24 Measurement and Test Equipment

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When considering the revision for the 4th edition of *Test Equipment for the Radio Amateur* [1] and for this chapter it became even more obvious that there was a wealth of information on the Internet about measurements and test equipment, both in the amateur and professional fields. There are various kits and projects worth considering as well as those contained within this chapter. Internet auction sites such as eBay are a good source for buying test equipment as well as the familiar radio rallies. Have a look around the web using some of the section headings in this chapter and it is surprising what can be uncovered. This chapter can only try to give some of the fundamentals for measurements and test gear.

Correct operation of amateur radio equipment involves measurements to ensure optimum performance in order to monitor how equipment is functioning, comply with any 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 some items such as multi-meters (analogue and digital), frequency counters and oscilloscopes cost less than commercially bought amateur radio equipment. It is the intent of this chapter to show how some of the cheaper equipment (and maybe home-built) can be used to good effect.

For further information, equipment and a more detailed discussion of some of the topics in this chapter, the reader should consult [1]. Useful material is also contained in [2] and [3]. The chapter in this book on Low Frequencies may also contain useful information.

One word of caution - whilst most of 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 in as a search on the Internet.

CURRENT AND VOLTAGE MEASUREMENTS

Many electrical measurements 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.

Fig 24.1: Typical analogue and digital multimeters

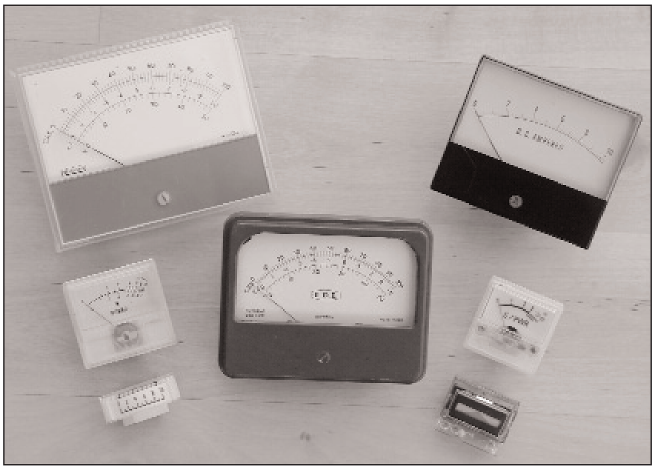


Fig 24.2: Various moving-coil and moving-iron meters

The ubiquitous multimeter (Fig 24.1) 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 necessary to have meters dedicated to a single function or a group of functions. This is the aspect considered in the following sections, together with how the meters can be adapted to the ranges that need to be used.

ANALOGUE METERS

These are of electromechanical design and consist (amongst others) of the moving-coil and moving-iron type meters (Fig 24.2). 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. The moving-iron instrument, however, is AC/DC with a response up to about 60Hz. The modern analogue meter tends to be rectangular, older types often being round.

The sensitivity of analogue meters is defined by the current that must flow through them in order to provide *full-scale*

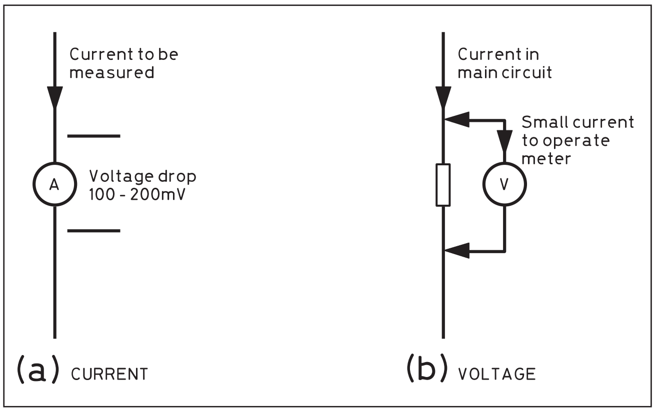


Fig 24.3: The use of meters for measurement

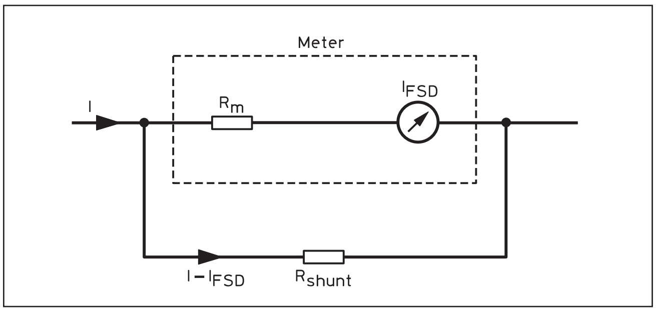


Fig 24.4: Arrangement for current shunts

deflection (FSD). The moving-coil range starts at about 50µA FSD while the moving iron range begins at about 100mA FSD.

These analogue meters do draw current from the circuit under test to operate because the coil is made from wire, it has resistance (R_m). When used for measuring current, Fig 24.3(a), the meter is placed in series with the circuit and so there is a voltage drop - typically 100 to 200mV across the meter. When used as a voltmeter, Fig 24.3(b), the current drawn depends on the basic meter movement (FSD current) and, if this takes more than 10% of what is flowing in the circuit, the circuit conditions are being progressively affected.

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 - it is possible to use transfers, or if you are skilful, to redraw it. There is also software available for making dial scales - see later in the Software-Based Test Equipment section. The scale plate can usually be removed.

An analogue meter requires current to operate; consider the measurement of current initially. The current for 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 in Fig 24.4.

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 max 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.

When large currents are to be measured, the shunt resistor becomes very small in value and may be fractions of an ohm for large currents.

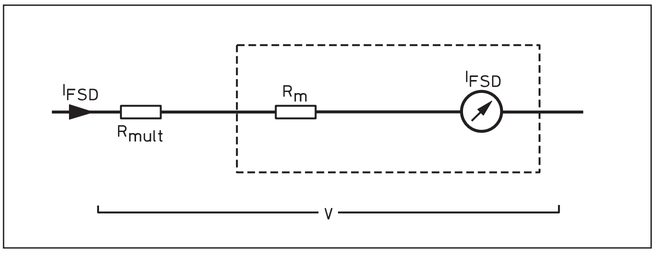


Fig 24.5: Arrangement for voltage multipliers

Example: A 1mA FSD meter with coil resistance of 1000 ohms is to be used to measure a current of up to 10A.

R_{shunt} = 0.1Ω with a power rating of 10W

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 in Fig 24.5. 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

When high voltages need to be measured, the series resistor becomes very high in value and the value of the coil resistance can be ignored.

Example: A 100µA FSD meter with coil resistance of 2000 ohms is to be used in a valve amplifier to measure up to 2kV.

R_{mult} = 20MΩ with a power rating of 0.2W

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

One problem with all measuring instruments is how they affect the circuit they are measuring; with these analogue meters it is the current they take in order to operate.

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

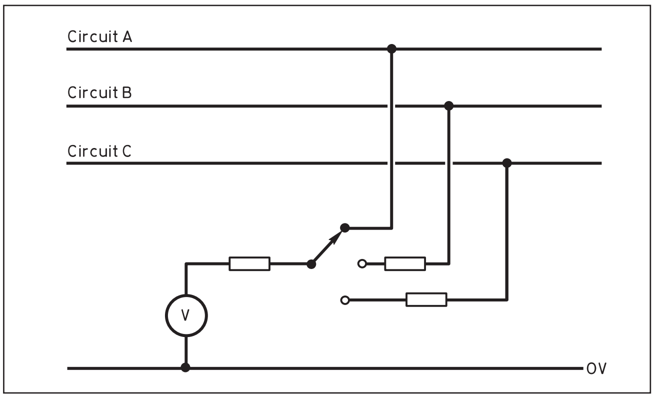


Fig 24.6: Switched voltage measurements

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 to 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, and often different polarities - meaning that the connections to the meter need to be reversed to give the correct needle movement. 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. The switching in of different multiplier resistors is of little consequence as these tend to have high-value resistances.

Voltage measurements are normally made with respect to 0V or earth. This means that one end of the voltmeter is fixed - see **Fig 24.6**. 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

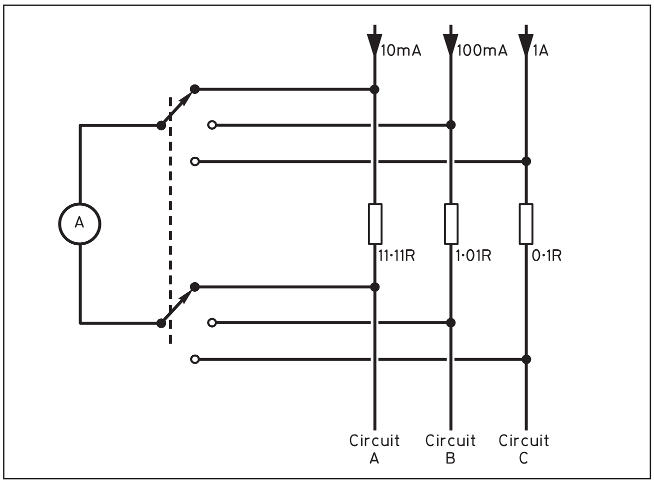


Fig 24.7: Switched current measurements

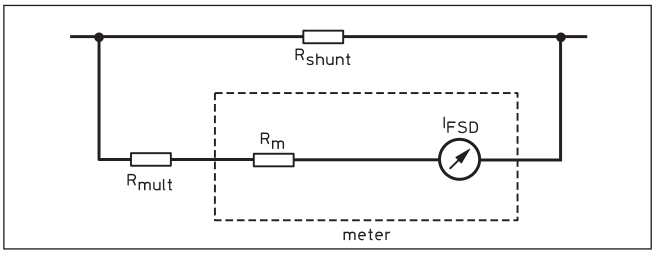


Fig 24.8: Current measurement by volt-drop method

that circuit A in Fig 24.6 has the lowest voltage to be measured, 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 24.7** 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, 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 24.8**.

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 24.9**, 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 have very high resistance until the anode is some 400mV above the cathode, at which point they just start to conduct and the resistance begins to fall. 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 in Fig 24.9 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.

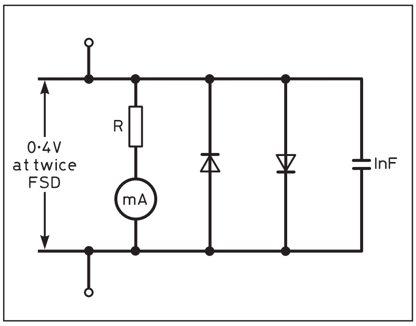


Fig 24.9: Meter protection using diodes

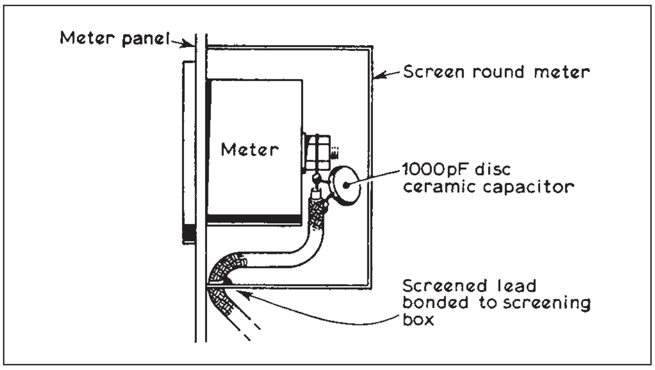


Fig 24.10: Screening and by-passing a meter in a transmitter

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 used 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 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 24.10. 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.

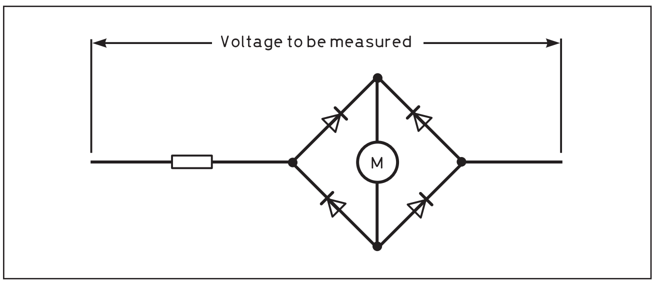


Fig 24.11: Typical arrangements for AC measurements

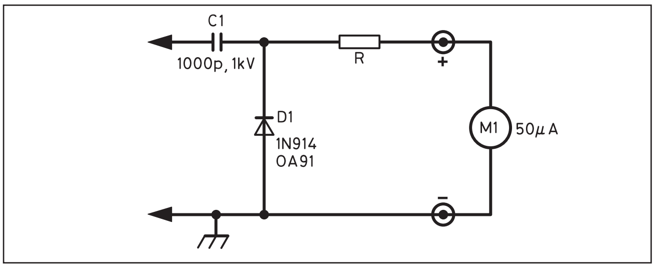


Fig 24.12: RF probe. For R = 270k + 12k, the meter scaling is 0-10V, and full-scale, power in 50 ohms is 2W. For R = 820k + 27k, the meter scaling is 0-30V, and full-scale, power in 50Ω is 18W

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 meters for AC sine-wave measurements therefore incorporate a rectifier (see **Fig 24.11** for a typical arrangement) and the scale is adjusted by the form factor (1.11) 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 measurements - AC current measurements pose additional problems and are not considered further. The typical frequency range extends to between 10kHz 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, LT1967, 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 (for example the 1N914 is 100V, and the OA202 is 150V with slightly poorer RF capabilities). The approach in measuring RF voltages is to rectify as soon as possible and then use DC measuring circuits.

Fig 24.12 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

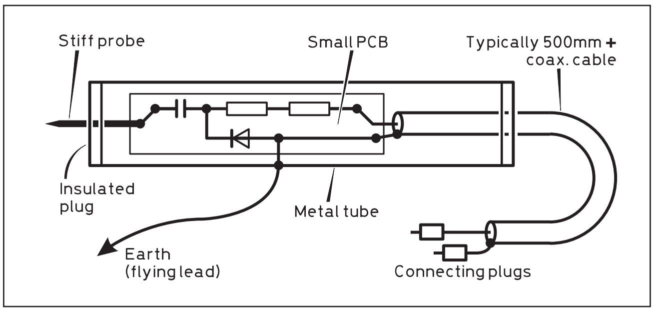


Fig 24.13: Typical construction of an RF probe

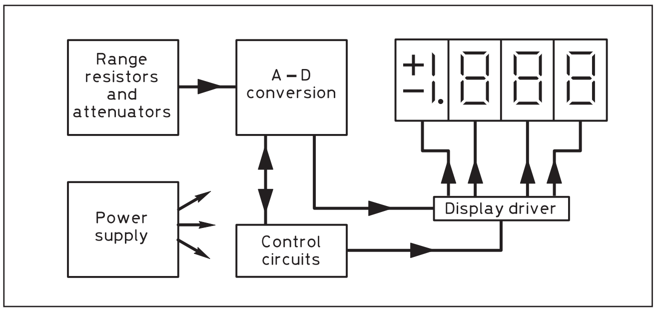


Fig 24.16: Block diagram of a digital meter (3½-digit display)

in Fig 24.10. Fig 24.13 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%.

Because of reverse-voltage limitations of the diodes, it is necessary to make modifications to take higher voltage readings. Fig 24.14 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 24.15. 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 24.16) 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.

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 3, 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 (depending on the 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*).

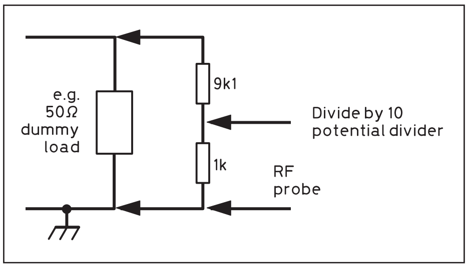


Fig 24.14: Suggested method for higher voltages

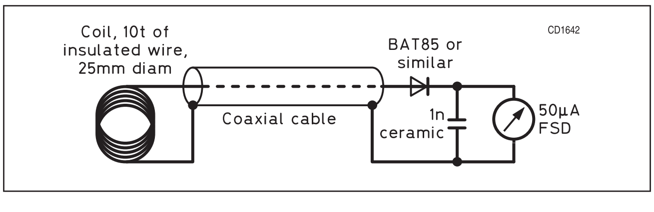


Fig 24.15: Pick-up loop with a diode

Fig 24.17: A typical digital panel meter

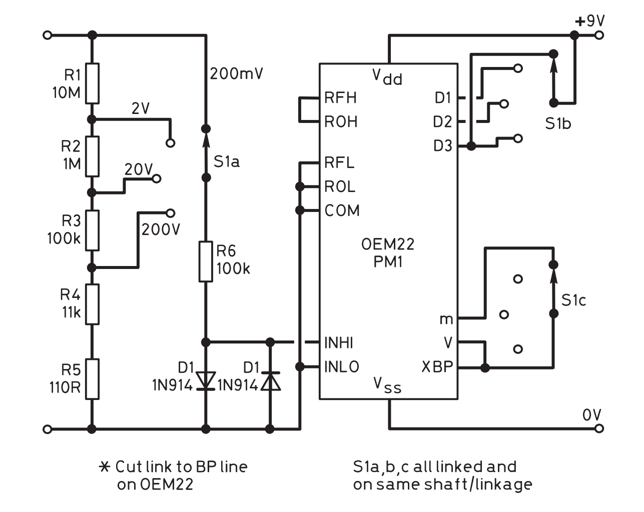


Fig 24.18: A practical digital voltmeter

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 24.1: Components for the practical digital voltmeter

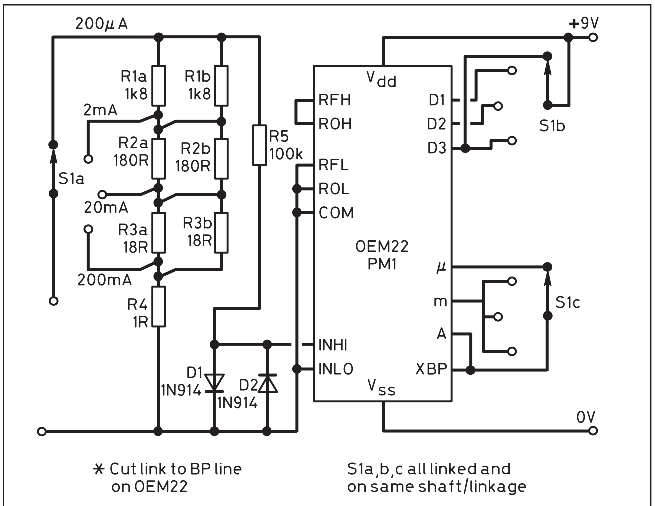


Fig 24.19: A practical digital ammeter

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 24.2: Components for the practical digital ammeter

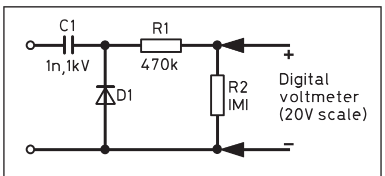


Fig 24.20: RF probe for digital voltmeter

They can be purchased with and without backlighting; a typical meter is shown in Fig 24.17. 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). Technical details comes with it or can be found on the Internet [4], and the pin designations are marked on the board. The principles explained can, however, be applied to modules available from other manufacturers.

