Chapter 14

Interfaces and filtering

Figure 12.1 on page 298 divided EMC control measures into three levels. This chapter deals with the secondary (interfaces and filtering) level. If your product has no interfaces – let's say it is a hand-held, battery operated device – then you already have a head start when it comes to EMC since only the enclosure port is relevant to you. But for the rest of us, power, signal and data interfaces are a fact of life and they offer a ready route for disturbances into and out of a product. The first thing that gets connected to an interface is a cable, so we need to consider how cables couple these disturbances, and if they are to be screened, how the screen connection needs to be dealt with in practice; and then if the interface is unscreened, how to specify and implement the necessary filtering.

14.1 Cables and connectors

Due to their length, external cables are more efficient at interacting with the electromagnetic environment especially in the HF and VHF range than are enclosures, PCBs or other mechanical structures. Cables, and their connectors which create the interface to the equipment, must be carefully specified. The main purpose of this is to ensure that differential mode signals are prevented from radiating from the cables, and that common mode cable currents are neither impressed on the cable by the signal circuit nor are coupled into the signal circuit from external fields via the cable.

In many cases you will have to use screened cables. Usual exceptions are the mains power cable (provided a mains filter is fitted), and low frequency interfaces which can be properly filtered to provide transient and RF immunity. An unfiltered, unscreened interface will provide a path for external emissions and for undesired inward coupling. The way that the cable screen is terminated at the connector interface is critical in maintaining the screening properties of the cable.

14.1.1 The mode of propagation

Although this subject was covered in section 11.1.5, it is particularly important to appreciate the difference between common mode and differential mode currents on cables since this difference directly affects many of the principles that apply to cable design for EMC.

Differential mode current, I_{DM}, is the current which flows in one direction along one cable conductor and in the reverse direction along another (the signal or power pair). It is normally equal to the signal or power current, and with shielded cable, is not present on the shield. It contributes little to the net radiation because the total loop area formed by the two conductors is small; the two currents tend to cancel each other.

Common mode current, I_{CM}, flows equally in the same direction along all conductors in the cable, including the shield if this is present, and may or may not be related to the signal currents. That part of the signal current which does not return via the cable but leaks out through stray coupling, does appear as a common mode component; this aspect is related to the longitudinal conversion loss of the cable, which is discussed later (section 14.1.9.1). The other major source is the noise voltage developed within the circuit and referred between the point of connection of the cable, and the circuit's ground reference. This is why it is good practice to couple the circuit to ground at the interfaces, which will minimize this noise voltage.

I_{CM} returns via the associated ground network and therefore the loop area involved in the coupling is large and uncontrolled (Figure 14.1). As a result, even a small I_{CM} can result in large emitted signals, or an impinging field causes large levels of I_{CM}. On the other hand, the total differential mode currents I_{DM} sum to zero in the cable and hence they create no net magnetic field around the whole cable.

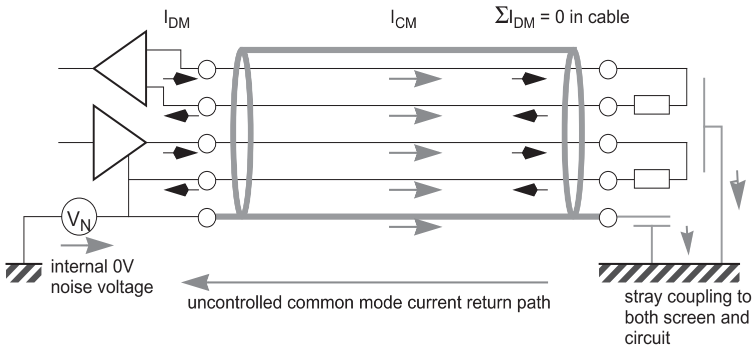


Figure 14.1 The distinction between differential and common mode cable currents

Notice that in a screened cable, we can also distinguish common mode currents that flow on all inner conductors together *and return down the inside of the screen*, from those which flow on the inner lines and outside of the screen together *in the same sense*. The first of these is largely benign since it is essentially still differential within the cable and therefore experiences cancellation outside the cable. This is a major reason for screening multi-conductor cables. The second, as shown by the grey arrows in Figure 14.1, is the dangerous mode.

14.1.2 Cable return currents

All returns should be closely coupled to their signal or power lines, preferably by twisting, as this reduces magnetic field coupling external to the circuit. Returns should never be shared between power and signal lines, and preferably not between individual signal lines either, as this leads to common impedance coupling (section 11.1.1).

It is not intuitively obvious that return currents will necessarily flow in the conductor which is local to the signal wires, when there may be several alternative return paths for them to take – for instance, an external ground connection at each end of the cable. At DC, the return currents are indeed shared between paths only by the ratio of conductor resistances. But as the frequency increases, the mutual inductance of the coupled pair (twisted, coaxial or simply adjacent in the cable bundle) tends to reduce the impedance presented to the return current by its local return compared to other paths, because the enclosed loop area is smallest for this path (Figure 14.2). The effect relies on the fact that the return current is equal in magnitude but opposite in sign to the signal current. Therefore their respective magnetic fields will tend to cancel, the cancellation being complete if the two wires are co-located. If the magnetic fields have cancelled, then by definition the inductance of the path is zero. In practice, co-location of conductors is naturally impossible, but it can be closely approximated by sheathing or twisting the pair together (note that this isn't the main purpose of twisting – see section 14.1.8.1), or perfection can be approached by adopting a coaxial geometry.

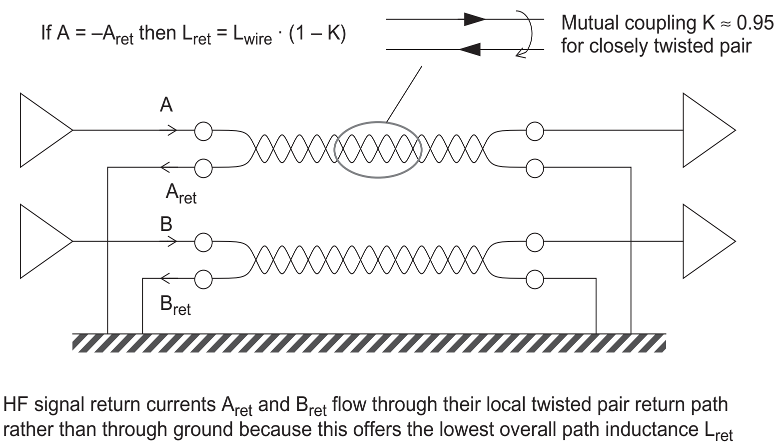


Figure 14.2 Signal return current paths

This effect is also responsible therefore for the magnetic shielding property of coaxial cable, and is the reason why current in a ground plane remains local to its signal track (compare section 12.2.2).

14.1.3 Crosstalk

The issue of crosstalk occupies the never-never land between EMC and signal integrity. Crosstalk within a cable is not a problem for external (inter-system) EMC, but it does have a bearing on *intra*-system EMC, that is, the ability of a system not to interfere with itself. The problem is essentially one of coupling between separate circuits in a cable loom.

Visualize two circuit pairs in a single cable (Figure 14.3). Along the length of the cable there is distributed capacitance between every conductor and each of the other three conductors. Similarly, there is mutual inductance linking every conductor to each of the others. At frequencies where the cable length is much shorter than a wavelength, the L and C can be simplified to a matrix of discrete reactances, but at higher frequencies it is necessary to assign elemental reactances to an infinitesimally short length and then integrate these over the length of the cable.

