5
Probes and input circuits

5.1 Introduction

An electronic instrument needs to be connected to whatever circuit or system is under test. Very wide bandwidth instruments, those used for signals above 100 MHz, usually have 50 Ω co-axial inputs. Instruments which accept signals of even higher frequency, in the microwave and optical range, will have waveguide or optical fibre inputs. The more conventional laboratory test equipment is usually supplied with some kind of probe input circuit which may be connected to the circuit under test.

The probes and input circuits used with oscilloscopes provide good examples of the techniques employed. These same probes and input circuits may, of course, be used in a variety of instruments: vector voltmeters, network analysers, spectrum analysers, and so on. Very high impedance probes are used for voltage measurement, while very low impedance sensor probes must be inserted for current measurement. These voltage and current probes may be passive or active, and both kinds will be considered in this chapter.

Voltage and current measurements are not the only ones that are called for in electronics. Measurements of incident, transmitted and reflected power may also be required. This is the approach often used for high frequency, wide-band circuits which work as part of a transmission line system; for example, a repeater amplifier in a cable television system. The input circuits needed for these power measurements are particularly interesting in that they may exhibit directional properties. Such directional circuits are considered at the end of this chapter.

To begin with, however, it is useful to look at the simplest of input circuits: the kind of circuit which lies behind the input socket of any instrument. It is at this point that the circuit designer must consider what

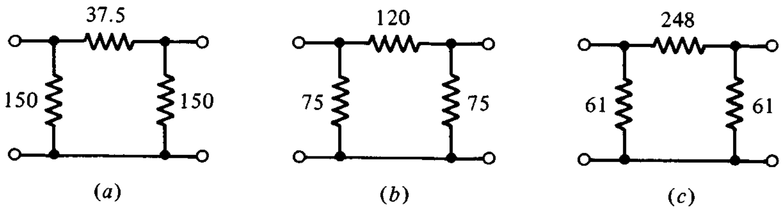

Fig. 5.1. *Showing the* π *networks that may be used to give attenuations of* (*a*) $\times 2$, (*b*) $\times 5$ *and* (*c*) $\times 10$, *in a* 50 Ω *transmission line system.*

the user of an instrument may, unexpectedly, connect to this input. Some protection should always be provided. The circuit which lies behind the input socket also determines what kind of voltage or current probe may be used.

5.2 Input circuits for 50 Ω systems

Fig. 5.1 shows three well-known 50 Ω π networks which may be used to give the 2, 5, 10 sequence of attenuation steps that is so often found.

Such simple π networks can be made, using discrete resistors, which will prove quite satisfactory 50 Ω input circuits for oscilloscopes with bandwidths up to 250 MHz. The reason is that unwanted capacitance in the networks, due to the finite size of the resistors and the connections that must be made between them, will present an impedance much higher than 50 Ω. A capacitance of 10 pF, which is really big for a stray capacitance, has an impedance of 50 Ω only when the frequency is just over 300 MHz.

The three networks shown in Fig. 5.1 are all that are needed to build up a switched 50 Ω input attenuator for any instrument. For example, an oscilloscope with a sensitivity switched between 10 mV/div and 1 V/div in the 2, 5, 10 sequence. Fig. 5.2 shows how this may be done.

At frequencies above 250 MHz, the same circuits may be used, but with thin film resistors and with quite advanced switching technology to overcome the problems of stray capacitance which were mentioned above.

At really high frequencies, the instrument usually has a fixed sensitivity at its input socket. A reduction in input sensitivity is then best achieved by means of fixed attenuator pads. These can be built, as distributed circuits, to have a well-defined, constant attenuation over a bandwidth of several gigahertz, even when the attenuation is as high as 20 db per pad.

5.3 High impedance input circuits

When a wide-band instrument, like an oscilloscope, is designed to have a high input impedance, the problem of unwanted capacitance is much

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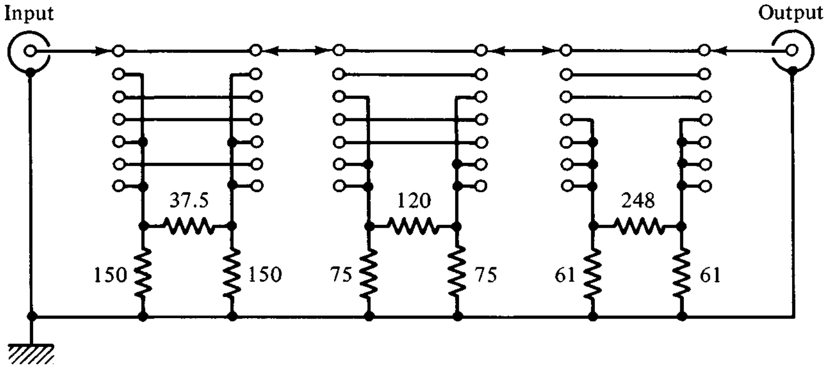

Fig. 5.2. The three networks shown in Fig. 5.1 *may be switched into combinations which give overall attenuations of* 0, × 2, × 5, × 10, × 50 *and* × 100.

more serious than in the case of 50 Ω input impedance instruments. It is virtually impossible to reduce the shunt capacitance across the input socket of the instrument below 20 pF. This capacitance is simply due to the input socket itself, the connection between this socket and the first stage of wide-band amplification, and then the input capacitance of the wide-band amplifier itself.