A Practical Digital Voltmeter

Fig 24.18 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 24.1.

Resistors R1-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 24.19 results in a meter measuring 200µA to 200mA in decade ranges.

Resistors R1-R4 form the load across which the voltage is developed from the current being measured. Resistors R1-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

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 24.3: Comparison of analogue and digital meters

combination R5/D1/D2 provides protection for the panel meter input. The unit can be powered from a PP3 battery or equivalent. **Table 24.2** 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 24.20**. 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 only. The construction should be similar to that shown in Fig 24.13.

Comparison of Analogue and Digital Meters

Table 24.3 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 the inclusion of input amplifiers and attenuators.

THE OSCILLOSCOPE

The oscilloscope (**Fig 24.21**) 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 also a good aid when fault finding. The display allows one to determine, for example:-

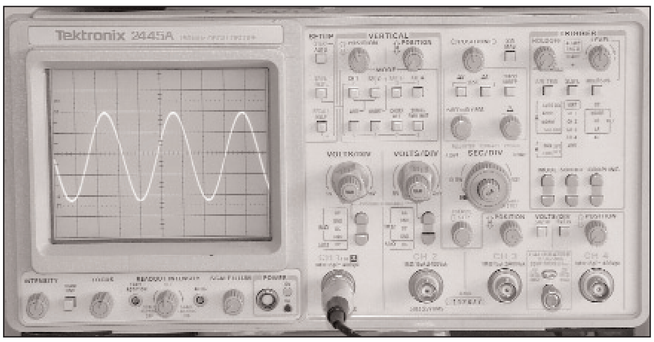
- the shape of a signal
- whether it is AC or DC or a mix
- the amplitude of a signal
- the frequency of a signal
- the time variation of a signal
- if a signal is distorted
- if a signal contains much noise

This section explains briefly how an oscilloscope functions and how it can be used for taking various measurements. Some of the limitations of the oscilloscope are also pointed out.

Reasonably priced oscilloscopes are available on the second-hand market but new ones are not particularly expensive - it all depends on the facilities required. Some oscilloscopes can display a single trace while others can display two traces or even more, sometimes with adapters.

Oscilloscopes covering frequencies up to about 30MHz are no more expensive than some dual band VHF/UHF transceivers! The CRT (cathode ray tube) based oscilloscopes probably are the best value for money but the newer LCD types are very good value - the battery powered ones are however expensive.

Fig 24.21: Oscilloscopes: (below) CRT type, and (right) LCD type



As the horizontal scale of the oscilloscope is in time, this instrument will give a display in what is known as the *time domain*.

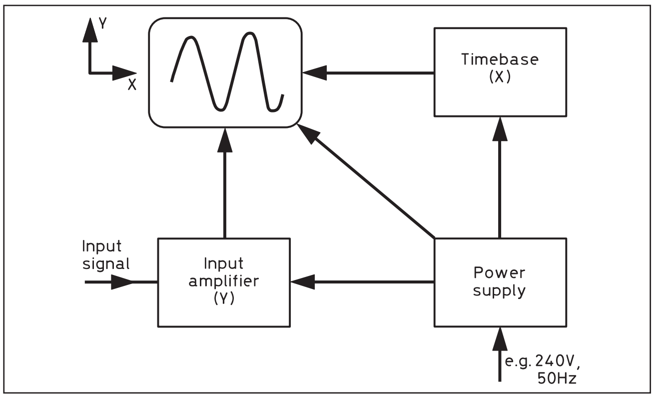


Fig 24.22: Basic block diagram of an oscilloscope

The Basic Oscilloscope

Fig 24.22 shows the absolute basics for an oscilloscope. It consists of a display 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. In addition a power supply is required to provide the amplifiers, timebase, display and other electronics with relevant voltages. The screen is split into two directions, the X (or horizontal) and Y (vertical direction).

With a CRT type, a single-trace oscilloscope has a single electron gun firing at the phosphor. Two-trace 'scopes 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 (less common) 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.

CRT type oscilloscopes are mainly analogue type with a few specialist units available using digital techniques.

LCD type oscilloscopes have input amplifiers but then generally use an ADC (analogue to digital converter) to convert the incoming signals into a digital value which can then be stored in

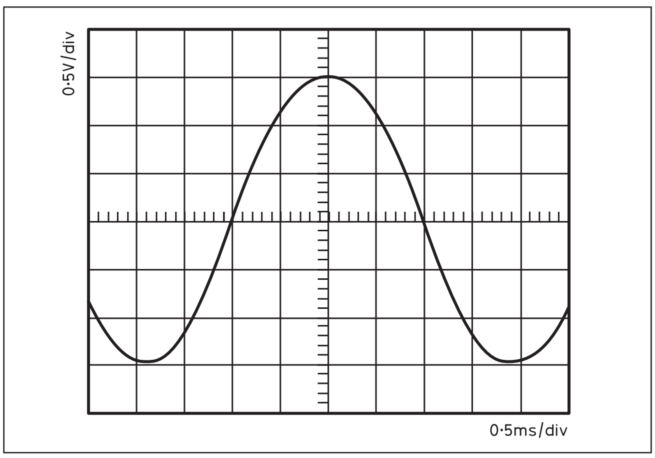


Fig 24.23: A typical displayed waveform

RAM and then plot the signal on a display under microprocessor control. These oscilloscopes can have several inputs, eg 2 or 4, be monochrome or may have coloured displays (each trace a different colour!). Most of these can easily store waveforms as well as communicate directly with a computer for further storage, printing and analysis. These types of oscilloscopes may also have incorporated digital voltmeters and frequency counters that will give digital values on the display as well.

The input to these oscilloscopes is essentially sampled by the ADC and so fast sampling speeds are normally required to provide wide bandwidths.

Voltage Measurements

Consider the oscilloscope screen display as depicted in **Fig 24.23**. 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 six 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. Don't expect accuracies any better than about 5-10%.

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 eight 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. Don't expect accuracies any better than about 5-10%.

Equipment Limitations

The Y-amplifiers (plus the CRT) or sampling rates limit the frequency response of the oscilloscope. This means that after a certain point the oscilloscope calibration is not valid or the display not accurate, but comparative measurements can still be made above this point provided 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.

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 individual items which make up radio equipment can be divided into passive and active components. Passive components consist largely of resistors (R), capacitors (C), inductors (L) and transformers. All of these can be readily purchased with chosen values but occasionally it is required to check something, especially if the colour coding has been worn or burnt off or if it is desired to make, for example, a coil of specific value. See later in this chapter for impedance measurements (ie combination of R, L and C at specific frequencies).

The question is to what accuracy does the measurement need to be? Most of the time capacitors will be bought with known accuracy and are usually labelled (except for those in the junk box).

When it comes to odd values (eg for filters) it is possible to make the value up from parallel and series combinations but it might still be a good idea to check the final value.

Inductors are more likely to be made from winding wire on a former and although there are good inductor design programs the values need to be checked before use.

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 semiconductor devices. There are also 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 or on Internet auctions at reasonable prices..

Resistors can most easily be checked with the ubiquitous multimeter or digital voltmeter.

A typical digital LCR tester will cost between £15 and £100 (2011 prices). However, you get what you pay for and the typical resolution for capacitance is 1pF, for inductance is 1μH and resistance 1 ohm. The measurement frequency depends on the model and varies between about 1kHz and 200kHz - not really RF frequencies and this is typical until you get to the more expensive types. The minimum value of inductance on the cheaper meters is not good enough for VHF and above.

There are articles and kits available on the Internet and it is worthwhile looking at [5], [6] and [7]. See also [1].

Diode and transistor testers again vary in price and are 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.

SIMPLE DIODE AND TRANSISTOR TESTER

The circuit of **Fig 24.24** 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

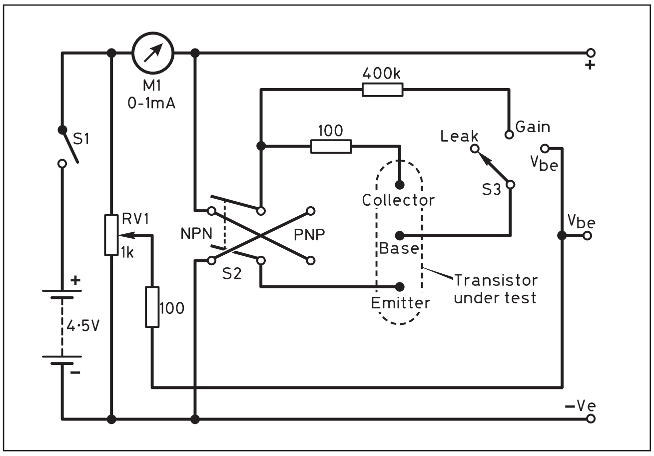


Fig 24.24: Simple transistor and diode tester

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.

IMPEDANCE MEASUREMENTS

Impedance measurement for the radio amateur probably means antenna or feed point impedance; however, it can also mean the input impedance of networks and amplifiers. The measurements are usually undertaken to see how well the impedance matches to a system value, eg 50 ohms. This is required to ensure maximum power transfer, minimising of reflected power and filter matching to achieve the correct characteristics.

You can purchase proprietary equipment designed to cover the amateur bands up to 470MHz and would be useful for other frequencies as well. These may well come under the heading of antenna analysers or network analysers. Equipment is available from about £180 to £500 (early 2011 prices) which measure various parameters; if you are serious about impedance and antenna measurements then this is probably the best route to follow.

The units cover differing frequency ranges and provide analysis such as impedance (R and X), SWR, transmission cable loss etc - look at the product specifications on the web as they are too varied to mention here. Some will interface to a PC and give graphical representation of the measurements. Typical of some of the units available are:-

- AEA VIA (0.1-54MHz) - serial connection to a PC using optional software
- AEA 140-525 (135-525MHz) - serial connection to a PC using optional software
- AIM4170: (0.1-170MHz) - RS232 connection to a PC for further data analysis
- Autek RF1 (1.2-34MHz)
- Autek RF5 (35-75 and 138-500MHz)
- Autek VA1 (0.5-32MHz)
- Feature Tech AW07A (Amateur bands 160m to 70cm)
- MFJ-259B (1.8-170MHz)
- MFJ-269 (1.8-450MHz)
- Micro908 kit (1-30MHz) - serial port to a PC for uploading any new software
- MiniVNA (0.1-180MHz) - USB connection to a PC for data analysis
- Kuranishi BR210 (1.8-170MHz)
- Kuranishi BR510A (1.8-170MHz and 300-500MHz)
- Kuranishi BR400 (100-170MHz and 300-500MHz)
- N2PK project/kit (0.05 to 60MHz)
- Palstar: ZM-30 (1-30MHz)
- TAPR VNA (0.2-120MHz) - USB connection to a PC for data analysis
- Times Technology T100 (100-170, 400-470MHz) - USB connection to PC for further displays

In addition there is an interesting project/kit described in the ARRL publication *QEX*, Jan/Feb 2009 - and now available as a kit [8], covering 1kHz to 1.3GHz, Also, there is the VK5JST Aerial Analyser (1.3-31MHz) [9], and the N2PK VNA (Vector Network Analyser) - 0.05 - 60MHz [10] and further described in [11].

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 reactive components separately, the second one will provide this in series form. A previous design in chapter 11 of [1] 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. 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, or from the Internet.

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 to the amateur workshop. In fact it is essential if experiments with antennas are undertaken.

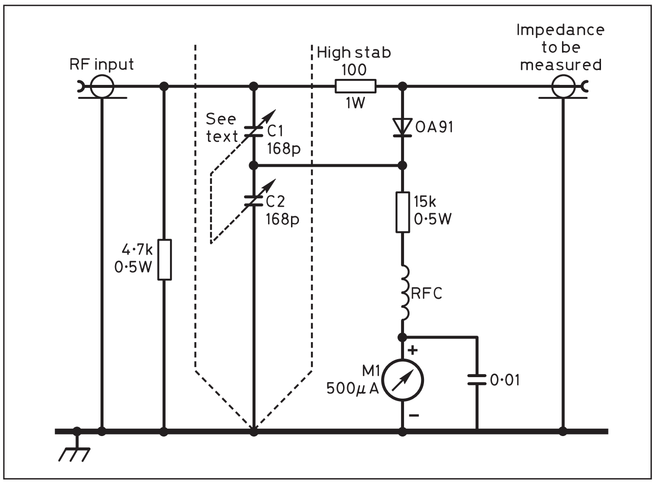


Fig 24.25: Simple RF bridge. Note that a BAT85 diode may be used instead of the OA91

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 in **Fig 24.25**. 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 24.25.

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 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 24.26**.

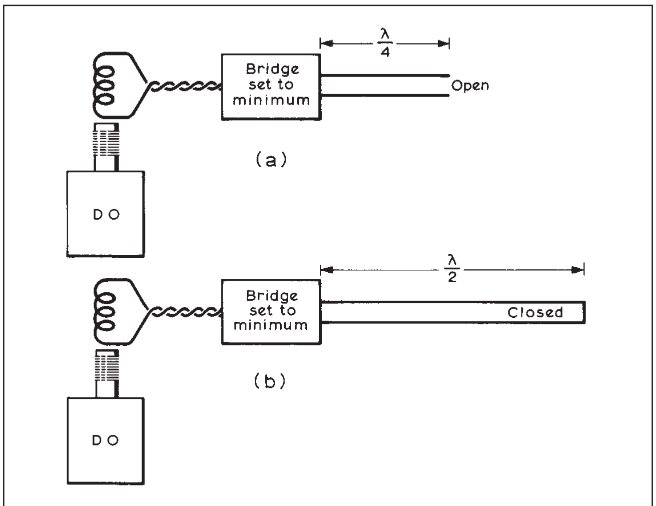


Fig 24.26: Use of the RF bridge with a dip oscillator

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.

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 24.27**, the circuit diagram is shown in **Fig 24.28** and parts list in **Table 24.4**. **Fig 24.29** 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

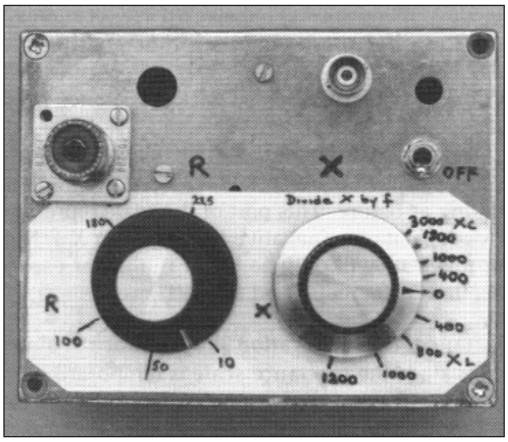


Fig 24.27: The G3ZOM noise bridge

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 - linear law if available
TR1,TR2	BC182, 2N3903 or similar	S1	SPCO switch
T1	Dust iron core FT50-43, FT50-5 or similar. See the circuit diagram for winding details		

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

Table 24.4: RF Noise bridge components list

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:

At 1MHz: 5000Ω capacitive to 1200Ω inductive

At 30MHz: l70Ω 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 24.28.

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 24.27 has a better than 50dB null at 10MHz.

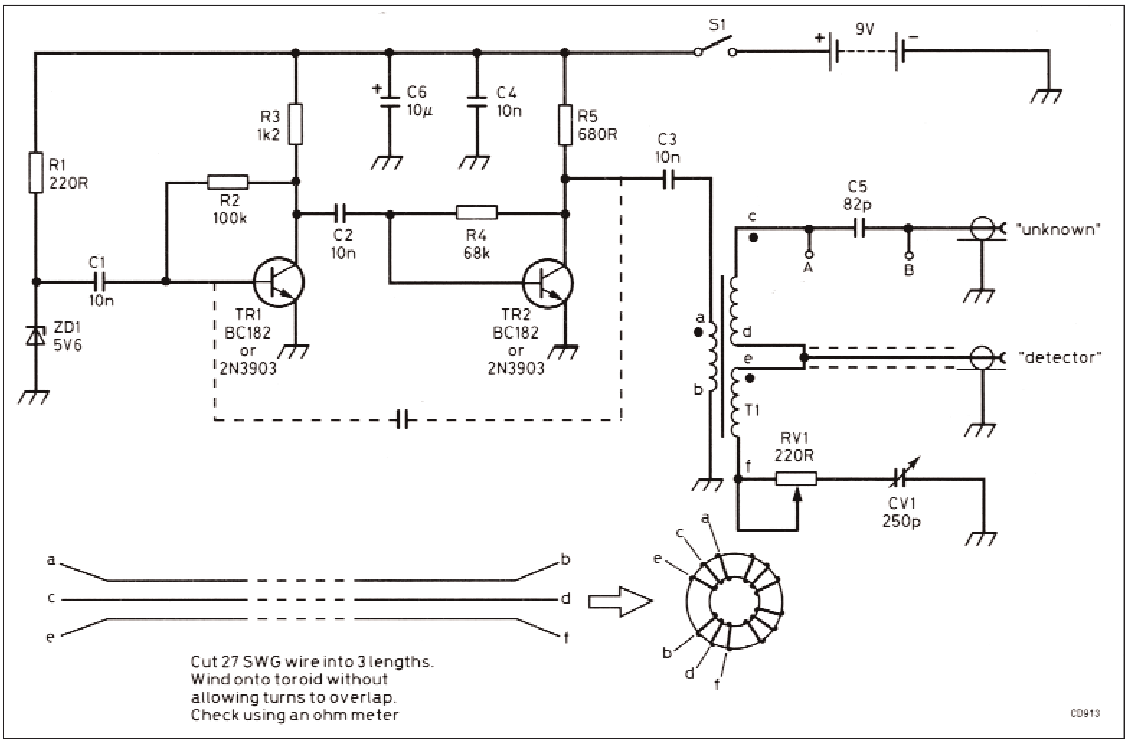


Fig 24.28: Circuit of the G3ZOM noise bridge

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.

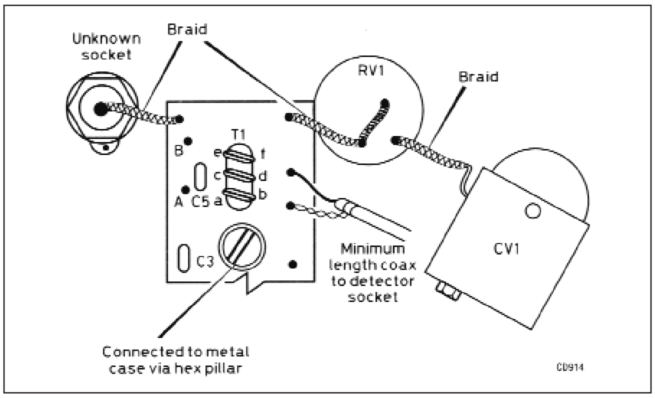


Fig 24.29: Layout of bridge components

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 24.30** 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).