The mutual capacitance and inductance between the two conductors that form each circuit pair are benign, and determine the characteristic impedance of that pair (from $Z_0 = \sqrt{L/C}$, assuming no losses). In conjunction with the circuit driving and load impedances they will determine the bandwidth capability of the cable/equipment

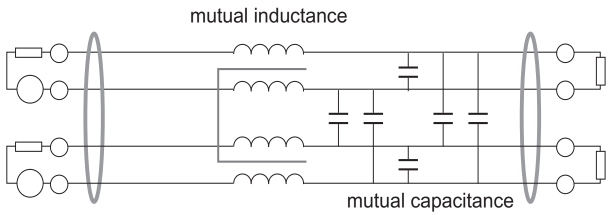


Figure 14.3 Intra-cable crosstalk

system. But the mutual impedances *between the pairs* are undesirable. These result in crosstalk interference between the two circuits.

14.1.3.1 Capacitive crosstalk

The voltages appearing on + and – of pair 1 are coupled by the mutual capacitances onto + and – of pair 2. The amplitude of the induced voltage is determined by the values of the capacitances and the circuit impedances, and the rate of change of source voltage (dv/dt). Balanced circuits, and a balanced cable construction which equalizes the capacitances, will minimize the effective crosstalk since voltages induced on or from the + conductor will be nearly cancelled by those induced on or from the – conductor. Unbalanced circuits with high dv/dt and high impedances will be the most susceptible to capacitive crosstalk. Screening each pair individually (Figure 14.4) will remove the capacitive crosstalk almost entirely, since the mutual capacitances between pairs are eliminated, to be replaced by mutual capacitance from each pair to its screen and mutual capacitance between screens. (Screens without 100% optical coverage, such as braids, will still allow a small amount of capacitance directly between conductors, through the gaps in the screen.) The screens must of course be connected to a fixed potential (which may be system earth, or sometimes circuit 0V); voltages will still be developed longitudinally along the screens as a result of their resistance, and this along with other screen imperfections is then the limiting factor in capacitive crosstalk suppression.

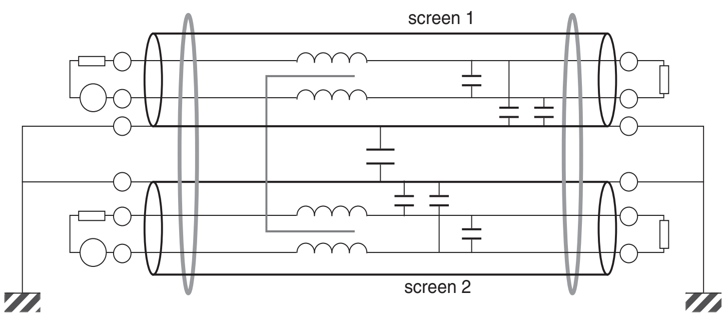


Figure 14.4 Screening against capacitive crosstalk

14.1.3.2 Inductive crosstalk

Currents flowing in each conductor will induce a longitudinal voltage in all other conductors as a result of mutual inductance within the cable. The amplitude of this voltage is proportional to the di/dt of the source current and the mutual inductance linking the conductors; circuit impedances do not affect it. If the mutual inductance from one source conductor to each conductor of the other circuit pair were to be equal, then the same voltage is induced in each and the net effect on that circuit is nil. By itself this is not usually the case; but when the opposite sense contribution from the other source conductor is included, in a cable with good symmetry the total contributions can cancel each other. For this reason inductive crosstalk by itself is rarely as serious a difficulty as capacitive.

14.1.3.3 Distributed crosstalk

At high frequencies the cable must be considered as a distributed structure and the mutual impedances of elemental lengths have to be integrated over the whole length. The phase differences along the length become significant, and the contributions of inductive and capacitive crosstalk result in constructive interference at each end at some frequencies, and destructive interference (nulls in the coupling) at each end at others. What is more, the sense of the wave travelling down the cable becomes significant, and it is necessary to talk of "near-end" crosstalk (NEXT) as distinct and different from "far-end" crosstalk (FEXT). This book does not attempt to go into the detail of distributed crosstalk analysis – see Chapter 10 of Paul [11] or section 4.3 of Tsaliovich [15] for an in-depth treatment.

14.1.4 Cable screens at low frequencies

Optimum screening requires different connection regimes for interference at low frequencies (audio to a few hundred kHz) and at radio frequencies. These two regimes may well be mutually exclusive – the best example being the oft-quoted advice to ground the screen at one end only, which may be correct for low frequencies but is definitely incorrect for high frequencies. Now that you have to meet RF emissions and immunity requirements for the EMC Directive, let alone for practical performance in the field, this advice is obsolete. If grounding the screen at both ends causes problems, then there is most probably a deficiency in the system grounding scheme. To see why this is so, consider the circuit of Figure 14.5.

14.1.4.1 Screen currents and magnetic shielding

An overall screen, grounded only at one end, provides good shielding from capacitively coupled interference (Figure 14.5(a)) but none at all from magnetic fields, which induce a noise voltage in the loop that is formed when both source and load are grounded. (Beware: different principles apply when either source or load is not grounded!) To shield against a magnetic field, *both* ends of the screen must be grounded. This allows an induced current (I_S in Figure 14.5(b)) to flow in the screen which will negate the induction effect in the centre conductor, as described in section 14.1.2. The effect of this current begins to become apparent only above the cable cut-off frequency F_C, which is a function of the screen inductance and resistance and is around 1−2kHz for braided screens or 7−10kHz for aluminium foil screens. Above about five times the cut-off frequency, the voltage induced in the centre conductor remains constant with frequency (Figure 14.5(c)) and therefore the shielding effect continues to increase with frequency.

The same principle applies when you are shielding a conductor to prevent magnetic field *emission*. The return current must flow through the screen, and this will only occur (for a circuit which is grounded at both ends) at frequencies substantially above the

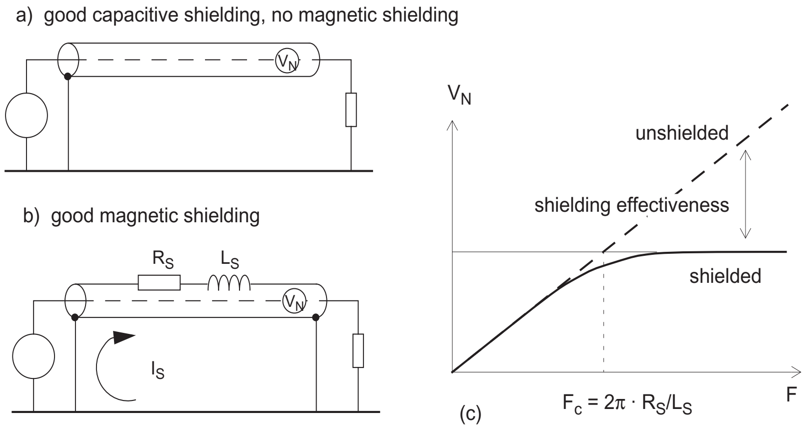


Figure 14.5 Magnetic shielding effectiveness versus screen grounding

shield cut-off frequency. Hence the difficulty in shielding against low frequency magnetic fields, which is explored further in section 15.1.2.