This means that, if the input impedance at low frequency is defined at 1 MΩ, which is a very common choice, the input impedance will begin to fall as $1/\omega$ once the input signal frequency exceeds $1/2\pi CR$, where $C = 20$ pF and $R = 1$ MΩ. This is a frequency of just over 7.5 kHz: not even beyond the audio range. It follows that a 1 MΩ, 20 pF, input point will present an impedance of only 800 Ω at 10 MHz, which is certainly not a very high frequency. In addition the oscilloscope must be connected to the circuit or system under test by means of a short length of cable, which, even if it is a really low capacitance cable, will have a capacitance of at least 50 pF/m. It is clear that something must be done to increase the impedance at the point where the oscilloscope is actually connected to the circuit under test.

5.4 Passive voltage probes

The simplest solution to the problem of low input impedance is the well-known 'high impedance probe' [1]. This is a passive device which increases the input impedance, at the actual point where the instrument is connected to the circuit under test, by the same factor with which it reduces the overall sensitivity. The high impedance probe is connected to the instrument by a convenient length of low capacitance cable.

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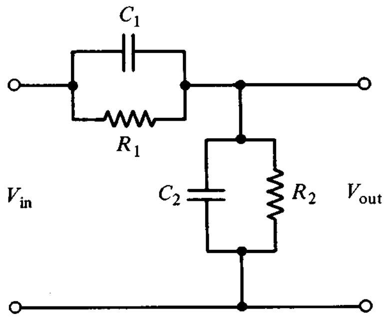

Fig. 5.3. *High impedance input circuits are usually based upon this all-pass network in which $C_1R_1 = C_2R_2$.*

For oscilloscopes, and other instruments, which can handle frequencies up to 10 MHz, the cable which connects the high impedance probe to the instrument may be treated as a lumped capacitance. This assumption allows the action of the high impedance probe to be understood by considering the all-pass network shown in Fig. 5.3. The transfer function of this is

$$V_{out}/V_{in} = R_2/[R_2 + R_1(1 + j\omega C_2 R_2)/(1 + j\omega C_1 R_1)] \qquad (5.1)$$

which shows that the attenuation may be made equal to $R_2/(R_2 + R_1)$ at all frequencies, provided the two time constants, C_1R_1 and C_2R_2, are made equal. The idea is applied by making C_2R_2 the input impedance of the oscilloscope itself, which means that $R_2 = 1\ \text{M}\Omega$ and $C_2 = 20$ pF, plus the capacitance of the cable which connects the probe to the oscilloscope: this can be 50 pF, making C_2 total at 70 pF. The components C_1R_1 are then connected to the input end of the cable to form the probe tip which is actually connected to the circuit under test. A $\times 10$ probe, for example, would call for $R_1 = 9\ \text{M}\Omega$ and $C_1 = 7.77$ pF, assuming the value of 70 pF for C_2 which was discussed above. Some way of adjusting either C_1 or C_2 would have to be arranged if the probe were to be used on different oscilloscopes.

This very simple model for the high impedance probe is quite inadequate at frequencies well above 10 MHz when the cable connecting the probe to the oscilloscope becomes an appreciable fraction of a wavelength. The design of high impedance probes for really wide bandwidth oscilloscopes has been dealt with in an important paper by McGovern [2], who deals with the fact that most commercially available high impedance probes use a special cable that has a high resistance inner conductor. McGovern gives a number of useful references.

