB material (FR4) and track widths are critical in maintaining transmission line impedance; components are mounted on both sides of the board. The choice of etching method is yours but it must be a double-sided board, one side forming a ground plane. After cutting the PCB to size, make sure that it fits under the lid of the box.

On the component side of the board, use 0.1in (2.5mm) tracks. The reason for doing this is that this size of stripline forms a 50Ω transmission line with the ground plane. It is known as microstrip. For power supply lines, the microstrip size makes no difference, but the same width is used for convenience. While you are working try to keep your fingers off the copper, as finger marks do not etch very well. Wash the board well and dry after etching, then drill.

The only difficult items to mount are the two filters (**Fig 25.52(b)**) because the lugs as well as the pins have to be accommodated. Holes are required for TR1 and TR2, and pin through holes for IC1 (note that there is no connection to pin 1 - clear the copper away from pins 1, 8 and 14). TR1 and TR2 should fit snugly into their holes, with their leads lying flat on the copper strips.

This just leaves the holes to be drilled for the circuit pins, otherwise known as half-track pins, Veropins or vias. Wherever the layout diagram shows a pad or a track as grounded, do this with a track pin. Push a pin through from the component side and, after soldering to track and ground plane, cut it off at board level. Note that three pins are not cut, namely those marked as 'S1a', 'S1b' and '+12V'. Clear the copper away from the base of these three pins; these are hard wired.

Next, mount the SMDs (surface mount devices). Tin one side of the strip or pad, then place the component in position and hold down firmly. You will need a steady hand. The tip of a small screwdriver or a toothpick is placed on the centre of the compo-

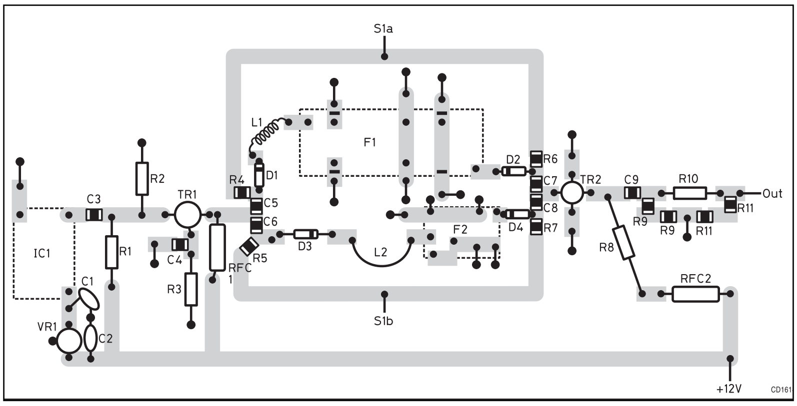


Fig 25.52 (a): Component placement for the signal source

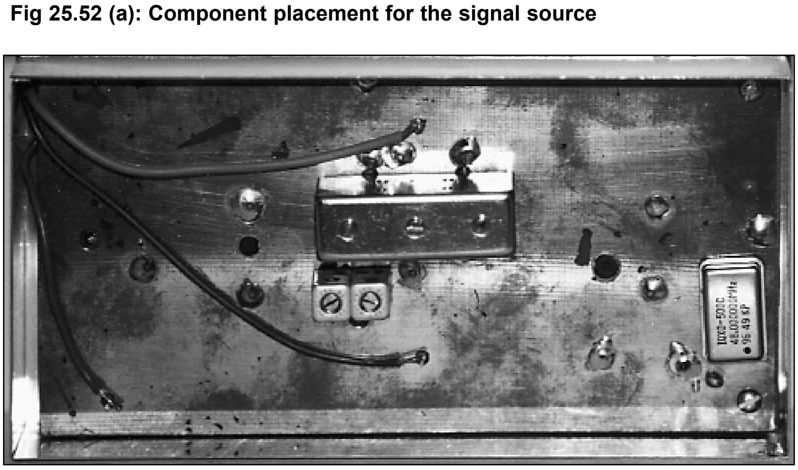


Fig 25.52 (b): The signal source showing the two filters and the crystal oscillator module

nent while a soldering iron with a small tip is applied to one end. It is very easy to play 'tiddlywinks' with these devices, and once in orbit, that's it. When one side has been soldered satisfactorily, solder the other side. Then re-solder the first connection if necessary. Finally, solder in the rest of the components, connecting the semiconductors last.

Take the lid of the box and fit the switch, the connections for the power supply and the BNC socket. The position of the BNC socket has to be exact, as the inner terminal of the socket must touch the pad marked 'out'. The wires from the switch and power supply are brought out around the sides of the board, which is held in position by screws.

When all appears to be in order, solder the tip of the BNC socket to the output pad, which is also connected to R10 and R11. Complete the wiring to the three uncut circuit pins.

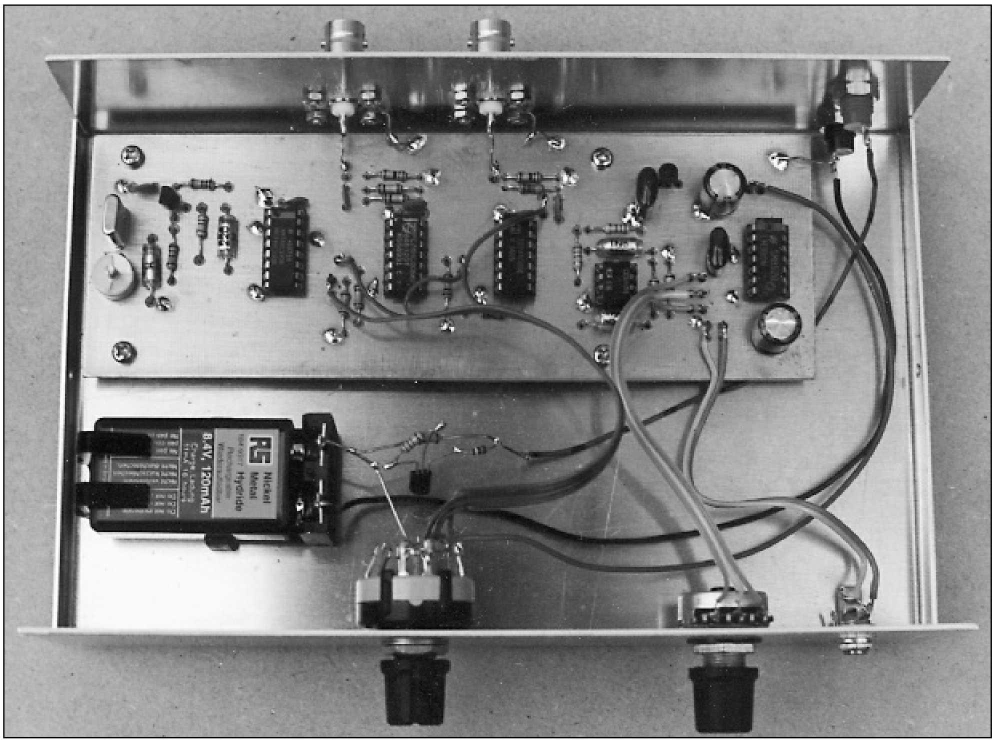


Fig 25.53: The receiver calibrator and transmitter monitor

You may find that the bottom lid of the box fits without trouble, but in G3DVV's case the protruding screws from the 2m Toko filter push against the lid, requiring two accommodation holes to be drilled.

Final Test

The components face the underside of the lid and are difficult to check in their final position. Therefore, check that the unit is working before finally fixing it in position. There is nothing in the circuit which requires adjustment. *Do not* try altering the filter settings.

The unit should work from switch-on, but if it doesn't try the following:

- Check that all the track pins are properly soldered.
- Check all the voltages.
- Use a probe to check the output from IC1, TR1 and TR2. The probe can be a diode, a link to the coil of a dip meter, or (probably) a 2m and 70cm receiver with a piece of coaxial cable and a wire loop as an RF sniffer.

Results

The output on 2m is 10mW, and on 70cm 1mW. On 2m the nearest harmonic is on 288MHz and 16dB down. On 70cm the nearest harmonic is on 366MHz and 24dB down.

Although there is a whole range of harmonics, the temptation to use the 27th for 1296MHz was not followed because of the difficulty of cramming the components onto the printed board.

RECEIVER CALIBRATOR & TRANSMITTER MONITOR

The receiver calibrator and transmitter monitor (**Fig 25.53**) described is by G4COL [11]. It offers the following facilities:

- Frequency calibration of receivers and transmitters from 50kHz to over 150MHz.
- Receiver sensitivity measurement from 50kHz to over 50MHz.
- Calibration of attenuators and receiver S meters.
- Monitoring transmissions for stability, speech and keying quality.

Frequency intervals of 5MHz, 500kHz and 50kHz are provided, and frequency can be set accurately against broadcast standards. Receiver measurements need an external attenuator. The unit is battery portable, and was designed to be reproducible with components readily available and minimal test equipment.

Principle of Operation

The heart of the unit is a harmonic comb generator, which produces very short duration pulses, quite commonly used for frequency calibration. Its use for providing a reasonably accurate signal level is less common. The main properties of relevance to this project are below. Some of them are not readily found and had to be calculated:

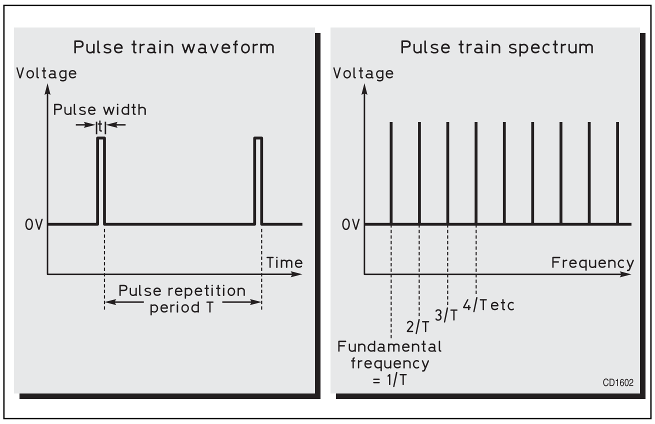


Fig 25.54: Waveform and spectrum of a train of narrow pulses

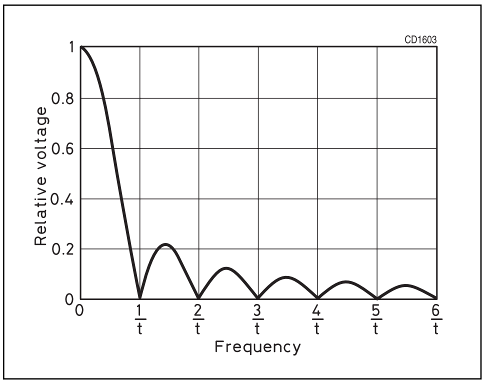


Fig 25.55: Enveolpe of the harmonic comb's spectrum, viewing a greater frequency than Fig 25.54

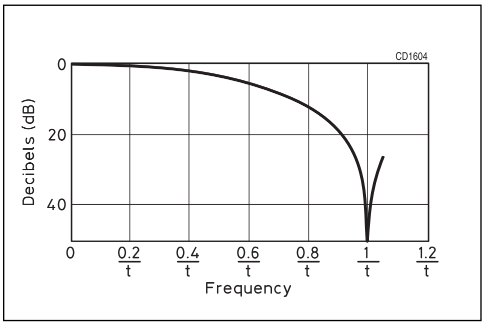


Fig 25.56: Lowest frequency lobe of the harmonic comb's spectrum. Note the logarithmic Y axis

- The term 'harmonic comb' describes the spectrum produced by very narrow pulses. As **Fig 25.54** shows, energy has shifted into harmonics to such an extent that the fundamental frequency component and many harmonics have similar amplitude. Note that the two voltage axes are not to scale. The individual 'teeth' of the harmonic comb have a much lower voltage than the peak voltage of the pulse - the energy of the pulse has been spread among its many components.
- The pulse width determines how the amplitudes of the different harmonics vary with frequency. **Fig 25.55** shows the 'envelope' of the spectrum, looking much higher in frequency than Fig 25.54. The tip of each spectrum component lies on this curve.
- The lowest-frequency lobe is shown in more detail in **Fig 25.56**. It contains 90% of the total energy and is 3dB down with respect to low frequencies, at 44% of the first null frequency. 72% of the total power of the comb falls below this '3dB down' frequency.
- The lobe is 1dB down at 26% of the first null frequency. This means that if we are to use the harmonic comb for calibration only up to a frequency of 26% of the first null, the teeth of the comb fall within a 1dB range. This portion of the spectrum contains 49% of the total energy.
- If V_p is the peak voltage of the pulse, the DC component is V_p x t/T
- The RMS voltage of the low-frequency harmonics is the DC voltage x 1.414 (the square root of 2).
- The DC component can be measured relatively easily.

Don't worry if quite a bit of this seems hard to take in. Some of the points are discussed later, and understanding is likely to improve with use of the test gear for making real measurements. If you want to delve further, consult an engineering maths textbook.