We can note at this point that, if we are only interested in shielding against low frequency *capacitively* induced interference, there is nothing wrong with the single-ended grounding method of Figure 14.5(a) and it is, of course, widely used in instrumentation systems for precisely this purpose.

14.1.4.2 Where to ground the cable screen

There are two problems with grounding the screen at both ends in the circuit of Figure 14.5(b). One is that it becomes a circuit conductor and any voltage dropped across the screen impedance will be injected in series with the signal. Whenever a *circuit* is grounded at both ends, only a limited amount of magnetic shielding is possible because of the large interference currents induced in the screen-ground loop, which develop an interference voltage along the screen. To minimize low frequency magnetic field pick-up, one end of the circuit should be isolated from ground, the circuit loop area should be small, and the screen should not form part of the circuit. You can best achieve this by using shielded twisted pair cable with the screen grounded at only one end. The screen then takes care of capacitive coupling while the twisting (section 14.1.8.1) minimizes magnetic coupling.

For a circuit with an ungrounded source the screen should be grounded at the input common, whereas if the input is floating and the source is grounded then the screen should be grounded to the source common. These arrangements (Figure 14.6) minimize capacitive noise coupling from the screen to the inner conductor(s), since they ensure the minimum voltage differential between the two. Notice though that as the frequency increases, stray capacitance at the nominally ungrounded end reduces the efficiency of either arrangement by allowing undesired ground and screen currents to flow.

Undesired screen currents

The second problem is that if there is a significant voltage difference between the grounds at the two ends of the screen, connecting both ends will encourage the screen to carry a resulting current. This will be limited only by the impedance of the screen and

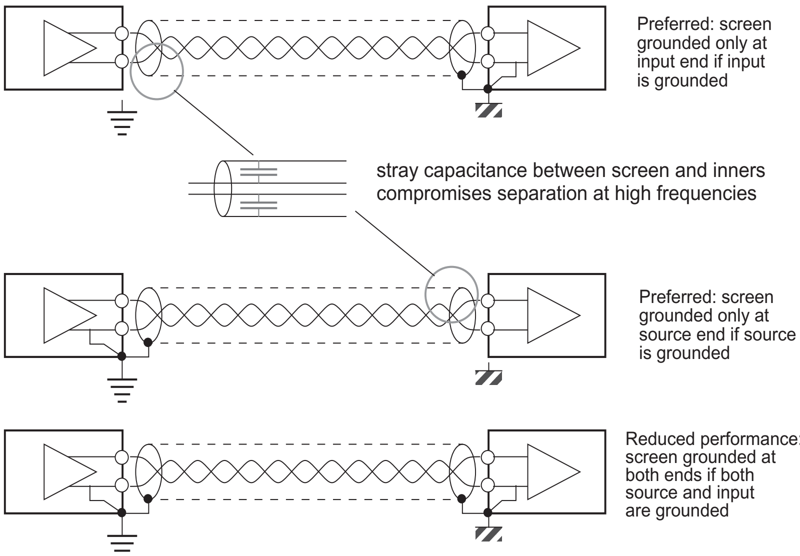


Figure 14.6 Screen grounding arrangements versus circuit configuration

the source impedances of the voltage differential; both of these could be very low, and the current that flows could be enough to damage the cable. Such voltage differentials are not unusual on large sites or between buildings. This is often the practical reason cited for the installation engineer's mantra, don't connect the screen at both ends.

The preferred solution here is to bite the bullet and connect a parallel earth conductor (PEC, as discussed in Chapter 15, and [19] and [173]) along the length of the installed cable so that it takes the brunt of the earth currents that may flow. This solution has historically been resisted by installation engineers on grounds of cost, but if in fact the PEC is a suitably bonded metallic cable tray or conduit, the actual cost can be a lot less than is imagined. This allows the cable screen itself to be bonded at both ends, or not, as EMC circumstances dictate.

But as we are about to see, if you use a screened cable with no other precautions and leave the screen open at one end, your RF shielding at the unconnected end is non-existent. The above discussion applies only for *low frequency* shielding. Not everybody appreciates this state of affairs: even standards writers can be caught out, as the following quote from the NMEA 2000 specification for marine electronic data communications [107] illustrates:

> 2.4.1 It is required that shielded cables be used to facilitate meeting radio frequency interference requirements.
>
> 2.4.1.1 The shield shall not be electrically connected within the interface to the electronic device chassis or ground.

These two statements are mutually exclusive. In the marine industry, it's necessary to be very careful of low frequency and DC currents circulating between different parts of a ship because of the danger of electrochemical corrosion; so the purpose of 2.4.1.1 in the above quote is to prevent these, but that automatically also prevents the intended purpose of 2.4.1.

14.1.5 Cable screens at RF

14.1.5.1 Electrical length

Physical dimensions are not of themselves important in determining the ability of a source to couple to a victim. The electrical dimensions are the significant factor. In free space, the wavelength λ and frequency F are related by the speed of light:

$$\lambda = (c/F), \text{ where } c = 300 \cdot 10^6 \text{ m/s} \quad (14.1)$$

But more generally, the relationship replaces c with v, where v is the velocity of propagation in a medium. v is determined by the permittivity ε and permeability μ of the medium:

$$c = v_0 = (1/\sqrt{\varepsilon_0 \mu_0}) \text{ where } \varepsilon_0 = 1/36\pi \cdot 10^{-9} \text{ F/m}, \mu_0 = 4\pi \cdot 10^{-7} \text{ H/m} \quad (14.2)$$

Other media than free space are characterised by their relative permeability and permittivity. So in such media – particularly insulating dielectrics – the velocity of propagation is slower:

$$v = c/\sqrt{\varepsilon_r \mu_r} \quad (14.3)$$

The significance of this is that the electrical length of a structure (such as a cable) with $\varepsilon_r > 1$ is longer than it would be without the dielectric medium; and so a wavelength in such a structure is shorter than in free space. The same holds for a permeable medium.

A structure is said to be *electrically short* if its electrical dimension is much less than a wavelength – typically λ/10 is taken as a criterion. Conversely, structures approaching λ/4 are *electrically long*. This has implications for how the circuits they carry are treated analytically. Table 14.1 gives data for some typical dielectrics.

Table 14.1 Dielectric properties of materials and their electrical length

	ε_r	Wavelength @ 100MHz (free space = 3m)
Paper	3	1.73
Mica	7	1.13
Neoprene	5.7	1.26
Nylon	3.5	1.60
Perspex	3.45	1.62
Polyethylene	2.3	1.98
Polypropylene	2.2	2.02
Polystyrene	2.6	1.86
PVC	3	1.73
PTFE	2.1	2.07
Quartz	4.43	1.43
Butyl Rubber	2.4	1.94
Silicon	11.7	0.88
Epoxy resin	3.6	1.58

14.1.5.2 Standing waves and skin effect

Once the cable length approaches a quarter wavelength at the frequency of interest†, screen currents due to external fields become a fact of life. An open circuit at one end of the cable becomes transformed into a short circuit a quarter wavelength away, and screen currents flow in a standing wave pattern whether or not there is an external connection (Figure 14.7). The magnitude of the current is related to the characteristic impedance of the transmission line formed by the cable and the ground reference (this behaviour is discussed in section 11.3.1.1). Even below resonant frequencies, stray capacitance can allow screen currents to flow.