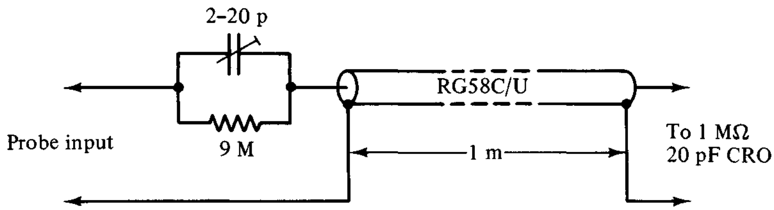

Fig. 5.4. The circuit for an experimental × 10, 10 *MΩ, passive probe.*

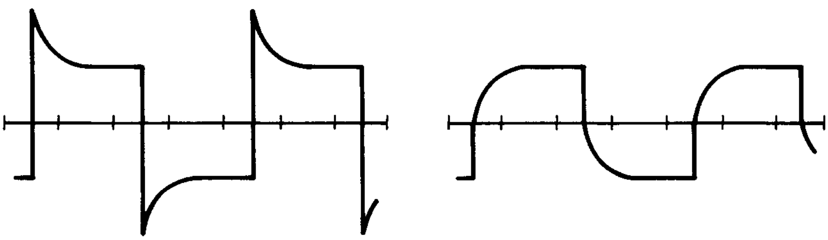

Fig. 5.5. When the probe shown in Fig. 5.4 *is connected to a low frequency square wave, from a low impedance source, the waveform* (*a*) *is observed when the* 2–20 *pF trimmer is adjusted to too large a value, and the waveform* (*b*) *is observed when the trimmer is adjusted to too small a value. When adjusted correctly, the probe reproduces the square wave perfectly, attenuating all frequency components by a factor of* 10.

5.5 An experimental high impedance probe

Even at frequencies below 10 MHz, the problem of designing a high impedance probe is not quite so simple as equation (5.1) might suggest, and it is worthwhile doing an experiment to show this. Fig. 5.4 shows what is needed. Take a 1 m length of ordinary 50 Ω co-axial cable with BNC connectors at either end. Because this cable will have a capacitance of, say, 100 pF, the input end will look like 120 pF in parallel with 1 MΩ when the cable is connected to a 1 MΩ, 20 pF, oscilloscope input.

Take a BNC panel socket, bolt a solder tag on one of the four holes and solder on a stiff 2 cm length of copper wire to act as a ground probe. Solder the 9 MΩ and the 2–20 pF trimmer capacitor directly on to the centre conductor of the socket, but note that the side of the trimmer which makes connection to the adjusting screw should be on the input side, not the socket side.

When this probe circuit is connected to a square wave, from a low impedance source and at a frequency of about 1 kHz, the waveforms shown in Figs. 5.5(*a*) and (*b*) will be observed, depending upon whether the trimmer capacitor is adjusted above or below the critical value which

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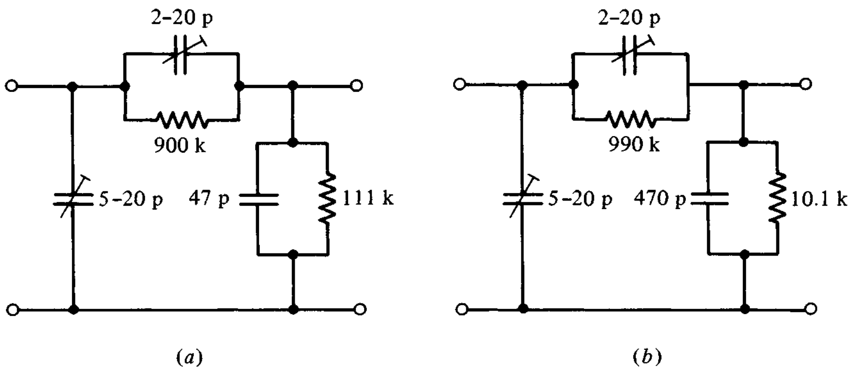

Fig. 5.6. The input attenuators typically found inside an analog oscilloscope with a 1 MΩ input impedance. Network (a) gives ×10 attenuation and network (b) gives ×100. The ×2 and ×5 factors, which are also needed, would be provided by switching the feedback networks associated with the input amplifier.

makes $C_1R_1 = C_2R_2$. The time constant involved, shown by the exponential rise or fall in Fig. 5.5, is around 120 μs because this is the order of both C_1R_1 and C_2R_2.

However, this simple test circuit will never be made to perform like a high impedance probe of the kind that is supplied with a commercial analog oscilloscope. There will always be a small error in the high frequency attenuation when the probe circuit is adjusted to give a good low frequency square wave response. The reason for this lies in the kind of input attenuator circuits found inside a typical analog oscilloscope with a 1 MΩ input impedance. These are shown in Fig. 5.6.

5.6 The problem of input capacitance

Fig. 5.6 shows that the same all-pass network idea, first shown in Fig. 5.3, is again being used in the input attenuator of the analog oscilloscope with a 1 MΩ input impedance. The 2–20 pF trimmer capacitor, shown in both Figs. 5.6(*a*) and (*b*), is adjusted to give a perfect low frequency response when the instrument is connected to a low impedance source.