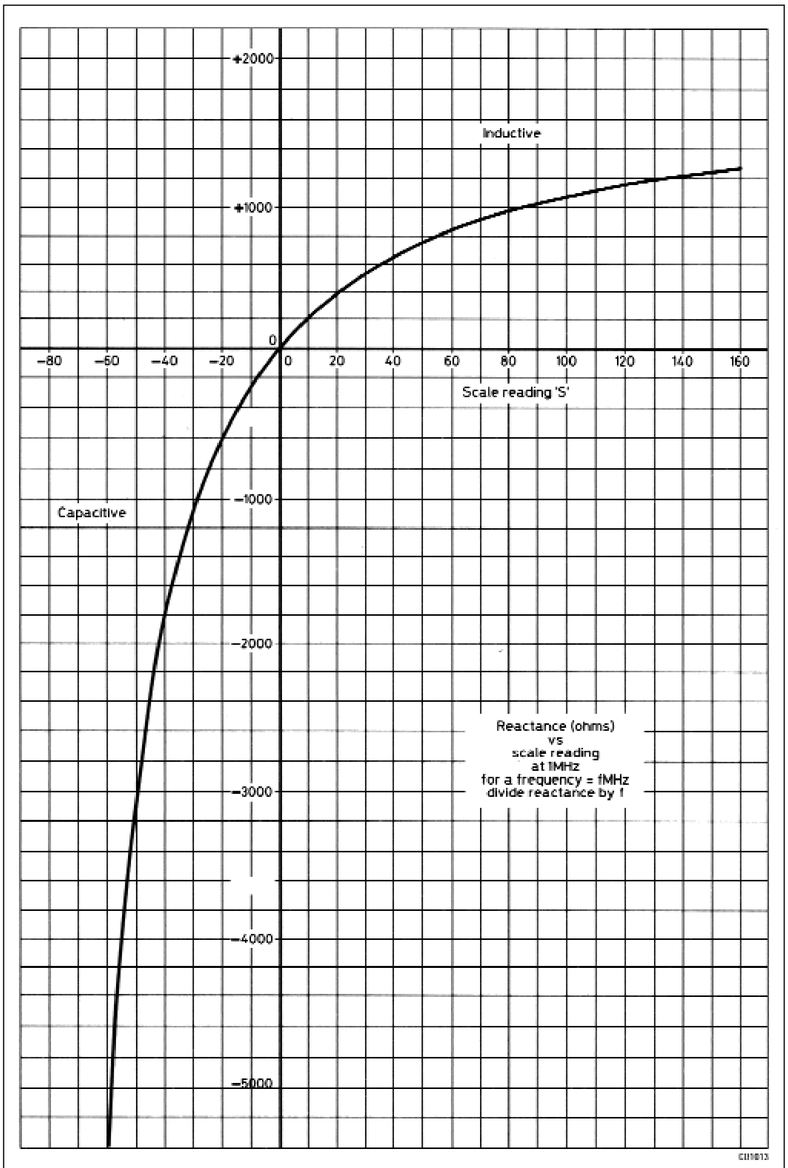


Fig 24.30: Calibration graph for converting capacitance values to reactance (+ or - j)

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 24.31**. 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 24.31(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 24.31(b)**. The adjustment, and the measurement of the 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

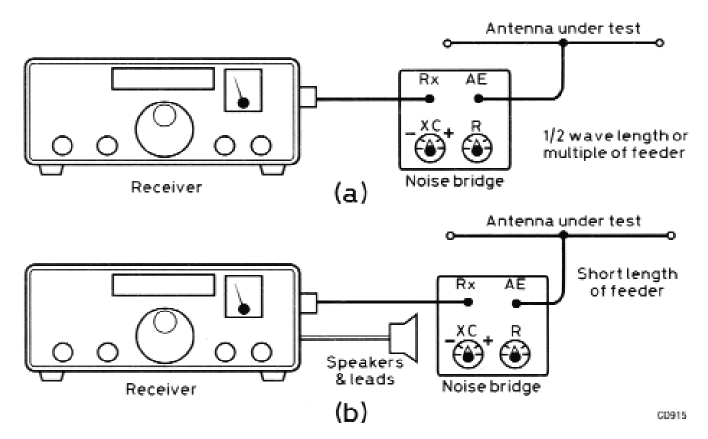


Fig 24.31: Noise bridge and receiver connections for antenna impedance measurements

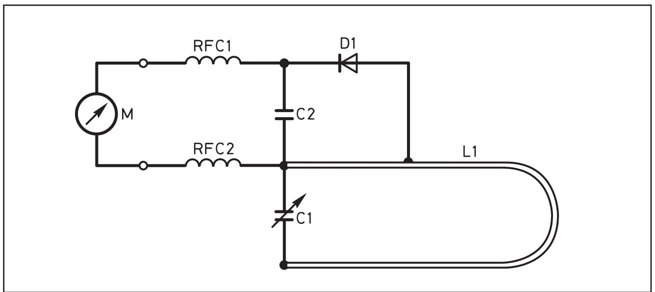


Fig 24.32: A simple absorption wavemeter for 65-230MHz

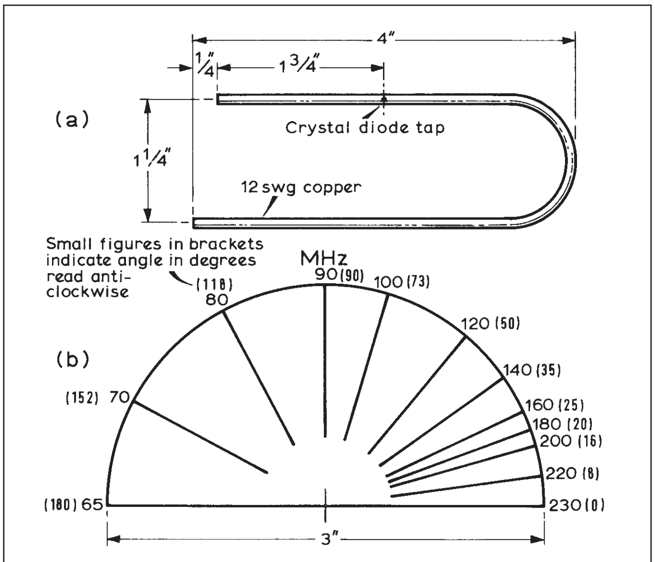


Fig 24.33: Constructional details of simple absorption wavemeter: (a) inductor; (b) dial plate. ¼in = 6.3mm, 1¼in = 31.8mm, 1¾in = 44.5mm, 3in = 76.2mm, 4in = 101.6mm

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 24.36(a)	C2	470p ceramic
D1	OA91, BAT85 or similar		

Table 24.5: Components list for the simple absorption wavemeter

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.*

A SIMPLE ABSORPTION WAVEMETER FOR 65 - 230MHZ

The absorption wavemeter shown in Fig 24.32 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 in Fig 24.33(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 24.5. 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 24.33(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 you can build them - see later and reference [1] and [12], or the Internet. In the past, these will have been called grid dip oscillators when valves were used.

Although the dip oscillator has a wide range of uses for measurements on both complete equipment and individual components [13, 14], 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 24.34.

If the tuned circuit being investigated is well shielded magnetically (eg a coaxial line) it may be difficult to use inductive coupling. In such cases it may be possible to use capacitive coupling by placing the open end of the line near to one end of the dip oscillator coil.

A completely enclosed cavity is likely to have some form of coupling loop and the dip meter coil can usually be coupled inductively by means of a low-impedance transmission line such as a twisted pair with a coupling loop.

When used as 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.

For uses such as finding the resonant frequency of VHF/UHF antennas and measuring feeder length see [13].

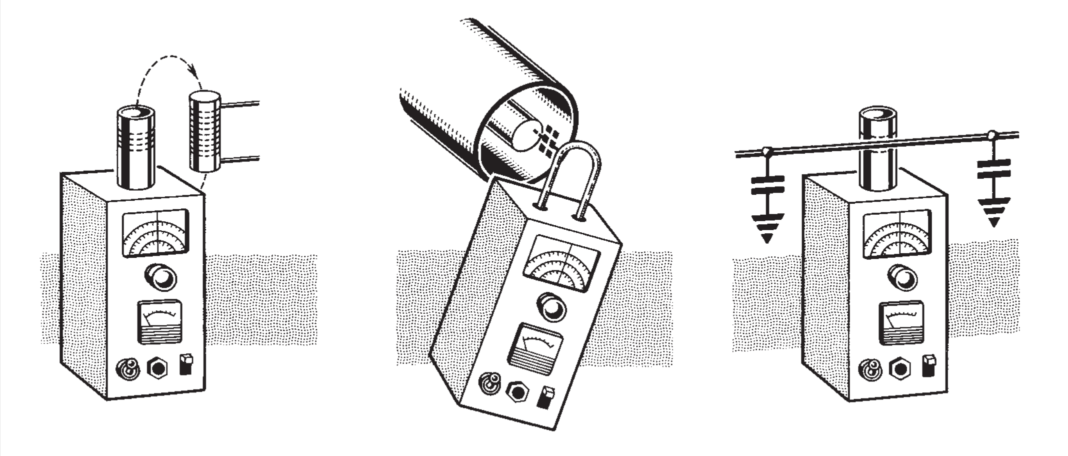


Fig 24.34: Using a dip oscillator

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 [14] 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 24.6. 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 24.6 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 24.6, 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 24.6, the value of the unknown capacitor is given by:

$$C\text{ pF} = 2533 \div F^2 = 2533 \div 6.1 \div 6.1 = 68\text{pF}$$

Example 2: An unknown coil in parallel with an known 47pF capacitor, produces a dip at 12.8MHz, hence F = 12.8. From Table 24.6, the value of the unknown inductance is given by:

$$L\ \mu\text{H} = 539 \div F^2 = 539 \div 12.8 \div 12.8 = 3.3\mu\text{H}$$

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.

Table 24.6: Determination of L and C using known values

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

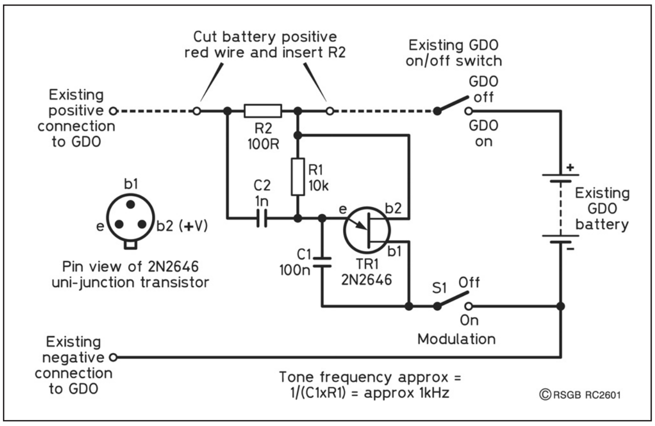


Fig 24.35: Add-on audio 1kHz oscillator for a dip oscillator (sometimes referred to as a GDO)

Tone Modulation

The following is also by G3BIK [14]: Sometimes it is also useful to be able to hear an audio tone when using the dip oscillator 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 24.35**. 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 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

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* [15, 16]. **Fig 24.36** shows 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

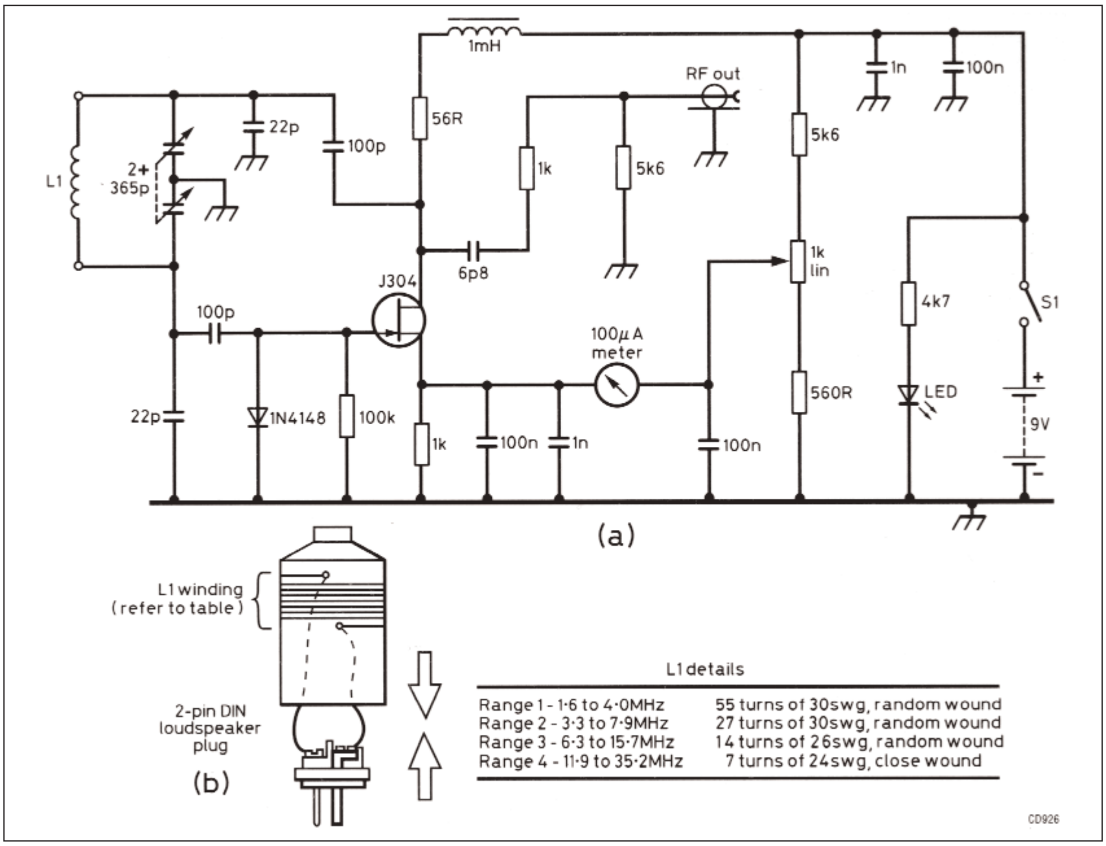


Fig 24.36:A simple dip oscillator. The J304 can be replaced by a similar N channel FET

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.

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 24.36 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.

There are also oscillators based on direct digital synthesis (DDS) - for example see [17]. These will need to be followed by a high quality attenuator.

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 equipment as well.

For less onerous requirements, simple oscillators can be constructed for tuning over a limited range, such as VFOs and VXOs. There are many designs on the Internet as well as from manufacturer's data sheets. See also the oscillators chapter in this book.

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.

Below is a selection of signal generators for various frequencies, see also the chapter on microwave receivers and transmitters for a microwave signal source.

LOW-FREQUENCY SINEWAVE OSCILLATOR FOR 10Hz-100kHz

The circuit diagram for this oscillator is shown in **Fig 24.37**. 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:
1: 10Hz-100Hz
2: 100Hz-1kHz
3: 1kHz-10kHz
4: 10kHz-100kHz

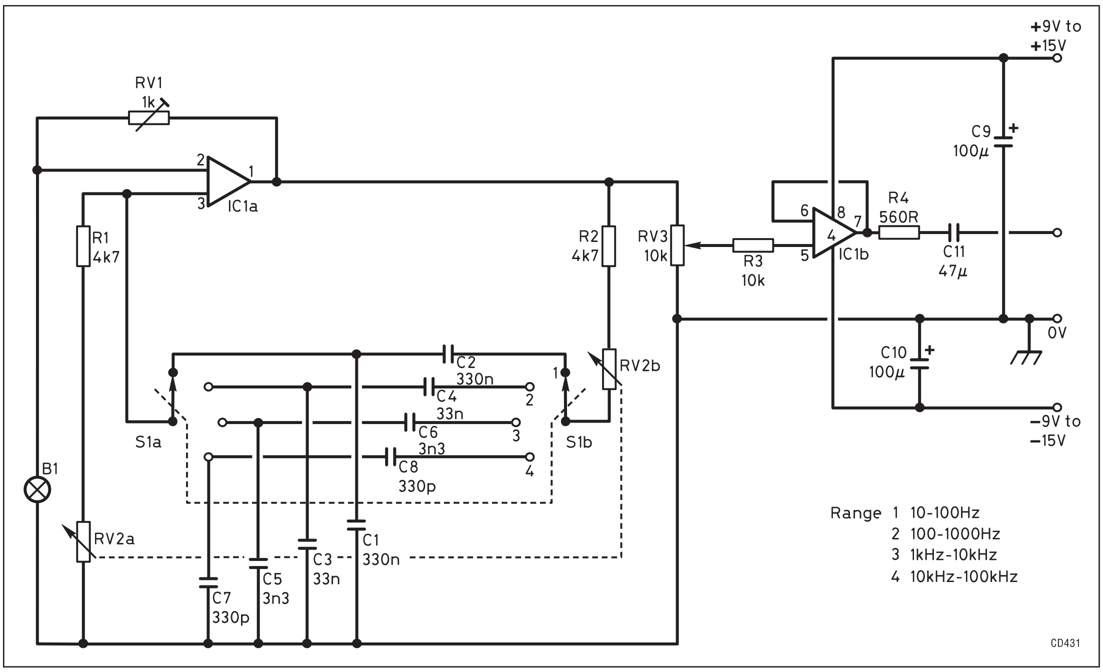


Fig 24.37: Circuit of the low frequency oscillator

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 log pot.	C9,C10	100µ, 25V
RV3	10k lin. pot.	C11	47µ, bipolar
B1	28V, 40mA bulb	IC1	LM358

Table 24.7: Low frequency oscillator components list

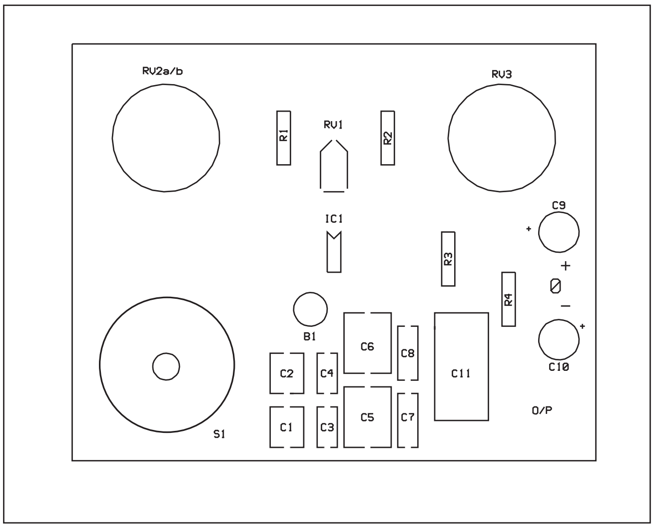


Fig 24.39: Component layout of the low frequency oscillator (not to scale)

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

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

Construction

A components list is given in **Table 24.7**. The layout of the circuit is not critical but a PCB pattern (**Fig 24.38**) is given in Appendix B and a component layout in **Fig 24.39**. 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 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 24.40**. The signal is derived from a 1MHz crystal-controlled oscillator formed by XL1 and IC1 plus various components. Capacitor VC1 allows a slight variation of the crystal frequency for calibration as described later. This