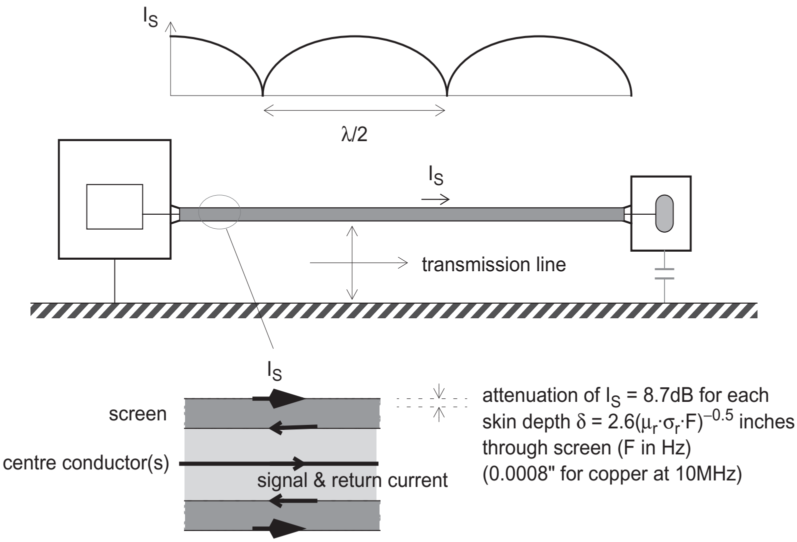


Figure 14.7 The cable screen at RF

Separation of inner and outer screen currents

However, at high frequencies the inner and outer of the screen are isolated by skin effect, which forces currents to remain on the surface of the conductor. Signal currents on the inside of the screen do not couple with interference currents on the outside. The skin depth as shown in Figure 14.7 is the depth at which surface currents are attenuated by 8.7dB and depends on the reciprocal of the square root of the material properties and frequency. At 10MHz in copper it is 0.8 thou or 0.021mm. So a solid screen thickness of 0.25mm would give an attenuation from outer to inner surface of 103dB. Thus multiple grounding of the screen, or grounding at both ends, although it deliberately allows screen currents to flow, does not introduce interference voltages on the inside to the same extent as at low frequencies. This effect is compromised by a braided screen due to its incomplete optical coverage – that is, there are small gaps between braid strands – and because the strands are continuously woven from inside to out and back again. It is also more seriously compromised by the quality of the screen ground connection at either end, as is discussed in section 14.1.7.

† Bearing in mind that the "frequency of interest" extends up to and beyond 1GHz for RF immunity purposes, practically all cables will be longer than $\lambda/4$.

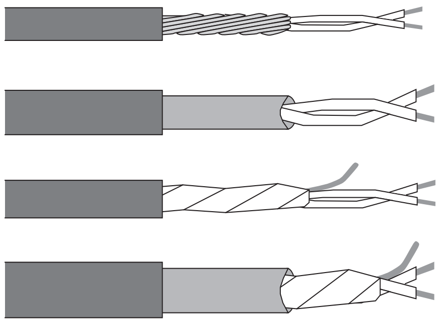


Figure 14.8 Common screen types

14.1.6 Types and performance of cable screens

The performance of cable screens depends on their construction. Figure 14.8 shows some of the more common types of screen available commercially at reasonable cost; for more demanding applications specialized screen constructions such as optimized or multiple braids are available at a premium. Of course, you can also run unscreened cable in shielded conduit, in a separate braided screen or wrap it with screening or permeable material. These options are most useful for systems or installation engineers.

- *Lapped wire* screens consist of wires helically wound onto the cable. They are very flexible, but have poor screening effectiveness and are noticeably inductive at HF, so are restricted to audio use. Steel Wire Armoured (SWA) cable, widely used for mechanical protection of external cable runs, can be regarded as a form of lapped wire and the armouring will provide a degree of screening, but only if it is electrically terminated to ground.
- *Single braid* screens consist of wire woven into a braid to provide a metallic frame covering the cable, offering 80–95% coverage and reasonable HF performance. The braid adds significantly to cable weight and stiffness.
- *Laminated tape or foil* with drain wire provides a full cover but at a fairly high resistance and hence only moderate screening efficiency. Light weight, flexibility, small diameter and low cost are retained. Making a proper termination to this type of screen is difficult; screen currents will tend to flow mainly in the drain wire, making it unsuitable for magnetic screening since the geometry doesn't allow for proper field cancellation, although its capacitive screening is excellent.
- *Composite foil and braid* combines the advantages of both laminated foil and single braid to optimize coverage and high frequency performance.
- *Multiple braid* screens improve the performance of single braids by deliberately separating the inner and outer current flows, and allowing the screens to be dedicated to different (low and high frequency) purposes.
- *Superscreened* cables [71] enhance the multiple braid construction by adding one or more layers of mu-metal over the braid, which extends good screening performance to very low frequencies.

14.1.6.1 Transfer impedance

Cable screening performance is best expressed in terms of transfer impedance. This is denoted by Z_T and is a measure of the voltage induced per unit length on the inner conductor(s) of the cable by an interference current flowing down the cable outer shield, which will vary with frequency and is normally expressed in milliohms per metre. (The same parameter was introduced in section 12.1.2.2 to describe the performance of grounding structures in general.) A perfect screen would not allow any voltage to be induced on the inner conductors and would have a Z_T of zero, but practical screens will couple some energy onto the inner via the screen impedance. At low frequencies it is equal to the DC resistance of the screen.