Both Figs. 5.6(*a*) and (*b*) show a 5–20 pF trimmer capacitor connected across the input. The value of this trimmer is quite irrelevant when the input is connected to a low impedance source, and the reason it is there is that it can be adjusted to make the input capacitance of the instrument itself a constant, regardless of the setting of the internal attenuator. Unfortunately, this means that it is very difficult to get a high impedance probe of the kind shown in Fig. 5.4, one in which C_1 is a variable, to work

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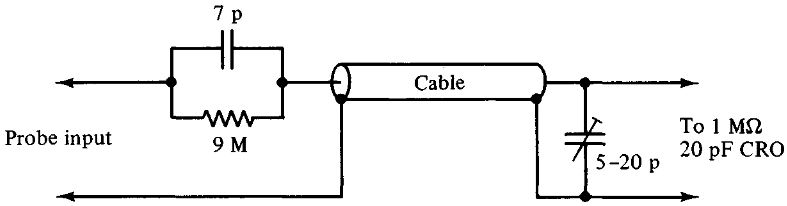

Fig. 5.7. When the adjustable capacitor in a high impedance probe is connected across the oscilloscope end of the probe cable, it is much easier to get really good high frequency performance.

correctly at high frequency. The setting of the 5–20 pF trimmer shown in both Figs. 5.6(*a*) and (*b*) may not be correct for both conditions: constant instrument input capacitance *and* all-pass properties for the combination of probe and input attenuator.

This is why nearly all high impedance probes adopt a circuit like the one shown in Fig. 5.7, in which the input components, 7 pF and 9 MΩ in this particular case, are fixed, and the probe adjustment is made by means of a trimmer capacitor at the oscilloscope end of the probe cable. It is then possible to get the correct $\times 10$ probe attenuation at both low and high frequency. There is a further advantage in that the trimmer capacitor shown in Fig. 5.7 has one end grounded and is also located in a much more convenient place.

5.7 Errors caused by passive voltage probes

Errors of adjustment, discussed in the previous section, are not the only errors to be cautious of when using high impedance voltage probes on analog oscilloscopes. Whenever the probe is connected to a finite source impedance, there may be very serious errors introduced.

For example, suppose a $\times 10$ probe has an input impedance of 10 MΩ in parallel with about 7 pF, as would be the case for the arrangement shown in Fig. 5.7. Just as in the case discussed in section 5.3, this 10 MΩ input impedance begins to fall as $1/\omega$ when the signal frequency is only in the kilohertz region. When the frequency is 10 MHz, this 'high impedance probe' looks like about 2.5 kΩ. This may cause considerable errors in amplitude, phase and rise time measurements [3]. It is fairly safe to say that measurements made with oscilloscopes using 10 MΩ, $\times 10$, probes are accurate only at frequencies below 10 MHz, and then only when the source impedance being probed is below 100 Ω. As 10 MHz is by no means a high frequency in today's electronics, this problem is one which must be borne in mind.

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Very wide bandwidth oscilloscopes, both analog and digitising, usually have only a 50 Ω input impedance, although a 1 MΩ input facility is often provided so that these very wide bandwidth instruments may also be used to make simple low frequency measurements. When a 50 Ω input impedance is unsuitable for some measurement which needs to be made at the really high frequencies, then probing is best done with an *active* voltage probe. These are described in the next two sections.

5.8 Classical active voltage probes

An active voltage probe is one, as the name implies, which uses active devices to provide a high impedance at the probe tip. There are two distinct kinds of active probe. The classical variety uses a microcircuit actually inside the probe tip, which is then connected, via a cable which must carry both signal and power supply lines, to some kind of connector box which presents the signal to the oscilloscope, and also picks up the power supply needed. A more modern kind of active probe is one which puts all the active devices into the connector box so that the probe tip can be made extremely small. Both kinds of active probe use interesting circuit shapes.

Fig. 5.8 shows an example of the kind of circuit shape found in the classical variety of active voltage probe [4]. This active probe is intended to provide an overall voltage gain of unity, a bandwidth of several hundred megahertz, and to be used with a very wide bandwidth oscilloscope having an input impedance of 50 Ω.

The probe tip, shown in Fig. 5.8, contains an amplifier which must provide the power gain needed for a voltage gain of unity and an impedance transformation from R_1, the high input impedance of the probe, down to the 50 Ω input impedance of the oscilloscope. This amplification is done with Q_1, a junction FET connected as a source follower, followed by a microcircuit with a voltage gain of 2, which can be matched, by means of R_4, to the 50 Ω cable that carries the signal on to the oscilloscope.

Note that this probe tip amplifier only handles the a.c. part of the input signal. The time constant $C_1 R_2$ decides the low frequency cut-off point for the probe tip, and would be chosen to lie in the kilohertz region. With $R_2 = 10$ MΩ, C_1 need then be only 100 pF, and this would mean that all the components associated with the gate of Q_1 would be of very small size and the probe tip input capacitance could be made very small indeed: only 2–3 pF.

