1.1 Grounding

A fundamental property of any electronic or electrical circuit is that the voltages present within it are referenced to a common point, conventionally called the ground. (This term is derived from electrical engineering practice, when the reference point is often taken to a copper spike literally driven into the ground.) This point may also be a connection point for the power to the circuit, and it is then called the 0V (nought-volt) rail, and ground and 0V are frequently (and confusingly) synonymous. Then, when we talk about a five-volt supply or a minus-twelve-volt supply or a two-and-a-half-volt reference, each of these is referred to the 0V rail.

At the same time, ground is not the same as 0V. A ground wire connects equipment to earth for safety reasons, and does not carry a current in normal operation. However, in this chapter the word “grounding” will be used in its usual sense, to include both safety earths and signal and power return paths.

Perhaps the greatest single cause of problems in electronic circuits is that 0V and ground are taken for granted. The fact is that in a working circuit there can only ever be one point which is truly at 0V; the concept of a “0V rail” is in fact a contradiction in terms. This is because any practical conductor has a finite non-zero resistance and inductance, and Ohm’s Law tells us that a current flowing through anything other than a zero impedance will develop a voltage across it. A working circuit will have current flowing through those conductors that are designated as the 0V rail and therefore, if any one point of the rail is actually at 0V (say, the power supply connection) the rest of the rail will not be at 0V. This is illustrated in Figure 1.1.

Assume the 0V conductor has a resistance of 10mΩ/inch and that points A, B, C and D are each one inch apart. The voltages at points A, B and C referred to D are

\[ V_C = (I_1 + I_2 + I_3) \cdot 10 \text{m}\Omega \quad = 400 \mu\text{V} \]
\[ V_B = V_C + (I_1 + I_2) \cdot 10 \text{m}\Omega \quad = 700 \mu\text{V} \]
\[ V_A = V_B + I_3 \cdot 10 \text{m}\Omega \quad = 900 \mu\text{V} \]

Figure 1.1 Voltages along the 0V rail
Now, after such a trenchant introduction, you might be tempted to say well, there are millions of electronic circuits in existence, they must all have 0V rails, they seem to work well enough, so what’s the problem? Most of the time there is no problem. The impedance of the 0V conductor is in the region of milliohms, the current levels are milliamps, and the resulting few hundred microvolts drop doesn’t offend the circuit at all. 0V plus $500\mu\text{V}$ is close enough to 0V for nobody to worry.

The difficulty with this answer is that it is then easy to forget about the 0V rail and assume that it is 0V under all conditions, and subsequently be surprised when a circuit oscillates or otherwise doesn’t work. Those conditions where trouble is likely to arise are

- where current flows are measured in amps rather than milli- or microamps
- where the 0V conductor impedance is measured in ohms rather than milliohms
- where the resultant voltage drop, whatever its value, is of a magnitude or in such a configuration as to affect the circuit operation.

*When to consider grounding*

One of the attributes of a good circuit designer is to know when these conditions need to be carefully considered and when they may be safely ignored. A frequent complication is that you as circuit designer may not be responsible for the circuit’s layout, which is handed over to a layout draughtsman (who may in turn delegate many routing decisions to the software package). Grounding is always sensitive to layout, whether of discrete wiring or of printed circuits, and the designer must have some knowledge of and control over this if the design is not to be compromised.

The trick is always to be sure that you know where ground return currents are flowing, and what their consequences will be; or, if this is too complicated, to make sure that wherever they flow, the consequences will be minimal. Although the above comments are aimed at 0V and ground connections, because they are the ones most taken for granted, the nature of the problem is universal and applies to any conductor through which current flows. The power supply rail (or rails) is another special case where conductor impedance can create difficulties.

**1.1.1 Grounding within one unit**

In this context, “unit” can refer to a single circuit board or a group of boards and other wiring connected together within an enclosure such that you can identify a “local” ground point, for instance the point of entry of the mains earth. An example might be as shown in Figure 1.2. Let us say that printed circuit board (PCB) 1 contains input signal conditioning circuitry, PCB2 contains a microprocessor for signal processing and PCB3 contains high-current output drivers, such as for relays and for lamps. You may not place all these functions on separate boards, but the principles are easier to outline and understand if they are considered separately. The power supply unit (PSU) provides a low-voltage supply for the first two boards, and a higher-power supply for the output board. This is a fairly common system layout and Figure 1.2 will serve as a starting point to illustrate good and bad practice.

**1.1.2 Chassis ground**

First of all, note that connections are only made to the metal chassis or enclosure at one point. All wires that need to come to the chassis are brought to this point, which should
be a metal stud dedicated to the purpose. Such connections are the mains safety earth (about which more later), the 0V power rail, and any possible screening and filtering connections that may be required in the power supply itself, such as an electrostatic screen in the transformer. (The topic of power supply design is itself dealt with in much greater detail in Chapter 7).

The purpose of a single-point chassis ground is to prevent circulating currents in the chassis. If multiple ground points are used, even if there is another return path for the current to take, a proportion of it will flow in the chassis (Figure 1.3); the proportion is determined by the ratio of impedances which depends on frequency. Such currents are very hard to predict and may be affected by changes in construction, so that they can give quite unexpected and annoying effects: it is not unknown for hours to be devoted to tracking down an oscillation or interference problem, only to find that it disappears when an inoffensive-looking screw is tightened against the chassis plate. Joints in the chassis are affected by corrosion, so that the unit performance may degrade with time, and they are affected by surface oxidation of the chassis material. If you use multi-point chassis grounding then it is necessary to be much more careful about the electrical construction of the chassis.

† But, when RF shielding and/or a low-inductance ground is required, multiple ground points may be essential. This is covered in Chapter 8.
1.1.3 The conductivity of aluminium

Aluminium is used throughout the electronics industry as a light, strong and highly conductive chassis material – only silver, copper and gold have a higher conductivity. You would expect an aluminium chassis to exhibit a decently low bulk resistance, and so it does, and is very suitable as a conductive ground as a result. Fortunately, another property of aluminium (which is useful in other contexts) is that it oxidises very readily on its surface, to the extent that all real-life samples of aluminium are covered by a thin surface film of aluminium oxide (Al₂O₃). Aluminium oxide is an insulator. In fact, it is such a good insulator that anodised aluminium, on which a thick coating of oxide is deliberately grown by chemical treatment, is used for insulating washers on heatsinks.

The practical consequence of this quality of aluminium oxide is that the contact resistance of two sheets of aluminium joined together is unpredictably high. Actual electrical contact will only be made where the oxide film is breached. Therefore, whenever you want to maintain continuity through a chassis made of separate pieces of aluminium, you must ensure that the plates are tightly bonded together, preferably with welding or by fixings which incorporate shakeproof serrated washers to dig actively into the surface. The same applies to ground connection points. The best connection (since aluminium cannot easily be soldered) is a force-fit or welded stud (Figure 1.4), but if this is not available then a shakeproof serrated washer should be used underneath the nut which is in contact with the aluminium.

Other materials

Another common chassis material is cadmium- or tin-plated steel, which does not suffer from the oxidation problem. Mild steel has about three times the bulk resistance of aluminium so does not make such a good conductor, but it has better magnetic shielding properties and it is cheaper. Die-cast zinc is popular for its light weight and strength, and ease of creating complex shapes through the casting process; zinc’s conductivity is 28% of copper. Other metals, particularly silver-plated copper, can be used where the ultimate in conductivity is needed and cost is secondary, as in RF circuits. The advantage of silver oxide (which forms on the silver-plated surface) is that it is conductive and can be soldered through easily. Table 1.1 shows the conductivities and temperature coefficients of several metals.
1.1.4 Ground loops

Another reason for single-point chassis connection is that circulating chassis currents, when combined with other ground wiring, produce the so-called “ground loop”, which is a fruitful source of low-frequency magnetically-induced interference. A magnetic field can only induce a current to flow within a closed loop circuit. Magnetic fields are common around power transformers – not only the conventional 50Hz mains type (60Hz in the US), but also high-frequency switching transformers and inductors in switched-mode power supplies – and also other electromagnetic devices: contactors, solenoids and fans. Extraneous magnetic fields may also be present. The mechanism of ground-loop induction is shown in Figure 1.5.