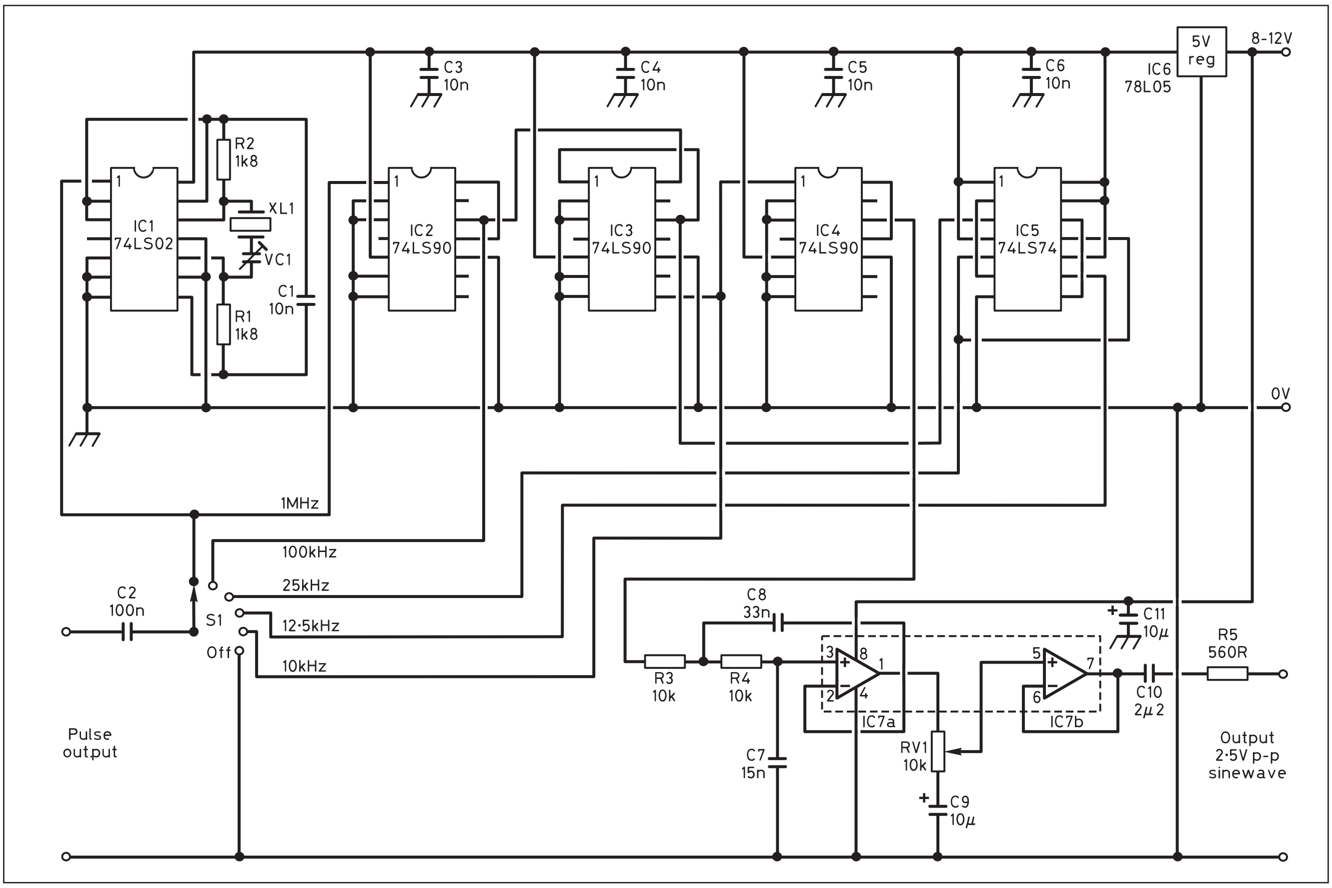


Fig 24.40: Circuit of the crystal-based frequency marker

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 24.8: Crystal-based frequency marker components list

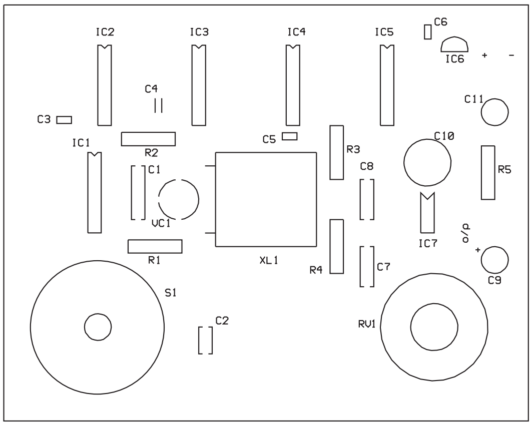


Fig 24.42: Component placement details for the crystal-based frequency marker (not to scale)

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 24.8**. 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 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 24.41**) is in Appendix B and component placement in **Fig 24.42**.

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.

A COMBINED 2M/70CM SIGNAL SOURCE

This signal source was developed by John Brown, G3DVV [18]. The project (**Fig 24.43**) uses one of the low-cost 48MHz crystal oscillator modules that are available - 48 x 3 = 144MHz and 48 x 9 = 432MHz, ie 2m and 70cm. The unit can be followed by a well-shielded attenuator.

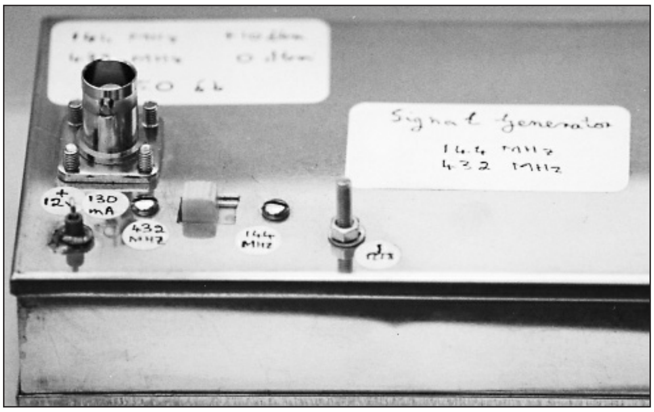


Fig 24.43: The combined 2m and 70cm signal source showing the BNC output socket, power supply terminals and frequency selector circuit

R1	3k9
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
XCO1	48MHz crystal oscillator module
TR1	BFR96
IC1	78L05
IC2	MMIC. See text. Original used MSA0304
D1 - D4	1N914 or any suitable switching diode
FL1	Toko CBT3
FL2	Toko 7HW
S1	2-way single pole switch
Double sided glass-fibre PCB type FR4, 1.6mm thick	
Tin-plate box, 74 x 148 x 30 mm	
Track pins	

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

Table 24.9. Components for combined 2m and 70cm signal source

NOTE: The original device for IC2 (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 and Avago Semiconductors (eg Mar-3 or MAV-3). Several of the devices match the specification but the bias resistor R8 must be changed - see technical data at [19] or [20] or put 'MMIC' in as a search on the Internet.

Preview

The component list is given in **Table 24.9**. The circuit, shown in **Fig 24.44**, starts with a 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

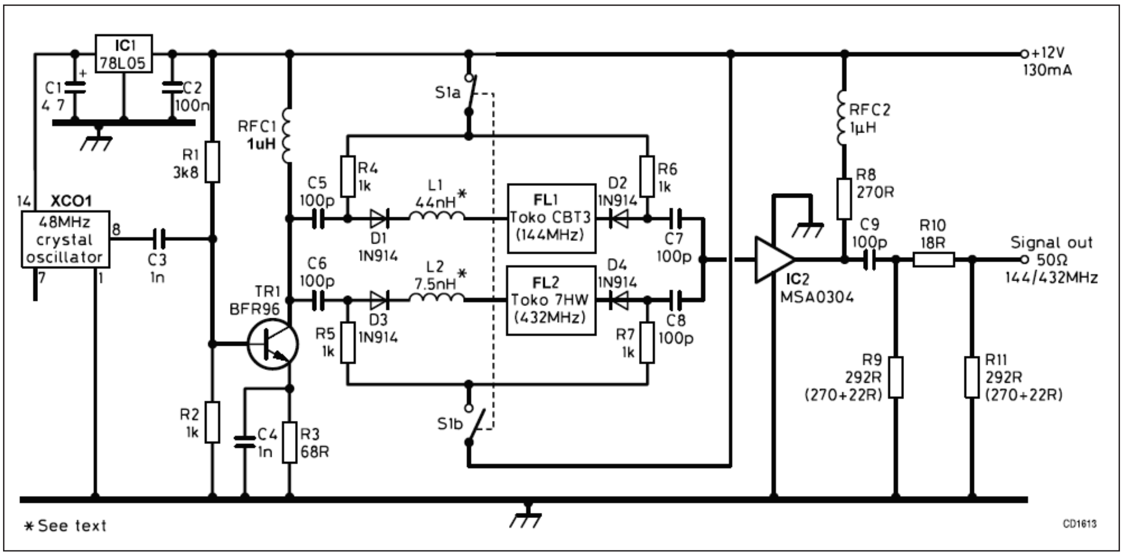


Fig 24.44: Circuit of the combined 2m and 70cm signal source. Steering diodes are used to switch between filters FL1 and FL2

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 with IC1 providing 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.

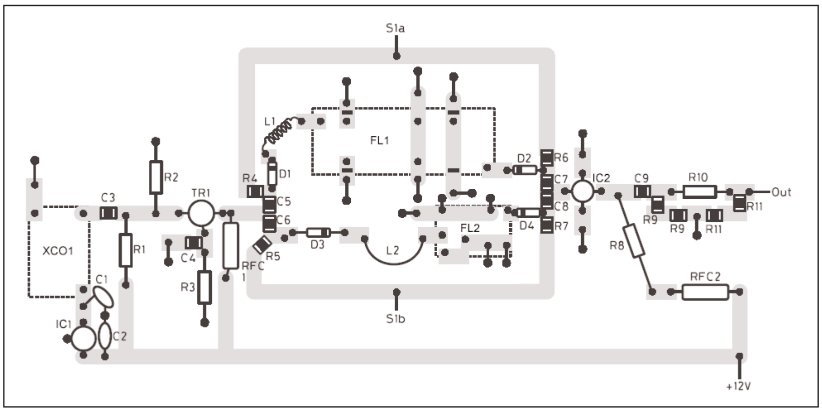


Fig 24.46 (a): Component placement for the signal source

The Circuit in Detail

The output from the crystal oscillator module (XCO1) 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 the two filters (FL1 and FL2). 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.

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- and short-circuits.

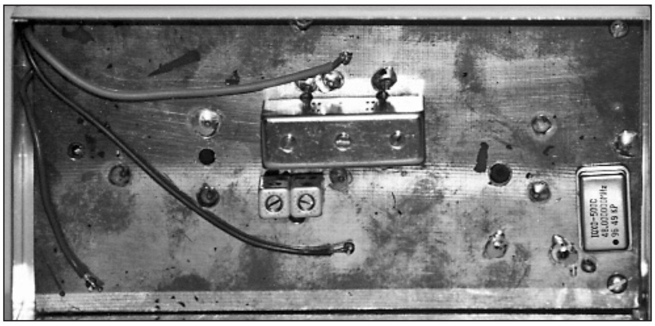


Fig 24.46 (b): Signal source showing the two filters and the oscillator module

Construction

The PCB pattern is in **Fig 24.45** (in Appendix B). 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 component layout is in **Fig 24.46(a)**. The only difficult items to mount are the two filters (**Fig 24.46(b)** because the lugs as well as the pins have to be accommodated. Holes are required for TR1 and IC2, and pin through holes for XCO1 (note that there is no connection to pin 7 - clear the copper away from pins 1, 8 and 14). TR1 and IC2 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 component 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.

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 IC2. 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.

RF SAMPLERS

It is often necessary to obtain a 'sample' of an RF signal for frequency measurements, modulation measurements etc. When running more than a few watts the magnitude of the RF voltage may damage test equipment and a smaller sample of the RF signal is adequate to drive test equipment. An attenuator of some form must be used that has minimal disturbance of the main RF system. The following gives some basic details. On safety grounds use a different connector such as BNC or SMA for the 'sampled' output to prevent inadvertent wiring/cabling and a sad outcome. Further discussion is provided in [1] and a simple sampler described in [21].

Resistive Sampling

Fig 24.47 shows a typical resistive sampler whilst the circuit in **Fig 24.48** gives some variations and additional protection. The input resistance of the 'sampling' network should be at least ten times that of the characteristic impedance of the through-line system to prevent unwanted loading, the resistors should be carbon if at all possible and the power (voltage) on the through-line will determine the power rating of the series resistor. The series capacitor will isolate any DC voltage from test equipment (there could be a preamplifier DC feed on the coaxial cable) and the back to back diodes will try to limit input voltage to the test equipment.

In a 100W system, the power rating of the 9.1kΩ resistor is 0.5W and for the 1kΩ system (**Fig 24.48**) it needs to be 5W for continuous ratings. One can get away with less than this for SSB because of the nature of the signal. The exact attenuation needs a knowledge of the input impedance of the equipment being attached to the output labelled RF sample.

It is also possible to get known values of attenuation by using Pi or T attenuator circuits. In this case it is assumed that the loading of the attenuator on the through-line must be minimal - say 1kΩ input and the through-line is terminated in a 50Ω load. It is also assumed that a 50Ω input test equipment is used at the other. In calculating the attenuator circuits it is assumed that the test equipment wants a voltage drive, hence the figure of 1V on the through-line provides 10mV at the test equipment terminals (ie a voltage reduction of 100:1). This is equivalent to an

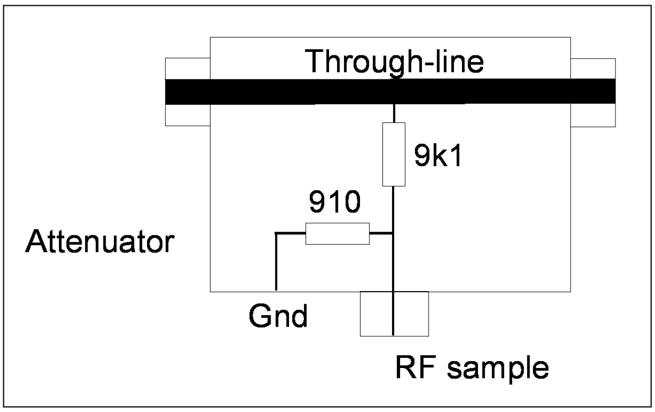


Fig 24.47: Resistive sampler

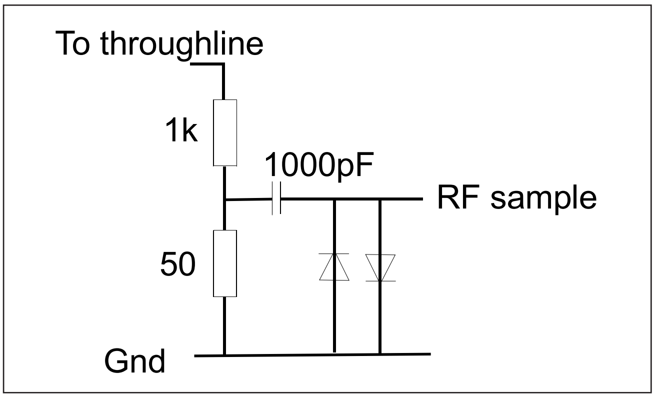


Fig. 24.48: Typical circuits for resistive sampler

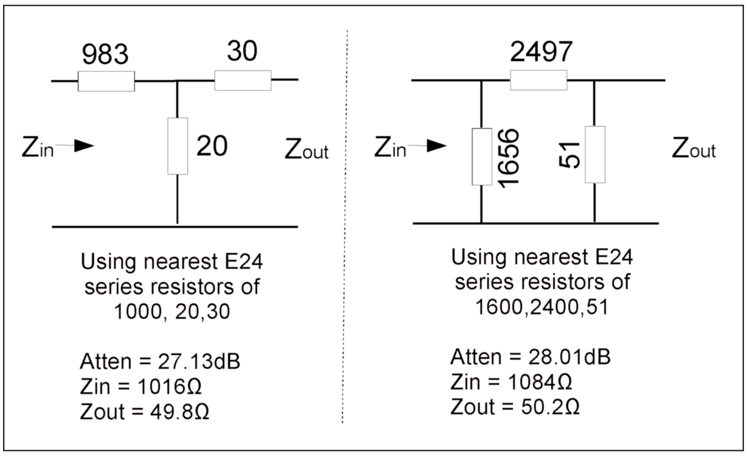


Fig 24.49: Resistive attenuator sampler circuits

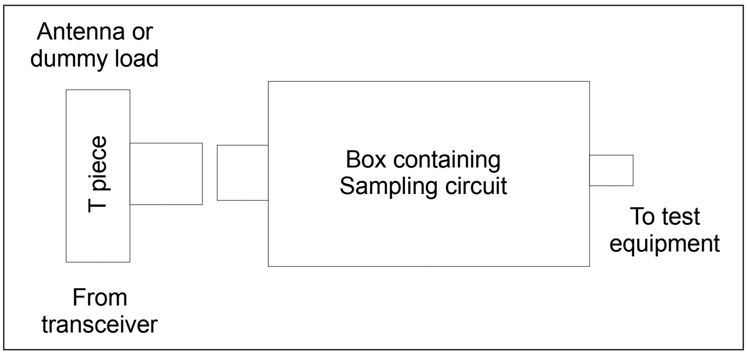


Fig. 24.50: General arrangement of T piece sampler

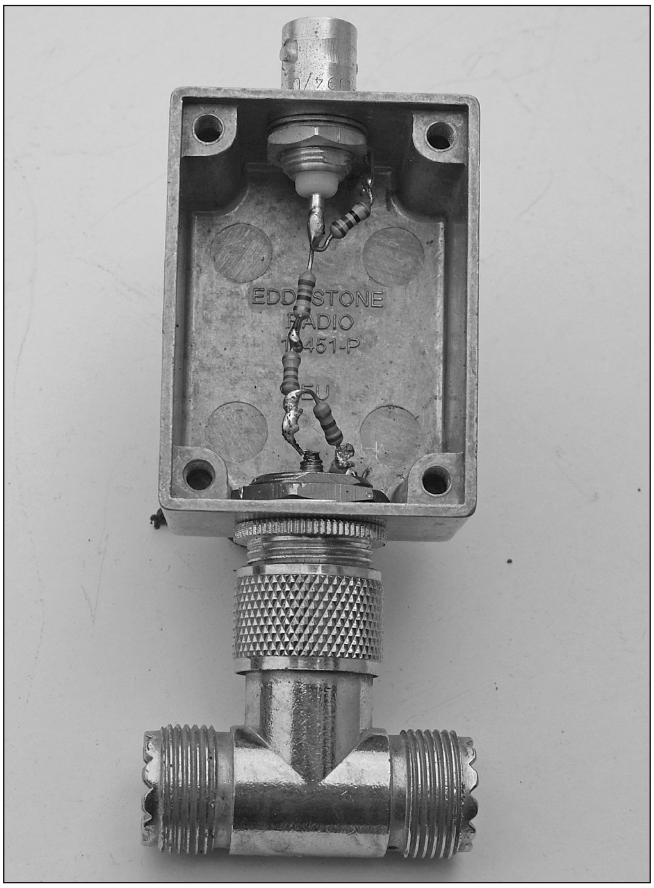


Fig. 24.51: A practical T piece sampler

attenuation of approximately 27dB. Typical T and Pi circuits with theoretical components values are given in **Fig 24.49** - the attenuator calculator at [22] can be used. The input resistance of these networks is approximately 1kΩ and the output 50Ω. Power calculations *must* be made for the resistors used. The protective components of Fig 24.48 can also be employed.