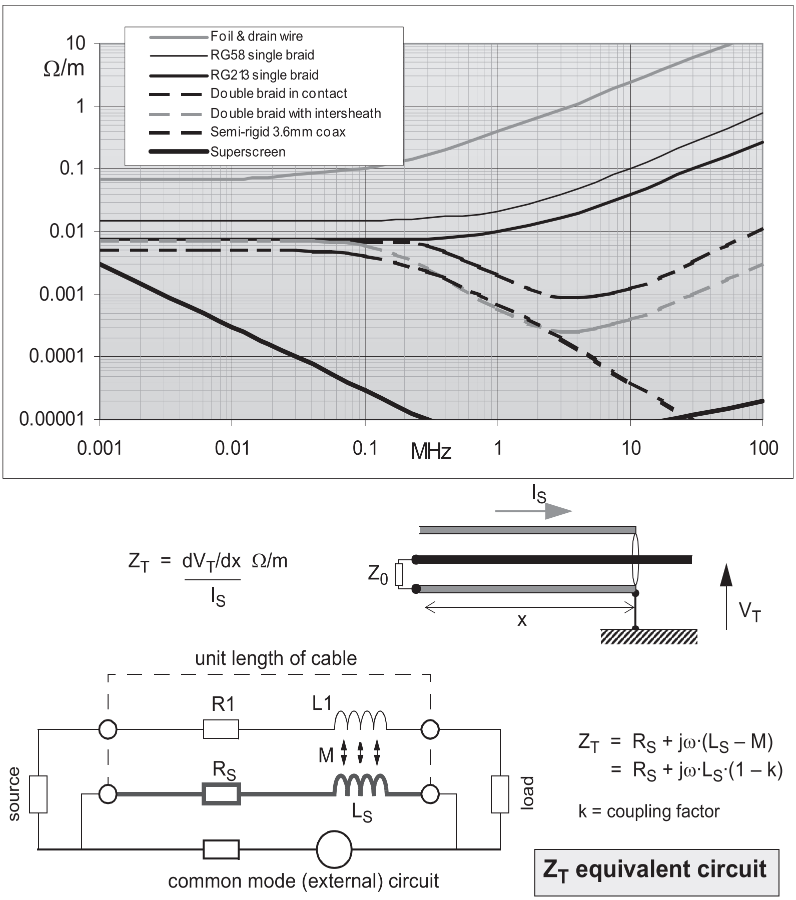


Figure 14.9 Transfer impedance

Figure 14.9 compares Z_T versus frequency for various types of cable screen construction. The initial decrease in Z_T with frequency for the better performance screens is due to the skin effect separating signal currents on the inside of the screen from noise currents on the outside. The subsequent increase is due to field distortion by the holes and weave of the braid, which manifests as leakage inductance. Once the frequency approaches cable resonance then Z_T figures become inapplicable to real situations. This is because the condition of constant screen current I_S per unit length cannot in general be met, because of standing waves. Only when the cable *and its external ground reference* form a transmission line of known characteristic impedance, and this transmission line is properly terminated at each end, will the RF current be constant along the length. A special jig which ensures this condition is one method for measuring Z_T; standardized methods have been published in IEC 62153-4 [177] and EN 50289-1-6 [164], and an introduction to the subject can be found in IEC 62153-4-1.

Note that the inexpensive types have a worsening Z_T with increasing frequency, and that the laminated foil screen is approximately 20dB worse than a single braid, due to its higher resistance and to the field distortion introduced by the drain wire, which carries the major part of the longitudinal screen current. A solid copper screen, on the other hand, just gets better and better with increasing frequency because its skin effect is not compromised. The superscreened cable in the graph has a single tape of mu-metal laid between an inner and an outer braid; the effect of this is to increase the inductive coupling between the inner circuit and the inner braid, which substantially improves the low frequency screening. Coupled with an “optimized” braid weave this gives much better performance across the spectrum. If you want even better performance, [71] states that

> a double sandwich construction of three braids and two magnetic tapes is calculated to fall off at 12dB/octave from the same frequency as the simpler cable... the low frequency measured performance accurately follows that predicted down to a Z_T value of $0.1\mu\Omega$/m and above 100kHz it is not possible to detect any interference penetrating the screen.

Such cables are available as standard designs for use in extreme screening applications; they are not cheap, they are not as flexible as simple braid or foil, there is a weight penalty and you will need exceptionally good connectors. But they work.

14.1.6.2 Z_T versus screening effectiveness

The transfer impedance is a function of the cable only, and therefore is useful for comparing different constructions, but it doesn't directly relate to the screening effectiveness of the cable. The screening effectiveness can be defined as:

$$SE\ (\text{dB}) = 10 \log_{10} (|P_{feed}/P_{rad,max}|) \tag{14.4}$$

where P_{feed} = feed power into cable, $P_{rad,max}$ = power radiated from outside of cable

This is affected by the difference in effective dielectric constants between the outer circuit – outside the screen with respect to the environment – and the inner circuit, within the screen. It is also a function of the characteristic impedances of the two circuits. Therefore to make an exact prediction of the screening performance of a particular cable set, you need to know the detail of the outer environment through which the cable passes. This is rarely controlled, and even more rarely is it exactly known.

When a cable's shielding effectiveness is quoted this is expressed as the ratio of the powers as in equation (14.4), measured using an absorbing clamp. As Goedbloed [5] has pointed out, this value is of little use in EMC calculations; instead, “what is actually done is to use a more or less defined measurement procedure that 'yields a number'; experience has shown that a certain value of this number limits the number of interference complaints to an acceptable level”.

A practical, simple but inexact expression for the conversion [79] is

$$SE\ (dB) = 36 - 20\log_{10}(Z_T) - 20\log_{10}(L) \tag{14.5}$$

where Z_T is in ohms per metre and L is cable length in metres

The number 36dB represents the combination of internal and external impedances of the test system "modified by a factor which expresses the impedance discontinuities seen in a practical system".

Strictly speaking, a cable screen also has a “transfer admittance” which represents the capacitive coupling through a braid with poor coverage. It is also the case that for frequencies of a few kHz and below, magnetic fields will couple directly (*cf* section 14.1.4.1). Nevertheless, for most conditions the direct coupling effects are swamped by coupling via the transfer impedance [71].

14.1.7 Screened cable connections

14.1.7.1 How to ground the cable shield

The over-riding requirement for terminating a cable screen is a connection direct to the metal chassis or enclosure ground which exhibits the lowest possible impedance. This ensures that interference currents on the shield are routed to ground without passing through or coupling to other circuits. The best connection in this respect is one in which the shield is extended up to and makes a solid 360° connection with the ground plane or chassis (Figure 14.10). This is best achieved with a hard-wired cable termination using a conductive gland and ferrule which clamps over the cable screen. A connector will always compromise the quality of the screen-to-chassis bond, but some connectors are very much better than others.

Connector types

Military-style connectors allow for this construction, as do the standard ranges of RF coaxial connectors such as N type or BNC. Of the readily available commercial multi-way connectors, those with a connector shell that is designed to make positive 360° contact with its mate are best. Most commercial connectors compromise on this requirement to some extent; examples are the subminiature D range with dimpled tin-plated shells, or the standard types of USB, DVI and screened RJ. Two- or four-point spring finger connections are inherently less effective than continuous contact.

The importance of the backshell

The cable screen must make 360° contact with a screened conductive backshell which must itself be positively connected to the connector shell. The 360° contact is best offered by an iris or ferrule arrangement although a well-made conductive clamp to the backshell body is an acceptable alternative. Also, for mass-produced cable sets, copper tape soldered both to the cable screen and the connector shell, covering the entire assembly and subsequently overmoulded, is fine if a little labour-intensive. A floating cable clamp, or a backshell which is not tightly mated to the connector shell are not adequate. The backshell itself can be conductively coated plastic rather than solid metal with little loss of performance, because the effect of the 360° termination is felt at the higher frequencies where the skin depth allows the use of very thin conductive surfaces. On the other hand, the backshell is *not* primarily there to provide electric field screening; simply using a metal or conductively coated shell without ensuring a proper connection to it is pointless.