Lenz’s law tells us that the e.m.f. induced in the loop is

\[ V = -10^{-8} \cdot A \cdot n \cdot dB/dt \]

where
- \( A \) is the area of the loop in cm\(^2\)
- \( B \) is the flux density normal to it, in microTesla, assuming a uniform field
- \( n \) is the number of turns (\( n = 1 \) for a single-turn loop)

As an example, take a 10\( \mu \)T 50Hz field as might be found near a reasonable-sized mains transformer, contactor or motor, acting at right angles through the plane of a 10cm\(^2\) loop that would be created by running a conductor 1cm above a chassis for 10cm and grounding it at both ends. The induced emf is given by

\[ V = -10^{-8} \cdot 10 \cdot d/dt(10 \cdot \sin 2\pi \cdot 50 \cdot t) \]

### Table 1.1 Conductivity of metals

<table>
<thead>
<tr>
<th>Metal</th>
<th>Relative Conductivity (( Cu = 1, ) at 20ºC)</th>
<th>Temperature coefficient of resistance (ºC at 20ºC)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Aluminium (pure)</td>
<td>0.59</td>
<td>0.0039</td>
</tr>
<tr>
<td>Aluminium alloy:</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Soft-annealed</td>
<td>0.45-0.50</td>
<td></td>
</tr>
<tr>
<td>Heat-treated</td>
<td>0.30-0.45</td>
<td></td>
</tr>
<tr>
<td>Brass</td>
<td>0.28</td>
<td>0.002-0.007</td>
</tr>
<tr>
<td>Cadmium</td>
<td>0.19</td>
<td>0.0038</td>
</tr>
<tr>
<td>Copper:</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Hard drawn</td>
<td>0.895</td>
<td>0.00382</td>
</tr>
<tr>
<td>Annealed</td>
<td>1.0</td>
<td>0.00393</td>
</tr>
<tr>
<td>Gold</td>
<td>0.65</td>
<td>0.0034</td>
</tr>
<tr>
<td>Iron:</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Pure</td>
<td>0.177</td>
<td>0.005</td>
</tr>
<tr>
<td>Cast</td>
<td>0.02-0.12</td>
<td></td>
</tr>
<tr>
<td>Lead</td>
<td>0.7</td>
<td>0.0039</td>
</tr>
<tr>
<td>Nichrome</td>
<td>0.0145</td>
<td>0.0004</td>
</tr>
<tr>
<td>Nickel</td>
<td>0.12-0.16</td>
<td>0.006</td>
</tr>
<tr>
<td>Silver</td>
<td>1.06</td>
<td>0.0038</td>
</tr>
<tr>
<td>Steel</td>
<td>0.03-0.15</td>
<td>0.004-0.005</td>
</tr>
<tr>
<td>Tin</td>
<td>0.13</td>
<td>0.0042</td>
</tr>
<tr>
<td>Tungsten</td>
<td>0.289</td>
<td>0.0045</td>
</tr>
<tr>
<td>Zinc</td>
<td>0.282</td>
<td>0.0037</td>
</tr>
</tbody>
</table>
Magnetic field induction is usually a low-frequency phenomenon (unless you happen to be very close to a high-power radio transmitter) and you can see from this example that in most circumstances the induced voltages are low. But in low-level applications, particularly audio and precision instrumentation, they are far from insignificant. If the input circuit includes a ground loop, the interference voltage is injected directly in series with the wanted signal and cannot then be separated from it. The cures are:

- open the loop by grounding only at one point
- reduce the area of the loop (A in the equation above) by routing the offending wire(s) right next to the ground plane or chassis, or shortening it
- reduce the flux normal to the loop by repositioning or reorienting the loop or the interfering source
- reduce the interfering source, for instance by using a toroidal transformer.

### 1.1.5 Power supply returns

You will note from Figure 1.2 that the output power supply 0V connection (0V(B)) has been shown separately from 0V(A), and linked only at the power supply itself. What happens if, say for reasons of economy in wiring, you don’t follow this practice but
instead link the 0V rails together at PCB3 and PCB2, as shown in Figure 1.6?

The supply return currents $I_{0V}$ from both PSB/PCB3 and PSA/PCB2 now share the same length of wire (or track, in a single-pcb system). This wire has a certain non-zero impedance, say for dc purposes it is $R_S$. In the original circuit this was only carrying $I_{0V(2)}$ and so the voltage developed across it was

$$V_S = R_S \cdot I_{0V(2)}$$

but, in the economy circuit,

$$V_S = R_S \cdot (I_{0V(2)} + I_{0V(3)})$$

This voltage is in series with the supply voltages to both boards and hence effectively subtracts from them.

Putting some typical numbers into the equations,

$$I_{0V(3)} = 1.2A \quad \text{with a VB+ of 24V because it is a high-power output board,}$$

$$I_{0V(2)} = 50mA \quad \text{with a VA+ of 3.3V because it is a microprocessor board with some CMOS logic on it.}$$

Now assume that, for various reasons, the power supply is some distance remote from the boards and you have without thinking connected it with 2m of 7/0.2mm equipment wire, which will have a room temperature resistance of about 0.2Ω. The voltage $V_S$ will be

$$V_S = 0.2 \cdot (1.2 + 0.05) = 0.25V$$

which will drop the supply voltage at PCB2 to 3.05V, less than the lower limit of operation for 3.3V logic, before allowing for supply voltage tolerances and other voltage drops. One wrong wiring connection can make your circuit operation borderline! Of course, the 0.25V is also subtracted from the 24V supply, but a reduction of about 1% on this supply is unlikely to affect operation.

**Varying loads**

If the 1.2A load on PCB3 is varying – say several high-current relays may be switched at different times, ranging from all off to all on – then the $V_S$ drop at PCB2 would also vary. This is very often worse than a static voltage drop because it introduces noise on
the 0V line. The effects of this include unreliable processor operation, variable set
threshold voltage levels and odd feedback effects such as chattering relays or, in audio
circuits, low-frequency “motor-boating” oscillation.

For comparison, look at the same figures but applied to Figure 1.2, with separate
0V return wires. Now there are two voltage drops to consider: \( V_{S(A)} \) for the 3.3V supply
and \( V_{S(B)} \) for the 24V supply. \( V_{S(B)} \) is 1.2A times 0.2\( \Omega \), substantially the same (0.24V)
as before, but it is only subtracted from the 24V supply. \( V_{S(A)} \) is now 50mA times 0.2\( \Omega \)
or 10mV, which is the only 0V drop on the 3.3V supply to PCB2 and is negligible.

The rule is: always separate power supply returns so that load currents for each
supply flow in separate conductors (Figure 1.7).

![Figure 1.7 Ways to connect power supply return](image)

Note that this rule is easiest to apply if different power supplies have different 0V
connections (as in Figure 1.2) but should also be applied if a common 0V is used, as
shown above. The extra investment in wiring is just about always worth it for peace of
mind!

**Power rail feed**

The rule also applies to the power rail feed as well as to its return, and in fact to any
connection where current is being shared between several circuits. Say the high-power
load on PCB3 was also being fed from the +5V supply VA+, then the preferred method
of connection is two separate feeds (Figure 1.8).

![Figure 1.8 Separate power supply rail feeds](image)

The reasons are the same as for the 0V return: with a single feed wire, a common
voltage drop appears in series with the supply voltage, injected this time in the supply
rail rather than the 0V rail. Fault symptoms are similar. Of course, the example above
is somewhat artificial in that you would normally use a rather more suitable size of wire
for the current expected. High currents flowing through long wires demand a low-
resistance and hence thick conductor. If you are expecting a significant voltage drop
then you will take the trouble to calculate it for a given wire diameter, length and
current. See Table 1.3 for a guide to the current-carrying abilities of common wires. The
point of the previous examples is that voltage drops have a habit of cropping up when you are not expecting them.

Conductor impedance

Note that the previous examples, and those on the next few pages, tacitly assume for simplicity that the wire impedance is resistive only. In fact, real wire has inductance as well as resistance and this comes into effect as soon as the wire is carrying ac, increasing in significance as the frequency is raised. A one-metre length of 16/0.2 equipment wire has a resistance of \(38\, \Omega\) and a self-inductance of \(1.5\, \mu\text{H}\). At 4A dc the voltage drop across it will be 152mV. An ac current with a rate of change of 4A/\(\mu\text{s}\) will generate 6V across it. Note the difference! The later discussion of wire types includes a closer look at inductance.

1.1.6 Input signal ground

Figure 1.2 shows the input signal connections being taken directly to PCB1 and not grounded outside of the pcb. To expand on this, the preferred scheme for two-wire single-ended input connections is to take the ground return directly to the reference point of the input amplifier: see Figure 1.9(a).

The reference point on a single-ended input is not always easy to find: look for the point from which the input voltage must be developed in order for the amplifier gain to act on it alone. In this way, no extra signals are introduced in series with the wanted signal by means of a common impedance. In each of the examples in Figure 1.9 of bad input wiring, getting progressively worse from (b) to (d), the impedance X-X acts as a source of unwanted input signal due to the other currents flowing in it as well as the input current.

Connection to 0V elsewhere on the pcb

Insufficient control over pc layout is the most usual cause of arrangement (b), especially if auto-routing layout software is used. Most CAD layout software assumes that the 0V rail is a single node and feels itself free to make connections to it at any point along the track. To overcome this, either specify the input return point as a separate node and connect it later, or edit the final layout as required. Manual layout is capable of exactly the same mistake, although in this case it is due to lack of communication between designer and layout draughtsman.

Connection to 0V within the unit

Arrangement (c) is quite often encountered if one pole of the input connector naturally makes contact with the metal case, such as happens with the standard BNC coaxial connector, or if for reasons of connector economy a common ground conductor is shared between multiple input, output or control signals that are distributed among different boards. With sensitive input signals, the latter is false economy; and if you have to use a BNC-type connector, you can get versions with insulating washers, or mount it on an insulating sub-panel in a hole in the metal enclosure. Incidentally, taking a coax lead internally from an uninsulated BNC socket to the pcb, with the coax outer connected both to the BNC shell and the pcb 0V, will introduce a ground loop (see section 1.1.4) unless it is the only path for ground currents to take. But at radio frequencies, this effect is countered by the ability of coax cable to concentrate the signal and return currents within the cable, so that the ground loop is only a problem at low frequencies.
External ground connection

Despite being the most horrific input grounding scheme imaginable, arrangement (d) is unfortunately not rare. Now, not only are noise signals internal to the unit coupled into the signal path, but also all manner of external ground noise is included. Local earth differences of up to 50V at mains frequency can exist at particularly bad locations such as power stations, and differences of several volts are more common. The only conceivable reason to use this layout is if the input signal is already firmly tied to a remote ground outside the unit, and if this is the case it is far better to use a differential amplifier as in Figure 1.9(e), which is often the only workable solution for low-level signals and is in any case only a logical development of the correct approach for single-ended signals (a). If for some reason you are unable to take a ground return connection from the input signal, you will be stuck with ground-injected noise.