Block Diagram

Referring to **Fig 25.57**, the reference frequency is provided by a 5MHz crystal oscillator, the output of which is divided by 10 and 100 to produce 500kHz and 50kHz. One of these three frequencies is selected as the pulse repetition frequency, to drive a pulse generator which produces very narrow pulses, giving the spectral comb. The pulses are fed to the output socket via an attenuator and also drive the sampling gate, which is essentially a switch that acts as the input stage of the transmitter monitor.

The sampling gate's input comes from a panel socket. Its output feeds a buffer amplifier, volume control and output amplifier (which drives headphones or a small loudspeaker). The sampling gate behaves as the mixer of a direct-conversion receiver, in which an audible beat signal can be produced by a signal close to any of the harmonics of the pulse repetition frequency. Suppose we are switching the sampling gate at 500kHz; a 1kHz beat note will be produced by an input signal present at any of 499, 501, 999, 1001, 1499, 1501, 1999, 2001, 2499, 2501kHz and so on.

Circuit Description

Most of the circuit, shown in **Fig 25.58**, operates from a 5V supply derived from the 9V battery by three-terminal regulator IC1. The 5MHz crystal reference frequency oscillator is a Colpitts, based around TR1. IC2 divides this continuously by 10 and 100 to produce 500kHz and 50kHz. The pulse repetition frequency is selected by three sections of analogue switch IC3, controlled by wafer switch S1b. IC4 generates the narrow pulses, using AC series TTL.

The pulse width is set by the delay section R7 with IC4b. Negative-going pulses from IC3c feed an output attenuator consisting of R11 and R12. An inverter stage IC4d drives the sampling gate IC3d with positive-going pulses. The attenuator gives the unit an output resistance very close to 50Ω. That is the extent of the calibrator, the rest of the circuitry being the monitor.

A plot of the negative-going output pulse is shown in **Fig 25.59**. The original of this was captured using a fast digital storage oscilloscope and shows the pulse width to be within the 5 nanosecond per division horizontal graticule. The corresponding spectrum was found to have its first null at 270MHz, indicating a pulse width of 3.7 nanoseconds.

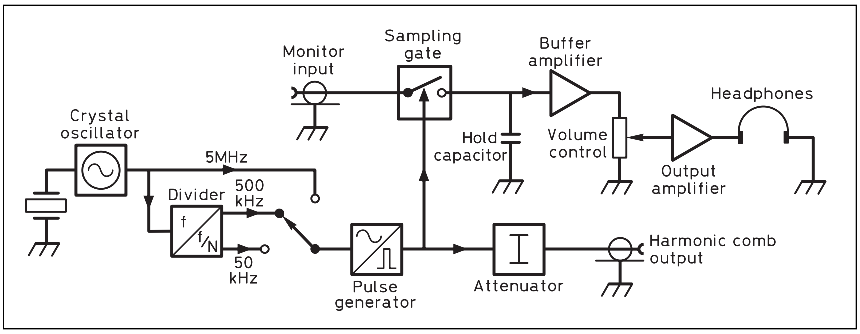


Fig 25.57: Block diagram of the receiver calibrator and transmitter monitor

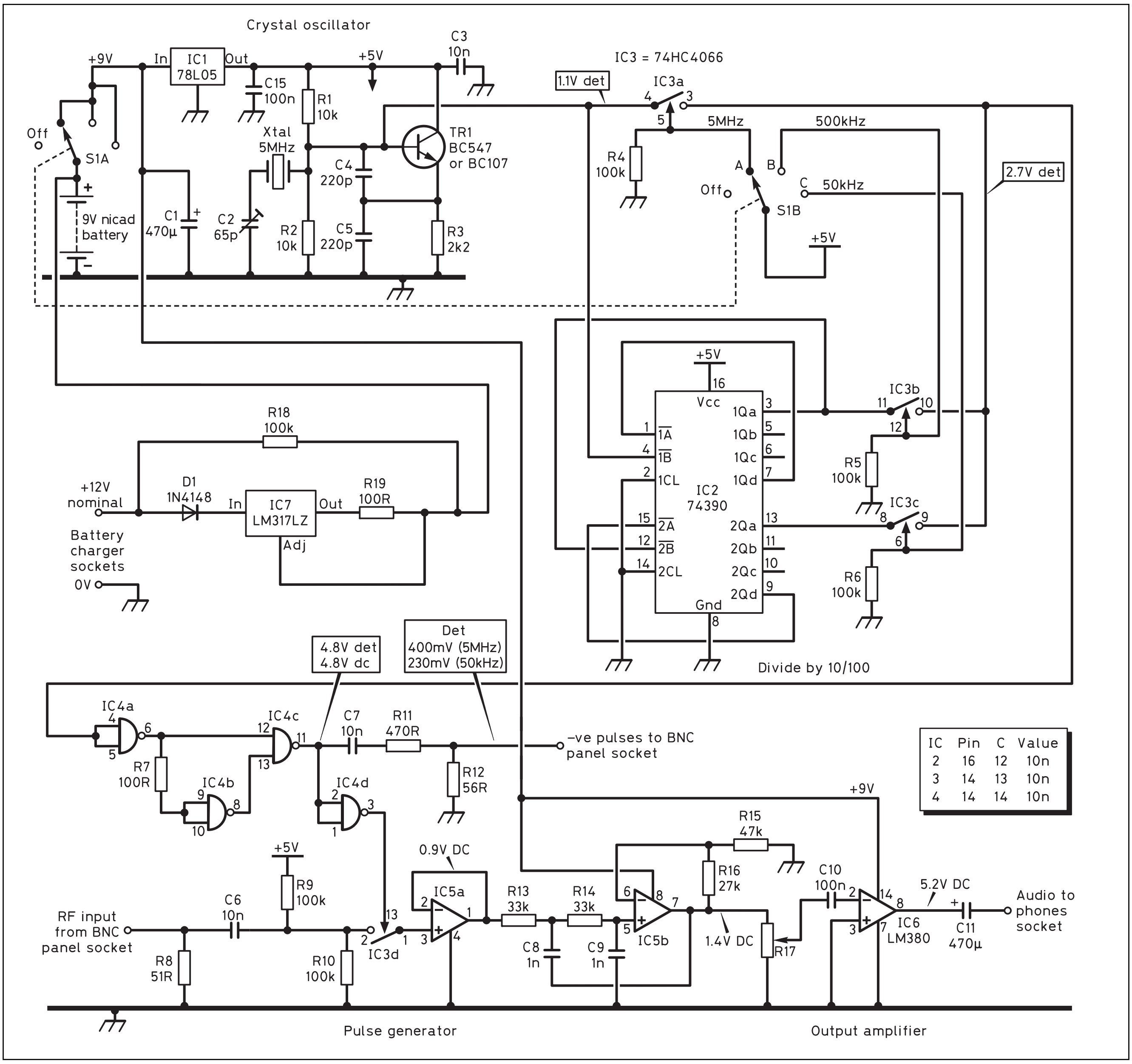


Fig 25.58: Circuit diagram of the receiver calibrator and transmitter monitor. S1 is used to control the 'comb' of frequencies at 5MHz, 50kHz or 50kHz intervals

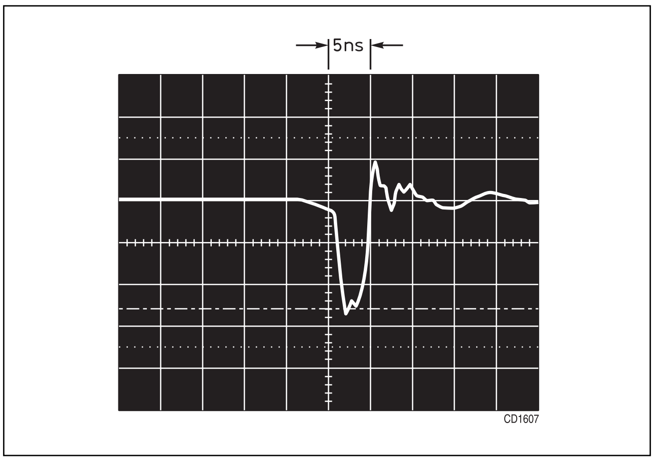


Fig 25.59: Oscilloscope trace of negative-going calibrator output pulse

R8 sets the monitor input resistance. R9 and R10 bias the sampling gate switch IC3d into its operating range, between 0 and 5V. The sampling gate's 'hold' capacitance is not shown on the circuit diagram.

It consists of stray capacitance plus the input capacitance of IC5a, a high-impedance, unity-gain buffer. The buffer is followed by a low-pass filter to attenuate the sampling pulses which inevitably feed through the gate. After the volume control, IC6 provides audio power output.

Part of Fig 25.58 is a battery charger circuit, for use with nickel cadmium (NiCd or 'nicad') or nickel metal hydride (NiMH) batteries. It is based on IC7, which is configured as a current source of about 12mA (set by R19). The charger allows the internal battery to be recharged by connecting a power source of 12 to 24V to the terminals.

D1 prevents damage from a reversed supply and R18 allows the battery voltage (when not being charged) to be monitored by a high-impedance voltmeter (eg digital multimeter) at the charger terminals.

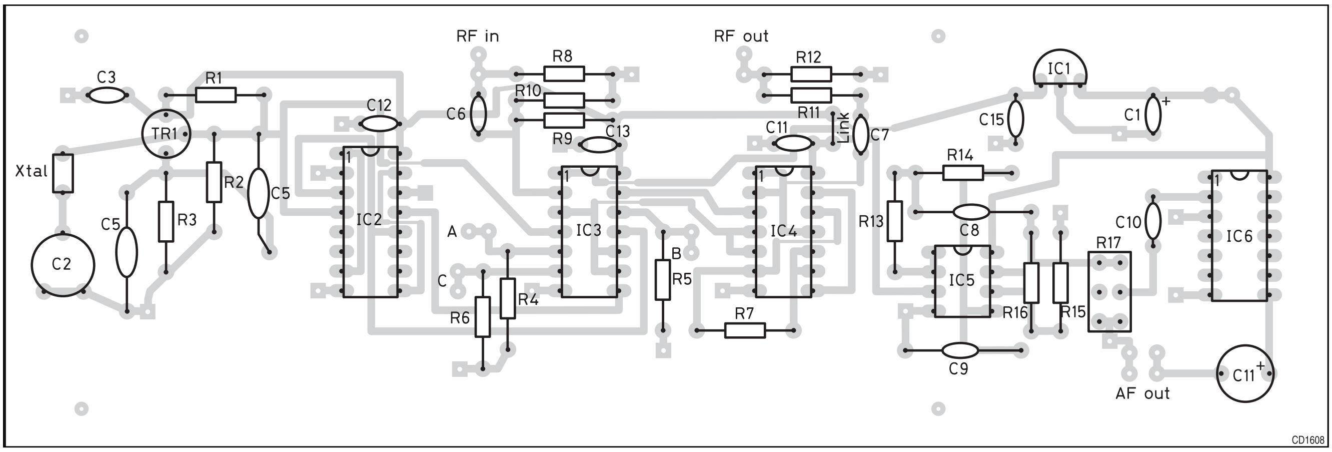


Fig 25.61: Component placement for the receiver calibrator and transmitter monitor

Construction

Most of the salient features of the construction can be gleaned from the photograph, Fig 25.53. The layout is not critical and the component list is given in **Table 25.12**. A printed circuit board layout is given in Appendix B (**Fig 25.60**) whilst the component placement is given in **Fig 25.61**. Note that the PCB is double-sided.

The square pads denote connection to the copper plane on the component side; wire links are soldered to the pad on one side and the copper plane on the other.

The remaining component holes should be carefully cleared of copper from the component side using a 4mm drill bit to clear a diameter of 2mm or so. The four isolated pads are for the mounting holes, which were opened up to 4mm. Single-sided construction is unlikely to work well for this unit, because of the long earth tracks. Note: the battery charger circuitry is not included on the PCB.

R1, R2	10k	C3	10n, 63V ceramic
R3	2k2	C4, C5	220p, 160V polystyrene
R4, R5, R6	100k	C6, C7	10n, 63V ceramic
		C8, C9	1n, 160V polystyrene
R7	100R	C10	100n, 100V mylar
R8	51R	C11	470µ, 16V electrolytic
R9, R10	100k	C12,C13, C14	10n, 63V ceramic
R11	470R		
R12	56R	C15	100n, 100V mylar
R13, R14	33k	IC1	78L05ACZ
R15	47k	IC2	74HC390N
R16	27k	IC3	74HC4066P
R17	100k log pot, carbon or similar	IC4	74AC00
		IC5	CA3240E
R18	100k	IC6	LM380N
R19	100R	IC7	LM317LZ
C1	470µ, 16V electrolytic	TR1	BC547, BC107
		D1	1N4148
C2	5p5 to 65p trimmer	X1	5MHz crystal, HC-49/U

S1 Rotary switch, 3pole 4 way
3.5mm jack socket
BNC square flange chassis socket, 50Ω
Case, knobs, PCB battery holder

All resistors 0.6W metal film, 1% tolerance unless specified otherwise

Table 25.12: Components list for receiver calibrator and transmitter monitor

Testing

Verifying that the unit is operating correctly can be done with a diode detector (see **Fig 25.62**), a high-impedance multimeter (most digital multimeters should be fine) and a signal source. As the calibrator is not tuneable, the source needs to be able to be set to produce a beat note with one of the comb teeth. A crystal oscillator operating close to a multiple of 500kHz could be used.