The d.c. part of the input signal is dealt with at the far end of the probe

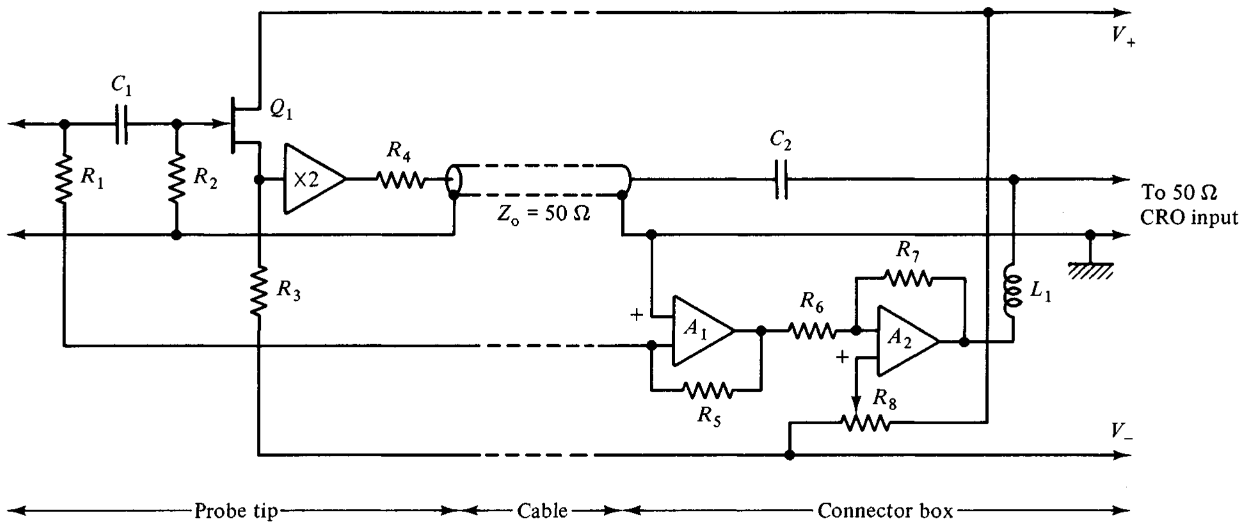

Fig. 5.8. An active voltage probe which has a microcircuit in the probe tip and an overall gain of unity.

cable, inside the connector box. This allows the probe tip to be made smaller because there are less components inside it. The d.c. signal path is simply an amplifier, A_1 with feedback components, $R_1 = R_5$, to give unity gain. The real part of the probe input impedance is R_1, and this is made 100 kΩ.

Because A_1 is an inverting amplifier, a second inversion is needed before the d.c. signal can be passed to the oscilloscope input. This is done by means of A_2, with feedback components $R_6 = R_7$. This second amplifier can also be used to introduce a d.c. offset facility, and this is shown as R_8.

An interesting general point should be made about the relative frequency responses of the a.c. and d.c. paths. The a.c. path, through Q_1, the $\times 2$ amplifier, and then the cable, *via* C_2, into the oscilloscope, will have its gain rising at 12 db/octave if $C_1 R_2$ and $C_2 \times 50\ \Omega$ are made equal to one another. As stated above, the low frequency cut-off point would be made in the kilohertz region. The d.c. signal path, R_1 through A_1 and then A_2, only needs to have unity gain up to a frequency marginally above the low frequency cut-off point of the a.c. path. Because the a.c. and d.c. signal paths are in *parallel*, it does not then matter how the gain of this d.c. path falls as the frequency increases.

Finally, a very small inductor, L_1, is added to ensure that the output impedance of A_2 does not load the high frequency signal circuit in the megahertz region.

5.9 Recent developments in active probes

The most recent digitising oscilloscopes have effective bandwidths out in the gigahertz region, and, of course, have 50 Ω input impedance. To make the equivalent of a high impedance probe for instruments with this kind of very wide bandwidth calls for some radically new circuit ideas and constructional techniques. The aim is to get the capacitance at the probe tip down to as small a value as possible.

Fig. 5.9 shows the kind of circuit shape that may be found in these very wide bandwidth active probes [5]. The probe tip contains no active components, and can be made extremely small for this reason. All it does contain is the parallel combination of 10 kΩ and 0.5 pF, which gives a low frequency attenuation of 46 db into the 50 Ω cable, when this is terminated with 50 Ω at the far end. At a frequency close to 30 MHz, which is $1/(2\pi \times 0.5\ \text{pF} \times 10\ \text{k}\Omega)$, the attenuation introduced by the probe tip network begins to fall at 6 db/octave.

It follows that the connector box needs to include an amplifier that, ideally, has a 50 Ω input impedance and a gain of 46 db at low frequency

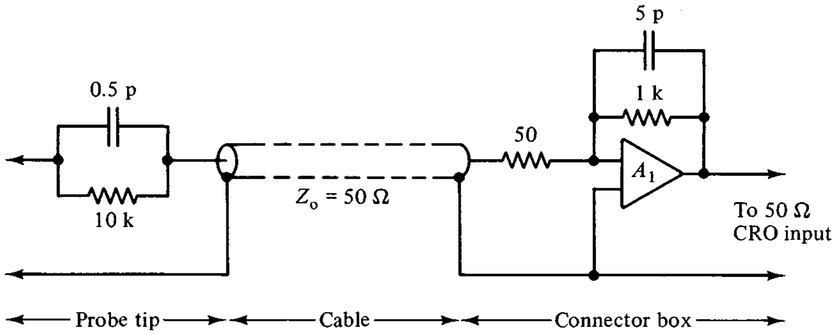

Fig. 5.9. An active voltage probe for very wide bandwidth.