Figure 1.9 Input signal grounding
All of the schemes of Figure 1.9(b) to (d) will work perfectly happily if the desired input signal is several orders of magnitude greater than the ground-injected interference, and this is frequently the case, which is how they came to be common practice in the first place. If there are good practical reasons for adopting them (for instance, connector or wiring cost restrictions) and you can be sure that interference levels will not be a problem, then do so. But you will need to have control over all possible connection paths before you can be sure that problems won’t arise in the field.

### 1.1.7 Output signal ground

Similar precautions need to be taken with output signals, for the reverse reason. Inputs respond unfavourably to external interference, whereas outputs are the cause of interference. Usually in an electronic circuit there is some form of power amplification involved between input and output, so that an output will operate at a higher current level than an input, and there is therefore the possibility of unwanted feedback.

The classical problem of output-to-input ground coupling is where both input and output share a common impedance, in the same way as the power rail common impedances discussed earlier. In this case the output current is made to circulate through the same conductor as connects the input signal return (Figure 1.10(a)).

![Figure 1.10 Output to input coupling](image)

A tailor-made feedback mechanism has been inserted into this circuit, by means of $R_S$. The input voltage at the amplifier terminals is supposed to be $V_{in}$, but actually it is

$$V_{in}' = V_{in} - (I_{out} \cdot R_S)$$

Redrawing the circuit to reference everything to the amplifier ground terminals (Figure 1.10(b)) shows this more clearly. When we work out the gain of this circuit, it turns out to be

$$V_{out}/V_{in} = A/(1 + [A \cdot R_S/(R_L + R_S)])$$

which describes a circuit that will oscillate if the term $[A \cdot R_S/(R_L + R_S)]$ is more negative than $-1$. In other words, for an inverting amplifier, the ratio of load impedance to common impedance must be less than the gain, to avoid instability. Even if the circuit
remains stable, the extra coupling due to $R_S$ upsets the expected response. Remember also that all the above terms vary with frequency, usually in a complex fashion, so that at high frequencies the response can be unpredictable. Note that although this has been presented in terms of an analogue system (such as an audio amplifier), any system in which there is input-output gain will be similarly affected. This can apply equally to a digital system with an analogue input and digital outputs which are controlled by it.

**Avoiding the common impedance**

The preferable solution is to avoid the common impedance altogether by careful layout of input and output grounds. We have already looked at input grounds, and the grounding scheme for outputs is essentially similar: take the output ground return directly to the point from which output current is sourced, with no other connection (or at least, no other susceptible connection) in between. Normally, the output current comes from the power supply so the best solution is to take the return directly back to the supply. Thus the layout of PCB3 in Figure 1.2 should have a separate ground track for the high-current output as in Figure 1.11(a), or the high-current output terminal could be returned directly to the power supply, bypassing PCB3 (b).

![Figure 1.11 Output signal returns](image)

If PCB3 contains only circuits which will not be susceptible to the voltage developed across $R_S$, then the first solution is acceptable. The important point is to decide in advance where your return currents will flow and ensure that they do not affect the operation of the rest of the circuits. This entails knowing the ac and dc impedance of any common connections, the magnitude and bandwidth of the output currents and the susceptibility of the potentially affected circuits.

### 1.1.8 Inter-board interface signals

There is one class of signals we have not yet covered, and that is those signals which pass within the unit from one board to another. Typically these are digital control signals or analogue levels which have already been processed, so are not low-level enough to be susceptible to ground noise and are not high-current enough to generate significant quantities of it. To be thorough in your consideration of ground return paths, these signals should not be left out: the question is, what to do about them?

Often the answer is nothing. If no ground return is included specifically for inter-board signals then signal return current must flow around the power supply connections
and therefore the interface will suffer all the ground-injected noise $V_n$ that is present along these lines (Figure 1.12). But, if your grounding scheme is well thought out, this may well not be enough to affect the operation of the interface. For instance, 100mV of noise injected in series with a CMOS logic interface which has a noise margin of 1V will have no direct effect. Or, ac noise injection onto a dc analogue signal which is well-filtered at the interface input will be tolerable.

**Figure 1.12** Inter-board ground noise

**Partitioning the signal return**

There will be occasions when taking the long-distance ground return route is not good enough for your interface. Typically these are

- where high-speed digital signals are communicated, and the ground return path has too much inductance, resulting in ringing on the signal transitions;
- when interfacing precision analogue signals which cannot stand the injected noise or low-voltage dc differentials.

If you solve these headaches by taking a local inter-board ground connection for the signal of interest, you run the risk of providing an alternative path for power supply return currents, which nullifies the purpose of the local ground connection. A fraction of the power return current will flow in the local link (Figure 1.13), the proportion depending on the relative impedances, and you will be back where you started.

**Figure 1.13** Power supply return currents through inter-board links

If you really need the local signal return, but are in trouble with ground return currents, there are two options to pursue:

- separate the ground return (Figure 1.14) for the input side of the interface from the rest of the ground on that pcb. This has the effect of moving the ground noise injection point inboard, after the input buffer, which may be all that you need. A development of this scheme is to include a “stopper” resistor of a few ohms in the gap X-X. This prevents dc ground current flow because its impedance is high relative to that of the correct ground path, but
it effectively ties the input buffer to its parent ground at high frequencies and prevents it from floating if the inter-board link is disconnected.

Figure 1.14 Separating the ground returns

- use differential connections at the interface. The signal currents are now balanced and do not require a ground return; any ground noise is injected in common mode and is cancelled out by the input buffer. This technique is common where high-speed or low-level signals have to be communicated some distance, but it is applicable at the inter-board level as well. It is of course more expensive than typical single-ended interfaces since it needs dedicated buffer drivers and receivers.

1.1.9 Star-point grounding

One technique that can be used as a circuit discipline is to choose one point in the circuit and to take all ground returns to this point. This is then known as the “star point”. Figure 1.2 shows a limited use of this technique in connecting together chassis, mains earth, power supply ground and 0V returns to one point. It can also be used as a local sub-ground point on printed circuit layouts.

When comparatively few connections need to be made this is a useful and elegant trick, especially as it offers a common reference point for circuit measurements. It can be used as a reference for power supply voltage sensing, in conjunction with a similar star point for the output voltage (Figure 1.2 again). It becomes progressively messier as more connections are brought to it, and should not substitute for a thorough analysis of the anticipated ground current return paths.

1.1.10 Ground connections between units

Much of the theory about grounding techniques tends to break down when confronted with the prospect of several interconnected units. This is because the designer often has either no control over the way in which units are installed, or is forced by safety-related or other installation practices to cope with a situation which is hostile to good grounding practice.

The classic situation is where two mains powered units are connected by one (or more) signal cable (Figure 0.15). This is the easiest situation to explain and visualise; actual set-ups may be complicated by having several units to contend with, or different and contradictory ground regimes, or by extra mechanical bonding arrangements.

This configuration is exactly analogous to that of Figure 1.12. Ground noise, represented by $V_n$, is coupled through the mains earth conductors and is unpredictable and uncontrollable. If the two units are plugged in to the same mains outlet, it may be very small, though never zero, as some noise is induced simply by the proximity of the
live and neutral conductors in the equipment mains cable. But this configuration cannot be prescribed: it will be possible to use outlets some distance apart, or even on different distribution rings, in which case the ground connection path could be lengthy and could include several noise injection sources. Absolute values of injected noise can vary from less than a millivolt rms in very quiet locations to the several volts, or even tens of volts, mentioned in section 1.1.6. This noise effectively appears in series with the signal connection.

In order to tie the signal grounds in each unit together you would normally run a ground return line along with the signal in the same cable, but then

- noise currents can now flow in the signal ground, so it is essential that the impedance of the ground return \((R_s)\) is much less than the noise source impedance \((R_n)\) – usually but not invariably the case – otherwise the ground-injected noise will not be reduced;
- you have created a ground loop (Figure 1.16, and compare this with section 1.1.4) which by its nature is likely to be both large and variable in area, and to intersect various magnetic field sources, so that induced ground currents become a real hazard.

\[
V_g = V_n \cdot \left(\frac{R_s}{R_s + R_n}\right)
\]

![Figure 1.15 Inter-unit ground connection via the mains](image)

![Figure 1.16 Ground loop via signal and mains earths](image)
Breaking the ground link

If the susceptibility of the signal circuit is such that the expected environmental noise could affect it, then you have a number of possible design options.

- float one or other unit (disconnect its mains ground connection), which breaks the ground loop at the mains lead. This is already done for you if it is battery-powered and in fact this is one good reason for using battery-powered instruments. On safety-class I (earthed) mains powered equipment, doing this is not an option because it violates the safety protection.

- transmit your signal information via a differential link, as recommended for inter-board signals earlier. Although a ground return is not necessary for the signal, it is advisable to include one to guard against too large a voltage differential between the units. Noise signals are now injected in common-mode relative to the wanted signal and so will be attenuated by the input circuit’s common mode rejection, up to the operating limit of the circuit, which is usually several volts.