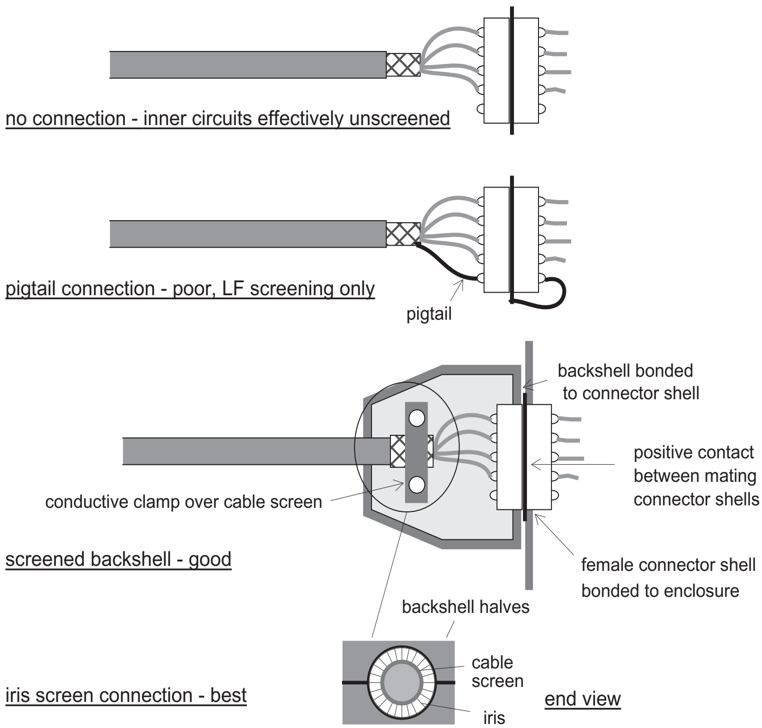


Figure 14.10 Cable screen connection methods at RF

14.1.7.2 The effect of the pigtail

A pigtail connection is one where the screen is brought down to a single wire and perhaps extended through a connector pin to the ground point. Because of its ease of assembly it is very commonly used for connecting the screens of data cables. Indeed, some older data interface standards explicitly provide a pin in the connector for this purpose. Unfortunately, it may be almost as bad as no connection at high frequencies because of the pigtail inductance [91][117]. This can be visualized as being a few tens of nanohenries in series with the cable screen connection (Figure 14.11), which develops a common mode voltage on the screen at the interface as a result of the interference current I_S flowing down the screen. This voltage then couples readily onto the inner conductors, or vice versa, noise voltages on the inner conductors couple readily out onto the screen.

The equivalent transfer impedance of such a connection rises rapidly with increasing frequency until it is dominated by the pigtail inductance, and effectively negates the value of a good HF screened cable. At higher frequencies resonances with the stray capacitances around the interface limit the impedance, but they also make the actual performance of the connection unpredictable and very dependent on construction and movement. If a pigtail connection is unavoidable then it must be as short as possible, and preferably doubled and taken through two pins on opposite ends of the connector so that its inductance is halved. Even with this precaution, it is best to regard

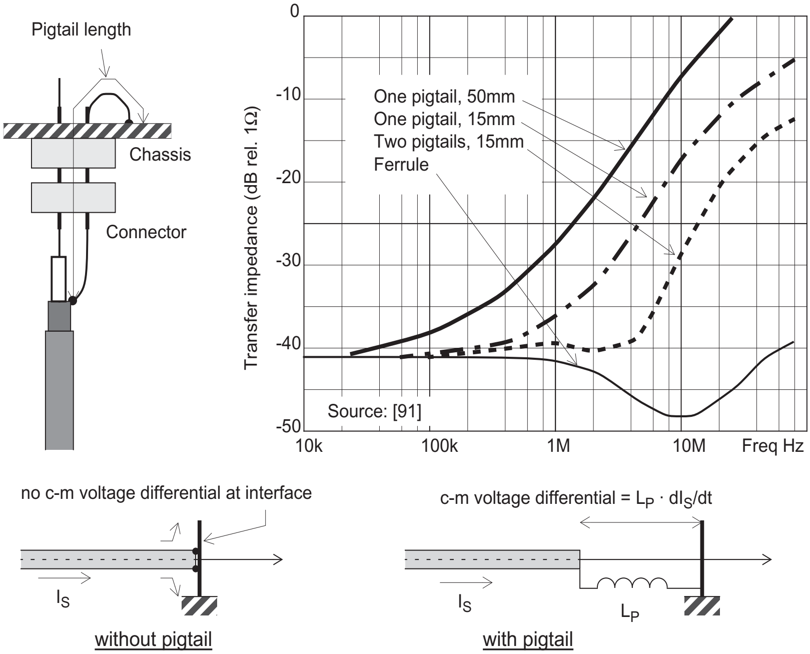


Figure 14.11 The bane of the pigtail

interfaces at which you know that a pigtail is involved as essentially screened only for low frequencies, and to apply good practice HF filtering as well.

The connector shell to chassis link

The effective length of the pigtail extends from the end of the cable screen through the connector and up to the point of the ground plane or chassis connection. The common practice of mounting screened connectors on a PCB with the screening shell taken to ground via a length of track – which in the worst case travels the length of the board – is equivalent to deliberate insertion of a pigtail on the opposite side of the connection. Screened connectors should always be mounted so that their shells are bonded directly to chassis, sometimes even needing a conductive gasket to ensure the bond. A lower performance compromise that is sometimes acceptable is to take the shell to a "quiet ground" (not circuit 0V) on the PCB, which is then taken via pillars to the chassis (see section 12.2.4).

14.1.7.3 Terminating screened cables without connectors

Systems builders are frequently faced with the requirement to bring many field cables into a cabinet and wire these to terminal blocks, whence the signals are routed to the appropriate electronics modules. If the incoming cables are screened, these screens should be terminated to the cabinet frame, and traditionally this has been done via a pigtail to a tag bolted to the frame, perhaps some distance from where the screen ends. For proper RF screening this is not acceptable. The simplest and cheapest method is to bring the cable in under a metal clamp, exactly like a strain relief clamp, which is bolted directly to the frame, and to fold back the cable screen underneath this clamp so that a direct, low-inductance connection is made to the frame. A better method is to use a metal gland ferrule through which the cable passes on its way into the cabinet and to which the screen is terminated. There are other more specialized methods which are used where assured screening is mandatory, for instance in military or Tempest-qualified installations, which involve metal boxes full of copper shavings or compressible gasket modules, either of which is designed to make direct contact to the bare screen as it passes through.

14.1.8 Unscreened cables

You are not always bound to use screened cable to combat EMC problems. The various unscreened types offer major advantages in terms of cost and the welcome freedom from the need to terminate the screen properly. In situations where the cable carries signal circuits that are not in themselves susceptible or emissive, and where common mode cable currents are inoffensive or can be controlled at the interface by other means such as filtering, unscreened cables are quite satisfactory.

14.1.8.1 Twisted pair

Twisted pair is a particularly effective and simple way of reducing both magnetic and capacitive interference pick-up. Twisting the wires tends to ensure a uniform distribution of capacitances to structures outside the cable. 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 provided that each end of the circuit is also balanced.

Twisting is most useful in reducing low frequency magnetic pick-up because it reduces the effective magnetic loop area to almost zero. Each half-twist reverses the direction of induction so, assuming a uniform external field, the wires' interaction with the field is cancelled on successive twists. Effective loop pick-up 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. If the termination area is included in the field, the number of twists per unit length is secondary [47][54]. Clearly, the untwisted termination area or length should be minimized. If as is more usual the field is localized along the cable, performance improves as the number of twists per unit length increases. Inter-pair magnetic crosstalk is reduced by randomizing the twist rate or twisting adjacent pairs in the opposite sense.

The crosstalk or external coupling to any unscreened cable, whether or not it is twisted pair, has both capacitive and inductive components. The twisted pair equivalent circuit (Figure 14.12) depicts the capacitive coupling as a current source I_C from external sources onto each conductor half-twist, while the inductive coupling is a voltage source V_C in series with each conductor.