Having carefully inspected the unit for correct connections with an absence of dry joints, apply power. Set the multimeter to DC volts and, with the detector-plus-multimeter combination, measure the voltages as marked on the circuit diagram, Those marked 'det' involve the detector. The voltages may differ by a few percent.

Although not part of the testing sequence, it is instructive to measure IC4d output. Provided that the multimeter input resistance is of the order of megohms, measure DC with a resistor of 10kΩ to 100kΩ in series with the meter, so as not to disturb the pulse.

The reading should be close to zero. The detected voltage, on the other hand, should be greater than 4.5V. These high peak (detected) and low average (DC) voltages are characteristic of a very narrow pulse.

Connect the signal source (level around 100mV RMS) to the BNC input connector, with headphones plugged into the audio output socket and the volume control turned above minimum. A beat note should be audible, the pitch of which should be able to be varied by tuning the source or adjusting C2. The higher the frequency of the source, the greater the effect of C2. If the source is close to an exact number of megahertz or multiple of 500kHz, a beat note should be produced with the unit switched to 50kHz or 500kHz. If a source at a multiple of 5MHz is available, this too should give a beat.

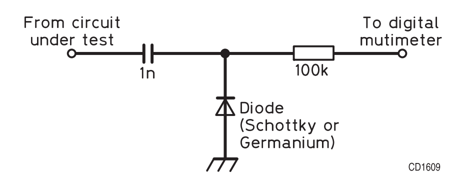


Fig 25.62: The diode detector used for testing

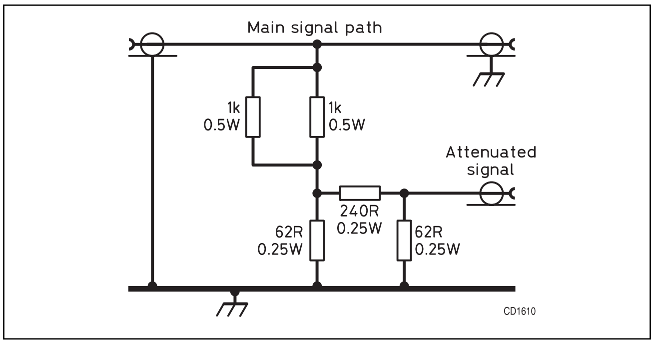


Fig 25.63: Signal pick-off

Operation

Some additional items are needed:

- For monitoring transmitters a high-power attenuator or signal pick-off plus dummy load;
- For receiver measurements a low-power attenuator will suffice and the signal pick-off can be used. A suitable signal pick-off is shown in **Fig 25.63**.

Fig 25.64 shows the measurement set-ups. Where 'calibrator' is shown, connect the coaxial cable to the pulse output, and for 'monitor', to the monitor input. Similarly 'receiver' and 'transmitter' can be a transceiver in receive and transmit modes respectively.

Frequency Calibration

See **Figs 25.64(a) and (b)**. If a receiver is available covering 10MHz, the calibrator crystal oscillator frequency can be adjusted for zero beat with the WWV frequency and time standard signal.

Measuring Receiver Sensitivity

See Figs 25.64(a) and (b). The harmonic comb's principal advantage for this task is that the signal power in each 'tooth' is already low. One of the reasons professional-quality signal generators are expensive is that it is difficult to produce a pure signal with an accurate low level. Much careful screening and filtering is involved, as well as an accurate switched output attenuator of up to 130dB total attenuation, which itself demands extreme isolation from input to output.

The output level is given by:

pulse height x pulse width ' repetition frequency x attenuation factor x √2

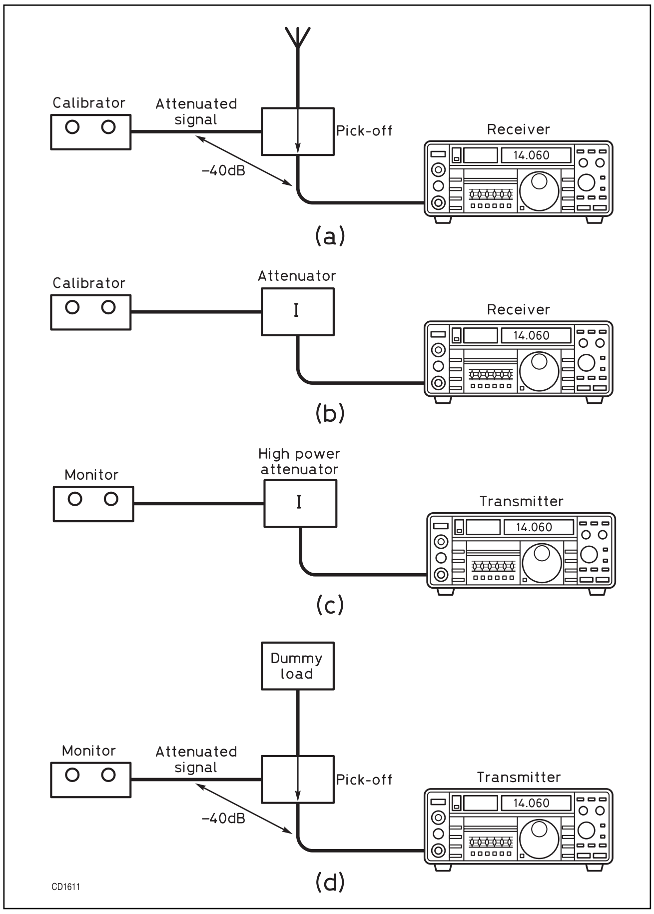
The pulse width is nominally 4 nanoseconds. If the pulse height is 5V, the repetition frequency 5MHz, and the output attenuation factor into a 50Ω load is 0.05, then the level of the lowest frequency teeth is:

$$5 \times 4 \times 10^{-9} \times 5 \times 10^{6} \times 0.05 \times \sqrt{2}$$
$$= 7.07\text{mV or } -30\text{dBm}$$

Similarly, for 500kHz and 50kHz repetition frequency, the corresponding outputs are 707mV and 70.7mV respectively.

At 50kHz repetition frequency, a 40dB attenuator will present the receiver with 0.7mV signal. A 40dB attenuator is not unduly difficult to make, or may be purchased, sometimes in the form of a switched attenuator or perhaps a pair of 20dB coaxial attenuators used in series. The theoretical -1dB frequency of the comb is 65MHz. A spectrum analyser plot of the comb at 5MHz repetition frequency is shown in **Fig 25.65**. The agreement with the predicted level is really quite good (the spectrum analyser used has much better amplitude accuracy than most). The reason for the difference between teeth is not clear but nevertheless the plot shows that there is relatively little variation up to 50MHz. The main sources of uncertainty will be the height and width of the pulse. A very rough calculation indicates that the signal level should be within ±3dB of nominal.

Beware: when testing receivers with broad-band front-ends (no preselector), it is advisable to use at least 20dB attenuation between calibrator and receiver. This is because the receiver mixer is subjected to many teeth of the harmonic comb. The peak pulse voltage, which includes all the teeth, is 250mV (5V x the output attenuation factor of 0.05). Although this is not likely to damage the mixer, most will be driven outside their normal operating range.



24.64: Measurement set-ups

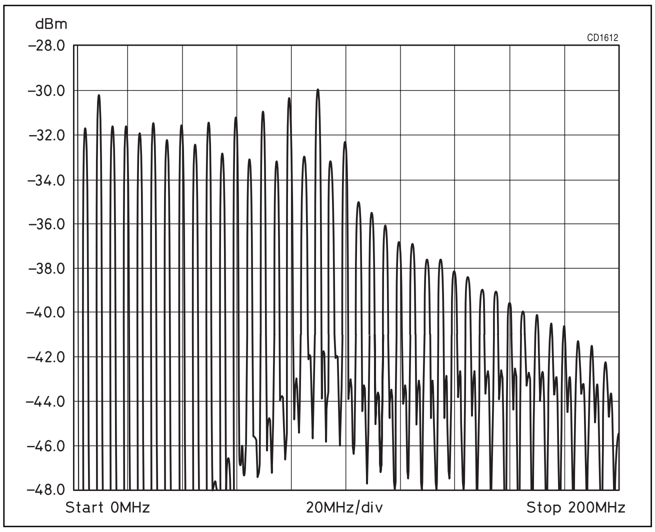


Fig 25.65: Spectrum analyser plot of 5MHz pulse output

Calibrating Attenuators and Checking S-meters

Referring to Fig 25.64(b), 20dB attenuators can be checked very accurately using the receiver's S-meter. The receiver itself does not require calibration. Couple the calibrator to the receiver. Set the calibrator to 500kHz and tune the receiver to a harmonic of 5MHz (this may not be possible on an amateur-bands-only receiver), obtaining a beat from the calibrator signal. Note the S-meter reading.

Now insert the attenuator and set the calibrator to 5MHz. The S-meter reading should return to exactly the same position if the attenuator is an accurate 20dB, any change indicating an error.

Once an accurate attenuator is available, the receiver S-meter can be checked by comparing its readings with and without the attenuator in the signal path.

To give an idea of the accuracy which can be obtained, G4COL measured a nominal 20dB attenuator (using a spectrum analyser as the receiver) to be 19.3dB. The figures produced by the calibration company were then consulted and showed their measurement to be 19.22dB!

Monitoring Transmissions

Referring to Fig 25.64(c) and (d), make sure that adequate attenuation is used between the transmitter and the monitor, and that the attenuator or dummy load can handle the transmitter power. The input to the monitor should be kept to 0dBm (1mW or 224mV RMS), but can be driven up to 10dBm without undue distortion. This means that for up to 10W (40dBm) transmitter output, a 40dB attenuator or the signal pick-off shown are suitable.

Tune the transmitter for an audible beat note for CW and intelligible speech for SSB. With headphones which are fairly acoustically opaque, it is quite feasible to assess one's own speech quality speaking into the microphone, though if another source is available so much the better.

For CW signals, if you have an oscilloscope, monitoring the audio output with the beat note around or below 1kHz will show the shape of the actual radio frequency waveform and the effect of adding filters. Harmonic distortion was evident on the output of G4COL's HP606A signal generator at 25MHz and was removed by adding a low-pass filter to the signal path.

By listening you should be able to detect such defects as key clicks, frequency drift and chirp, speech clipping and splatter. Frequency calibration can also be checked.

Note that the audio bandwidth of the monitor, actually set by the sampling process, is around 5kHz on the 50kHz repetition frequency setting. Since most transceivers contain steep cut-off filters, this is not likely to have a significant effect.

Suggestions for Experimentation and Improvements

Here are a few suggestions that can be used as a basis for improvement and experiment:

(a) The calibrator could be built on its own without the monitor receiver.

(b) The monitor noise could be reduced. The aim here has been simplicity. An earlier, more complex, version was quieter.

FREQUENCY COUNTERS

The cost of frequency counters has decreased over the last few years and handheld types operating in excess of 1GHz cost less than £100, with some units going up to 3GHz not much more expensive. It is the number of digits shown which is important. It has come to a point where the cost of construction and availability of components makes construction at home not worthwhile.

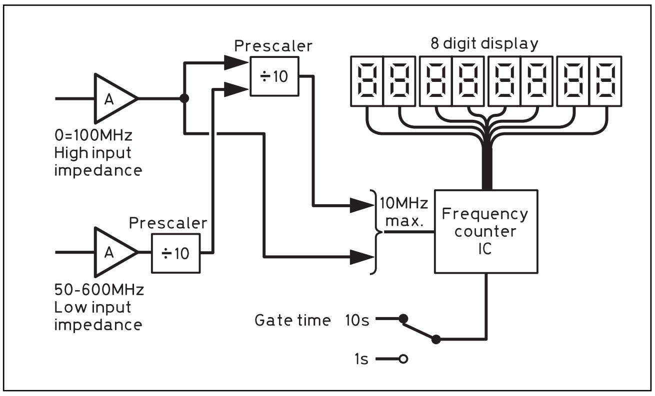


Fig 25.66: Block diagram of a frequency counter

The purpose of this section is to explain briefly how a frequency counter works and gives tips for making measurements. **Fig 25.66** shows the typical block diagram of a frequency counter but there will be variations on the input arrangements. There is usually a high (1MΩ) and low input impedance input - the low input is typically 50Ω for the higher frequencies. The problem with a high input impedance is the effect of the shunt capacitance, eg 10pF at 144MHz has a reactance of 110Ω. The actual inputs can either be from a test probe or direct pick-up off-air using an antenna. The input to a frequency counter is fairly sensitive and it may only require 50mV of RF in order to obtain a digital readout. Consider also the maximum signal the frequency counter can cope with and any DC voltage present - above this you are likely to damage an expensive piece of test equipment.