- electrically isolate the interface. This entails breaking the direct electrical connection altogether and transmitting the signal by other means, for instance a transformer, opto-coupler or fibre optic link. This allows the units to communicate in the presence of several hundred volts or more of noise, depending on the voltage rating of the isolation; alternatively it is useful for communicating low-level ac signals in the presence of relatively moderate amounts of noise that cannot be eliminated by other means.

1.1.11 Shielding

Some mention must be made here of the techniques of shielding inter-unit cables, even though this is more properly the subject of Chapter 8. Shielded cable is used to protect signal wires from noise pickup, or to prevent power or signal wires from radiating noise. This apparently simple function is not so simple to apply in practice. The characteristics of shielded cable are discussed later (see section 1.2.4); here we shall look at how to apply it.

At which end of a cable do you connect the shield, and to what? There is no one correct answer, because it depends on the application. If the cable is used to connect two units which are both contained within screened enclosures to keep out or keep in RF energy, then the cable shield has to be regarded as an extension of the enclosures and it must be connected to the screening at both ends via a low-inductance connection, preferably the connector screen itself (Figure 1.17). This is a classic application of EMC principles and is discussed more fully in sections 8.5 and 8.7. Note that if both of the unit enclosures are themselves separately grounded then you have formed a ground loop (again). Because ground loops are a magnetic coupling hazard, and because magnetic coupling diminishes in importance at higher frequencies, this is often not a problem when the purpose of the screen is to reduce hf noise. The difficulty arises if you are screening both against high and low frequencies, because at low frequencies you should ground the shield at one end only, and in these cases you may have to take the expensive option of using double-shielded cable.

The shield should not be used to carry signal return currents unless it is at RF and you are using coaxial cable. Noise currents induced in it will add to the signal, nullifying the effect of the shield. Typically, you will use a shielded pair to carry high-impedance low-level input signals which would be susceptible to capacitive pickup. (A
cable shield will not be effective against magnetic pickup, for which the best solution is twisted pair.

Which end to ground for LF shielding

If the input source is floating, then the shield can be grounded at the amplifier input. A source with a floating screen around it can have this screen connected to the cable shield. But, if the source screen is itself grounded, you will create a ground loop with the cable shield, which is undesirable: ground loop current induced in the shield will couple into the signal conductors. One or other of the cable shield ends should be left floating, depending on the relative amount of unavoidable capacitive coupling to ground ($C_c$) that exists at either end. If you have the choice, usually it is the source end (which may be a transducer or sensor) that has the lower coupling capacitance so this end should be floated.

If the source is single-ended and grounded, then the cable shield should be grounded at the source and either left floating at the (differential) input end or connected through a choke or low value resistor to the amplifier ground. This will preserve dc and low-
frequency continuity while blocking the flow of large induced high-frequency currents along the shield. The shield should not be grounded at the opposite end to the signal. Figure 1.18 shows the options.

**Electrostatic screening**

When you are using shielded cable to prevent electrostatic radiation from output or inter-unit lines, ground loop induction is usually not a problem because the signals are not susceptible, and the cable shield is best connected to ground at both ends. The important point is that each conductor has a distributed (and measurable) capacitance to the shield, so that currents on the shield will flow as long as there are ac signals propagating within it. These shield currents must be provided with a low-impedance ground return path so that the shield voltages do not become substantial. The same applies in reverse when you consider coupling of noise induced on the shield into the conductors.

![Figure 1.18](image)

**Figure 1.18** Electrostatic screening

**Surface transfer impedance**

At high frequencies, the notion of surface transfer impedance becomes useful as a measure of shielding effectiveness. This is the ratio of voltage developed between the inner and outer conductors of shielded cable due to interference current flowing in the shield, expressed in milliohms per unit length. It should not be confused with characteristic impedance, with which it has no connection. A typical single braid screen will be ten milliohms/m or so below 1MHz, rising at a rate of 20dB/decade with increasing frequency. The common aluminium/mylar foil screens are around 20dB worse. Unhappily, surface transfer impedance is rarely specified by cable manufacturers.

**1.1.12 The safety earth**

A brief word is in order about the need to ensure a mains earth connection, since it is obvious from the preceding discussion that this requirement is frequently at odds with anti-interference grounding practice. Most countries now have electrical standards which require that equipment powered from dangerous voltages should have a means of protecting the user from the consequences of component failure. The main hazard is deemed to be inadvertent connection of the live mains voltage to parts of the equipment with which the user could come into contact directly, such as a metal case or a ground terminal.

Imagine that the fault is such that it makes a short circuit between live and case, as shown in Figure 1.20. These are normally isolated and if no earth connection is made the equipment will continue to function normally – but the user will be threatened with a lethal shock hazard without knowing it. If the safety earth conductor is connected then the protective mains fuse will blow when the fault occurs, preventing the hazard and alerting the user to the fault.

For this reason a safety earth conductor is mandatory for all equipment that is designed to use this type of protection, and does not rely on extra levels of insulation.
The conductor must have an adequate cross-section to carry any prospective fault current, and all accessible conductive parts must be electrically bonded to it. The general requirements for earth continuity are

- the earth path should remain intact until the circuit protection has operated;
- its impedance should not significantly or unnecessarily restrict the fault current.

As an example, EN 60065 requires a resistance of less than 0.5 Ω at 10A for a minute. Design for safety is covered in greater detail in section 9.1.

### 1.2 Wiring and cables

This section will look briefly at the major types of wire and cable that can be found within typical electronic equipment. There are so many varieties that it comes as something of a surprise to find that most applications can be satisfied from a small part of the range. First, a couple of definitions: **wires** are single-circuit conductors, insulated or not; **cables** are groups of individual conductors, separately insulated and mechanically contained within an overall sheath.

#### 1.2.1 Wire types

The simplest form of wire is tinned copper wire, available in various gauges depending on required current carrying capacity. Component leads are almost invariably tinned copper, but the wire on its own is not used to a great extent in the electronics industry. Its main application was for links on printed circuit boards, but the increasing use of double-sided and multilayer plated-through-hole boards makes them redundant. Tinned copper wire can also be used in re-wirable fuselinks. Insulated copper wire is used principally in wound components such as inductors and transformers. The insulating coating is a polyurethane compound which has self-fluxing properties when heated, which makes for ease of soldered connection, especially to thin wires.

Table 1.2 compares dimensions, current capacity and other properties for various sizes of copper wire. In the UK the wires are specified under BS EN 13602 for tinned copper and BS EN 60182 (IEC 60182-1) for enamel insulated, and are sold in metric sizes. Two grades of insulation are available, Grade 1 being thinner; Grade 2 has roughly twice the breakdown voltage capability.

**Wire inductance**

We mentioned earlier that any length of wire has inductance as well as resistance. The
The approximate formula for the inductance of a straight length of round section wire at high frequencies is

\[ L = K \cdot l \cdot (2.3 \log_{10}(4l/d) - 1) \text{ microhenries} \]

where \( l \) and \( d \) are length and diameter respectively, \( l >> d \) and \( K \) is 0.0051 for dimensions in inches or 0.002 for dimensions in cm.

This equation is used to derive the inductance of a 1m length (note that this is not quite

<table>
<thead>
<tr>
<th>Wire size (mm dia)</th>
<th>1.6</th>
<th>1.25</th>
<th>0.71</th>
<th>0.56</th>
<th>0.315</th>
<th>0.2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Approx. standard wire gauge (SWG)</td>
<td>16</td>
<td>18</td>
<td>22</td>
<td>24</td>
<td>30</td>
<td>35</td>
</tr>
<tr>
<td>Approx. American Wire Gauge (AWG)</td>
<td>14</td>
<td>16</td>
<td>21</td>
<td>23</td>
<td>28</td>
<td>32</td>
</tr>
<tr>
<td>Current rating (Amps)</td>
<td>22</td>
<td>12.2</td>
<td>3.5</td>
<td>2.5</td>
<td>0.9</td>
<td>0.33</td>
</tr>
<tr>
<td>Fusing current (Amps)</td>
<td>70</td>
<td>45</td>
<td>25</td>
<td>17</td>
<td>9</td>
<td>5</td>
</tr>
<tr>
<td>Resistance/metre @ 20°C (W)</td>
<td>0.0085</td>
<td>0.014</td>
<td>0.043</td>
<td>0.069</td>
<td>0.22</td>
<td>0.54</td>
</tr>
<tr>
<td>Inductance of 1 metre length (µH)</td>
<td>1.36</td>
<td>1.41</td>
<td>1.53</td>
<td>1.57</td>
<td>1.69</td>
<td>1.78</td>
</tr>
</tbody>
</table>

**Table 1.2 Characteristics of copper wire**

<table>
<thead>
<tr>
<th>Wire size (no. of strands/mm dia)</th>
<th>1/0.6</th>
<th>7/0.2</th>
<th>16/0.2</th>
<th>24/0.2</th>
<th>32/0.2</th>
<th>63/0.2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistance (Ω/1000m at 20°C)</td>
<td>64</td>
<td>88</td>
<td>38</td>
<td>25.5</td>
<td>19.1</td>
<td>9.7</td>
</tr>
<tr>
<td>Current rating at 70°C (A)</td>
<td>1.8</td>
<td>1.4</td>
<td>3.0</td>
<td>4.5</td>
<td>6.0</td>
<td>11.0</td>
</tr>
<tr>
<td>Current rating at 25°C (A)</td>
<td>3.0</td>
<td>2.0</td>
<td>4.0</td>
<td>6.0</td>
<td>10.0</td>
<td>18.0</td>
</tr>
<tr>
<td>Voltage drop/metre at 25°C current</td>
<td>192mV</td>
<td>176mV</td>
<td>152mV</td>
<td>153mV</td>
<td>191mV</td>
<td>175mV</td>
</tr>
<tr>
<td>Voltage rating</td>
<td>1KV</td>
<td>1KV</td>
<td>1KV</td>
<td>1.5KV</td>
<td>1.5KV</td>
<td>1.5KV</td>
</tr>
<tr>
<td>Overall diameter (mm)</td>
<td>1.2</td>
<td>1.2</td>
<td>1.55</td>
<td>2.4</td>
<td>2.6</td>
<td>3.0</td>
</tr>
<tr>
<td>Near equivalent American Wire Gauge (not direct equivalent)</td>
<td>23</td>
<td>24</td>
<td>20</td>
<td>18</td>
<td>17</td>
<td>15</td>
</tr>
</tbody>
</table>