If only approximate values of attenuation are required then use the nearest preferred values of resistance found in the junk box or combinations of series and parallel circuits. See later in this chapter for typical methods of construction.

The T-connector Sampler

This concept can be found on various Internet sites. It is based on using a connector T piece and connecting to it a small metal box containing the sampling elements - see **Fig 24.50** for the general arrangement and the photograph in **Fig 24.51**. It is suggested that for safety reasons the connector to the test equipment is different to that at the 'power' end - eg a BNC connector. Use the connectors which are for your system but remember that the PL259/SO239 series is a non-constant impedance type and not really suitable for use above about 150MHz.

There are many variations on the sampling circuit and different boxes can have varying sample networks such as different attenuators and connectors - the choice is yours. The one shown here uses the pi circuit from above using the E24 series resistors. The advantage of not including a capacitor in series is that it can be checked at DC but remember to load with a 50Ω resistor at the BNC.

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. The purpose of this section is to explain briefly how a frequency counter works and gives tips for making measurements.

Fig 24.52 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

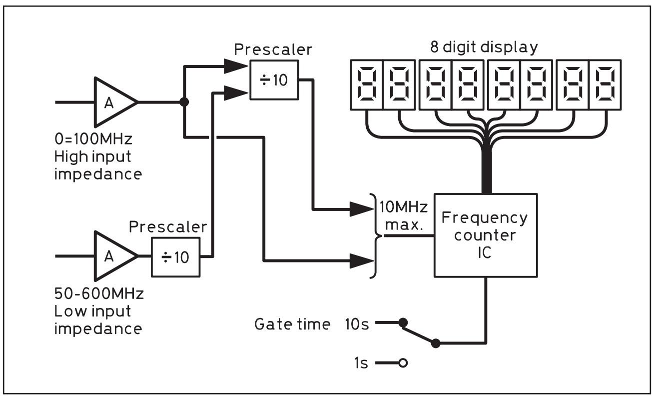


Fig 24.52: Block diagram of a frequency counter

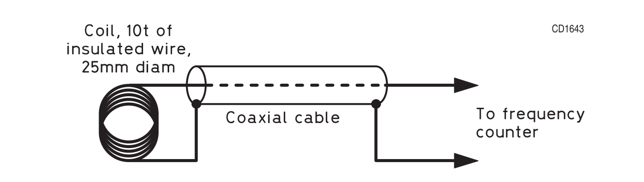


Fig 24.53: Pick-up loop

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 in Fig 24.53. 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.

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.

FREQUENCY STANDARDS

It is fine having test equipment but how accurate is it? Has it been calibrated recently? It is unlikely that an atomic frequency standard is owned or the cost of periodic calibration can be justified. So how can items such as frequency counters be calibrated?

In the past the best frequency standard available to most amateurs was a crystal oscillator carefully adjusted to zero-beat with a station of known frequency, such as one providing a standard frequency service - see Table 24.10.

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}. See the web sites at [23].

ATA	India	10,000kHz
BPM	China	2500, 5000, 10000, 15000kHz
CHU	Canada	3300, 7850, 14670kHz
DCF77	Germany	77.5kHz
HLA	Korea	5000kHz
MSF	Cumbria, UK	60kHz
LOL2	Argentina	5000, 10,000, 15,000kHz
RWM	Russia	4996, 9996, 14,996kHz
TDF	France	162kHz
WWV	Colorado, USA	2500, 5000, 10,000, 15,000, 20,000kHz
WWVB	Colorado, USA	60kHz
WWVH	USA	2500, 5000, 10,000, 15,000kHz
WWVH	Hawaii, USA	2500, 5000, 10000, 15000kHz

Table 24.10: A selection of standard frequency transmissions

An additional method that may be available is the line timebase of an analogue TV (providing the TV radiates well!) 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.

In the past, accuracies of the order of 0.1ppm were good enough to ensure keeping transmissions within band edges but for EME and weak signal VHF work this is not really good enough. Now there is available a high accuracy clock available from space - the GPS satellites. All that is required is equipment and a method to decode the signal and produce a usable output. There are various articles written on this theme and rather than re-produce them here, reference is made to some of them [24]. They all use various GPS modules which extract and output the 1pps signal and in some cases a 10kHz signal. These can often be found on eBay. The final project usually has a 10MHz output with better than 1 part in 10^9 stability and against which equipment can be calibrated. It is also worth looking at the Microwave equipment chapter earlier in this book.

SPECTRUM ANALYSERS

The spectrum analyser is an invaluable item of electronic test equipment used during the design, test and maintenance of RF circuits and associated equipment. Unfortunately, spectrum analysers tend to be quite expensive, although some can be picked up on the second-hand market. As time goes by some of the earlier models come within the pocket of the keen amateur.

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. However, where oscilloscopes look at signals in the time domain (horizontal axis is time), spectrum analysers look at signals in the frequency domain - a spectrum analyser displays the amplitude of signals on the vertical scale (as with an oscilloscope) but the horizontal scale is frequency.

Fig 24.54 shows a typical display and depicts a 30MHz signal with harmonics at 60, 90, 120 and 150MHz. The third harmonic (for example) is shown 48dB down on the fundamental.

The horizontal axis is linearly calibrated in frequency (in Fig 24.54 it is 1MHz/div) with the higher frequency being at the

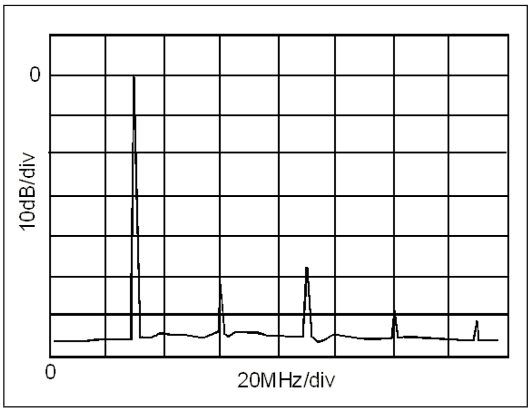


Fig 24.54: Typical spectrum analyser display

right hand side of the display. The plot of Fig 24.54 does not start at zero - the display can be offset by a fixed amount or alternatively centred on a specific frequency. The vertical axis is amplitude and can be either linear (V/div) or logarithmic dB/div). For most applications a logarithmic scale is chosen because it enables signals over a much wider range to be seen on the spectrum analyser. Typically a value of 10dB/div is used as in Fig 24.54. The scale is normally calibrated in dBm (decibels relative to 1 milliwatt) and therefore it is possible to see absolute power levels as well as comparing the difference in level between two signals.

The spectrum analyser can only deal with a small input power level so when dealing with transmitters a sample is taken of the output through a sampling device [1] that has a flat frequency response over the frequency range considered. Typically this may produce an output for the spectrum analyser some 30dB down (one thousandth) of the normal output power. Also *beware*, there may be a maximum permissible input DC voltage.

You can purchase a brand new one or if you are lucky 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 additional units to couple to PCs.

As an alternative it is possible to mix the output with a carrier and produce an output in say the 0 to 24kHz range that can be displayed on a PC using a soundcard. This does not allow the full examination of a 150MHz segment as shown in Fig. 24.54 but would allow, say, the signals about a carrier or harmonic to be examined and show a typical sideband structure or results of a two-tone test. A block diagram for such a system is shown in **Fig.24.55**. A good example of this approach is given at [25]. Possible software can be found in the software-based test equipment section of this chapter and [25].

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. It has the following definitions:

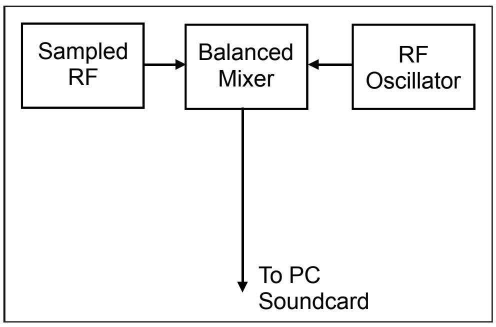


Fig 24.55: Typical arrangement for PC sound card spectrum analyser

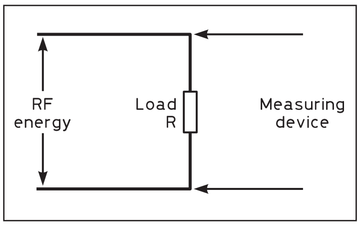


Fig 24.56: Power measurement arrangement

(a) *dBW is the power level in dB relative to one Watt.*
(b) *Peak envelope power is 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.*
(c) *Effective radiated power (ERP) (in a given direction) is the product of the power supplied to the antenna and its gain relative to a half-wave dipole in a given direction.*

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.

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 in **Fig 24.56**, 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; 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

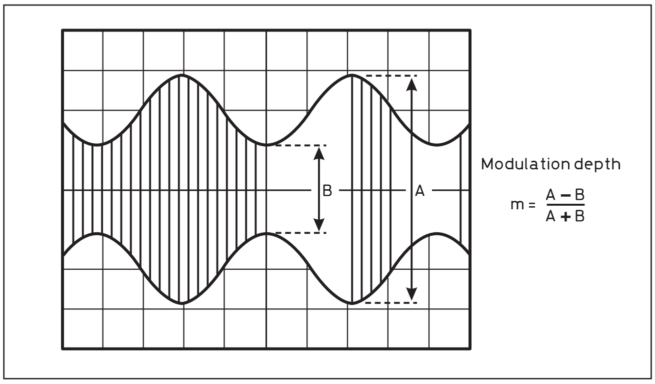
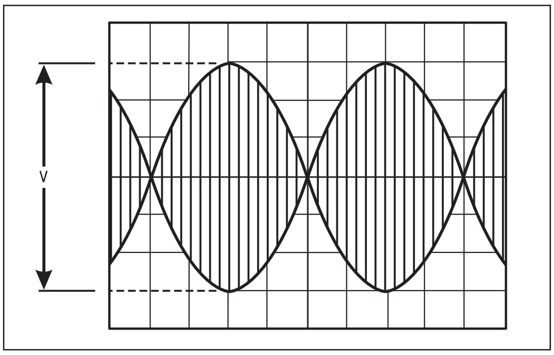
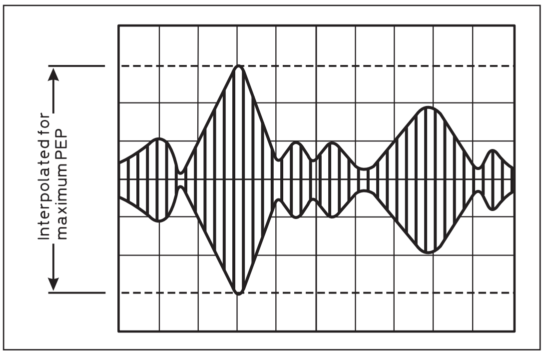


Fig 24.57: Modulation depth measurement



(left) Fig 24.58: Two-tone test display

(right) Fig 24.59: Speech waveform and interpolated maximum PEP level



increases. The power is divided between the sidebands and the carrier component. With 100% modulation the average 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 24.57** and calculate the modulation depth m. The output power is then given by:

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

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

For test purposes a constant amplitude 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 24.58 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:

$$P_{out} = \frac{V_{pk-pk}{}^2}{8R}$$

The equivalent peak-to-peak voltage reading can then be interpolated for the maximum allowable PEP and the position noted on the display. **Fig 24.59** 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 24.11** 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.

Circuit description

The circuit is shown in **Fig 24.60** and is a basic but tried and tested design using 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 24.61**.

The primary of the toroidal transformer is formed by threading a ferrite toroid onto 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.

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 24.11: Guideline for various VSWRs

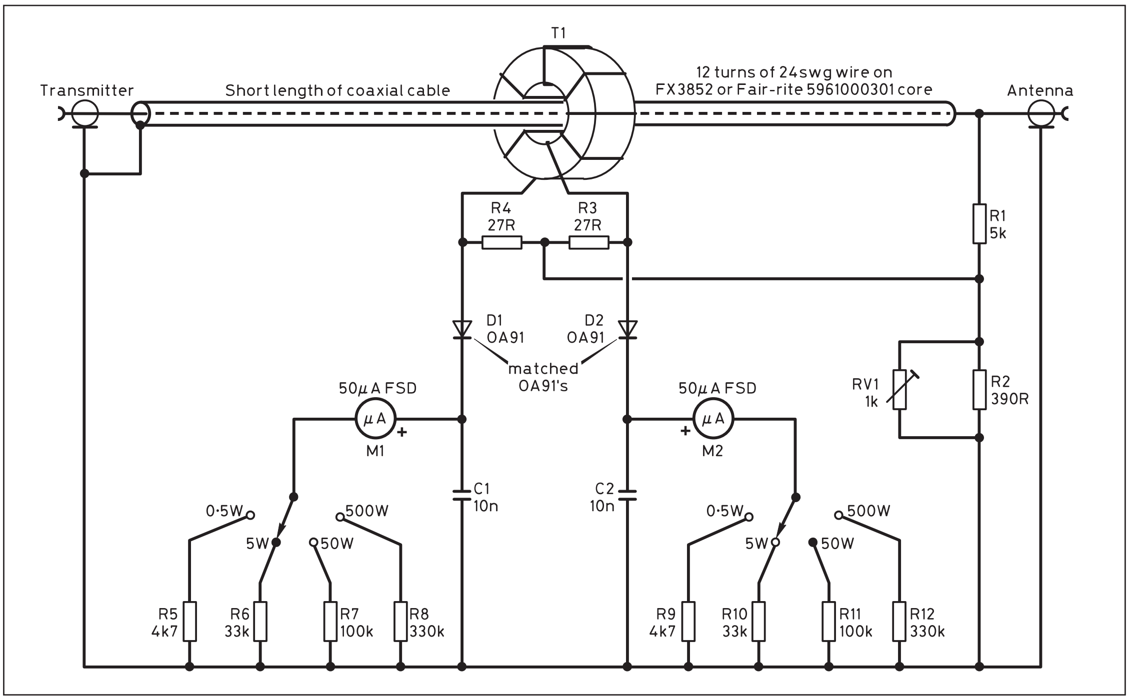


Fig 24.60: Circuit of the frequency-independent VSWR meter

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.

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, 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!).

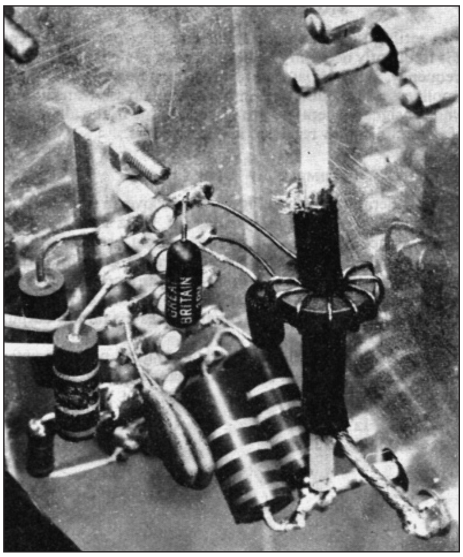


Fig 24.61: Construction of the frequency-independent VSWR meter

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 24.12: VSWR meter components list

A components list for this project is given in **Table 24.12**.

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 50Ω. 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.

POWER METERS

Not only are power meters needed to monitor output transmissions, but RF power measurements may be made when delivering power into another circuit (eg filter or amplifier). The serious experimenter may also need to measure very low powers from oscillators and driver stages. Further details are contained in [1].

The first project considers low power measurement and this is then followed by a Precision Peak-following Power Meter. The reader may also like to consider units in reference [1], and the "Crawley Power Meter" by G3GRO and G3YSX [27].

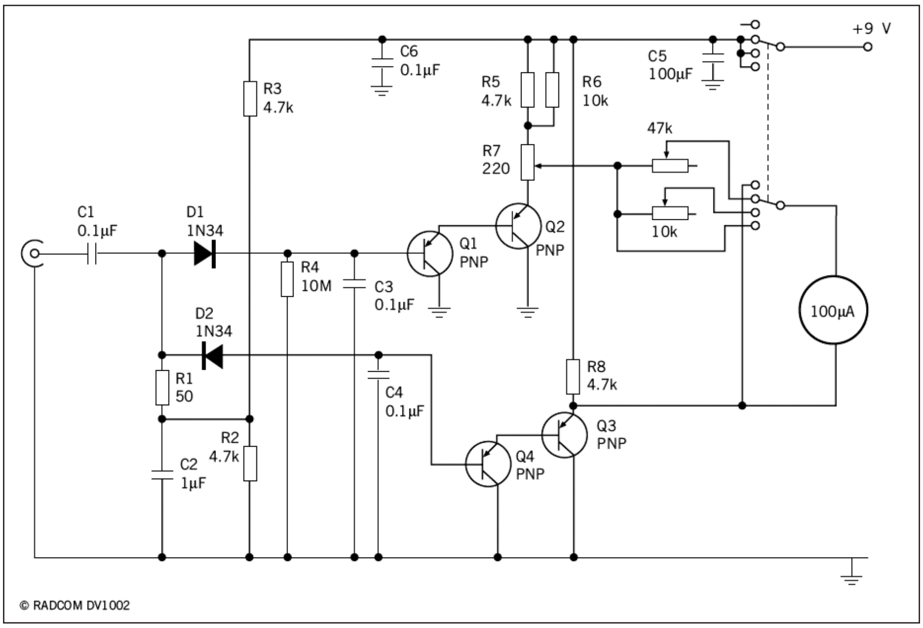


Fig 24.62: Circuit of the milliwatt meter

Fig 24.63: The meter in use

Low Power Measurements

Provided the power levels are greater than a few tens of milliwatts the voltage across a load (eg 50 ohms) can be measured by an RF probe and power determined by a familiar formula.

When the voltage drops below about 1V the forward voltage drop of any diode used becomes appreciable and the results become erroneous. One approach is to use an amplifier initially to raise the level of the AC voltage level and then rectify, a possible alternative was suggested by DF3GJ in an article in *RadCom* [28]. The circuit diagram for this is shown in Fig.24.62. The PNP transistors are general-purpose silicon types such as 2N3906. Fig.24.63 shows the milliwatt meter in use and the dBm meter scale. Three ranges of sensitivity were chosen, with full scales of -10, 0 and +10 dBm. Calibration needs to be made with a known power source and a stepped attenuator. The author states that calibration was made at 5MHz and the finished instrument gave satisfactory results at 144MHz and even 450MHz. Switching between ranges gave consistent readings. Signals as low as -20dBm (10µW) have been measured and the unit can positively detect signals to less than -25dBm.