When using a probe with physical connection to a circuit, care should be taken that the probe does not affect circuit conditions. The main problem is the probe impedance (especially the shunt capacitance) changing the frequency of operation of the circuit, eg 'pulling' of the oscillator. Also the probe may affect the DC operating conditions. An oscilloscope probe is a good unit to use. If trying to measure frequency in a 50Ω system, make sure the probe does not cause a mismatch or overload the circuit. These points are summarised as follows:

- Do not attach a probe to frequency-determining elements.
- Measure oscillator frequency after the buffer amplifier.
- Use AC coupling, especially in higher-voltage valve circuits.
- Put a 1kΩ resistor in series with the probe.
- Do not allow a 50Ω system to become mismatched.

An alternative is to use a pick-up loop as shown in **Fig 25.67**. Ensuring only loose coupling, this overcomes some of the inherent problems in trying to minimise influence on the circuit and damage to the counter. Another similar way is to use an antenna as the input device which is coupled straight to the frequency counter input.

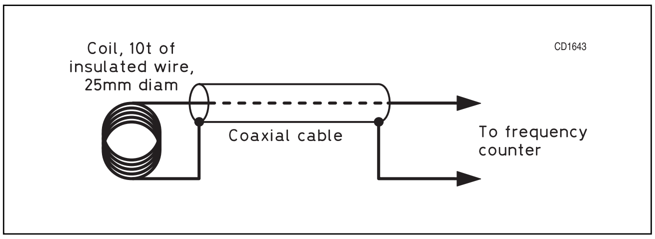


Fig 25.67: Pick-up loop

When taking frequency measurements, especially off-air, only an unmodulated carrier should be used. The following guidelines should be used:

- With AM and FM radios, *Do not* speak into the microphone or provide modulation.
- With SSB sets, the CW position should be used with key down.
- With digital transmissions, no superimposed data should be transmitted, just a carrier.

More expensive frequency counters will allow the period to be measured in usually microseconds or milliseconds. This is useful for frequencies less than about 100Hz and will generally give a more accurate result but you will need to use a calculator to revert to frequency.

SPECTRUM ANALYSERS

What is the purpose of a spectrum analyser? It is a piece of equipment that can receive a signal (or group of signals) and give a display of the frequency components present and the relative amplitudes.

It is possible to purchase a brand new one or perhaps a second-hand can be found, but the cost will still be high! There are add-on units for oscilloscopes which will function to 1GHz and cost about £600 at 2005 prices.

A Simple Spectrum Analyser (SSA), designed by Roger Blackwell, G4PMK (see [12] for the original article) can be found at http://www.qsl.net/g3pho/ssa.html. It offers reasonable performance over the approximate range 1-90MHz, is fairly cheap to build and utilises almost any oscilloscope for its display.

POWER OUTPUT MEASUREMENTS

The UK Amateur Licence requires that you should be able to measure transmitter output power in order to comply with the licence conditions.

The following information is taken from the Amateur Radio Licence Terms and Limitations Booklet BR68. 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 be specified by the peak envelope power (pep).*

(e) *Interpretation*

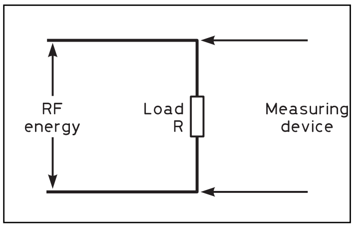
(i) Effective Radiated Power (erp): *The product of the power supplied to the antenna and its gain in the direction of maximum radiation.*

(ii) Gain of an antenna: *The ratio, usually expressed in decibels, of the power required at the input of a loss free reference antenna to the power supplied to the input of the antenna to produce, in a given direction, the same field strength or the same power flux-density at the same distance. When not otherwise specified, the gain refers to the direction of maximum radiation. The gain may be considered for a specified polarisation. The reference antenna is usually a half-wave dipole. The gain may be referred to as decibels relative to a half-wave dipole (dBd).*

(iii) 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 oscilloscope can be used up to about 30MHz to monitor modulated waveforms and measure output power, but above this it becomes an expensive item and may provide unwanted loading effects on the equipment being monitored. The familiar VSWR meter monitors forward and reflected signals and the scale can be made to represent power in a 50Ω line. It is possible to use an RF voltmeter across a given load to measure power. The higher you go in frequency, the more difficult it becomes to measure the modulation and power with relatively cheap equipment. Yet it is a condition of the licence that these parameters can be monitored. It is more difficult to measure PEP than average carrier power.

Fig 25.68: Power measurement arrangement



Constant-amplitude Signals

In a carrier-wave situation (CW, FM or unmodulated AM), the output is of constant amplitude and so it is relatively easy to measure the output power. To measure these signals using the circuit as shown on **Fig 25.68**, the power output is given by:

$$P_{out} = V^2 / R$$

where V must be the RMS value of the voltage.

This voltage measurement can be carried out using an oscilloscope or RF voltage probe. The SWR meter described later can also provide this value.

Amplitude-modulated Signals

These pose more of a problem and two cases are dealt with below.

Amplitude modulation (A3E)

With no modulation, the problem reverts to the measurement of power of a constant carrier as described above. If the carrier is amplitude modulated (A3E) 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 being one-quarter that in the carrier. It is suggested that for this form of modulation the carrier power is measured (ie no modulation) and multiplied by 1.5 to give the maximum output power.

If an exact value for the output power is required it is necessary to determine the modulation index. This can be carried out using an oscilloscope of adequate frequency response. Set the oscilloscope as shown in **Fig 25.69** and calculate the modula-

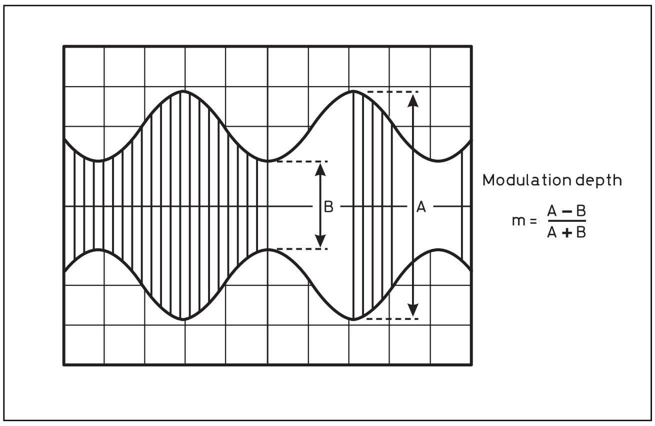
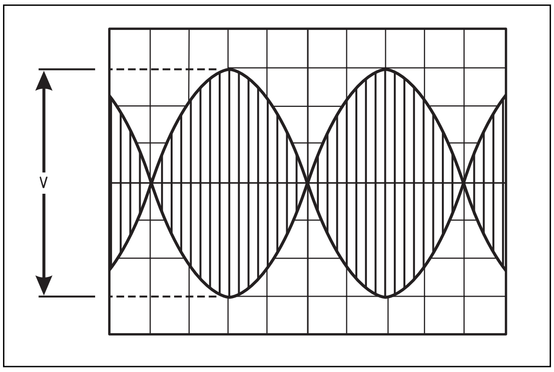
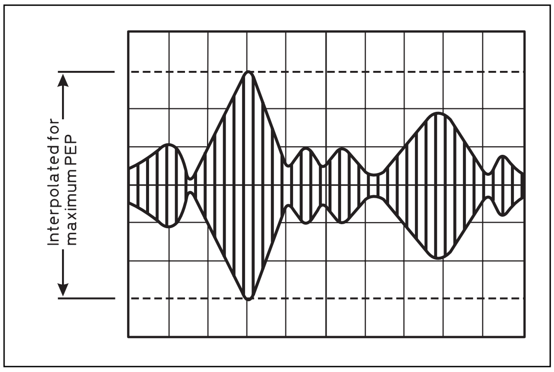


Fig 25.69: Modulation depth measurement



(left) Fig 25.70: Two-tone test display

(right) Fig 25.71: Speech waveform and interpolated maximum PEP level



tion depth m. The output power is then given by:

$$P_{out} = \frac{V^2}{R}\left(1 + \tfrac{1}{2}m^2\right)$$

where V is the RMS value of the unmodulated carrier and R the load.

For test purposes an audio signal can be fed in at the microphone socket of the transmitter using the low-frequency oscillator described earlier.

Single sideband (J3E)

With single sideband, 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 UK licensing authority. This can be accomplished using an oscilloscope of suitable frequency response as described below or using a SWR meter that will respond to peak envelope power - see later in this chapter or reference [1].

Fig 25.70 shows the display when a two-tone test signal is fed in via the microphone socket. If the peak-to-peak voltage V at the crest of the envelope is measured across a load of value R, then the PEP is given by:

$$P_{out} = \frac{V^2}{8R}$$

The equivalent peak-to-peak voltage reading can then be interpolated for the maximum allowable PEP and the position noted on the display. **Fig 25.71** shows a typical display for a speech waveform and the interpolated maximum PEP level.

VSWR

Every station should have a VSWR meter somewhere in its line-up. These can be bought commercially or made. When building, the higher the frequency range required, the more careful the constructor should be in placing the components so that the forward and reverse measuring circuits are symmetrical.

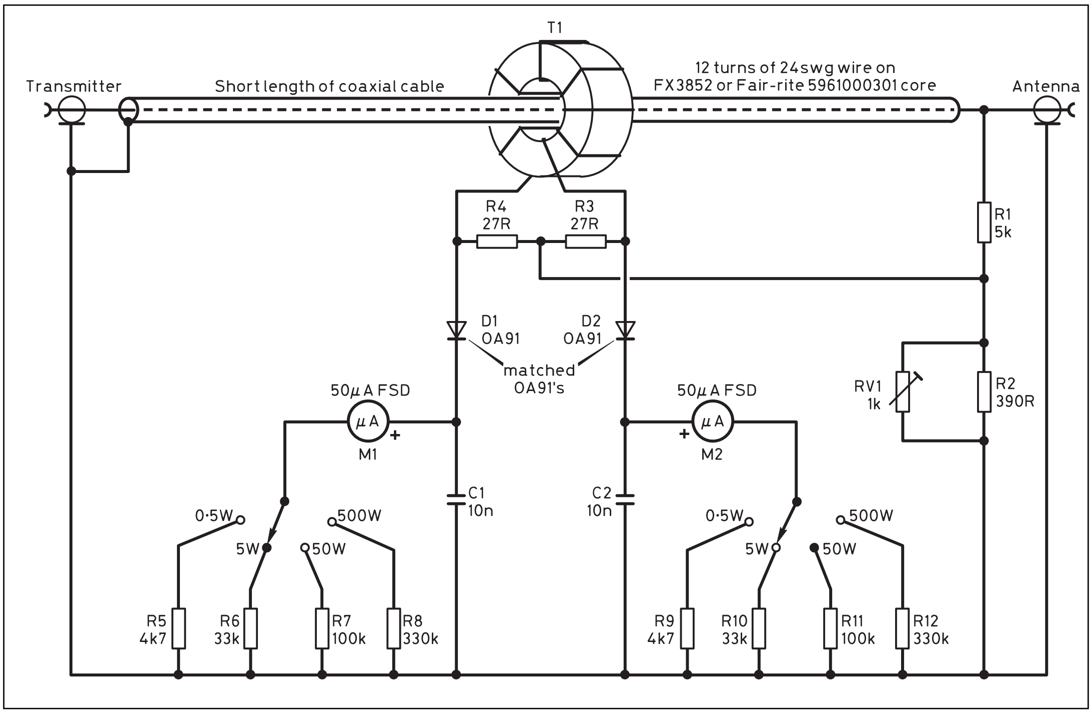
In theory a 1:1 VSWR is desirable but this is a condition that is often impossible to achieve. Looking at the problem from a practical viewpoint, it is worth trying to get a VSWR of better than 2:1 (equivalent to 11% reflected power). The guidelines shown in **Table 25.13** are suggested practical conditions and the actions that should be taken.

A VSWR METER

Reflectometers designed as VSWR indicators have normally used sampling loops capacitively coupled to a length of transmission line. This results in a meter deflection that is roughly proportional to frequency and they are therefore unsuitable for power measurement unless calibrated for use over a narrow band.

By the use of lumped components this shortcoming can be largely eliminated and the following design may be regarded as independent of frequency up to about 70MHz.

Fig 25.72: Circuit of the frequency-independent VSWR meter



VSWR	% reflected power	Comment
1 - 2.5	0 - 18	Solid-state transmitter SWR protection starts to operate, try looking for an improvement at the higher SWR value
2.5 - 5	18 - 45	Valve equipment probably OK but start looking for a problem or improve the SWR to get closer to 2:1
5 - ∞	45 - 100	Check the feed/antenna system; there is a problem!