**Table 1.3 Characteristics of BS4808 equipment wire**

<table>
<thead>
<tr>
<th>Kynar: 30AWG</th>
<th>26AWG</th>
<th>Tefzel: 30AWG</th>
<th>26AWG</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conductor dia (mm)</td>
<td>0.25</td>
<td>0.4</td>
<td>0.25</td>
</tr>
<tr>
<td>Maximum service temperature °C</td>
<td>105</td>
<td>105</td>
<td>155</td>
</tr>
<tr>
<td>Resistance/m @ 20°C (W)</td>
<td>0.345</td>
<td>0.136</td>
<td>0.345</td>
</tr>
<tr>
<td>Voltage rating (V)</td>
<td>-</td>
<td>-</td>
<td>375</td>
</tr>
<tr>
<td>Current rating @ 50°C (A)</td>
<td>-</td>
<td>-</td>
<td>2.6</td>
</tr>
</tbody>
</table>

**Table 1.4 Characteristics of wire-wrap wire**
the same as inductance per metre) in Table 1.2 and you can see that inductance is only marginally affected by wire diameter. Low values of inductance are not easily obtained by adding cross-section and the reactive component of impedance dominates above a few kiloHertz whatever the size of the conductor. A useful rule of thumb is that the inductance of a one inch length of ordinary equipment wire is around 20nH and that of a one centimetre length is around 7nH. This factor becomes important in high speed digital and RF circuits where performance is limited by physical separation, and also in circuits where the rate-of-change of current (di/dt) is high.

Equipment wire

Equipment wire is classified mainly according to its insulation. This determines the voltage rating and the environmental properties of the wire, particularly its operating temperature range and its resistance to chemical and solvent attack. The standard type of wire, and the most widely available, is PVC insulated to BS4808 which has a maximum temperature rating of 85˚C. As well as current ratings at 25˚C you will find specifications at 70˚C; these allow for a 15˚C temperature rise, to the maximum rated temperature, at the specified current. Temperature ratings of 70˚C for large conductor switchgear applications and 105˚C to American and Canadian UL and CSA standards are also available in PVC. PTFE is used for wider temperature ranges, up to 200˚C, but is harder to work with. Other more specialised insulations include extra-flexible PVC for test leads and silicone rubber for high temperature (150˚C) and harsh environments. Many wires carry military, telecom and safety authority approval and have to be specified on projects that are carried out for these customers.

Table 1.3 is included here as a guide to the electrical characteristics of various commonly-available PVC equipment wires. Note that the published current ratings of each wire are related to permitted temperature rise. Copper has a positive temperature coefficient of resistivity of 0.00393 per ˚C, so that resistance rises with increasing current; using the room temperature resistance may be optimistic by several per cent if the actual ambient temperature is high or if significant self-heating occurs.

Wire-wrap wire

A further specialised type of wire is that used for wire-wrap construction. This is available primarily in two sizes, with two types of insulation: Kynar, trademark of Pennwalt, and Tefzel®, trademark of Du Pont. Tefzel is the more expensive but has a higher temperature rating and is easier to strip. Table 1.4 lists the properties of the four types.

1.2.2 Cable types

Ignoring the more specialised types, cables can be divided loosely into three categories:

- power
- data and multicore
- RF

1.2.3 Power cables

Because mains power cables are inherently meant to carry dangerous voltages they are subject to strict standards: in the UK the principal one is BS6500. International ones are IEC 60227 for PVC insulated or IEC 60245 for rubber insulated. These standards have been harmonised throughout the CENELEC countries in Europe so that any
equipment which uses a cable with a harmonised code number will be acceptable throughout Europe. BS6500 specifies a range of current ratings and allows a variety of sheath materials depending on application. The principal ones are rubber and PVC; rubber is about twice the price of PVC but is somewhat more flexible and therefore suitable for portable equipment, and can be obtained in a high-temperature HOFR (heat and oil resisting, flame retardant) grade. The current-carrying capacities and voltage drops for dc and single-phase ac, and supportable mass are shown in Table 1.6.

Unfortunately, American and Canadian mains cables also need to be approved, but the approvals authorities are different (UL and CSA). Cables manufactured to the European harmonised standards do not meet UL/CSA standards and vice versa. So, if you intend to export your mains-powered equipment both to Europe and North America you will need to supply it with two different cables. The easy way to do this is to use a CEE-22 6 Amp connector on the equipment and supply a different cable set depending on the market. This practice has been adopted by virtually all of the large-volume multinational equipment suppliers with the result that the CEE-22 mains inlet is universally accepted. There are also several suppliers of ready-made cable sets for the different countries!

The alternative, widely used for information technology and telecoms equipment, is to use a “wall-wart” plug top power supply and provide different ones for each market, so that the cable carries low voltage dc and no approved mains cable is needed.

### 1.2.4 Data and multicore cables

Multicore cables are used when you need to transport several signals between the same source and destination. They should never be used for mains power because of the hazards that could be created by a cable failure, nor should high-power and signal wires be run within the same cable because of the risks of interference. Conventional multicore is available with various numbers of conductors in 7/0.1mm, 7/0.2mm and 16/0.2mm, with or without an overall braided screen. As well as the usual characteristics of current and voltage ratings, which are less than the ratings for individual wires because the conductors are bunched together, inter-conductor capacitance is an important consideration, especially for calculating crosstalk (to which we return shortly). It is not normally specified for standard multicore, although nominal conductor-to-screen capacitances of 150–200pF/m are sometimes quoted. For a more complete specification you need to use data cable.

<table>
<thead>
<tr>
<th>Cross-sectional area (mm²)</th>
<th>0.5</th>
<th>0.75</th>
<th>1.0</th>
<th>1.25</th>
<th>1.5</th>
<th>2.5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Current-carrying capacity (A)</td>
<td>3</td>
<td>6</td>
<td>10</td>
<td>13</td>
<td>16</td>
<td>25</td>
</tr>
<tr>
<td>Voltage drop per amp per metre (mV)</td>
<td>93</td>
<td>62</td>
<td>46</td>
<td>37</td>
<td>32</td>
<td>19</td>
</tr>
<tr>
<td>Maximum supportable mass (kg)</td>
<td>2</td>
<td>3</td>
<td>5</td>
<td>5</td>
<td>5</td>
<td>5</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Correction factor for ambient temperature</th>
</tr>
</thead>
<tbody>
<tr>
<td>60°C rubber and PVC cables:</td>
</tr>
<tr>
<td>Temp.</td>
</tr>
<tr>
<td>CF</td>
</tr>
<tr>
<td>85°C HOFR rubber cables:</td>
</tr>
<tr>
<td>Temp.</td>
</tr>
<tr>
<td>CF</td>
</tr>
</tbody>
</table>

Table 1.5 Characteristics of BS6500 mains cables
Source: IEE Wiring Regulations 15th Edition
Data communication cables

Data cables are really a special case of multicore, but with the explosion in data communications they now deserve a special category of their own. Transmitting digital data presents special problems, notably

- the need to communicate several parallel channels at once, usually over short distances, which has given rise to ribbon cable;
- the need to communicate a few channels of high-speed serial data over long distances with a high data integrity, which has given rise to cables with multiple individually-screened conductor pairs in an overall sheath which may or may not be screened.

Inter-conductor capacitances and characteristic impedances (which we will discuss when we come to transmission lines) are important for digital data transmission and are quoted for most of these types. Table 1.6 summarises the characteristics of the most common of them.
Structured data cable

One particular cable application which forms an important aspect of data communications is so-called “structured” or “generic” cabling. This is general-purpose datacomms cable which is installed into the structure of a building or campus to enable later implementation of a variety of telecom and other networks: voice, data, text, image and video. In other words, the cable’s actual application is not defined at the time of installation. To allow this, its characteristics, along with those of its connectors, performance requirements and the rules for acceptable routing configurations, are defined in ISO/IEC 11801 (the US TIA/EIA-568 covers the same ground).

Equipment designers may not be too interested in the specifications of this cable until they come to design a LAN or telecom port interface; then the cable becomes important. The TIA/EIA-568 (both ISO/IEC 11801 and EN 50173 have similar specifications) parameters for the preferred 100Ω quad-pair cable are shown in Table 1.8. The standard allows for a series of categories with increasing bandwidth. Cat 5 and Cat 5e are popular and have been widely installed.