See the original article for a fuller description of the project.

A Precision Peak-following Power Meter

The following is an abridged description of a power meter by G3GKG [29] 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 24.64.

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.

Using the correct type of toroid core (which must be of high permeability ferrite), this is a precision circuit (Fig 24.65) 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.

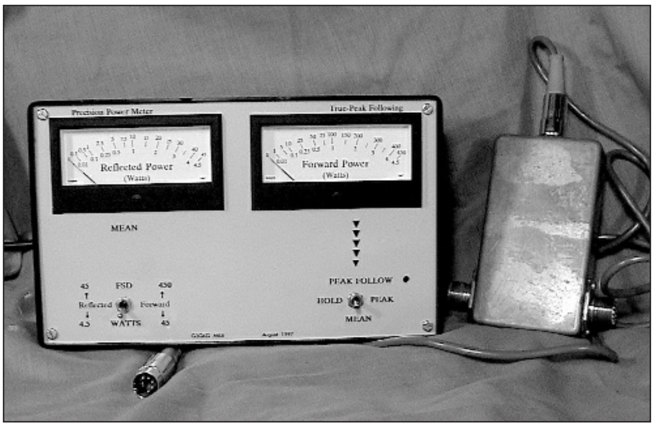


Fig 24.64: Precision peak-following power meter

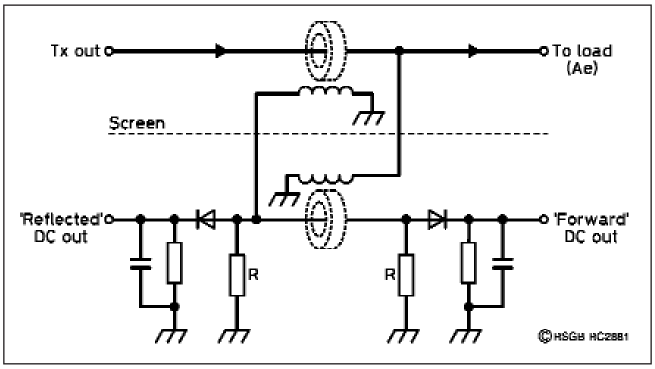


Fig 24.65: The Tandem Match - symmetrical between input and output

It is readily apparent from Fig 24.65 that the circuit is completely symmetrical and this is indeed borne out by its performance.

Reversing the transmitter and antenna 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.

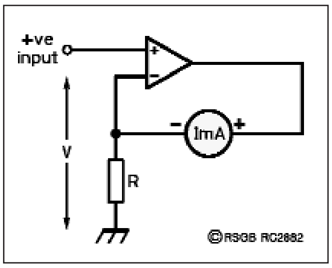


Fig 24.66: The basic meter circuit, showing the meter in the feedback loop of the op-amp driver

RF to DC

The display unit (Fig 24.64) 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, 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 24.66**. 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 24.65 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

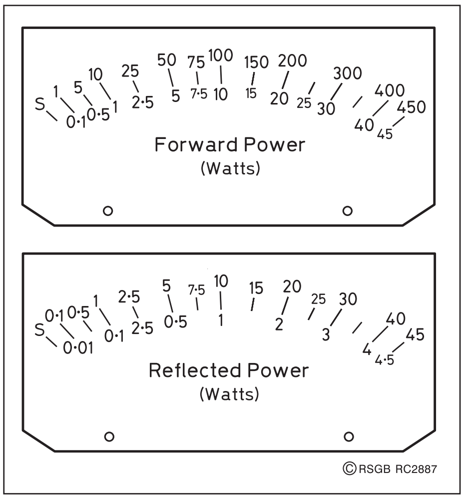


Fig 24.67: Meter scales showing the common calibration marks for the two power ranges

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 may be:

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

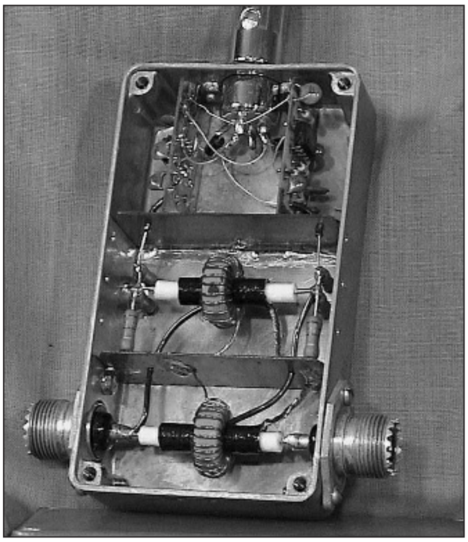
thus enabling a single set of scale markings to be calculated and used for all ranges, with different figures on the two meters such as those shown on **Fig 24.67**.

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.

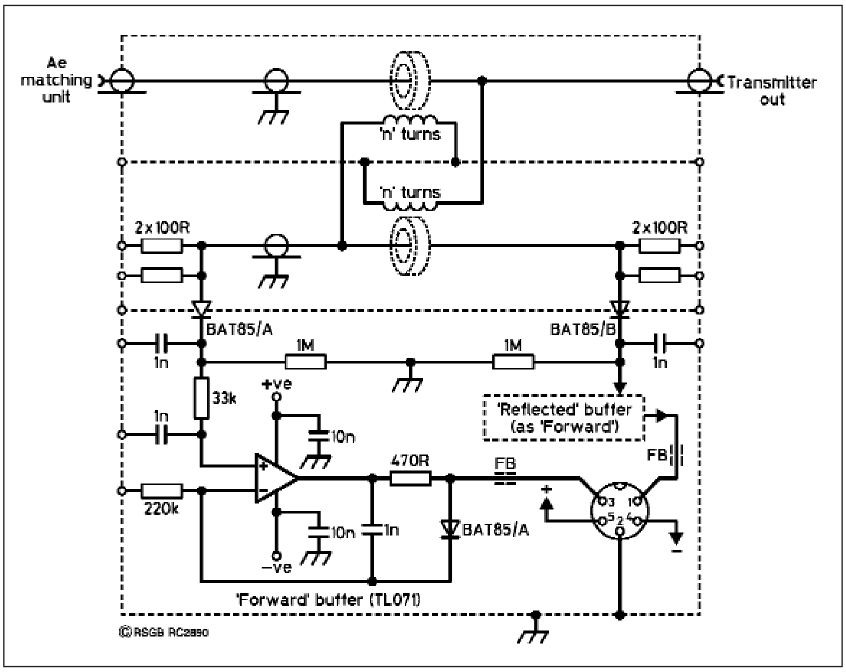
Peak Reading

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



(above) **Fig 24.68: Inside the RF head unit**



(right) **Fig 24.69: Circuit of the head unit**

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 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).

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 24.68**) is built into a die-cast aluminium box measuring about 115 x 62 x 29mm and includes the tandem match and detection circuitry (**Fig 24.69**). It is connected to the main display unit (Fig 24.64) 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 power ranges are required, could be accomplished by connecting an experimentally determined, high-value resistor from the negative supply to 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 24.70**) apart from possibly the power supply unit.

(*NOTE:* A negative rail is needed 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 24.70). 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!

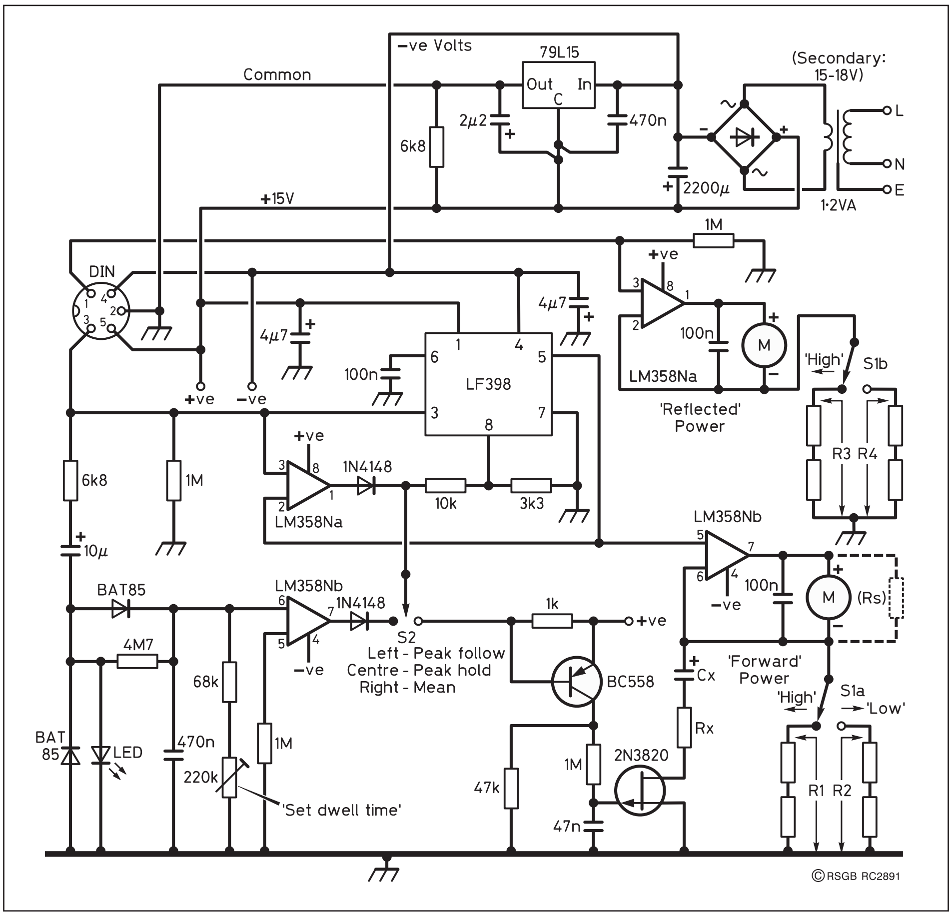


Fig 24.70: Complete circuit of the main display

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 also had a variant on this design using a digital bargraph instead of meters.

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 and suitable units are described earlier. 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 already been dealt with at the transmitter!).

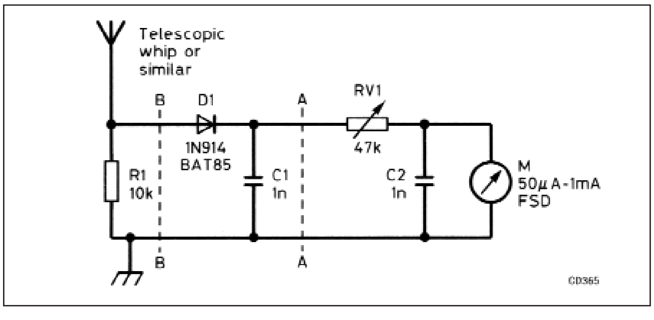


Fig 24.71: Simple untuned field strength meter

The alternative is to use a simple type of system such as that shown on **Fig 24.71**. 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 24.71 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 24.72 shows a circuit based around a MAR8 monolithic amplifier produced by Mini Circuits. The circuit has a response from DC to

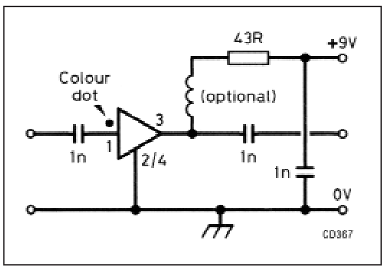


Fig 24.72: Broad-band amplifier based around a MAR8

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.

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

A Calibrated Field Strength Meter

This is an abridged version of the *RadCom* article by G3PVH [30].This equipment (**Fig 24.73**) uses a small 'known antenna', the characteristics of which can be easily established. This is followed by a very sensitive diode-detector and meter amplifier.

To obtain high sensitivity from the diode detector, it is used in its square law (or small-signal) region. The output current (DC) is proportional to the square of the input RF voltage, or in other words, to the power. To ensure a true square law, the amplifier has a low input-impedance so that the diode characteristic is not made linear by the load resistance. Square law operation allows power to be read from the scale in linear fashion, but voltages must be converted using a table or graph. *NOTE:* a suitable meter scale could be drawn using a computer program (see the software section of this chapter).

The circuit must work down to zero volts DC, thus a symmetrical supply is required. This can be achieved by using two batteries or as in **Fig 24.74** a single battery with an op-amp to form the mid-voltage point. The meter amplifier uses an op-amp in a circuit known as a current-to-voltage converter and uses a single feedback resistor to set the gain; it has a low input resistance. If the feedback resistor is 1MΩ, then 1μA into the input terminal will produce 1V across the output. A feedback resistor of 1MΩ will give an RF sensitivity of about 250mV and 30MΩ will give 30mV. The circuit shown incorporates a zero-setting control which may be needed when very high gain is used. Unfortunately, it is not possible to calculate the overall sensitivity of the circuit due to the variations in diode characteristics.

The project can be built on matrix board or similar and housed in a small box (certainly for a UHF version use a metal box and good quality RF connectors).

Two types of antenna can be used, a frame antenna or a 2m-long dipole - in each case the 'effective height' (or effective length) is calculated to enable conversion from the antenna terminal voltage to field strength. 'Effective height' is an old term that was used when antennas were mostly inverted-L or T-type, sometimes it is now called antenna factor. It is the conversion factor from field strength to antenna voltage (strictly speaking, antenna EMF).

The frame antenna can be varied in size depending on how much sensitivity is needed. For instance, a 235mm (9in) diameter frame antenna gave full scale deflection at 1km from a 1kW medium wave station. A 1m square frame gave full-scale deflection at 10km from the same station and at 500km from Allouis (ERP 2MW) on 162kHz. Typical constructional details are given in **Fig 24.75**. For operation above 15MHz, it is better to use the smaller antenna.

The frame antenna is tuned by a 500+500pF variable capacitor. At HF the two capacitor sections are normally in series and the rotor left floating. For low frequencies the two sections can be wired in parallel. At VHF a smaller capacitor will be needed.

To find the effective height, Q needs to be determined. Couple a little energy from a signal generator by means of a 150mm (6in) coupling loop and find the bandwidth B (kHz) for half deflection. This is the half-power point - Q can then be found from the equation

$$Q = F / B$$

where F is the centre frequency in the same units as B

The effective height (H) in metres can now be found from the formula

$$H = 2\pi N A Q / \lambda$$

where N is the number of turns, A is the loop area in square metres, Q is as described above and λ is the wavelength in metres.

The dipole antenna is the best for use above 30MHz but it will still work even at LF. It is 2m long, made from a pair of telescopic whips, and has an effective height of 2m. It should be mentioned that the field strength meter has very high input impedance to RF as a result of the diode resistance. The low resistance input of the DC amplifier is swamped by the diode resistance when working with very small signal currents.

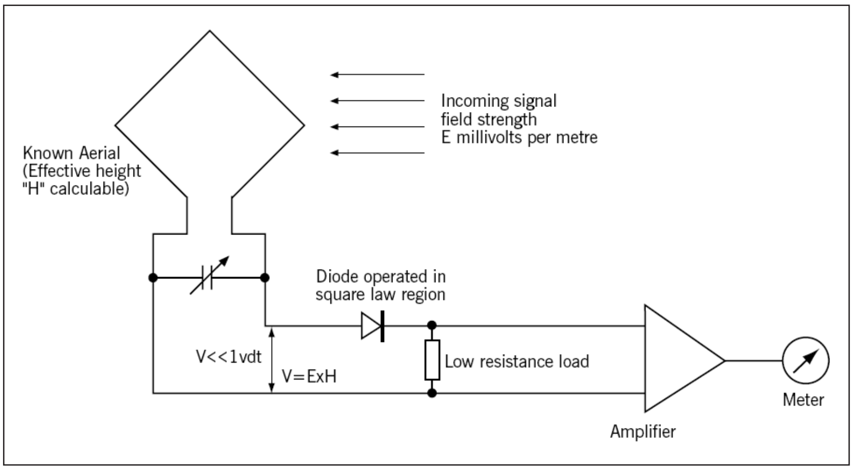


Fig 24.73: Concept of the field strength meter

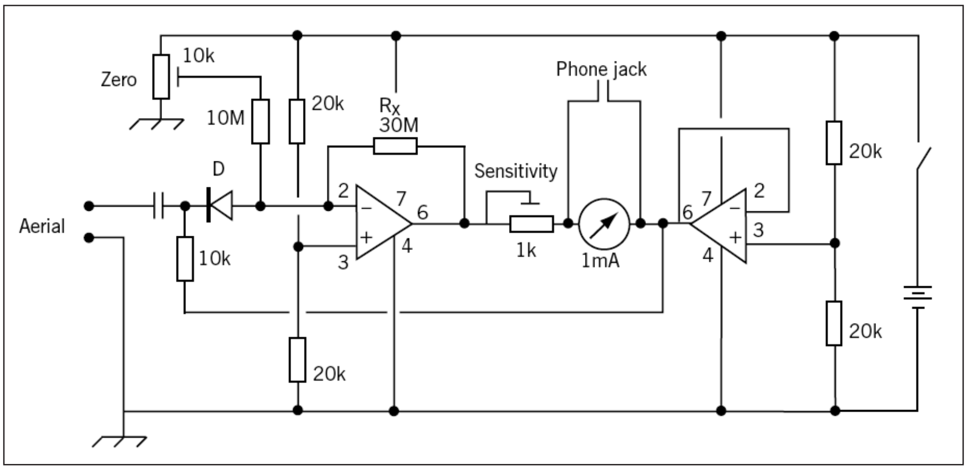


Fig 24.74: Sensitive circuit using a second op-amp to provide mid rail voltage

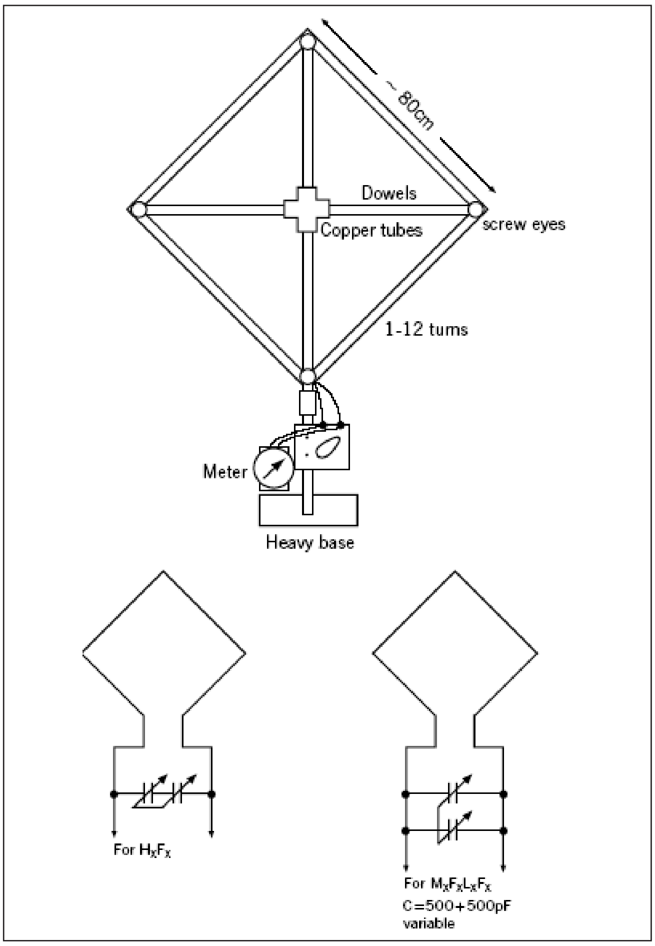


Fig 24.75: Making a frame antenna

The high input impedance avoids drawing current from the antenna, so that the terminal voltage is equal to the EMF. In other words, the field strength meter is a true voltmeter which does not unduly load the device under test. Provided the frame can tune to the required frequency there is no need to adhere to the dimensions given here, as the calibration procedure will take care of any variations in efficiency.