which then falls at 6 db/octave beginning at the same critical frequency close to 30 MHz. Now this would call for the feedback components across A_1, in Fig. 5.9, to be 10 kΩ and 0.5 pF, identical to the values in the probe tip. Such an arrangement would give an overall gain of unity up to, and above, 30 MHz. However, a gain of 46 db over a bandwidth of 30 MHz is not an easy specification to meet, and when it is coupled with the requirement that the gain falls off at 6 db/octave from 30 MHz to around 1 GHz, it is virtually impossible, even with today's most advanced gallium arsenide microcircuit technology. For this reason the specification is relaxed to make the active probe shown in Fig. 5.9 have an overall attenuation of 20 db. It is, in fact, a '$\times 10$' active probe. Even with this relaxed specification, the amplifier, A_1 in Fig. 5.9, is a real challenge to the circuit designer. The design described by Rush, Escovitz and Berger [5] used four discrete microwave transistors in a thick film hybrid circuit [6].

5.10 Passive current probes

While the measurement of voltages, in a circuit under test, is most conveniently done with some kind of high impedance probe, the measurement of current calls for a very low impedance to be *inserted* in series with the current carrying conductor. For fairly high frequency work, the simplest way of doing this is to use a clip-on current transformer probe. These first became available around 1960 [7] and consist of a simple wound U-shaped core which has a sliding I-shaped core over the top. In this way the core may be opened, the lead carrying the current can be inserted, and the core then closed.

The design of such a passive current probe is best considered without the complication of needing to open and close the core. In fact, this kind of probe is of little use when currents in a printed circuit board are being

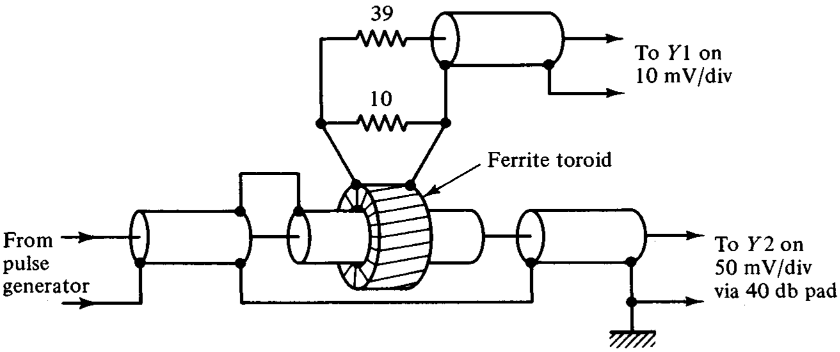

Fig. 5.10. *An experimental wide bandwidth current transformer. Full details are given in the text. Note that the* Y1 *and* Y2 *input impedances must be made* 50 Ω *and that the pulse generator must have* 50 Ω *output impedance.*

studied. It follows that some circuits may need to have wide bandwidth current transformers built into them during development. A good example would be the measurement of the power dissipation in a transistor as a function of time. It would be necessary to measure the collector to emitter voltage with a high impedance, active or passive, voltage probe, and also to measure the emitter current by including a wide bandwidth current transformer around the emitter lead. A modern digitising oscilloscope may have sufficient computing power to display instantaneous power as a function of time by taking the product of these two measurements. Otherwise, an analog multiplier could be used, or the two waveforms simply displayed as *Y*1 and *Y*2, in the traditional way, and the experimentalist would note the time at which, for example, the power dissipation was at a maximum.

A valuable paper by Ritson and Wood [8] deals with the most straightforward kind of wide bandwidth current transformer, and the problem may be best considered by referring to Fig. 5.10. A ferrite toroid is wound with a single layer winding of several turns, and the conductor carrying the current to be measured passes through the centre of this toroid. Between the winding and the current carrying conductor, some kind of screen must be arranged, and in the experimental set-up shown in Fig. 5.10, this is done by using a very short length of co-axial cable with the outer conductor grounded at only one end. All the co-axial cables shown in Fig. 5.10 are 50 Ω characteristic impedance.

The current transformation ratio of this kind of transformer is equal to the number of secondary turns, N. The most convenient size of ferrite toroid for the experiment shown in Fig. 5.10 is 9 mm [9]. This can carry a uniform single layer winding of 50 turns, using 200 μm polyurethane coated wire, and then be a sliding fit on the outside of standard RG58C/U co-axial cable. Choosing a low value of secondary load resistor, R_L, not only reduces the insertion impedance of the transformer, which ideally is R_L/N^2, but also improves the bandwidth. Ritson and Wood [8] show that the low frequency cut-off frequency is approximately $R_L/2\pi L_s$, where L_s is the secondary inductance, and that the high frequency cut-off occurs at $1/2\pi C_s R_L$, where C_s is the stray capacitance across the secondary winding.