### Table 1.8 Characteristics of TIA/EIA-568 (ISO/IEC 11801) 100Ω quad pair cable

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Cat 3</th>
<th>Cat 5</th>
<th>Cat 5e</th>
<th>Cat 6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandwidth</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Freq. MHz</td>
<td>16MHz</td>
<td>100MHz</td>
<td>250MHz</td>
<td></td>
</tr>
<tr>
<td>Characteristic impedance</td>
<td>0.1</td>
<td>75 – 150 Ω</td>
<td>N/A</td>
<td></td>
</tr>
<tr>
<td>≥1</td>
<td>100 ±15 Ω</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Attenuation dB per 100m</td>
<td>0.256</td>
<td>1.3</td>
<td>1.1</td>
<td>N/A</td>
</tr>
<tr>
<td></td>
<td>1.0</td>
<td>2.6</td>
<td>2.1</td>
<td>2.0</td>
</tr>
<tr>
<td></td>
<td>4.0</td>
<td>5.6</td>
<td>4.3</td>
<td>3.8</td>
</tr>
<tr>
<td></td>
<td>10.0</td>
<td>9.8</td>
<td>6.6</td>
<td>6.0</td>
</tr>
<tr>
<td></td>
<td>16</td>
<td>13.1</td>
<td>8.2</td>
<td>7.6</td>
</tr>
<tr>
<td></td>
<td>31.25</td>
<td>N/A</td>
<td>11.8</td>
<td>10.7</td>
</tr>
<tr>
<td></td>
<td>62.5</td>
<td></td>
<td>17.1</td>
<td>15.4</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td></td>
<td>22.0</td>
<td>19.8</td>
</tr>
<tr>
<td></td>
<td>200</td>
<td></td>
<td>N/A</td>
<td>29.0</td>
</tr>
<tr>
<td></td>
<td>250</td>
<td></td>
<td></td>
<td>32.8</td>
</tr>
<tr>
<td>Capacitance unbalance</td>
<td>1kHz</td>
<td>3400 pF/km</td>
<td>330pF/100m</td>
<td></td>
</tr>
<tr>
<td>DC loop resistance</td>
<td></td>
<td>19.2 Ω/100m, max unbalance 3%</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Return loss dB at 100m cable length</td>
<td>1 – 10</td>
<td>12</td>
<td>23</td>
<td>20 + 5log(ƒ)</td>
</tr>
<tr>
<td></td>
<td>10 – 20</td>
<td>12 – 10log(ƒ/10)</td>
<td>23</td>
<td>25</td>
</tr>
<tr>
<td></td>
<td>20 – 100</td>
<td>N/A</td>
<td>23 – 10log(ƒ/20)</td>
<td>25 – 7log(ƒ/20)</td>
</tr>
<tr>
<td></td>
<td>200</td>
<td>N/A</td>
<td></td>
<td>18.0</td>
</tr>
<tr>
<td></td>
<td>250</td>
<td>N/A</td>
<td></td>
<td>17.3</td>
</tr>
</tbody>
</table>

Other characteristics, particularly mechanical dimensions, crosstalk performance (extended for Cat 5e and 6), and propagation delay skew are also defined.
Shielding and microphony

Shielding of data and multicore falls into three categories:

- copper braid. This offers a good general-purpose electrical shield but cannot give 100% shield coverage (80–95% is typical) and it increases the size and weight of the cable.
- tape or foil. The most common of these is aluminised mylar. A drain wire is run in contact with the metallisation to provide a terminating contact and to reduce the inductance of the shield when it is helically-wound. This provides a fairly mediocre degree of shielding but hardly affects the size, weight and flexibility of the cable at all.
- composite foil and braid. These provide excellent electrostatic shielding for demanding environments but are more expensive – about twice the price of foil types.

For small-signal applications, particularly low-noise audio work, another cable property is important – microphony due to triboelectric induction. Any insulator generates a static voltage when it is rubbed against a dissimilar material, and this effect results in a noise voltage between conductor and screen when the cable is moved or vibrated. Special low-noise cable is available which minimises this noise mechanism by including a layer of low resistance dielectric material between braid and insulator to dissipate the static charge. When you are terminating this type of cable, make sure the low resistance layer is stripped back to the braid, otherwise you run the risk of a near short circuit between inner and outer.

1.2.5 RF cables

Cables for the transport of radio frequency signals are almost invariably coaxial, apart from a few specialised applications such as hf aerial feeder which may use balanced lines. Coax’s outstanding property is that the field due to the signal propagating along it is confined to the inside of the cable (Figure 1.21), so that interaction with its external environment is kept to a minimum. A further useful property is that the characteristic impedance of coax is easily defined and maintained. This is important for RF applications as in these cases cable lengths frequently exceed the operating wavelength.

![Coax cable diagram](image)

**Figure 1.21** Coax cable

The generic properties of transmission lines – of which coax is a particular type – will be discussed in section 1.3. The parameters that you will normally find in coax specifications are as follows:

- characteristic impedance ($Z_0$): the universal standard is $50\,\Omega$, since this results in a good balance between mechanical properties and ease of circuit application. $75\,\Omega$ and $93\,\Omega$ are other standards which find application in
video and data systems. Any other impedance must be regarded as a special.

- dielectric material. This affects just about every property of the cable, including \( Z_0 \), attenuation, voltage handling, physical properties and temperature range. Solid polythene or polyethylene are the standard materials; cellular polyethylene, in which part of the dielectric insulation is provided by air gaps, offers lower weight and lower attenuation losses but is more prone to physical distortion than solid. These two have a temperature rating of 85°C. PTFE is available for higher temperature (200°C) and lower loss applications but is much more expensive.

- conductor material. Copper is universal. Silver plating is sometimes used to enhance high-frequency conductivity through the skin effect, or copper can be plated onto steel strands for strength. Inner conductors can be single or stranded; stranded is preferred when the cable will be subject to flexing. The outer conductor is normally copper braid, again for flexibility. The degree of braid coverage affects high-frequency attenuation and also the shielding effectiveness. Solid outer conductor is available for extreme applications that don’t require flexing.

- voltage rating. A thicker cable can be expected to have a higher voltage rating and a lower attenuation. You cannot easily relate the voltage rating to power handling ability unless the cable is matched to its characteristic impedance. If the cable isn’t matched, voltage standing waves will exist which will produce peaks at distinct locations along the cable higher than would be expected from the power/impedance relationship.

- attenuation. Losses in the dielectric and conductors result in increasing attenuation with frequency and distance, so attenuation is quoted per 10 metres at discrete frequencies and you can interpolate to find the attenuation at your operating frequency. Cable losses can easily catch you out, especially if you are operating long cables over a wide bandwidth and forget to allow for several extra dB of loss at the top end.

Readily-available coax cables are specified to two standards, the US MIL-C-17 for the RG/U (Radio Government, Universal) series and the UK BS2316 for the UR-M (Uniradio) series. The international standard is IEC 60096. Table 1.7 gives comparative data for a few common 50\( \Omega \) types.

One word of warning: never confuse screened audio cable with RF coax. The braids and dielectric materials are quite different, and audio cable’s \( Z_0 \) is undefined and its attenuation at high frequencies is large. If you try to feed RF down it you won’t get much at the other end! On the other hand, RF coax can be used to carry audio signals.

1.2.6 Twisted pair

Special mention should be given to twisted pair because it is a particularly effective and simple way of reducing both magnetic and capacitive interference pickup. Twisting the wires tends to ensure a homogeneous distribution of capacitances. Both capacitance to ground and to extraneous sources are balanced. This means that common mode capacitive coupling is also balanced, allowing high common mode rejection. Figure 1.22 compares twisted and un-twisted pairs. But note that if your problem is already common mode capacitive coupling, twisting the wires won’t help. For that, you need shielding.

Twisting is most useful in reducing low-frequency magnetic pickup because it
Grounding and wiring

Each half-twist reverses the direction of induction so, assuming a uniform external field, two successive half-twists cancel the wires' interaction with the field. Effective loop pickup is now reduced to the small areas at each end of the pair, plus some residual interaction due to non-uniformity of the field and irregularity in the twisting. Assuming that the termination area is included in the field, the number of twists per unit length is unimportant: around $8\sim 16$ turns per foot (26$\sim$50 turns per metre) is usual. Figure 1.23 shows measured magnetic field attenuation versus frequency for twisted 22AWG wires compared to parallel 22AWG wires spaced at 0.032”.

A further advantage of twisting pairs together is that it allows a fairly reproducible characteristic impedance. When combined with an overall shield to reduce common-mode capacitive pickup, the resulting cable is very suitable for high-speed data communication as it reduces both radiated noise and induced interference to a minimum.

### 1.2.7 Crosstalk

When more than one signal is run within the same cable bundle for any distance, the mutual coupling between the wires allows a portion of one signal to be fed into another,
The phenomenon is known as crosstalk. Strictly speaking, crosstalk is not only a cable phenomenon but refers to any unwanted interaction between nominally uncoupled channels. The coupling can be predominantly either capacitive, inductive, or due to transmission-line phenomena.

The equivalent circuit for capacitive coupling at low-to-medium frequencies where the cable can be considered as a lumped component (in contrast to high frequencies where it must be considered as a transmission line) is as shown in Figure 1.24.

\[
C = C_C \cdot D
\]

for the case where circuit 1 is coupling into circuit 2

\[
\text{Crosstalk voltage } V_X = V_{S1} \cdot \left( \frac{(R_{S2}/R_{L2})}{[(R_{S2}/R_{L2}) + (R_{S1}/R_{L1}) + (1/\omega C)]} \right)
\]

**Figure 1.24** Crosstalk equivalent circuit

In the worst case where the capacitive coupling impedance is much lower than the circuit impedance, the crosstalk voltage is determined only by the ratio of circuit impedances.