The effectiveness of twisting a signal/return pair depends on the impedance and the balance or unbalance of the signal circuit. For *unbalanced* circuits, capacitive coupling dominates at high impedances and there is little reduction in overall coupling by twisting. As the circuit impedance drops so capacitive coupling reduces and the inductive part becomes dominant, so that twisting becomes progressively more beneficial. Twisting together power conductors (circuit impedances of a few ohms) is therefore good practice.

Balancing the circuit eliminates (to a first order) the effect of capacitive coupling, since the current sources sum equally in common mode in both halves of the differential

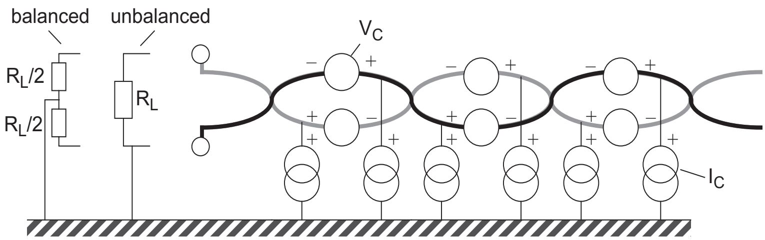


Figure 14.12 Equivalent circuit for coupling into twisted pair

circuit, and allows the crosstalk to be determined purely by residual inductive coupling. This will be sensitive to the uniformity of the twists and of the field through which the cable passes, but is unaffected by whether the circuit is or is not balanced.

14.1.8.2 Ribbon cable

Ribbon is widely used for parallel data transmission within enclosures. It allows mass termination to the connector and is therefore economical. Its route within the enclosure should be carefully considered, because an unshielded cable will allow both magnetic and capacitive coupling to nearby structures. Don't run a ribbon directly past, for instance, a microprocessor board with several VLSI packages or a switching power supply transformer, since it will capture the high frequency noise from these and distribute it around the rest of the box. Uncontrolled routing can mean that different builds of the same design can exhibit quite different EMC performance.

Ribbon should be shielded if it carries high frequency signals and is extended outside a screened enclosure, but you will find that proper termination of the shield is usually incompatible with the use of a mass-termination connector. Cable designs can be obtained with an integral ground plane underneath the conductors, or with full coverage screening [114]. However, the shielding performance figures for ground plane and shielded cables assume a low-inductance termination, which is difficult to achieve in practice; typical terminations via drain wires will worsen this performance, more so at high frequencies.

Ground configuration in ribbon

The performance of a ribbon cable carrying high frequency data is very susceptible to the configuration of its ground return(s). The cheapest configuration is to use one ground conductor for the whole cable (Figure 14.13(a)). This creates a large inductive loop area for the signals on the opposite side of the cable, and crosstalk and ground impedance coupling between signal circuits. It is highly undesirable, but if for other reasons you must have the minimum number of conductors, then at least place the ground conductor in the middle of the cable and place the most aggressive or most sensitive signals adjacent to it.

The preferred configuration is a separate ground return for each signal (b). This gives almost as good performance as a properly terminated ground plane cable, and is very much easier to work with. Crosstalk and common impedance coupling is virtually eliminated. Its disadvantage is the extra size and cost of the ribbon and connectors. An acceptable alternative is configuration (c), two signal conductors per return. This improves cable utilization by 50% over (b) and maintains a small inductive loop area, at the expense of possible crosstalk and ground coupling problems. For pin-limited applications, you need to analyse individual signals to determine where best to place the ground return pins. The optimum configuration of (b) can be improved even more by using twisted pair configured into the ribbon construction.

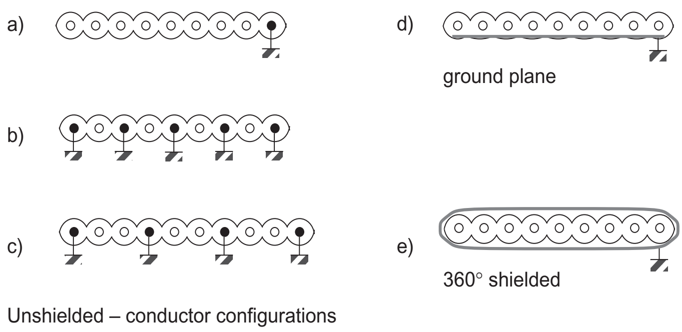


Figure 14.13 Ribbon cable configurations

14.1.8.3 Ground plane flexi

A particularly effective way to carry high frequency signals between boards within a product, and which provides better performance than a ground plane ribbon while being cheaper and easier to use, is the ground plane flexi connector assembly (Figure 14.14). The double-sided flexi has one side dedicated to a ground plane while the other side carries the signal tracks. Alternate pins on the surface-mount connector at each end take the 0V and are via'd through to the ground plane. This low-impedance ground return, very close to the signal tracks, ensures that the minimum of ground noise is developed between the two ends of the circuit.

This construction gives a simple, cheap and effective mass-termination connection system. Flexis can be designed for any physical configuration; some suppliers will do simple designs as standard parts. The method of construction and assembly is exactly like that of a standard PCB except that the substrate is thin and flexible rather than being rigid fibreglass.

The method is very suitable for wideband digital bus connections such as to TFT displays, off-board memories and communication ports, and any connection which carries high frequency clocks or buses.

14.1.8.4 Ferrite loaded cable

Common mode currents in cable screens are responsible for a large proportion of overall radiated emission. A popular technique to reduce these currents is to include a common mode ferrite choke around the cable, typically just before its exit from the enclosure – see section 14.2.2.1. Such a choke effectively increases the HF impedance of the cable to common mode currents without affecting differential mode (signal) currents.

An alternative to discrete chokes is to surround the screen with a continuous coating of flexible ferrite material. This has the advantage of eliminating the need for an extra component or components, and since it is absorptive rather than reflective it reduces discontinuities and hence possible standing waves at high frequencies. This is

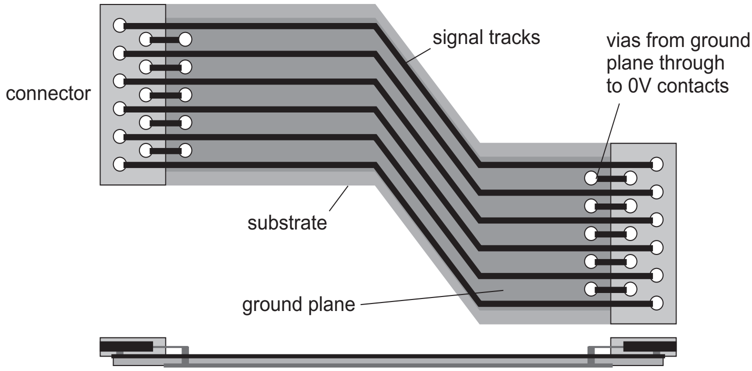


Figure 14.14 The ground plane flexi

particularly useful for minimizing the effect of the antenna cable when making radiated field measurements (section 7.5.2.4). It can also be applied to unscreened cables such as mains leads. Such "ferrite-loaded" cable is unfortunately expensive, not widely available and, like other ferrite applications, is only really effective at very high frequencies. Its use is more suited to one-off or ad hoc applications than as a production item. It can be especially useful when transients or ESD conducted along the cable are troublesome in particular situations.