Table 25.13: Guideline for various VSWRs

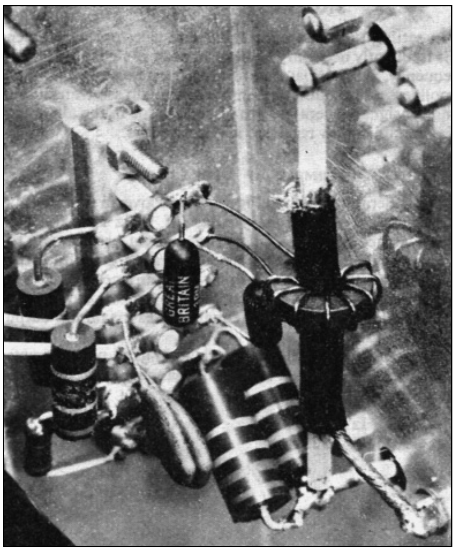


Fig 25.73: Construction of the frequency-independent VSWR meter

Circuit Description

The circuit is shown in **Fig 25.72** and uses a current transformer in which the low resistance at the secondary is split into two equal parts, R3 and R4. The centre section is taken to the voltage-sampling network (R1, R2, RV1) so that the sum and difference voltages are available at the ends of the transformer secondary winding.

Layout of the sampling circuit is fairly critical. The input and output sockets should be a few inches apart and connected together with a short length of coaxial cable. The coaxial cable outer must be earthed at one end only so that it acts as an electrostatic screen between the primary and secondary of the toroidal transformer. The layout of the sensing circuits in a similar instrument is shown in **Fig 25.73**.

The primary of the toroidal transformer is formed by threading a ferrite ring on to the coaxial cable. Twelve turns of 24SWG (0.56mm) enamelled copper wire are equally spaced around the entire circumference of the ring to form the secondary winding. The ferrite material should maintain a high permeability over the frequency range to be used: the original used a Mullard FX1596 which is no longer available but suggested alternatives are Philips FX3852 or 432202097180 and Fair-rite 5961000301, other types may also be suitable.

The remaining components in the sampling circuits should have the shortest possible leads. R1 and R2 should be non-inductive carbon types. For powers above about 100W, R1 can consist of several 2W carbon resistors in parallel. RV1 should be a miniature skeleton potentiometer in order to keep stray reactance to a minimum. The detector diodes D1 and D2 should be matched point-contact germanium types with a PIV rating of about 50V; OA91 diodes are suitable. The resistors R3 and R4 should be matched to 5% or better.

R1	5k carbon (see text)	C1, C2	10n ceramic
R2	390R carbon	T1	Philips FX3852, 4332202097180 or Fair-rite 596 1000301
R3, R4	27R, 2W carbon		
R5,9	4k7		
R6, R10	33k	D1, D2	OA91 (matched) - see text
R7, R11	100k		
R8, R12	330k	M1, M2	50μA FSD meters
RV1	1k skeleton pot., 0.5W	Switches	2 off, 1 p, 4w but good quality

All resistors are 0.25W, 5% tolerance unless specified otherwise

Table 25.14: VSWR meter components list

The ratio of the sampling resistors R1 and R2 is determined by the sensitivity of the current sensing circuit. As the two sampling voltages must be equal in magnitude under matched conditions, RV1 provides a fine adjustment of the ratio.

Germanium diodes as specified are essential if an instrument is to be used at low power levels, otherwise silicon diodes such as 1N914 or Schottky diodes such as the BAT85 may be substituted. To increase the sensitivity at low power levels, eg 1W, then the feed line could be looped through the toroid. It may then be necessary to use a large toroid or smaller coaxial cable (but this will not cope with high powers!).

A components list for this project is given in **Table 25.14**.

Calibration

Accurate calibration requires a transmitter and an RF voltmeter or possibly an oscilloscope. The wattmeter is calibrated by feeding power through the meter into a dummy load of 50wΩ. RV1 is adjusted for minimum reflected power indication and the power scale calibrated according to the RF voltage appearing across the load. The reflected power meter is calibrated by reversing the connections to the coaxial line.

The instrument has full-scale deflections of 0.5, 5, 50 and 500 watts, selected by the range switch. These should not normally be ganged since the reverse power will normally be much less than the forward power.

A SENSITIVE ANTENNA BRIDGE

The bridge (**Fig 25.74**) to be described (by G4COL [13]) enables antenna systems to be adjusted for minimum VSWR using single-frequency 'CW' signal powers as low as a few microwatts over a frequency range of 1.8 to greater than 60MHz (it should be

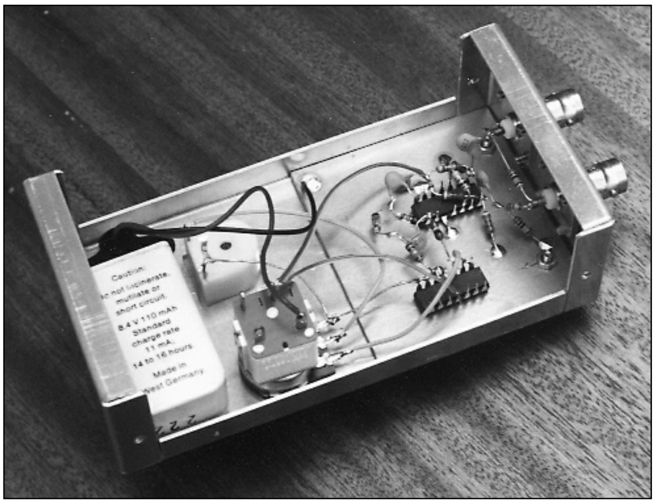


Fig 25.74: The frequency-independent VSWR meter

useable on 70MHz). It is most useful during extended periods of experimentation, where very-low-power transmissions should be used to avoid inconvenience to other band users and possible exposure of the experimenter to high radio frequency (RF) voltages or powers. This instrument is not intended to supplant the inline VSWR meter used for monitoring during normal transmissions. Its function is essentially the same as a 'resistance only' noise bridge but works with a low-power external signal source. It uses a built-in moving-coil meter to monitor the signal and does not require a receiver.

The bridge is so simple that it can be built in a few hours. Suitable signal sources include a signal generator, crystal oscillator, dip meter and transmitter plus attenuator. An outline specification is as follows:

- Operating signal power: 25 to +3dBm (3mW to 2mW) at 10MHz, for full-scale meter deflection
- Frequency response: less than 2dB variation from 1.8 to 65MHz
- Power supply: 9V PP3 battery with current drain of 18mA

Achieving Sensitivity

The bridge uses conventional diode detection. The most common method of driving a meter is to follow the diode detector by a DC amplifier as shown in **Fig 25.75(a)** which shows a 'system gain' G, which is the meter current produced by a given radio frequency input voltage (or power). In **Fig 25.75(b)**, the same system gain has been obtained by amplifying the RF signal prior to detection.

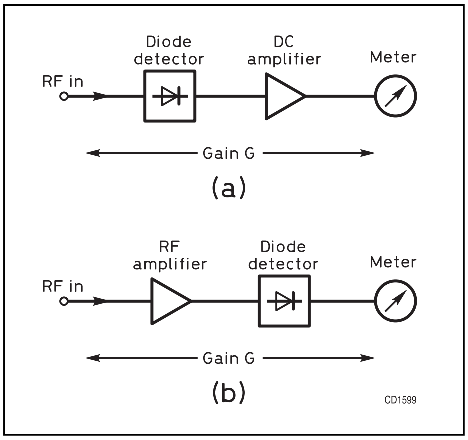


Fig 25.75: (a) Diode detector followed by DC amplifier. (b) RF amplifier followed by diode detector

Given enough RF input, the two methods produce the same result in principle. The first approach has a number of advantages: the DC amplifier does not affect the frequency response, is simpler and less dependent on layout.

However, at low RF levels, such as the operating range of this bridge, the diode exhibits a 'threshold effect' whereby its sensitivity (and so also the system gain) falls with signal level and the two approaches cease to be equivalent. The effect of this on an antenna bridge is to produce an erroneously wide null that masks the point of optimum match.

The bridge uses the method of Fig 25.75(b) and takes advantage of one of the low-cost, readily available RF (or video) integrated circuit amplifiers currently on the market. The amplifier boosts the RF signal level to the point where efficient detection can take place.

Circuit Description

The circuit is shown in **Fig 25.76**. Resistors R1 and R3, together with the impedance connected to SK2, form a bridge that will be perfectly balanced by a 51Ω resistive load. The bridge voltage is amplified by differential RF amplifier IC1 and detected by a full-wave rectifier using diodes D1 and D2. The differential amplifier obviates the need for the usual balun transformer. The rectified voltage is buffered at DC by IC2a/IC2b and at RF by R9/R10. Potentiometer R11 allows the meter sensitivity to be set.

Overall sensitivity can be further altered by changing the RF amplifier gain-setting resistor R8. Raising its value lowers the gain but increases the bandwidth, and conversely for a lower value. If working in a different system impedance, such as 75Ω, R1 to R3 should be made equal to this value.

Constructional Notes

The component list is given in **Table 25.15**. First prepare the case by drilling holes for the BNC connectors, potentiometer and panel meter (see photograph, Fig 25.74). The meter cut-out can be made by drilling a pattern of small holes and filing to final size, or by opening up a circular hole using small files. Next cut a piece of copper-clad board approximately 65mm square and hold it inside the case at the socket end. Drill two 3mm holes through the case and the board close to the BNC connectors. Once drilling and de-burring are complete, apply dry transfer let-

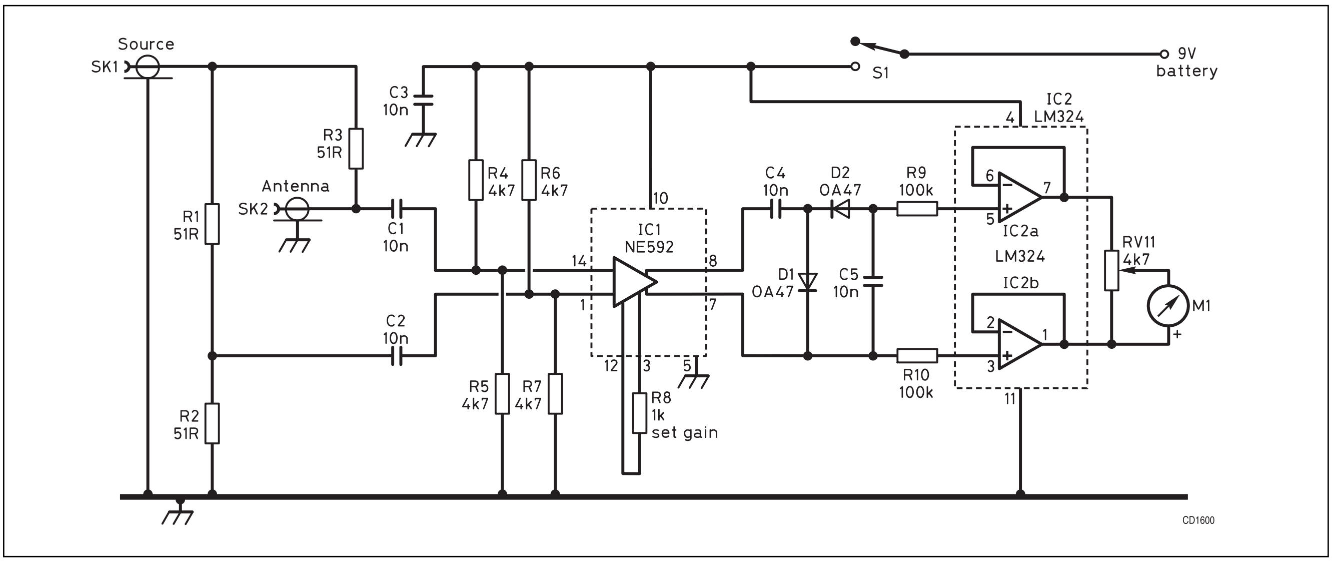


Fig 25.76: Circuit diagram of the bridge

R1, R2	51R	C1-C5	10n ceramic disc, 12V or greater
R3, R12	51R		
R4, R5	4k7	D1, D2	OA47, OA91 or similar germanium or BAR21, BAT42 Schottky diode
R6, R7	4k7		
R8	1k		
R9, R10	100k	IC1	NE592
R11	4k7 linear pot	IC2	LM324

M1 250µA FSD meter
SK1, SK2 BNC chassis sockets or other preferred type such as SO239
Aluminium box, suggested size 133xx70x38
PP3 battery and connector
Insulated wire
Knob
Dymo tape or dry transfer lettering or similar
Velcro for securing battery
Single-sided copper-clad circuit board
M2.5 x 6mm screws (10 off)
M2.5 nut (2 off)
Double-sided tape for securing meter
All resistors are 0.125W or greater, metal film

Table 25.15: Components list for sensitive antenna bridge

tering if desired, and spray the case lightly with lacquer to protect the surface. Clean the copper board and fix it in place with two M2.5 screws and nuts using the 3mm holes. Mount the potentiometer and BNC connectors. The meter can be fixed to the case with double-sided adhesive tape. Self-adhesive Velcro strips hold the 9V PP3 battery in place.