**Digital crosstalk**

Crosstalk is well known in the telecomms and audio worlds, for example where separate speech channels are transmitted together and one breaks through onto another, or where stereo channel separation at high frequencies is compromised. Although digital data might seem at first sight immune from crosstalk, in fact it is a serious threat to data integrity as well. The capacitive coupling is all but transparent to fast edges with the result that clocked data can be especially corrupted, as Figure 1.25 shows. If the logic noise immunity is poor, severe false clocking can result. A couple of worked examples will demonstrate the nature of the problem.

**Figure 1.25** Digital crosstalk effects
(a) Two audio circuits with 10kΩ source and load impedances are run in 2 metres of multicore cable with interconductor capacitances of 150pF/m. What is the crosstalk ratio at 10kHz?

The coupling capacitance $C_C$ is 2 metres of 150pF/m = 300pF. At 10kHz this has an impedance of 53kΩ.

The source and load impedances in the crosstalk circuit in each case are 10KΩ/10KΩ = 5kΩ.

So the crosstalk will be

$$\frac{5K}{(5K + 5K + 53K)} = 22\text{dB}$$

which is unacceptable in just about any situation!

If the output drive impedance is reduced from 10kΩ to 50Ω then the crosstalk becomes

$$\frac{49}{(49 + 49 + 53K)} = 60\text{dB}$$

which is acceptable for many purposes, though probably not for hi-fi.

(b) Two EIA-232 (RS-232) serial data lines are run in 16m of data cable (not individual twisted pair) which has a core/core capacitance of 108pF/m. The transmitters and receivers conform to the EIA-232 spec of 300Ω output impedance, 5kΩ input impedance, ±10V swing and 30V/µsec rise time. What is the expected magnitude of interference spikes on one circuit due to the other?

Coupling capacitance here is 16 x 108pF = 1728pF.

The current that will be flowing after t seconds in an RC circuit fed from a ramping voltage with a constant $dV/dt$ is

$$I = C \frac{dV}{dt} (1 - \exp[-t/RC])$$

which for our case with $dV/dt = 30V/\mu\text{sec}$ for 0.66 µsec and a circuit resistance of 567Ω is 25mA. This translates to a peak voltage across the load resistance of $\frac{300}{5K/5K}$ of

$$25 \times 10^{-3} \times 267 = 6.8V$$

This is one reason why EIA-232 isn’t suitable for long distances and high data rates!

Crosstalk can be combated with a number of strategies, which follow from the above examples. These are

- reduce the circuit source and/or load impedances. Ideally, the offending circuit’s source impedance should be high and the victim’s should be low. Low impedances require more capacitance for a given amount of coupling.
- reduce the mutual coupling capacitance. Use a shorter cable, or select a cable with lower core-to-core capacitance per unit length. Note that for fast or high-frequency signals this won’t solve anything, because the impedance of the coupling capacitance is lower than the circuit impedances. If you use ribbon cable, sacrifice some space and tie a conductor to ground between each signal conductor; another alternative is ribbon cable with an integral ground plane. Best of all, use an individual screen for each circuit. The screen must be grounded or you gain nothing at all from this tactic!
- reduce the signal circuit bandwidth to the minimum required for the data rate or frequency response of the system. As can be seen from (b) above, the coupling depends directly on the rise time of the offending signal. Slower
rise times mean less crosstalk. If you do this by adding a capacitance in parallel with the input load resistor (across \( RL_2 \) in Figure 1.24) this will act as a potential divider with the core-to-core capacitance, as well as reducing the input impedance for high-frequency noise.

- use differential transmission. The bogey of crosstalk is a major reason for the popularity of differential data standards such as EIA-422 (RS-422), and other more recent ones, at high data rates. Coupling capacitance is not necessarily reduced by using paired lines, but the crosstalk is now injected in common mode and so benefits from the common-mode rejection of the input buffer. The limiting factor to the degree of rejection that can be obtained is the unbalance in coupling capacitance of each half of the pair. This is why twisted pair cable is advised for differential data transmission.

1.3 Transmission lines

Electronics is not a homogeneous discipline. It tends to divide into set areas: analogue, digital, power, RF and microwave. This is a pragmatic division because different mathematical tools are used for these different areas and it is rare for any one designer to be proficient in all or even most of them. Unhappily for the designer, nature knows nothing of these civilised distinctions; all electrons follow the same physical laws regardless of who observes them and regardless of their speed.

When signal frequencies are low, it is possible to imagine that circuit operation is constrained by the laws of circuit theory: Thévenin, Kirchhoff et al. This is not actually true. Electrons do not read circuit diagrams, and they operate according to the rather grander and more universal laws of Electromagnetic Field Theory, but the difference at low frequencies is so slight that circuit-theoretical predictions are indistinguishable from the real thing. Circuit theory serves electronic engineers well.

As the speed of circuit operation rises, though, it breaks down. It is not that electrons change their behaviour at higher frequencies; there is no cut-off point beyond which everything is different. It is simply that the predictions of circuit theory diverge from those of Electromagnetic Field theory, and the latter, having the backing of nature, wins. One of the consequences of this victory is that perfectly ordinary lengths of wire and cable magically turn into transmission lines.

Transmission line effects

There is no straightforward answer to the question “when do I have to start considering transmission line properties?” The best response is, when the effects become important to you. One of the simplest electrical laws is that which relates frequency, wavelength and the speed of light:

\[ \lambda = 3 \cdot 10^8 / f \]

which is modified because of the reduction in velocity of propagation when a (lossless) dielectric medium is involved by the relative permittivity or dielectric constant of the medium,

\[ \lambda_d = \lambda / \sqrt{\varepsilon_r} \]

One rule of thumb is that a cable should be considered as a transmission line when the wavelength of the highest frequency carried is less than ten times its length. You may be embarrassed by transmission line effects at lengths of one fortieth the wavelength or
1. Side-by-side parallel strip

\[ Z_o = \frac{120}{\sqrt{\varepsilon_r}} \cdot \ln \left\{ \frac{h}{w} + \sqrt{\left(\frac{h}{w}\right)^2 - 1} \right\} \]

2. Face-to-face parallel strip

\[ Z_o = \begin{cases} \frac{377}{\sqrt{\varepsilon_r}} \cdot \frac{h}{w} & \text{if } h > 3t, w >> h \\ \frac{120}{\sqrt{\varepsilon_r}} \cdot \ln \frac{4h}{w} & \text{if } h >> w \end{cases} \]

3. Parallel wire

\[ Z_o = \begin{cases} \frac{120}{\sqrt{\varepsilon_r}} \cdot \ln \left\{ \frac{h}{d} + \sqrt{\left(\frac{h}{d}\right)^2 - 1} \right\} & \text{if } d << h \\ \frac{120}{\sqrt{\varepsilon_r}} \cdot \ln \frac{2h}{d} & \text{if } d >> h \end{cases} \]

(Zo of typical pvc-insulated pairs and twisted pairs is around 100Ω)

4. Wire parallel to infinite plate

\[ Z_o = \begin{cases} \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \left\{ \frac{2h}{d} + \sqrt{\left(\frac{2h}{d}\right)^2 - 1} \right\} & \text{if } d << h \\ \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \frac{4h}{w} & \text{if } h > 3w \end{cases} \]

5. Strip parallel to infinite plate

\[ Z_o = \begin{cases} \frac{377}{\sqrt{\varepsilon_r}} \cdot \frac{h}{w} & \text{if } w > 3h \\ \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \frac{8h}{w} & \text{if } h > 3w \end{cases} \]

6. Coaxial

\[ Z_o = \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \left( \frac{D}{d} \right) \]

<table>
<thead>
<tr>
<th>Dielectric constants of various materials</th>
<th>( \varepsilon_r )</th>
<th>Velocity factor ( \left(1/\sqrt{\varepsilon_r}\right) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Air</td>
<td>1.0</td>
<td>1.0</td>
</tr>
<tr>
<td>Polythene/Polyethylene</td>
<td>2.3</td>
<td>0.66</td>
</tr>
<tr>
<td>PTFE</td>
<td>2.1</td>
<td>0.69</td>
</tr>
<tr>
<td>Silicone Rubber</td>
<td>3.1</td>
<td>0.57</td>
</tr>
<tr>
<td>FR4 Fibreglass PCB</td>
<td>4.5 (typ)</td>
<td>0.47</td>
</tr>
<tr>
<td>PVC</td>
<td>5.0</td>
<td>0.45</td>
</tr>
</tbody>
</table>

Table 1.9 Characteristic impedance, geometry and dielectric constants
less if you are working with precision high-speed signals, or you may not care until the
length reaches a quarter wavelength – though by then you will certainly be getting some
odd results.

**Critical lengths for pulses**

If as a digital engineer you work in terms of rise times rather than frequency, then a
roughly equivalent rule of thumb is that if the shortest rise time is less than three times
the travelling time along the length of the cable you should be thinking in terms of
transmission lines. Thus for a rise time of 10ns in coax with a velocity factor \( \frac{1}{\sqrt{\varepsilon_r}} \) of
0.66 the critical length will be two thirds of a metre.