The small frame can be round, about 235mm diameter, and wound with thick wire, say 16SWG. It is then taped over for rigidity and the exit wires are formed into eyes as connection points. A tuning capacitor will be required, either in the field strength meter housing or in a separate box, maybe 500+500pF, or smaller for high frequencies. The winding details are shown in Table 24.13.

The large frame Fig 24.75 was made using wooden dowelling to form a diamond. The antenna will require a heavy piece of wood as a base. The dowelling can be grooved at the ends to hold the wire, or screw eyes can be used. It is, however, difficult to wind a lot of turns using screw eyes.

Determine the Q as described earlier at top, middle and bottom of the range - be very precise in making the measurements. Then calculate the effective height at these frequencies and write it on the front of the instrument. Usually the effective height changes only slowly with frequency.

To calibrate, set up a circuit such as Fig 24.76 to provide up to, say, 100mV of 50Hz AC. If necessary connect a 0.1µF capacitor across the input capacitor of the instrument. Do not use an antenna. Then vary the AC supply until each point on the meter scale is reached. Note the AC voltage applied. This will give you a result like Table 24.14.

Table 24.13: Construction details of the calibrated field strength meter

800mm square frame antenna

Number of turns	*Capacitor (pF)*	*Frequency*
12	2000	170kHz
12	1000	225kHz
12	50	660kHz
6	250	800kHz
6	25	1.8MHz
3	250	1.5MHz
3	25	4.5MHz
2	250	2.2MHz
2	25	7.1MHz
1	250	3.9MHz
1	25	15.5MHz

230mm round frame antenna

Number of turns	*Capacitor (pF)*	*Frequency*
4	1000	1.5MHz
4	250	2.75MHz
4	50	4.7MHz
4	25	6.7MHz
2	250	4.7MHz
2	25	7.5MHz
1	250	10.5MHz
1	25	28MHz

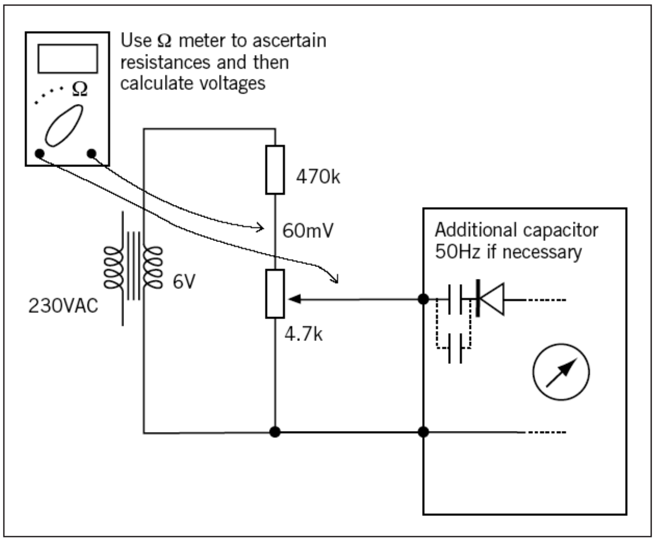


Fig 24.76: Calibration circuit for the field strength meter

Scale reading	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9	1.0
mV	10	13	16	19	21	23	25	27	28	30

Table 24.14: Calibrating the field strength meter

Scale reading	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9	1.0
mV	10	13	16	19	21	23	25	27	28	30
µW in 50Ω	2	3	5	7	9	11	13	15	16	18

Table 24.15: Adding a microwatt scale

If wanted, a microwatt power scale can be added (see **Table 24.15**) by using the formula $P = V^2/50$, where P is the power into 50Ω in microwatts and V is the voltage in millivolts. The microwatt scale applies when you measure across a 50Ω resistor.

To measure the field strength, set up the loop edge-on to the transmitter, tune the antenna, note the scale reading and convert it to voltage. Then divide the voltage (in millivolts) by the effective height in metres. This will give the field strength in millivolts per metre.

The easiest measurements are of vertically-polarised transmitters and low frequencies. For instance, measurements of 80m mobile transmitters would be interesting. UHF measurements over a distance of a few metres using dipoles etc also work very well. Excellent results are achieved at 1km from medium wave transmitters, at which distance the field strength is always 300mV/m for 1kW radiated. At other distances use the formula:

$$E = 300\sqrt{(P / D)}$$

where E is the field strength in mV/m, P is the power in kW and D is the distance in km.

If you are measuring dipole aerials in free-space conditions, a convenient formula is:

$$E = 7000\sqrt{(p / d)}$$

where E is the field strength in mV/m, p is the power in watts and d is the distance in metres.

Of course, you can find power if you know field strength and vice versa. Usually it is best to measure close to the transmitter, one or two wavelengths say, but if you want to measure large arrays, especially the depth of radiation pattern minima, you may need to measure at greater distances. Do not be overly concerned about induction- and near-field effects; you can always try doubling the distance to confirm that the field is halved.

In conclusion, actual measurements are more valuable than folklore, so perhaps this simple, calibrated field strength meter will further a worthwhile cause. It will enable very small voltages and powers to be measured and will enable the ERP of a station to be ascertained. The method is to use a known antenna such as a frame antenna whose characteristics can easily be found. Then use a diode as a sensitive square law detector and use the very simple current-to-voltage converter circuit as a meter amplifier.

ATTENUATORS

See the Building Blocks chapter earlier for further details on building and using attenuators.

A CLIP-ON RF CURRENT METER

The following is from 'In Practice' in *RadCom* [31], see that article for earlier references. The basic version of this handy device takes about 10 minutes to tack-solder together (**Fig 24.77**). 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 24.77. (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 24.77. are discussed in G0SNO's article which is reproduced on the 'In Practice' website [32]. 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 you will soon be thinking about something more permanent. The classic way to do this is using a clothes-peg **Fig 24.78** but there are now better alternatives.

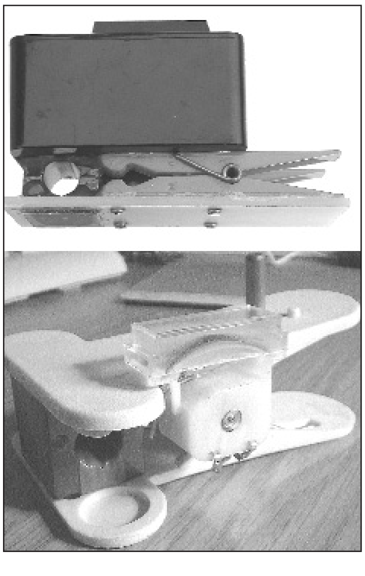


Fig 24.78: Two versions of the clip-on RF current meter

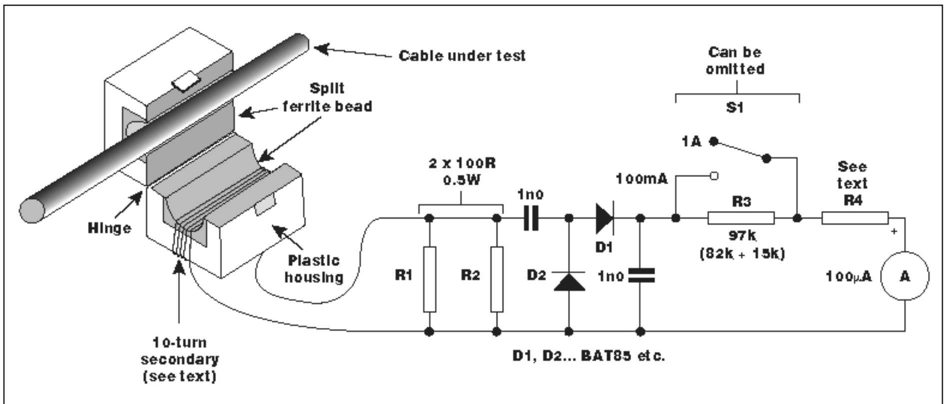


Fig 24.77: G0SNO's clipon RF current meter

Function	Brief Description	Filename
DTMF, CTCSS, tones	Produced by ComTekk. This is a trial version but nonetheless appears very useful, some functions are partly disabled. It allows DTMF, CTCSS, paging and other tones to be generated for testing radio equipment. The software allows auto-calibration and has an audio oscilloscope and spectrum analyser.	*ComTekkSetup106_EU.exe*
DTMF, CCIR, ZVEI and CTCSS decoder	This is a multi-tone decoder with recording log. Typical tonesets are DTMF, CCIR, ZVEI etc) plus CTCSS. It could be useful to those testing and repairing equipment. It is an older program and not so easy to find on the web but use filename as search word.	*Wintn22.zip*
Bench top calculators	1. A very useful calculator with many engineering and radio functions and conversions. Free. 2. Gcalc (Linux) at http://live.gnome.org/Gcalctool. Free	*Benchtop-calc.zip*
Intermodulation	1. A free download for calculating intermodulation products up to the fifth order. 2. Free program courtesy of TCS Consultants at www. tcstx.com/software/intermod/intermodulation.htm	*imodbeta.msi*
Meter dial maker	1. This is a basic version and will allow meter scales to be designed and printed. Useful when wanting to make your own scales to fit an analogue meter. 2. *Galva* from www.radioamateur.org/download/index.html	*MeterInstall221.exe* *Galva_185-3L.zip*
Oscilloscope	1. There are various programs that perform this function along with other facilities such as audio spectrum analysis. See www.dxzone.com/catalog/Software/Oscilloscope and www.tech-systems-labs.com/test-software.htm 2. *Xoscope* at http://xoscope.sourceforge.net. Linux. Free	various
Audio signal generator, including two tones for SSB testing	There are several programs that perform this function, see www.dxzone.com/catalog/Software/Signal_Generator and www.tech-systems-labs.com/test-software.htm	various
Polar plot	This program will measure and display the radiation pattern of a beam antenna. It relies on an antenna receiving a signal and the audio output (assuming AGC not operative) plotted with rotation.	*PolarPlotSetup.exe*
Real time audio spectrum analyser	Although the title suggests it is an audio spectrum analyser, it also functions as a complex signal generator and oscilloscope.	*TrueRTA_se.exe*
RF toolbox (1)	A DOS based program with various calculation options including noise and gain calculations, R, L and C calculations (including attenuators), resonance and VSWR.	*rftbox.zip*
RF toolbox (2)	This program allows various calculations to be made as well as antenna design. Trial version by Black Cat Systems.	*RFToolBoxSetup.exe*
SINAD measurement	A program from ComTekk using the latest DSP technology to measure some key receiver performance parameters by analysing the spectral content of received audio signal. A trial version.	*CTSinadSetup120.exe*
Spectrum analysers	Various programs perform this function in the audio range, the programs normally offer additional functions See www.dxzone.com/catalog/Software/Spectrum_analyzers and www.tech-systems-labs.com/test-software.htm	various
VSWR calculator	This is a calculator and contains calculators for Directivity Error, Mismatch Error and Ratio to dB.	*vna.exe*
RF system analysis	A program for modelling the noise, gain, intermodulation/spurious performance, and compression point of a system of cascaded components where each component is characterized by known parameters (as may have been determined by measurements). The Lite version of the program is freeware.	*Scw5Lite.exe*

Table 24.16: Typical software test equipment

For example, the first photograph (Fig 24.78) shows the rather heavy-duty version using two strong clothes-pegs, fibre-glass sheet and epoxy glue (more details at [32]). 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 basically 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!

SOFTWARE-BASED TEST EQUIPMENT

As in most walks of life, the use of the ubiquitous PC has been combined with software to provide alternatives. This is true with some type of test equipment. **Table 24.16** is only a list of suggestions, others are sure to be found. There are also programs that will run on PDAs and even mobile telephones. The free programs are generally only available for non-commercial use. Please read the agreements. Most programs are available from several sources and they can be easily found by using a search engine.

It is only possible to give a very brief overview of each program here, but see [1] for more details. Aternatively, search the Internet for further information or download a program and try it. Soundcard programs are generally limited by the card to a maximum frequency of about 20kHz. Many programs are *Windows* based, they have all been tried on *Windows XP*. Some of the programs may have problems with the latest offering from Microsoft - *Windows 7*. There are Mac and/or *Linux* versions of some the programs or equivalents. It is possible to run some *Windows* programs on *Linux* using an emulator such as *Wine*.

A word of warning. Be careful with the interface to the PC, high voltages or misuse may damage the computer. When using inputs and outputs from the soundcard they should at least be buffered and possibly isolated (eg transformers or opto-isolators).

TRANSMITTER / RECEIVER MEASUREMENT

The previous sections have dealt with the test equipment that can be used for making various measurements, this section looks specifically at the type of tests and measurements that can be performed on equipment. Many of these measurements require quality test equipment which only a few will either own or have access to. Some of this equipment is obtainable at radio rallies, surplus stores or Internet auction sites.

A good source of reference, and with further details, is Peter Hart's book, specifically the first chapter [33]. There are also some good articles on the Internet. The following can only be a short introduction to these measurements within the confines of this chapter.

Receiver Measurements

In very simple terms, there are broadly two sets of parameters which define the effectiveness of any receiver - how well it receives wanted signals and how well it rejects unwanted signals.

The types of receiver measurements can be expanded into various types as suggested below.

- Sensitivity and SINAD
- Noise figures
- AGC response
- Intermodulation and blocking
- Receiver reciprocal mixing

Sensitivity

Fig 24.79 shows the test arrangement for making measurements of sensitivity, spurious response rejection and selectivity. A suitable resistive load is connected to the external loudspeaker socket of the radio and the audio output monitored using a true RMS voltmeter. It is important that this indicator shows the true RMS level of the noise output as well as mixed noise and sine wave output. A dB scale is by far the most convenient. The audio output level should be low enough to avoid audio distortion at all times. The audio monitor shown in Fig 24.79 is just a simple amplifier and speaker so that the receiver audio can be heard.

As an alternative to the voltmeter, an audio spectrum analyser or FFT analyser can be used with software to compute signal-to-noise ratios directly. This can have a number of advantages with improved accuracy and ease of separating signals and noise.

Sensitivity measurements on SSB or CW are made with the RF generator set to give a 1kHz audio beat note. With the RF generator switched off the audio output level on noise alone is set to a convenient value. The generator is then turned on and the RF level set to give a 10dB increase in audio output. This level gives the sensitivity for 10dB (signal + noise) to noise ratio (s + n): n.

The receiver noise floor is 9.5dB lower than this figure. If the generator had been set to give a 3dB increase in audio output, the level would be equal to the noise floor of the receiver. Measurements of (s + n) : n for AM and FM signals are made, not by turning the RF generator on and off, but by switching the 1kHz modulation on and off: 30% modulation depth is usually used for AM and 60% of the maximum deviation is used FM (eg for a 25kHz channel spacing, which uses a maximum of 5kHz peak deviation, the modulation is set to 3kHz).

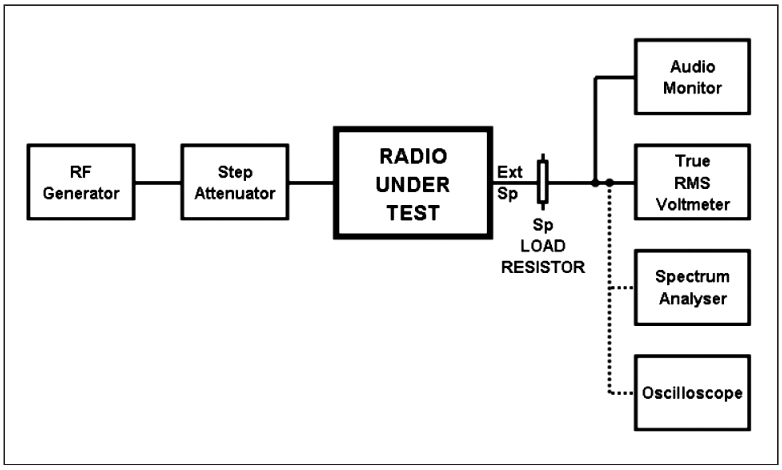


Fig 24.79: Sensitivity related receiver measurements

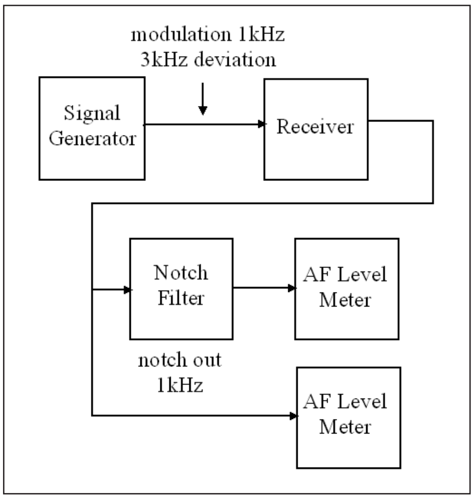


Fig 24.80: Arrangements for SINAD measurement

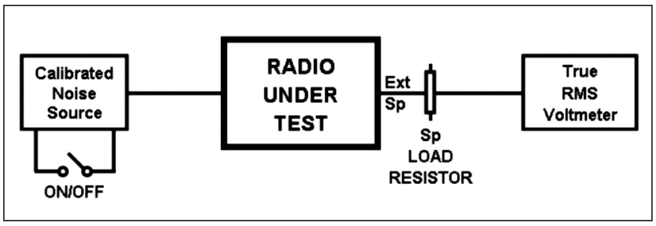


Fig 24.81: Noise figure measurements

SINAD

Another measurement that can be made to assess and specify the sensitivity performance of a radio receiver is SINAD (signal plus noise plus distortion to noise plus distortion). It is a common performance measurement in many applications including many two-way FM radio systems especially at VHF and above.

SINAD is a measurement that can be used with any radio communication equipment and examines the degradation of the signal by unwanted or extraneous signals including noise and distortion. The SINAD measurement is typically used for measuring and specifying the sensitivity of a radio receiver.

SINAD is defined as the ratio of the total signal power level (Signal + Noise + Distortion) to the unwanted signal power (Noise + Distortion). Hence the higher the figure for SINAD, the better the quality of the audio signal.

The SINAD figure is normally expressed in decibels (dB) and can be calculated from the formula:

$$\textbf{SINAD} = \textbf{10}\ \textbf{Log}_{10} \left\{ \frac{S+N+D}{N+D} \right\} \textbf{dB}$$

where: S+N+D = combined Signal + Noise + Distortion power level; N+D = combined Noise + Distortion power level. *NOTE:* SINAD is a power ratio and not a voltage ratio in this calculation.