As can be seen in the photo, the circuitry can be built up quickly and easily on the copper board using 'ugly' style construction. Start by positioning the ICs with their ground pins bent down to touch the copper and solder them down - place IC1 quite close to the BNC sockets. The other components can then be added at will, finishing off with the few connecting wires.

Operation

Other than the sensitivity panel potentiometer, there are no adjustments. After checking the wiring, connect the battery and switch on. Applying a CW signal within the specified range (see introductory paragraphs) to the source socket with nothing connected to the load socket should give a meter deflection which can be adjusted to full scale using the potentiometer. *Do not* connect a transmitter output directly to the bridge as this will damage the unit. If a 51Ω resistor is placed across the load socket, the meter reading should drop to zero.

The meter reading is proportional to reflection coefficient: for instance, half-scale corresponds to a reflection coefficient of 0.5. This is related to VSWR by:

$$\mathrm{VSWR} = \frac{1+\rho}{1-\rho}$$

where ρ is the reflection coefficient.

This means that the half-scale reading corresponds to a VSWR of 3. As the RF signal level falls, accuracy at low VSWRs will decline due to the fall-off in diode detector sensitivity. Calibration, if required, can be established with a set of resistors. For example, a 75Ω resistor presents a VSWR of 1.5, corresponding to a reflection coefficient of 0.2 which would ideally give a meter reading of 20% of full scale. Use carbon composition (check the resistance carefully), metal film or carbon film

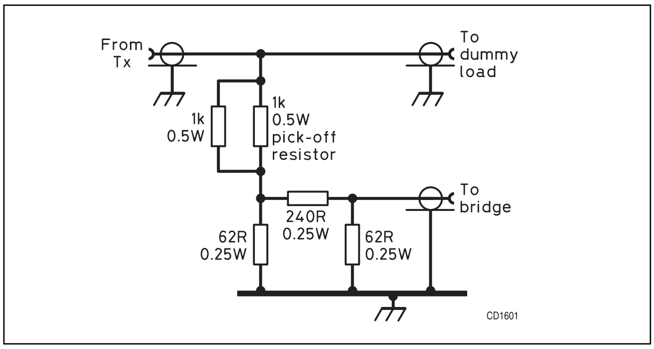


Fig 25.77: Signal pick-off

resistors for calibration and avoid wire-wound type since these have significant inductance.

Obtaining a sufficiently low power using a transmitter can be a problem and for this a signal pick-off, as shown in **Fig 25.77**, can be used. The values shown are suitable for up to 10W (40dBm), and attenuate the signals reaching the bridge by about 40dB. Make sure that the dummy load can dissipate the transmitter power. Below 100mW (20dBm) a low-power attenuator can be used directly without a dummy load and signal pick-off.

When making measurements, as opposed to simply using the bridge to indicate a dip for best match, remember to set the meter reading to full scale with an open-circuit or short-circuit at the load socket.

POWER METERS

Described below is a Precision Peak-following Power Meter but the reader may also like to consider units in chapter 6 of reference [1], and the 'Crawley Power Meter' by G3GRO and G3YSX [14].

A Precision Peak-following Power Meter

The following is an abridged description of a power meter by G3GKG [15] that requires no setting up and no adjustment during use. The meter covers the HF bands and copes with powers up to about 450W. A finished unit is shown in **Fig 25.78**.

The heart of this instrument is the type of coupler known variously as a Tandem Match, a bi-directional coupler and a 4-port hybrid transformer.

(NOTE: To quote from a reference by G4ZNQ:[16]:

"A hybrid is a very simple circuit - just two transformers and four connectors - with some amazing properties. The connectors or ports are best thought of as two pairs. If a signal is fed into one connector and out of the other

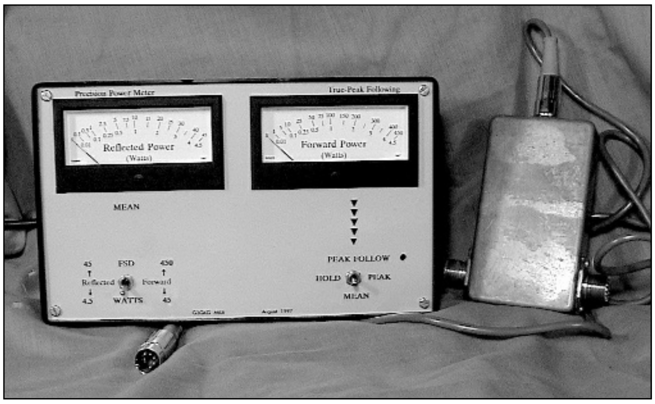


Fig 25.78: Precision peak-following power meter

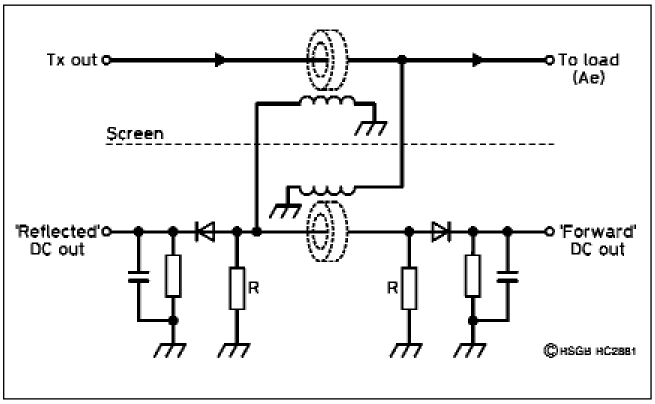


Fig 25.79: The Tandem Match - symmetrical between input and output

of a pair, into some unknown impedance load (say an antenna) then, if both the other connectors are terminated in the intended system impedance (say 50Ω), the hybrid feeds a fraction of the power passing forwards through the first pair of connectors into one of the terminations."
"It feeds an equal fraction of the reverse power flow into the other termination. Hybrids can be designed to have different sampling fractions, usually quoted in decibels, so that a 20dB hybrid diverts 1% of the flowing power to the appropriate terminated port".)

Various authors have suggested different coupling factors according to application and by juggling the figures around, G3GKG derived some simple formulas to produce an essentially standard output voltage which can be defined for any particular system. (Calculations made during the course of the development of the power meter are provided as spreadsheets on the RSGB Members Only web-site [17]. The spreadsheets may be freely downloaded and used to assist readers to develop their own circuits for use at different powers and with different meter sensitivities.)

Using the correct type of toroid core (which must be of high permeability ferrite), this is a precision circuit (Fig 25.79) which produces voltages at both the Forward and Reflected output ports which are strictly and predictably defined by the RF power, the designed load resistance and the number of turns on the secondary winding of the toroid. Used with the amplifier and display units to be described, the calibration is constant throughout (at least) the HF range of the amateur frequency bands and is accomplished completely and accurately just by using the calculated design parameters.

It is readily apparent from Fig 25.79 that the circuit is completely symmetrical and this is indeed borne out by its performance. Reversing the transmitter and aerial connections merely causes the Forward and Reflected output ports to interchange positions, as does reversing the connections to one or other of the toroidal windings.

RF to DC

The display unit (Fig 25.78) uses individual meters for Forward and Reflected power, so the outputs are brought out from the RF section separately, after rectification and buffering.

The Forward metering circuit has been designed to accept a DC voltage range close to an optimum of 10V FSD. To set the power range, it is therefore arranged for the coupler to produce this voltage from the designated maximum forward power (peak), by first finding the required number of turns on the toroids to produce 10V as closely as possible from that power,

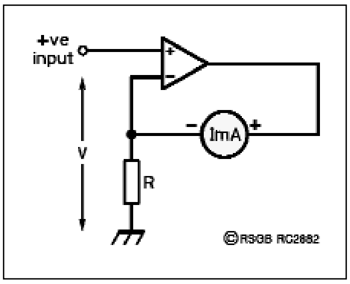


Fig 25.80: The basic meter circuit, showing the meter in the feedback loop of the op-amp driver

and then calculating the actual precise voltage for that number of turns. (Fractional turns cannot be wound on a toroid!)

With a 10V range and the employment of Schottky diodes for both RF rectification and for an op-amp linearising circuit, the errors are reduced to negligible proportions, the DC output tracking the RF voltage accurately down to about 30mV (representing a power level of 18μW with a 50Ω load) with little deviation well below that. It is important for this tracking that the pair of diodes in each of the detector/op-amp circuits is initially matched regarding forward voltage drop and also that they remain at the same ambient temperature during use.

The same operational amplifier also provides a convenient low impedance DC output from this part of the circuit to the main Display Unit, allowing the RF unit to be constructed in a separate housing which also caters for the requirement regarding ambient temperature. The RF Head Unit can then be installed in the direct coaxial line between transmitter and aerial matching unit, well away from the main measuring and display instrument, which can therefore be located in the optimum position for viewing.

Metering

Each meter circuit incorporates the meter itself in the feedback loop of an op-amp driver, Fig 25.80. The voltage range is determined only by the current range of the individual meter (irrespective of its inherent resistance) and the scaling resistor, R, the value of which is given simply by dividing the actual full scale voltage required, V, by the nominal full-scale deflection (FSD) current sensitivity, I, of the meter. For supreme accuracy, the FSD current can be individually measured and used in the calculation.

As the whole of the Head Unit circuitry is completely symmetrical (as Fig 25.79 shows), the reflected output voltage of the coupler for the same full-scale power would, of course, be the same, at 10 volts, but we can select a range for that metering circuit of something less - ie an FSD which is more commensurate with the maximum reflected power likely to be encountered - bearing in mind the square law relationship which dictates that half the full scale voltage represents a quarter of the power. If the Forward meter is calibrated so that 10V FSD represents 400W, a companion Reflected meter calibrated for 5V FSD will read up to 100W, which would be equivalent to an SWR of 3:1.

It is also convenient and easy with this degree of sensitivity to provide accurate, alternative, very low power ranges, so that the transmitter and aerial system can be tuned and matched with minimum chance of causing interference. This entails incorporating a two-way toggle switch to select different values of scaling resistors which set the scaling of both meters appropriately.

For a station using the full legal limit, convenient ranges might be -

Forward Power: 450W or 45W;
Reflected Power: 45W or 4.5W

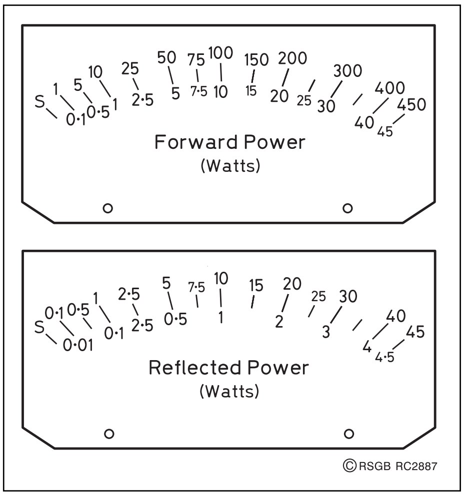


Fig 25.81: Meter scales showing the common calibration marks for the two power ranges

thus enabling a single set of scale markings to be calculated and used for all ranges, with different figures on the two meters such as those shown on **Fig 25.81**.

Consider now the actual scaling resistors. Because the voltage ranges are all calculable, these can be fixed components which, although unlikely to be easily available in the exact values, can be made up from suitable series or parallel combinations of 1% tolerance resistors. Alternatively, by using pre-set variable resistors, the completed instrument could be calibrated using precise values of DC voltage injected at the input socket to the main unit in place of the output from the head unit.

All the calculations are available on spreadsheets - see comment earlier. These are for determining the required number of turns on the toroids to suit a given power range, the actual output full-scale voltage, values of scaling resistors (R1 - R4) for all four chosen ranges with particular meter movement sensitivities (with the facility of calculating series combinations of preferred value resistors to produce the value required) and the meter needle deflections in degrees for the required calibration points.

Peak Reading

Many commercially-designed power meters include a function labelled 'Peak' or 'PEP'. Occasionally, such a meter will indicate something more or less close to the true peak power, at the expense of a very long decay time. Most of those which are manufactured as separate items, or included in an aerial matching unit, do not and cannot measure or indicate the instantaneous peaks of an SSB speech waveform because they do not include any active circuitry. It is frequently maintained that if the unit requires an external power supply it will most likely contain such a refinement and be capable of capturing these peaks, but this can be a snare and a delusion. The power supply frequently serves only to provide the illumination for the meter(s)!

Let's be sure what we are talking about. With a continuous carrier, which is of course just an RF sine wave, the peak power we want to measure is, in fact, what we know as the peak envelope power or PEP. The picture on a monitor scope will show a solid band (the envelope), the vertical width of which varies with the output power of the transmitter and clearly illustrates that, in this sense, the mean and peak powers are the same. Any of the instruments on the market should produce the same reading in either measuring mode. An SSB speech waveform, on the other hand, shows a band of power that is constantly varying at a syllabic rate and further illustrates that, without compression or other processing, the power only reaches local maxima for very brief periods of time. It is these brief peaks that we wish to measure when we refer to 'peak' power - more properly called the instantaneous peak power. When we switch to read 'mean' power, we want to know the average power output over a period of time, and we need the meters to read in this mode when tuning or adjusting the rig. With a properly adjusted transmitter in the SSB mode (ie no compression), the mean power will be quite small compared with the peak power.