### 1.3.1 Characteristic impedance

Characteristic impedance \( Z_o \) is the most important parameter for any transmission
line. It is a function of geometry as well as materials and it is a dynamic value
independent of line length; you can’t measure it with a multimeter. It is related to the
c conventional distributed circuit parameters of the cable or conductors by

\[
Z_o = \sqrt{\frac{(R + j\omega L)}{(G + j\omega C)}}
\]

where  
- \( R \) is the series resistance per unit length (\( \Omega/m \))
- \( L \) is the series inductance (\( H/m \))
- \( G \) is the shunt conductance (\( mho/m \))
- \( C \) is the shunt capacitance (\( F/m \))

\( L \) and \( C \) are related to the velocity factor by

\[
\text{velocity of propagation} = \frac{1}{\sqrt{LC}} = 3 \cdot 10^8 / \sqrt{\varepsilon_r}
\]

For an ideal, lossless line \( R = G = 0 \) and \( Z_o \) reduces to \( \sqrt{L/C} \). Practical lines have some
losses which attenuate the signal, and these are quantified as an attenuation factor for a
specified length and frequency (Table 1.7 shows these for coaxial cables). Table 1.7
summarises the approximate characteristic impedances for various geometries, along
with velocity factors of some common dielectric materials. The value 377 (120\(\pi\)) crops
up several times: it is a significant number in electromagnetism, being the *impedance
of free space* (in ohms), which relates electric and magnetic fields in free-field
conditions.

Driving a signal down a transmission line provides an important exception to the
general rule of circuit theory (for voltage drives) that the driving source impedance
should be low while the receiving load impedance should be high. When sent down a
transmission line, the signal is only received undistorted if both source and load
impedances are the same as the line’s characteristic impedance. This is said to be the
**matched** condition. It is easiest to consider the effects of matching and mismatching in
two parts: in the time domain for digital applications and in the frequency domain for
analogue radio frequency applications.

### 1.3.2 Time domain

Imagine a step waveform being launched into a transmission line from a generator
which is matched to the line’s characteristic impedance \( Z_o \). We can view the waveform
at each end of the line and, because of the finite velocity of propagation down the line,
the two waveforms will be different. The results for three different cases of open,
matched and short line terminating impedance (these are the easily-visualised special
cases) are shown in Figure 1.26. If you have a reasonably fast pulse generator, a wide bandwidth oscilloscope and a length of coax cable you can perform this experiment on the bench yourself in five minutes.

A matched transmission line is actually a simple form of delay line, with delays of the order of tens of nanoseconds achievable from practical lengths. Discrete-component delay lines are smaller but work on the same principle, with the distributed L and C values being replaced by actual components.

In all cases the long-term result is as would be expected from conventional circuit theory: an open circuit results in \( V_p \), a short circuit results in zero and anything in between results in the output being divided by the potential divider \( \frac{Z_L}{Z_{out} + Z_L} \), giving \( V_p/2 \) for the matched case. While the edge is in transit the driving waveform is different.

**Forward and reflected waves**

Transmission line theory explains the results in terms of a forward and a reflected wave, the two components summing at each end to satisfy the boundary conditions: zero current for an open circuit, zero voltage for a short. Thus in the short-circuit case, the forward wave of amplitude \( V_p/2 \) generates a reflected wave of amplitude \( -V_p/2 \) when it reaches the short, which returns to the driving end and sums with the already-existing \( V_p/2 \) to give zero. In the general case, the ratio of reflected to forward wave amplitude is

\[
\frac{V_r}{V_i} = \frac{Z - Z_o}{Z + Z_o}
\]

This explanation is most useful when you want to consider mismatches at both ends. Forward and reflected waves are then continually bounced off each mismatched end. Take as another example a drive impedance of \( Z_o/2 \) and an open-circuit load, which is a very crude approximation to an HCMOS logic buffer driving an unterminated HCMOS input. This is shown in Figure 1.27.

**Ringing**

The reflected wave from the open circuit end now gets reflected in turn from the mis-
The Circuit Designer’s Companion

matched driver end with a lower amplitude, which is reflected back by the open circuit which gets reflected again from the driver with a lower amplitude... Eventually the reflections die away and equilibrium is reached. The waveforms at both ends show considerable “ringing”. If you work with digital circuits you will be familiar with ringing if you have ever observed your signals over a few inches of pc track with a fast oscilloscope. The amplitude of the ringing depends entirely on the degree of mismatch between the various impedances, which are complex and for practical purposes essentially unknowable, and the period of the ringing depends on the transit time from driver to termination and hence on line length. A typical ringing frequency for a 0.6mm wide track over a ground plane on 1.6mm epoxy-glass pcb is 35MHz divided by the line length in metres.

The Bergeron diagram

An accurate determination of the amplitude of the reflections at both ends of a transmission line can be made using a Bergeron diagram. This shows the characteristic impedance of a transmission line as a series of load lines on the input and output characteristics of the line driver and receiver. Each load line originates from the point at which the previous load line intersects the appropriate input/output characteristic. To properly use the Bergeron diagram, you need to know the device characteristics both within and outside the supply rail voltage levels, since ringing carries the signal line voltage outside these points. Many manufacturers of high-speed logic ICs detail its use in their application notes.

Ringing in digital circuits is always undesirable since it leads to spurious switching, but it can be tolerated if the amplitudes involved are within the logic family’s noise immunity band, or if the transit times are faster than its response speed. In fact the idealised example in Figure 1.27 shows a step edge which is unrealistic, as practical rise
times will damp the response. The only way to avoid it completely is to consider every interconnection as a transmission line, and to terminate each end with its correct characteristic impedance. Very fast circuits are designed in exactly this way; designers of slower circuits will only meet the problem in severe form when driving long cables.

The uses of mismatching

Mismatching is not always bad. For instance, a very fast, stable pulse generator can be built by feeding a fast risetime edge into a length of transmission line shorted at the far end (Figure 1.28), and taking the output from the input to the line. A 1m length of coax with velocity factor 0.66 will give a 10ns pulse.

![Pulse generation with a shorted transmission line](image)

**Figure 1.28** Pulse generation with a shorted transmission line

### 1.3.3 Frequency domain

If you are more interested in radio frequency signals than in digital edges you want to know what a transmission line does in the frequency domain. Consider the transmission line of Figure 1.26 being fed from a continuous sine-wave generator of frequency \( f \) and matched to the line’s \( Z_0 \). Again, the energy can be thought of as a wave propagating along the line until it reaches the load; if the load impedance is matched to \( Z_0 \) then there is no reflection and all the power is transferred to the load.

If the load is mismatched then a portion of the incident power is reflected back down the line, exactly like an applied pulse edge. A short or open circuit reflects all the power back. But the signal that is reflected is a continuous wave, not a pulse; so the voltage and current at any point along the line is the vector sum of the voltages and currents of the forward and reflected waves, and depends on their relative amplitudes and phases. The voltage and current distribution down the length of the line forms a so-called “standing wave”. The standing wave patterns for four conditions of line termination are shown in Figure 1.29. You can verify this experimentally with a length of fairly leaky coax and a “sniffer” probe, connected to an RF voltmeter, held close to and moved along the coax.

**Standing wave distribution vs. frequency**

Note that the standing wave distribution depends on the wavelength of the applied signal and hence on its frequency. Standing waves at one frequency along a given length of line will differ from those at another. The standing wave pattern repeats itself at multiples of \( \lambda/2 \) along the line. The amplitude of the standing wave depends on the degree of mismatch, which is represented by the reflection coefficient \( \Gamma \), the ratio of reflected current or voltage to incident current or voltage. Standing wave ratio (s.w.r.) is the ratio of maximum to minimum values of the standing wave and is given by

\[
\text{s.w.r.} = \frac{1 + |\Gamma|}{1 - |\Gamma|} = \frac{R_L}{Z_0} \text{ for a purely resistive termination}
\]

Thus an s.w.r. of 1:1 describes a perfectly matched line; infinite s.w.r. describes a line
terminated in a short or open circuit. The generator source impedance has no effect on the s.w.r. It depends only on the nature of the load at the far end.

**Impedance transformation**

The variation of voltage and current along a mismatched transmission line is of great interest to the designers and operators of radio transmitters because it affects the efficiency of power transfer from the transmitter, through the feeder line to the antenna. It is also useful to high-frequency circuit designers as a means of making impedance transformations. Remembering that impedance is voltage divided by current, you can see from Figure 1.29 that the impedance at any given point along the line varies considerably depending on the distance from the termination. For each quarter wavelength, the impedance varies from minimum to maximum. In fact, for a quarter-wave transmission line transformer, the impedance transformation is given by

\[ Z_{in} = \frac{Z_o^2}{Z_L} \]

This useful property is of course frequency dependent; it only occurs at \( \lambda/4 \) and multiples thereof. If the frequency is changed then the line length departs from \( \lambda/4 \) and \( Z_{in} \) becomes reactive. A related property is that at even multiples of \( \lambda/4 \) (equivalent to saying any multiple of \( \lambda/2 \)) the original load impedance is regained, whatever the value

**Figure 1.29** Standing waves along a transmission line

**Figure 1.30** Quarter-wave transformer
of $Z_0$. Thus a shorted line will have a virtually zero impedance $\lambda/2$ away from the short, which property can be used to create a distributed tuned circuit.

**Lossy lines**

The preceding discussion assumes zero-loss lines, which in practice are unrealisable. For short line lengths the losses are usually insignificant – see Table 1.7 for typical coax losses. Note that these are quoted for the matched condition. If the line is operated with standing waves, the loss is greater than if it is matched because increased voltages and currents are present and the average heat loss is greater for the same power output. The effect of attenuation in long lines is to cause an improvement in s.w.r. towards the generator, since the effect of a mismatch is attenuated in both directions. In the limit, a long cable can make a very good power attenuator!