Fig 24.80 shows a typical arrangement for measuring SINAD. An RF signal modulated with a 1kHz audio tone is fed to the radio receiver. Two audio measurements are then made, one with the 1kHz modulation and one with it notched out. Once the figures for the signal plus noise plus distortion and the noise plus distortion are obtained it is then possible to calculate the value of SINAD for the radio receiver.

The notch filter characteristics for SINAD measurements are of importance and ETSI (European Telecommunications Standards Institute) defines such a notch filter in ETR 027. Briefly this specifies that for the standard modulating frequency of 1kHz, the 1kHz should be attenuated by at least 40dB, at 2kHz the attenuation should not exceed 0.6dB and the filter characteristic shall be flat within 0.6dB over the ranges 20Hz to 500Hz and 2kHz to 4kHz. Without modulation, the filter shall not cause more than 1dB attenuation of the total noise power of the audio frequency output of the receiver under test.

Whilst measurements of SINAD can be made using individual items of test equipment, there are a number of commercial SINAD meters. These meters connect directly to the radio receivers for the measurements. There is also a piece of software called *SINAD* using the PC soundcard (see software based test equipment section in this chapter).

Sensitivity performance of a radio receiver can be assessed by determining the RF antenna input level to achieve a SINAD figure of 12dB (a typical figure that is used commercially). This equates to a distortion factor of 25% with a modulating tone of 1kHz. (Note: the typical ETSI - European Telecommunications Standards Institute - specification states that a deviation level of 12.5% of the channel spacing should be used, with AM the modulation depth needs to be specified).

A typical specification for a narrow band FM receiver might state that a receiver has a sensitivity of 0.25μV for a 12dB SINAD. The lower the RF input voltage needed to achieve the given level of SINAD the better the receiver performance. A 12dB SINAD figure is considered the maximum acceptable level of noise that will not swamp intelligible speech.

Noise figures

As an alternative to measuring (s + n) : n or SINAD, the noise figure of a receiver may be measured using a suitable calibrated noise source. This technique gives more accurate results with low noise VHF or UHF receivers and has the advantage that noise figure is independent of receiver bandwidth. Noise figure, noise floor and bandwidth may be related for an SSB receiver as follows:

noise floor (dBm) = noise figure (dB) + 10 $\log_{10}$ (bandwidth) - 174
where bandwidth is in Hertz

Fig 24.81 shows the test configuration for making noise figure measurements. A detailed description is beyond the scope of this book but, essentially, two readings are taken, one of the receiver output with the noise source off and a second with the calibrated noise source switched on. It is important that the impedance presented to the receiver antenna connector does not change between on and off states. If the noise source is variable, it is set to give a 3dB increase in noise level and the receiver noise figure is simply read from the calibrated scale on the instrument. With fixed noise sources, the noise figure of the receiver is calculated from the on and off readings.

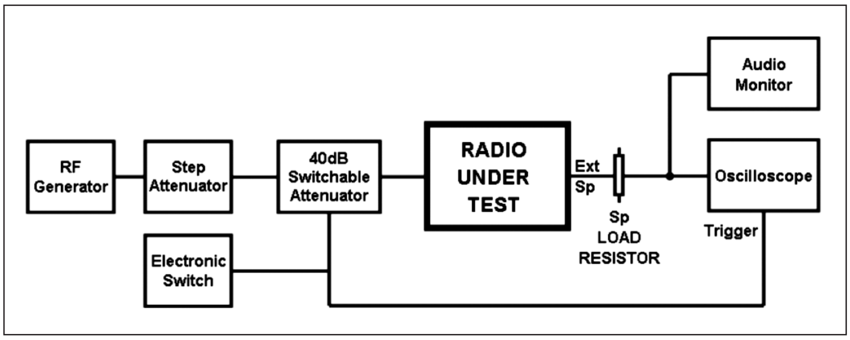


Fig 24.82: AGC measurements

It is worth reading an article by G8KBB in *RadCom* [34] on the Measurement of Noise. This also gives details of a noise source.

AGC response

Fig 24.82 shows the test configuration for making AGC measurements. The drive source from the generator includes a home-constructed 40dB attenuator which can be switched quickly and cleanly in or out of circuit by an electronic trigger signal. This signal is also used to trigger an oscilloscope scan to view the audio output from the receiver. The attack and decay characteristics for a 40dB change in level over a range of signal inputs can be observed together with the threshold level at which the AGC starts to operate

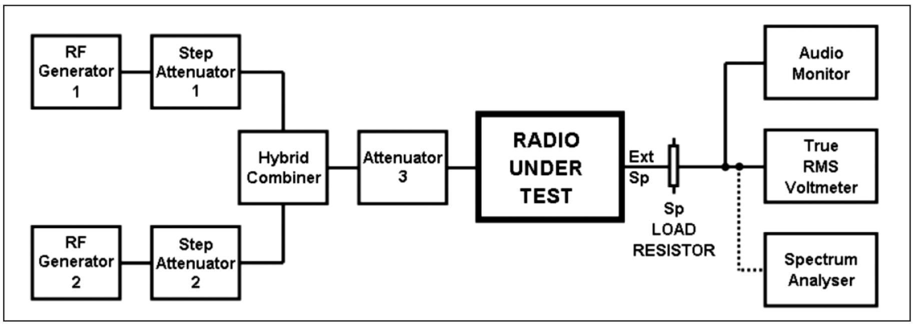


Fig 24.83: Intermodulation and blocking receiver measurements

Intermodulation and blocking

The test arrangement for making measurements on intermodulation and blocking is shown in **Fig 24.83**. This requires the use of two signal generators.

To make intermodulation measurements, it is most important to ensure that intermodulation products are not produced within the generators by one generator coupling into the other. A hybrid coupler and not just a resistive combiner is necessary as this will give some 30dB extra isolation. This extra isolation will only be achieved if the coupler is properly terminated at the output port and this requires that attenuator 3 in Fig 24.83 is not reduced too low in value. Typically use a 10dB attenuator.

Even with a suitable coupler, it is quite difficult to keep test equipment intermodulation within limits and get accurate results when receiver dynamic ranges exceed 100dB. Some generators are better than others and a careful balance between the generator output is often necessary to achieve best results.

To measure the third-order intermodulation, the levels of the two generators, are increased equally by using the step attenuators until the amplitude of the intermodulation product generated in the receiver gives a measured s + n : n ratio of 10dB.

NOTE: if the AGC is operative at this level, a spectrum analyser or notch filter needs to be used as for sensitivity measurements under this condition.

The difference between the amplitudes of either input signal as measured at the antenna input and the on-tune level or sensitivity figure for similar s + n : n is termed the intermodulation ratio. It is more convenient to quote intermodulation performance in terms of Intercept Point (IP), as this is independent of measurement signal-to noise ratio and bandwidth.

$$\text{3rd-order IP (dBm)} = \frac{(3S - I)}{2}$$

where: S is the amplitude in dBm of each input signal and I is the amplitude in dBm of intermodulation product generated when related to the receiver input.

Having measured the third order Intercept (IP3) and the noise floor (NF) of the receiver from the sensitivity measurement, the two-tone Spurious Free Dynamic Range or Intermodulation-Limited Dynamic Range is calculated from the expression

Dynamic Range (dB) = 0.667(IP3 - NF)

where IP3 and NF are expressed in dBm.

For every 1dB increase in generator levels, the third-order intermodulation products in the receiver increase by 3dB. If this does not happen, intermodulation may be occurring simultaneously in more than one stage in the receiver, front-end AGC may be operative, or intermodulation products may be generated within the measuring set-up.

The above expressions for IP3 and dynamic range assume that the 3:1 ratio holds and cannot be accurately applied if this is not the case.

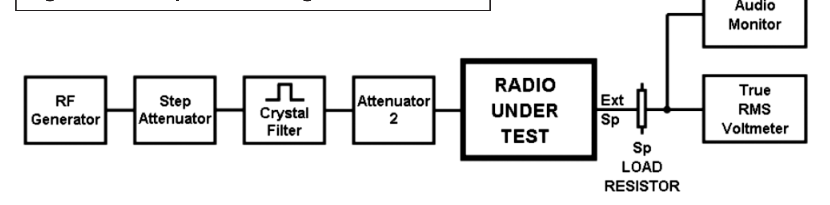
IP3 measurements are made over a range of frequency spacings, with and without the front-end preamplifiers. Second order measurements are also made to assess front-end filter effectiveness. In-band linearity is assessed by setting the generators 200Hz apart, centred in the receiver pass-band and observing the audio output on a spectrum analyser. The resulting intermodulation products observed are a good indication of the linearity of the total signal path right through to the loudspeaker. It is quite noticeable that a really clean-sounding receiver generally shows intermodulation products some 40 to 50dB down on either of the two tones and with a less-clean-sounding receiver these levels are only 20 to 30dB down. The results usually hold over a wide range of signal levels and are often improved by manually reducing the RF gain control.

Sometimes slow AGC gives better results than fast. Front-end blocking is caused by gain compression in the receiver front-end stages ahead of the main IF filters. Generator 1 is set on-tune at a defined S-meter level (typically S9). Generator 2 is offset from the on-tune frequency and the level increased until the S-meter reading drops by 1dB. The level of generator 2 related to the receiver input is taken as the blocking level. Measurements at different offset frequencies are usually made. If reciprocal mixing is poor then it may not be possible to measure blocking by this method, but then it is probably irrelevant as well. AGC is not normally applied to the front-end of modern receivers but, if it is, the measured blocking level will be dependent on the on-tune signal level. Measuring at lower on-tune levels may not be possible due to reciprocal mixing.

Receiver reciprocal mixing

Fig 24.84 shows the test arrangement for making reciprocal mixing measurements. It is most important that the sideband noise

Fig 24.84: Reciprocal mixing measurements



spectrum of the RF generator is considerably lower than that of the receiver local oscillator. In order to achieve this, measurements are made at a single frequency by inserting a narrow bandwidth crystal filter together with suitable matching components and attenuators between the generator and the receiver. For example, a 21.4MHz filter could be used and then all measurements relate to the 21MHz band.

The generator should be tuned to the pass-band of the filter and as close to the roll-off as possible on the measurement side. The total loss of the filter, matching components and attenuators are noted for calibration. The receiver is tuned away from the generator frequency, noting the generator level required to give a 3dB increase in noise output from the receiver. Hence the noise due to reciprocal mixing is then equal to the noise floor of the receiver.

The phase-noise-limited dynamic range is the difference between the generator level as seen at the receiver input and the receiver noise floor. Measurements are made in SSB bandwidths for compatibility with the other measurements, although CW bandwidths may really be better for close-in measurements. The phase noise of the receiver synthesiser in dBc/Hz can be calculated at the specified offset with the equation

$$\text{Oscillator noise (dBc/Hz)} = -(\text{PNDR} + 10 \log_{10} B)$$

where: PNDR is the phase noise limited dynamic range in dB; and B is the receiver noise bandwidth in Hz (typically 2500 for SSB).

TRANSMITTER MEASUREMENTS

With transmissions one is interested in spectral purity so as not to cause undue interference to other users of the spectrum and have what many call a 'clean' signal. This involves both the initial transmitter and any following amplifiers used.

The test arrangement for carrying out CW transmitter measurements is shown in Fig 24.85. The measurement of power, harmonic and spurious outputs is largely self-explanatory. It is important to keep input attenuation levels on the spectrum analyser as high as possible to avoid the possibility of inaccurate harmonic measurements due to harmonics being generated within the analyser itself.

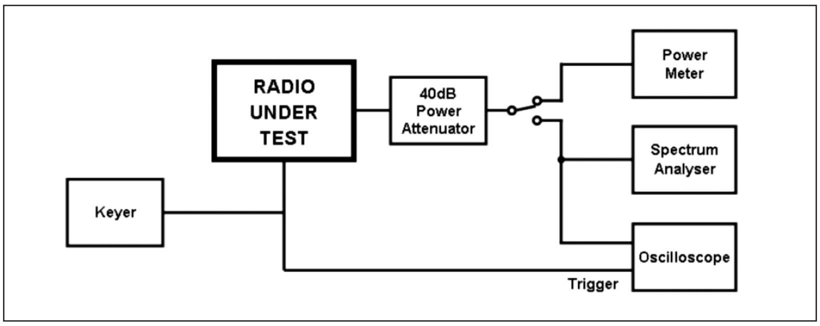


Fig 24.85: CW transmitter measurements

A critical test is to insert an additional 10dB attenuation and make sure that the relative levels do not change, just a change in the noise floor. CW keying is checked at 40 words per minute, which gives a 31ms dot length on the oscilloscope. Rise time, fall time, delays or distortions and any first-character differences are all noted, checking at both full break-in and semi-break-in.

The test arrangement for carrying out SSB transmitter measurements is shown in Fig 24.86. First, the transmitter is driven to full rated power output using a single audio tone. The amplitude of the waveform displayed on the oscilloscope is noted. Then the transmitter is driven by two equal-level audio tones (700Hz and 1700Hz) to the same peak amplitude level on the oscilloscope. The PEP level is then the same as the power output on a single tone. If an accurate and reliable PEP meter is available, the oscilloscope transfer method is not needed and the power level can be read directly from the meter scale. (*NOTE:* audio tones can be generated from software such as described in the software test equipment section).

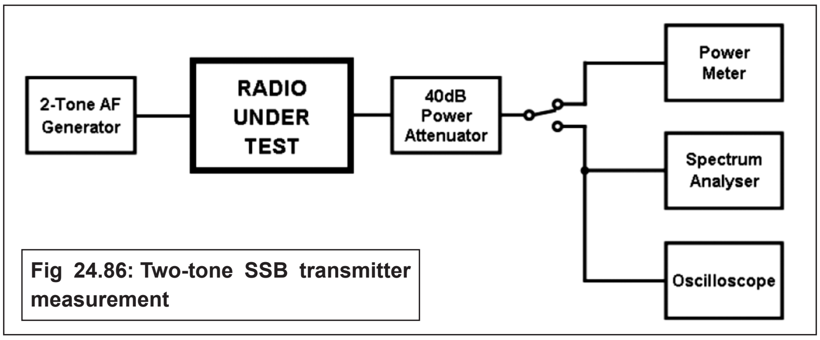


Fig 24.86: Two-tone SSB transmitter measurement

The level of the 3rd and 5th order intermodulation products is measured using a spectrum analyser. It is common to quote levels with respect to PEP as this is universally adopted for all amateur radio products, reviews and specifications. With the transmitter driven by a single audio tone the levels of carrier and unwanted sideband are measured using the spectrum analyser and an estimate made of audio distortion.

Linear Amplifiers

Fig 24.87 shows the specific test arrangement used to measure the two-tone distortion performance of SSB linear amplifiers. Two 100W HF transceivers are coupled together using a power hybrid coupler. This is a homebrew unit using ferrite-cored transformers, such designs are described in many books on RF design. This couples two 50Ω sources to a 25Ω load with half the power dissipated in a 100Ω resistor. This resistor needs to be air- blown hard by a suitable fan. The 25Ω load comprises the linear amplifier under test shunted by an additional 50Ω load.

Hence only one quarter of the original power is available to drive the linear, but this is sufficient, giving 40 - 50W mean power or 80 - 100W PEP. The residual level of intermodulation products is around -50dB to -60dB, so that the distortion level of

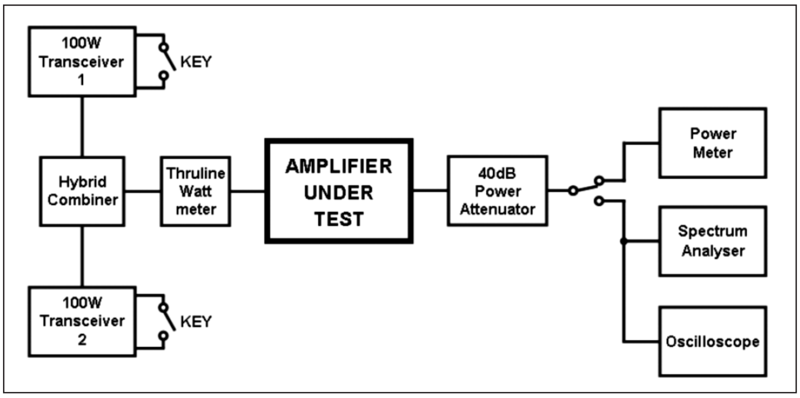
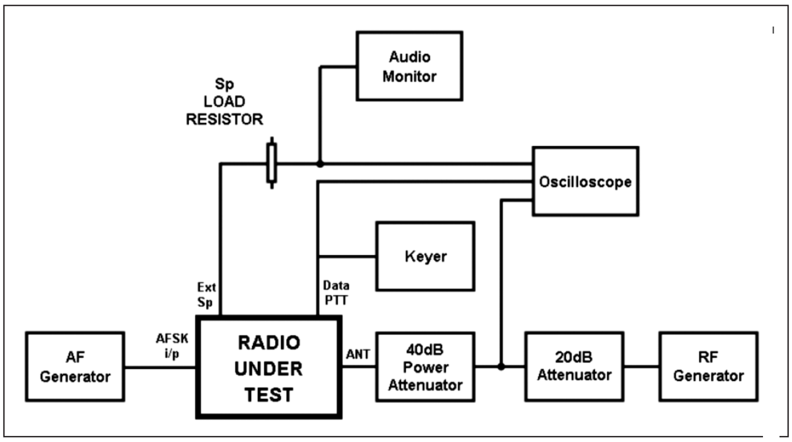


Fig 24.87: Two-tone linear amplifier transmitter measurements

Fig 24.88: T/R switching speed measurement



the drive source is significantly better than that of the amplifier to be measured. This test set-up must be used with great care. The transceivers are used on CW and both must be switched to transmit before either key is pressed. Also both keys must be released before either transceiver is switched back to receive. This is most important otherwise leakage through the hybrid coupler from one transmitter could damage the receiver in the other transceiver. The measurement of power output and distortion follows the same practice as for SSB transmitter measurements.

Transmit-Receive Switching Speed

A reasonably fast and a clean switchover between receive and transmit and back to receive is needed on some data modes.

The test arrangement is shown in **Fig 24.88**. An audio generator at 1kHz drives the transmitter in AFSK mode via the AFSK or microphone input. The receive signal generator is set to give a reasonable on-tune signal level and coupled into the antenna connector via 60dB of attenuation. This ensures that the amount of transmit signal entering the signal generator is insufficient to cause any damage. The data PTT line on the radio is keyed from a pulse generator and the resulting RF output, receiver audio and keying waveform observed on a multichannel oscilloscope. The times for the receiver to be muted, the transmitter to reach full power, the transmitter to be muted and the receiver to regain full sensitivity are all read from the oscilloscope display. Any anomalous behaviour can also be observed. Receiver and transmitter mute times are usually very fast. Transmit and receive enable times less than 20ms are quite fast and fully acceptable for all modes.

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Clive has written various articles for both Radcom *and* Practical Wireless. *Now partly retired, but still active in electronics, he is the Lead Instructor for the various courses at his local radio club.*