The peak-reading circuits normally encountered all employ the same sort of 'diode-pump-charging-a-capacitor' circuitry, with varying degrees of sophistication designed to overcome the inherent drawbacks of the circuit. In order to capture brief peaks accurately, the diode non-linearity and knee voltage must be

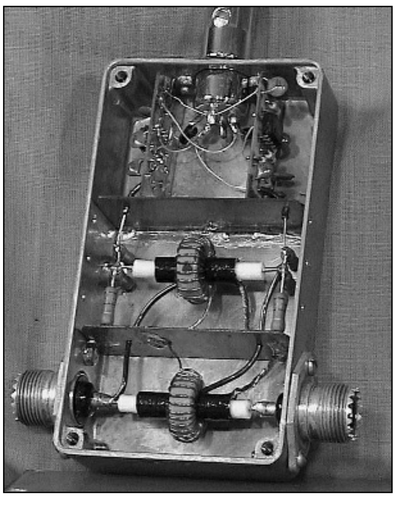


Fig 25.82: Inside the RF head unit

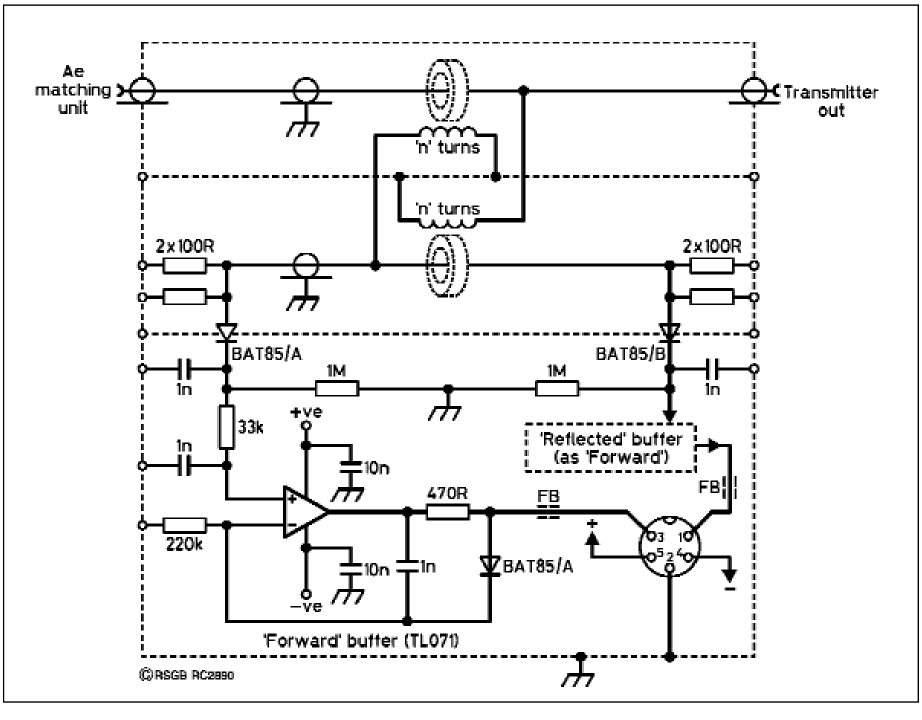


Fig 25.83: Circuit of the head unit

overcome, the charging circuit must have a fast attack (implying a very low source impedance) and it must have a decay time long enough to enable the peak to be read - often accomplished by incorporating a 'peak-hold' feature which removes the normal discharge resistor. In normal SSB use, most of them are inevitably very sluggish in their response.

In order to follow the peaks of an SSB signal, either at a syllabic rate or by capturing the peak amplitude in each short phrase of speech and still provide time to read the meter, it is necessary to provide a fast attack, a preset 'hang' period and a rapid decay, in order to capture the next peak. I have tried several approaches to this idea and the most successful is one using a sample-and-hold chip, type LF398 (of which the N version is the best in this application).

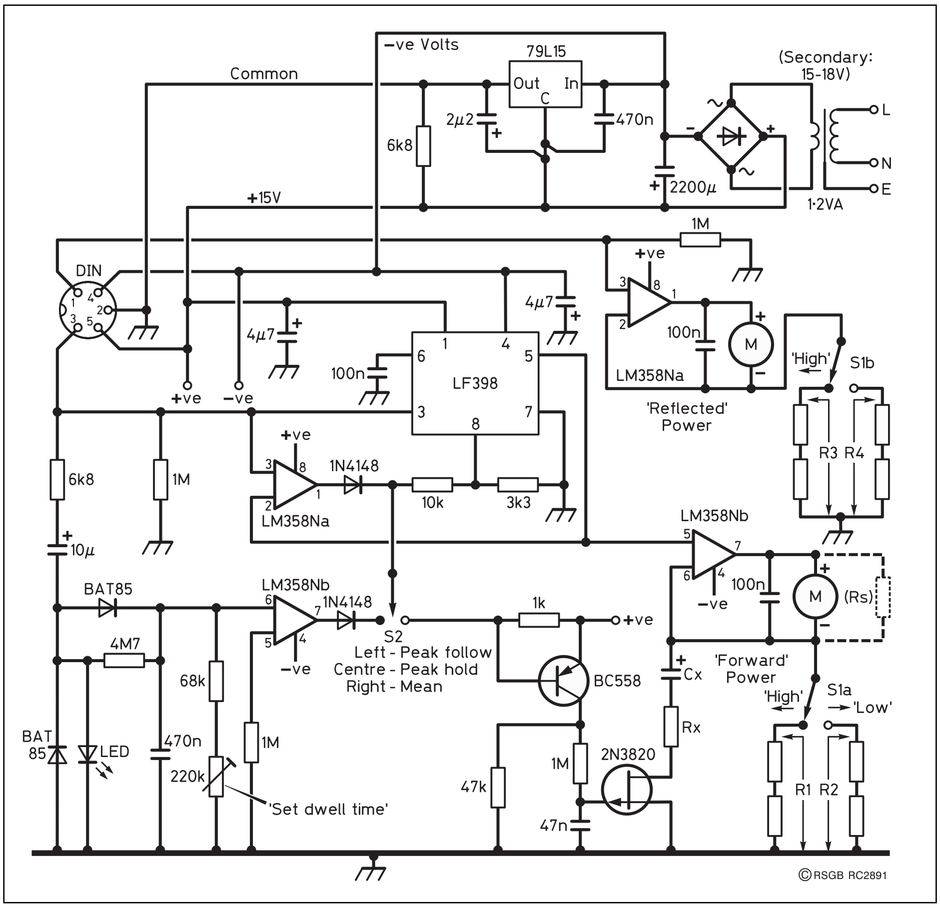


Fig 25.84: Complete circuit of the main display

Construction

It was not the intention that this description should provide full constructional details - everyone will have their own requirements and preferences. As can be seen in the photograph, the Head Unit (**Fig 25.82**) is built into a die-cast aluminium box measuring about 115 x 62 x 29mm and includes the tandem match and detection circuitry (**Fig 25.83**). It is connected to the main display unit (Fig 25.78) by a 5-pin DIN to 5-pin DIN with screened lead. Thin double-sided copper-clad fibreboard is used in the construction of the screened compartments and for the TL071 linearising buffer amplifiers (with the rectifier diodes passing through holes in the final screen).

The two ends of the second coax-toroid assembly are supported by a pair of orthogonally-mounted 100Ω, 2W metal film resistors. The toroids used so far have been generously proportioned with Al ratings in the 1800 to 2000 region (eg Electrovalue type B 64290K 632X27), but I suspect from more recent testing that rather smaller ones (...45X27) would serve equally well. Offset adjustment for the op-amps has not been found necessary but, especially if very low ranges of power are required, could be accomplished by connecting an experimentally determined, high-value resistor from the negative supply to either pin 1 or pin 5.

Obviously, the size and type of housing required for the display unit will be dictated largely by the choice of meters used. It should ideally include the rest of the circuitry (**Fig 25.84**) apart from possibly the power supply unit.

(*NOTE:* A negative rail is required by some of the ICs but, because all the actual signals are positive going, it is not required to be more than a few volts and need not be stabilised.)

Author's Note: From my experience with several different types of meter, most are over-damped (ie the response to a step change in input is too slow, with the needle creeping over the last few percent of its swing). This response is the easiest sort to improve, but the compensation must be done very carefully and precisely so as to prevent over-swing with consequent false readings. It is only required in the peak-following mode and only on the Forward power meter, as the Reflected meter always reads mean power. The required components form a series combination of resistance (Rx) and capacitance (Cx), switched into circuit in the appropriate mode, across the calibrating resistor of the op-amp (see Fig 25.84). Part of the resistive component is the 'on' resistance of the FET (which, with the associated transistor, performs the necessary part of the switching function from 'peak' to 'mean' reading), whilst the other values are determined for the particular meter, as follows.

Leave out these components until the instrument is completed and working. Then, remove the plug coming from the head unit and connect a signal from a rectangular-wave generator to the display unit, (between the forward voltage pin and common of the input socket). Use about 8V positive-going with a mark/space ratio about 1:1 and a repetition rate of about 1Hz. With the switches set to the High power range and Peak (follow), you will then need to adjust the values of both resistor (Rx) and capacitor (Cx) until the meter follows the amplitude excursions as fast as possible without over-swinging on either the upward or downward swings - it is easier than it sounds!

My own Mk4 instrument uses 1mA Sifam meters (from a surplus source) and the compensation consists of two 6.8µF capacitors in parallel together with just the 'on' resistance of the FET. Exceptions to my earlier statement, these meters are in fact basically under-damped so that the Forward one also had to be shunted (Rs) to achieve the required critical damping, thus altering its sensitivity and requiring changes to the scaling resistors. Whatever the type of meter, these values must be critically determined.

The original article [15] also had a variant on this design using a digital bargraph instead of meters.

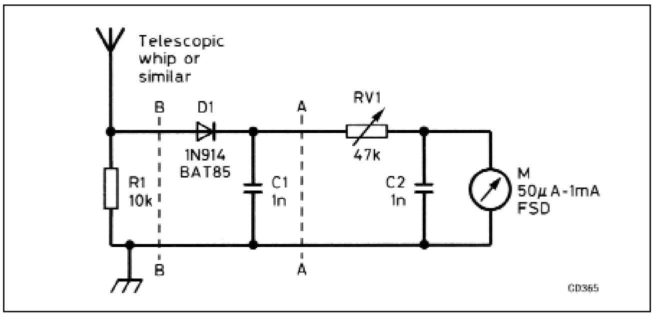


Fig 25.85: Simple untuned field strength meter

FIELD STRENGTH METERS

Field strength meters used by amateurs are normally used as indicators to maximise the radiated power and not to make an actual field strength measurement at a receiving site. The absorption wavemeter or dip oscillator (in absorption mode) can be used for this purpose. These units may need some form of external telescopic antenna to be fitted. However, the use of a tuned circuit is sometimes inconvenient as no attempt is being made to differentiate between wanted and unwanted transmitted signals (this should already have been dealt with at the transmitter!).

The alternative is to use a simple type of system such as that shown on **Fig 25.85**. A signal is picked up by the antenna, rectified and smoothed by D1/C1, the resulting DC signal is then indicated on the meter M1, with RV1 acting as a sensitivity control. Capacitor C2 provides an AC short across the meter for any unwanted RF signals.

Construct the unit in a box, using either a telescopic whip or a loop of wire. The unit can be used for relative field strength measurements at a given frequency. It should not be used for relative measurements between different frequencies as the efficiency of the antenna and rectifier will affect readings. It should be a useful device for tuning a transmitter to obtain maximum radiated power. It could also be used for adjusting an antenna for maximum radiated power, provided the unit is far enough away from the antenna. By splitting the circuit at AA, the antenna/rectifier combination could be used as a remote reading head, with the meter/sensitivity control being in the shack.

Higher Sensitivity Broadband Field Strength Meter

The concept here is to amplify the received signal first and then to detect it to drive a meter. This can be accomplished by using one of the relatively inexpensive broadband RF amplifiers such as the MAR series from Mini circuits; there may be alternatives on the surplus market.

These amplifiers would be placed in the circuit of Fig 25.85 at position BB. They should be constructed on circuit board with a good ground plane. All components should have short leads to minimise lead inductance and the capacitors carefully chosen for the frequency range envisaged. These devices generally have outputs of the order of +10dBm, it is therefore imperative that diode D1 is of a type with low forward volt drop such as Schottky type BAT85. If the field strengths being measured are very low, it may be possible to cascade two or more such amplifiers.

Fig 25.86 shows a circuit based around a MAR8 monolithic amplifier produced by Mini Circuits. The circuit has a response from DC to 1GHz with a quoted gain at 100MHz of 33dB and 23dB at 1GHz; the maximum output is about +10dBm. The device requires 7.5V at 36mA, the circuit shows a series resistor for operation from a 9v DC supply.