By choosing the 10 Ω and 39 Ω ($\approx$ 40 Ω) shown in Fig. 5.10, the 50 turn winding supplies only 1/500th of the primary current to the 50 Ω termination at the input to the *Y*1 amplifier on the oscilloscope. This is particularly convenient because the use of a 40 db attenuator pad, and the sensitivity ranges shown in Fig. 5.10, then allow a direct comparison of the input current to the transformer on *Y*2, and its reproduction of this current as a waveform on *Y*1. The pulse generator should be set for 5–10 V pulse amplitude.

The importance of the screen between primary and secondary may be checked easily with the set-up shown in Fig. 5.10 by temporarily disconnecting the connection to the outer conductor of the short length of co-axial cable that passes through the toroid. If the whole assembly is made in between two BNC panel sockets, it is possible to see the residual problems that remain, even when the screen is properly grounded, by reversing the direction of the assembly relative to the rest of the equipment. This changes the sign of the pulse signal observed on *Y*1, which is due to the magnetic coupling, but it does not change the sign of the spurious signal due to the unwanted capacitive coupling.

Such a passive current probe cannot be made to work up to very high frequencies. This is because of the poor permeability of the ferrite at higher frequencies, and also because of the length of the secondary winding itself. At very high frequencies the transformer must be designed as a distributed circuit. The secondary winding must be made in the form of a transmission line and terminated with the correct impedance. When this is done, it is possible to make wide bandwidth current transformers, for insertion into 50 Ω transmission line systems, which operate satisfactorily up to 1 GHz [10, 11].

5.11 Active current probes

One severe disadvantage of the simple current transformer, described in the previous section, is its inability to handle direct current and very low frequencies. This has been a problem in power engineering for a considerable time, and is solved by some kind of current comparator technique [12] in which the magnetic field in a toroidal core is nulled by a sensor and feedback arrangement. This causes the secondary ampere-turns to be exactly equal and opposite to the primary ampere-turns. Some active current probes of this kind can operate up to the high audio frequencies [13].

For really wide bandwidths, it is possible to combine a Hall sensor and the kind of wide bandwidth current transformer described in the previous section, all in one unit [14], and make the Hall sensor look after the d.c. and low frequency end of the spectrum. Such active current probes need very careful setting up and are sensitive to stray magnetic fields. Models are available which cover d.c. to 50 MHz [15].

5.12 Input circuits for power measurements

Voltage and current measurements are not at all easy at very high frequencies. For noise figure and gain measurements, the measurement of power, instead of voltage or current, may be a much more accurate approach.

This chapter closes with an example of one of the input circuits that can be of great use in power measurement: the directional coupler or directional bridge. This is a device that can be used to measure the incident or the reflected power at any input or output port of a network, device or system. Such a device also makes the measurement of input or output impedance, at high frequencies, particularly simple, and an example will be given in chapter 10. Incident and reflected power measurements are also needed when the magnitude of a network's S-parameters [16] are to be measured.

The function of the directional coupler, or directional bridge, may be seen at a glance from Fig. 5.11. The device has three ports. One is connected to a source of high frequency power, P_{in}, such as a signal generator or a swept frequency source. This source, of course, has an output impedance, usually 50 Ω, that matches the co-axial transmission line being used to interconnect the system under test.

The device under test, for example, a wide-band amplifier, is then connected to the port labelled 'load' in Fig. 5.11. A sensitive detector, with a 50 Ω input impedance, is connected to the third port. As shown in Fig. 5.11, any power which is reflected from the load, due to it not being a perfect match, is coupled back to the detector port where it can be measured.

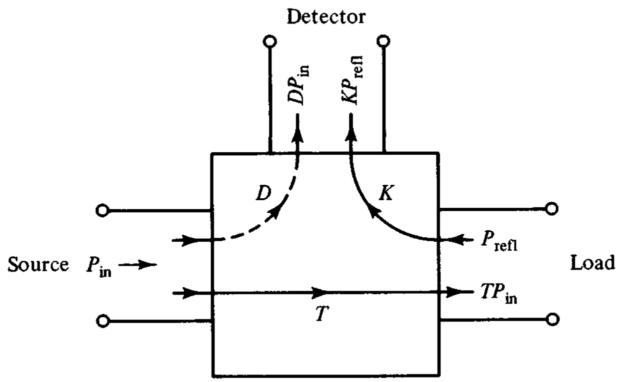

Fig. 5.11. *The directional coupler. The transmission loss, T, coupling factor, K, and the directivity, D, are discussed in the text.*

The directional coupler is a passive device so that T, K and D, shown in Fig. 5.11, must all be attenuations and these are usually expressed in decibels. Ideally, the transmission loss will be negligible, the directivity as high as possible, and the coupling factor only a few decibels. However, to get really high directivity, it is usually necessary to make the coupling factor, K, about 20 db. This is a nuisance when the directional coupler is being used to measure reflected power (the mode illustrated in Fig. 5.11) but, of course, 20 db is a sensible coupling factor when the same directional coupler is turned around and used to measure incident power.