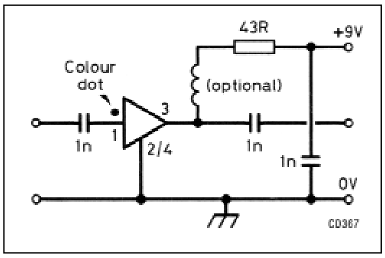


Fig 25.86: Broad-band amplifier based around a MAR8

It is also worth consulting reference [2] where other circuits are suggested as well as the use of a communications receiver.

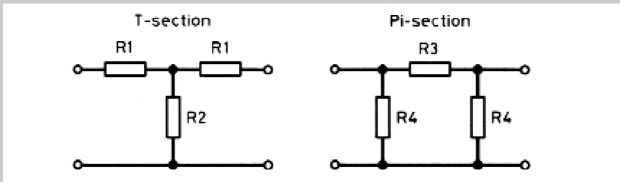
ATTENUATORS

Attenuators are useful for receiver measurements, especially when testing from signal generators that have minimal or no attenuators within them. The greatest problem 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. An attenuator might also be useful between a transmitter and transverter.

Attenuators are normally made from Pi or T networks. For this exercise it is assumed that load and source impedances are equal. See **Table 25.16** for component values for 50 and 75-ohm source/loads.

The circuit shown on **Fig 25.87** is a low power switched attenuator using a combination of Pi and T networks, this enables preferred value resistors to be used. A 1-2-4-8-...dB switching sequence is used so that the maximum attenuation range is obtained for a given number of sections.

The switches used are a standard wafer type panel mounting slide switches, for which the effective transfer capacitance in the circuit used is only 0.8pF. With 5% carbon film resistors the attenuator accuracy is ±0.5dB on the 1,2,4 and 8dB positions



	R1 (Ω)		R3 (Ω)	
dB	(50)	(75)	(50)	(75)
1	2.9	4.35	5.7	8.5
2	5.73	8.6	11.6	17.4
4	11.3	17.0	24.0	36.0
8	21.5	32.3	52.9	79.2
16	36.3	54.4	154.0	231.0
32	47.5	71.1	1.0k	1.5k
64	50.0	75.0	39.6k	59.3k

	R2 (Ω)		R4 (Ω)	
dB	(50)	(75)	(50)	(75)
1	434.0	650.0	870.0	1.3k
2	215.0	323.0	436.0	635.0
4	104.0	156.0	213.0	320.0
8	47.3	71.0	116.0	174.0
16	16.25	24.4	69.0	103.5
32	2.5	3.75	53.0	79.5
64	0.075	0.113	50.0	75.0

Table 25.16: Resistors for 50-ohm and 75-ohm attenuators based on pi and T sections

Fig 25.87: Circuit of a 50Ω attenuator using preferred values of resistor

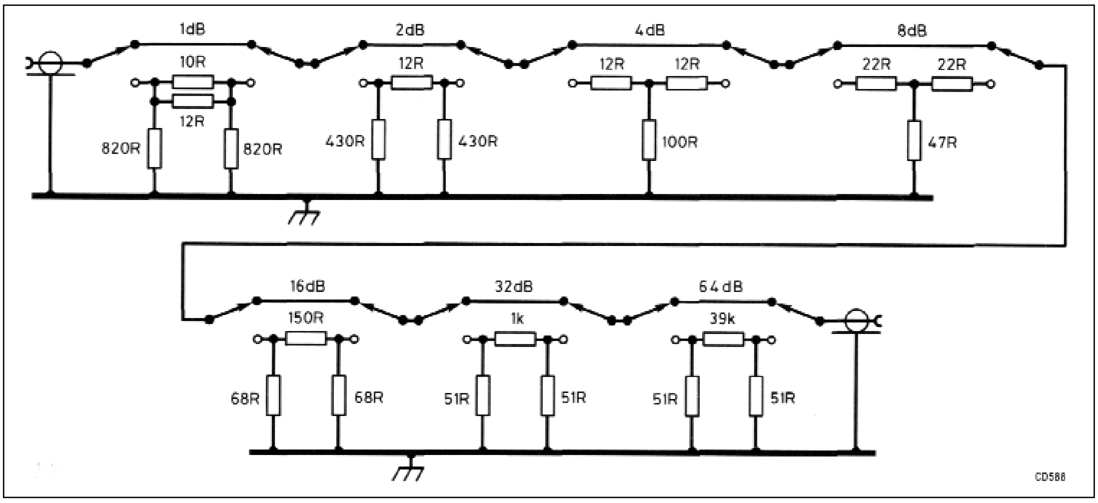
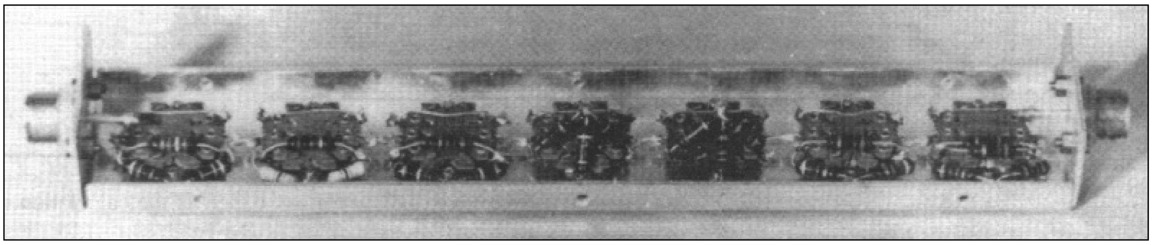


Fig 25.88: Construction of the 50Ω attenuator



up to 500MHz. The 16dB position is 1dB low at 500MHz, the 32dB position 1dB low at 30MHz and the 64dB position 1dB low at 750kHz. The photograph on **Fig 25.88** shows the construction of the unit.

An alternative design is given on page 51 of reference [2].

A CLIP-ON RF CURRENT METER

This article is from 'In Practice' in *RadCom* [18], see this article for earlier references. The basic version of this handy device takes about 10 minutes to tack-solder together (**Fig 25.89**). When you're convinced how useful it is, you can then go on to build a more permanent version. The clip-on RF current meter has a long history, early versions involved breaking a ferrite ring into two equal pieces - which takes some doing! The constructional breakthrough was G0SNO's idea to use a large split ferrite bead intended for HF interference suppression. This clamps around the conductor under test, to form the one-turn primary of a wideband current transformer. The secondary winding is about 10 turns, and is connected to a load resistor, R1-R2, and the diode detector.

The load resistor, R1-R2, is important because it creates a low series impedance when the current transformer is effectively inserted into the conductor under test. For the values shown in Fig 25.89. (10-turn secondary, 2 x 100Ω) this is $50/10^2 = 0.5\Omega$. Some circuits omit this resistor, but that creates a high insertion impedance - exactly the opposite of what is needed. Also, more secondary turns create a lower insertion impedance, but at the expense of HF bandwidth.

The other components in Fig 25.89. are discussed in G0SNO's article which is reproduced on the 'In Practice' website [19]. Component types and values are critical only if you want to make a fully calibrated meter with switchable current ranges. However, for a first try, and for most general RFI investigations, the meter is almost as useful without any need for calibration. Make R4 about 4.7-10kΩ, and omit R3 and S1. If the meter is either too sensitive or not sensitive enough, either change R4 or change the HF power level.

Just about any split ferrite core intended for RFI suppression will do the job, but there are a few practical points. Choose a large core, typically with a 13mm diameter hole. This allows you to clip the core onto large coax, mains and other multi-core cables while still leaving enough space for the secondary winding (which should be made using very thin enamelled or other insulated wire). It is important that the core closes with no air gap - and that can be a problem. A major disadvantage of the basic split ferrite core in its plastic housing is that the housing is not meant to be repeatedly opened and closed, so the hinge will soon break. By all means try out this gadget in the basic form but it is likely that you will soon be thinking about something more permanent. The classic way to do this is using a clothes-peg **Fig 25.90** but there are now better alternatives.

For example, the first photograph (Fig 25.90) shows the rather heavy-duty version using two strong clothes-pegs, fibreglass sheet and epoxy glue (more details at [19]). The second photograph shows GI0XAC's neat and simple version using a giant plastic paper-clip, with a small plastic-cased meter stuck on the side. The only requirement of the clip is that it must be basi-

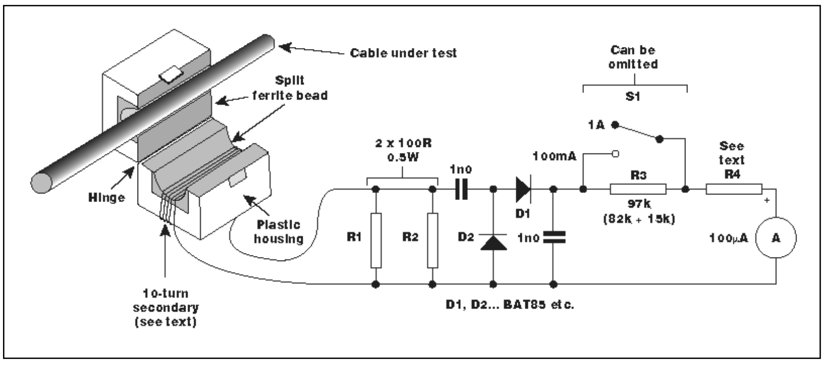


Fig 25.89: G0SNO's clipon RF current meter

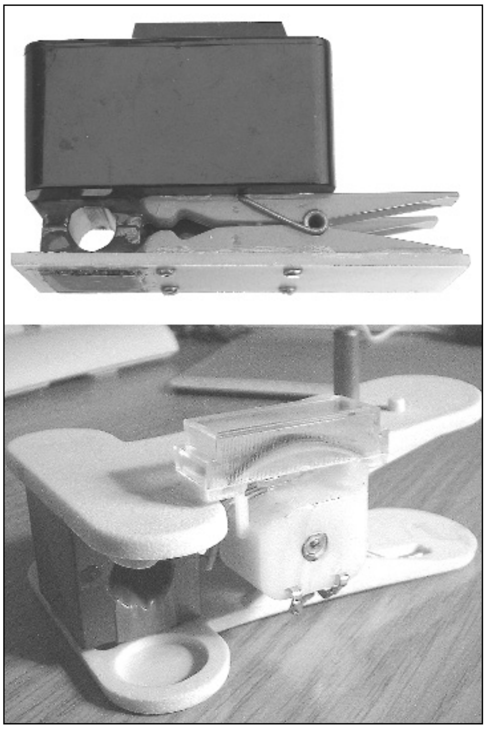


Fig 25.90: Two versions of the clip-on RF current meter

ATA	India	10,000kHz
DCF77	Germany	77.5kHz
CHU	Canada	3330, 7335, 14,670kHz
HLA	Korea	5000kHz
MSF	UK	60kHz
LOL2	Argentina	5000, 10,000, 15,000kHz
RWM	Russia	4996, 9996, 14,996kHz
TDF	France	162kHz
WWV	USA	2500, 5000, 10,000, 15,000, 20,000, 25,000kHz
WWVB	USA	60kHz
WWVH	USA	2500, 5000, 10,000, 15,000, 20,000, 25,000kHz

Table 25.17: A selection of standard frequency transmissions

cally non-metallic, and that it can hold the two halves of the core accurately together while the whole weight of the meter is dangling from the cable. Another option worth investigating would be the pliers-style plastic work clamps that are sold in a range of sizes by hobby shops. Whatever you use, it is vital that you glue the two halves of the core to the clip in such a way that they always close tightly together with no air gap. Hint: glue one half of the core to one side of the clip first, and let that side set; don't try to glue the second half until the first is good and solid.

A clip-on RF current meter could hardly be simpler to build. It's an ideal project for beginners and clubs. Once upon a time, every amateur station was required to have an absorption wavemeter, which achieved almost nothing; if every amateur station today had a clip-on RF current meter, we'd see a lot less RFI and a lot more confidence about going on the air!

STANDARD FREQUENCY SERVICES

There are various standard frequencies transmitted throughout the world and these can be harnessed in order to check other equipment against them. Typical of these transmissions are those shown in **Table 25.17**.

These standard frequencies are maintained to an accuracy of typically one part in 10^{11}. However, if the sky-wave is used there could be a large error in reception due to Doppler shift and there will be fading of the signal. These problems can be avoided by using a low-frequency transmission such as those from MSF or WWVB. Timing information is also impressed on the signals in either GMT or UTC.

In addition in the UK the BBC maintains the accuracy of the Droitwich 198kHz (formerly 200kHz) carrier to high accuracy - on a long-term basis being 2 parts in 10^{11}.

An additional method that may be available (providing the TV radiates well!) is the line timebase of a TV and the associated harmonics. The line timebase is at 15.625kHz and harmonics may well be present way up into the HF bands - the 64th harmonic is 1MHz and this would be accurate to ±0.01%. However, newer generations of TV sets may not provide such a good source if additional screening has been added to help minimise extraneous radiations.

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