Ideal directional couplers can only be made over a fairly limited bandwidth, and then only at quite high frequencies, because all must involve coupled waveguides or transmission lines at least one wavelength long. For really wide bandwidth and useful properties down to frequencies of only a few megahertz, directional couplers must make use of the kind of transformer which has been discussed above in section 5.9 and illustrated in Fig. 5.10. Spaulding [17] has published a very interesting design for such a directional coupler which will work over the range 1–1000 MHz, and the circuit shape of this coupler is shown in Fig. 5.12.

When the load port in Fig. 5.12 is terminated with 50 Ω, it is clear that no signal will appear at the detector port when power flows from source to load. This follows because power flow in this direction implies current flow from left to right during the positive half cycle of source and load voltage. The two transformers then produce equal and opposite voltages across the detector port which cancel one another out.

Power flow from load to source, however, implies current flow from right to left in Fig. 5.12 during the same positive half cycle of voltage. The signals from the two transformers now add across the detector port, and

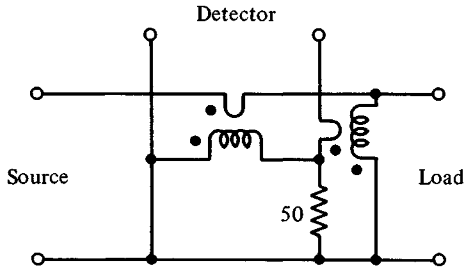

Fig. 5.12. *A wide bandwidth directional coupler using two transformers of the kind shown in Fig.* 5.10. *With turns ratio* 1:10, *the coupling factor, K, is* 20 *db.*

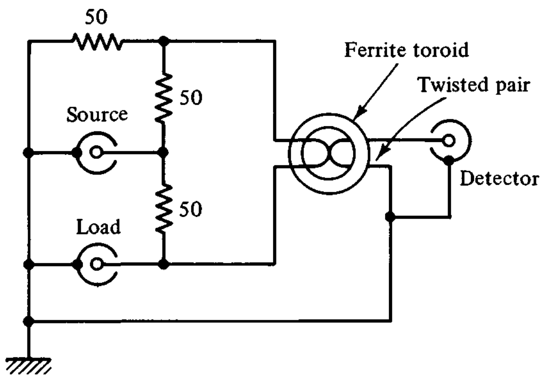

Fig. 5.13. *An experimental directional bridge for work on* 50 Ω *systems. For clarity, only one turn of the twisted pair around the ferrite toroid is shown.*

a 20 db fraction of this reverse power is available at the detector port when this is terminated correctly with 50 Ω.

The kind of wide bandwidth directional coupler which is shown in Fig. 5.12 can be made to have very small transmission loss and excellent directivity, but the construction of the transformers calls for very advanced techniques [17]. For many applications, a simpler device known as a 'directional bridge' [18] may be useful, and an example is shown in Fig. 5.13. This is the final experimental exercise for this chapter.

In Fig. 5.13 a classical bridge network is formed from three 50 Ω resistors and the load. Because one side of the load is grounded, one node of the bridge network becomes grounded, and the source can only be connected to the node which is opposite this grounded node. This leaves the other two nodes of the bridge free, and it is between these that the detector output can be taken. Because one side of the detector is grounded, the connection between the detector and bridge must be made *via* a balanced to unbalanced transformer of the kind discussed in chapter 3, sections 3.8 and 3.9. In this experimental directional bridge, the balun is again made using a 500 mm length of twisted pair, 0.4 mm polyurethane coated wire twisted to 5 mm pitch, wound around a 25 mm ferrite toroid [19].

Despite its name, the circuit shown in Fig. 5.13 does not have directional properties in the same way that the directional coupler shown in Fig. 5.12 has: source and load cannot be interchanged in Fig. 5.13 to obtain an incident power measurement. Source and detector ports in Fig. 5.13 may be interchanged, however. There are other ways of making these so-called directional bridges [20] so that they do have real directional properties, but these designs again call for wide bandwidth transformers with large turns ratios.

The experimental directional bridge shown in Fig. 5.13 should have a transmission loss of 6 db and a coupling factor of 6 db. The directivity of the bridge should be 60 db at 5 MHz, and fairly constant over the band 1–10 MHz. It is a very useful device for checking impedance in circuits and devices which are intended for use in 50 Ω coaxial systems.

Notes

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19 Mullard FX3312 Ferroxcube toroids are a good choice. So are the Siemens R25 toroids in type N30 Siferrit.
20 See the *Radio Amateur's Handbook*, ARRL, Newington, Conn., USA, 1981, pp. 16.11 and 16.31.