14.1.9 Structured cabling: UTP versus STP

A common issue for equipment with data communication ports is how it will interface with cabling that has already been installed within a building, or "structured cabling". International and European specifications for such cabling have been published as ISO/IEC 11801 and EN 50173 [159]. These documents classify cables by their performance, as Category 3, 5, 6, 6_A, 7 or 7_A. The most important specifications from the operational perspective are near-end crosstalk, attenuation and characteristic impedance, and these are laid down in the standards. Unshielded twisted pair (UTP) is in principle capable of meeting these requirements, and there is now a large installed base of such cabling in commercial premises to allow the widespread adoption of local area networks.

There is no clear preference as to whether the data cable should be shielded or unshielded. A good quality shielded cable can ensure minimum coupling with its environment, but only if it is properly terminated (see section 14.1.7) and if the shield is maintained unbroken along its length – not an easy matter if it includes joints or patch panels. If this is not the case, then a well-specified UTP cable is likely to be better overall, if the equipment to which it is connected has good common mode rejection.

If you are designing a product with a LAN or telecom data interface then you will need to decide which type of cable to use. If shielded, then you must provide for a correctly terminated shielded connector, and ensure that the installation uses this connector with the proper cable in the right way. If unshielded, then the connector is less important, but the interface must be actively designed for good common mode rejection, which will mean ensuring that the physical layout is balanced and, usually, incorporating a wideband common mode choke (section 14.2.4). It will also be necessary to specify the Longitudinal Conversion Loss (LCL) of allowable connected cables – typically by restricting them to one or other of the ISO/IEC 11801 categories.

14.1.9.1 Longitudinal Conversion Loss

The LCL of a balanced cable system – or indeed any one- or two-port network – is a measure of the mode conversion exhibited by the system, that is the degree to which an inadequately balanced termination will develop an unwanted transverse (differential) signal when excited by a longitudinal (common mode) signal. It is measured as shown

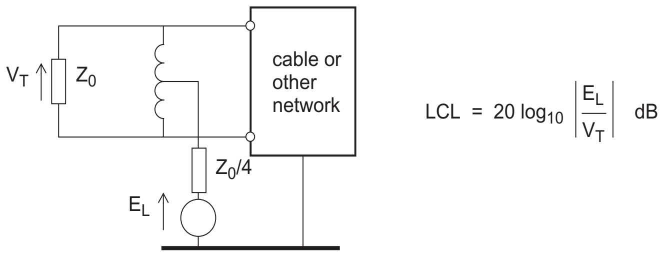


Figure 14.15 Longitudinal conversion loss

in Figure 14.15 [208]. Although this diagram shows a differential mode signal generated by a common mode input, the principle is reciprocal and can be used to describe unwanted common mode signals developed by intended differential signals.

If limits are placed on the common mode emissions at a particular port, then the LCL specified for that port can be used to determine the maximum allowable differential signal amplitude that can be transmitted. Annex E of CISPR 22 (EN 55022) gives the following expression for estimating the common mode current I_{CM} caused by a differential signal voltage U_T:

$$I_{CM}\,(\text{dB}\mu\text{A}) = U_T(\text{dB}\mu\text{V}) - \text{LCL (dB)} - 20 \log_{10} |\{2Z_0 \cdot (Z_{cm} + Z_{ct})/(Z_0 + 4Z_{cm})\}| \quad (14.6)$$

where Z_0 is the signal characteristic impedance, Z_{cm} is the common mode impedance of the item with the worst (lower) LCL and Z_{ct} is that of the item with the better LCL

So for a Class B current limit of 30dBµA, as per CISPR 22 third edition, an LCL for category 3 cable of 50dB, a Z_0 of 100Ω, and $Z_{cm} = Z_{ct} = 25\Omega$, the maximum permissible signal level at any given frequency is 114dBµV or 0.5V. The higher the common mode impedances Z_{cm} and Z_{ct}, the more differential signal can be allowed; but more importantly, the lower the LCL, the greater the level of interference that is created. This has important implications both for testing conducted emissions from telecom ports, as discussed in section 7.2.2.5, and also for using poorly specified cables for passing broadband data – such as for broadband over power lines (section 1.3.5).

14.2 Filtering and suppression

You cannot completely eliminate noise being conducted out of or into equipment along connecting leads. The purpose of filtering is to attenuate such noise to a level either at which it meets a given specification, for exported noise, or at which it does not result in system malfunction, for imported noise. If a filter contains lossy elements, such as a resistor or ferrite component, then the noise energy may be absorbed and dissipated within the filter. If it does not – i.e. if the elements are purely reactive – then the energy is reflected back to its source and must be dissipated elsewhere in the system. This is one of the features which distinguishes EMI filter design from conventional signal filter design, that in the stop-band the filter should be as resistive as possible.

14.2.1 Filter configuration

In EMC work, "filtering" almost always means low-pass filtering. The purpose is normally to attenuate high frequency components, which are assumed to be interference, while passing low frequency ones which are assumed to be wanted. Various simple low-pass configurations are shown in Figure 14.16, and filter circuits are normally made up from a combination of these. The effectiveness of the filter configuration depends on the impedances seen at either end of the filter network.

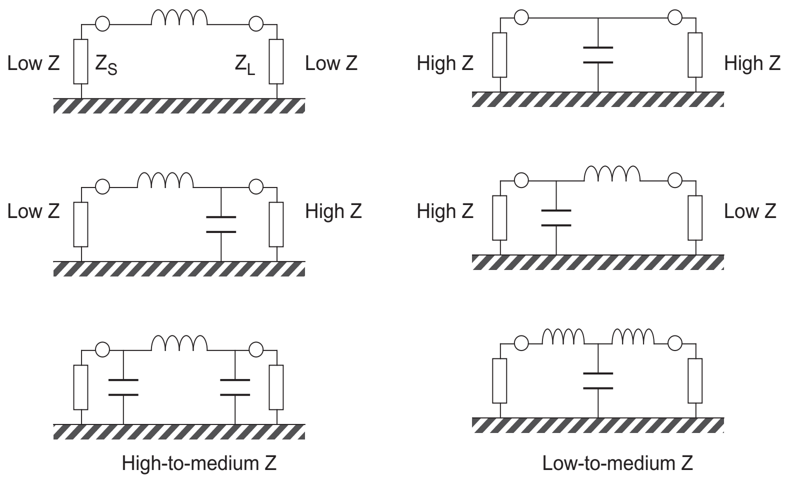


Figure 14.16 Filter configuration versus impedance

14.2.1.1 Source and load impedances

The simple inductor circuit will give good results – better than 40dB attenuation – in a low-impedance circuit but will be useless at high impedances. The simple capacitor will give good results for high impedances but will be useless with low ones. The multi-component filters will give better results provided that they are configured correctly; the capacitor should face a high impedance and the inductor a low one.

Conventionally, filters are specified for terminating impedances of 50Ω at each end because this is convenient for measurement and is an accepted RF standard. If you are sourcing filters specified like this, it is a convenient way to compare the performance of different units but it won't help you much in knowing how much attenuation to expect in the actual circuit. In the real application, Z_S and Z_L are complex and perhaps unknown at the frequencies of interest for suppression. Resonances can be created which may convert an insertion loss into an insertion gain at some frequencies.

Differential mode impedances may be predictable if the components which make up the source and load are well characterized at RF, but common mode impedances such as are presented by cables or the stray reactances of mechanical structures are essentially unpredictable. Practically, cables have been found to have common mode impedances in the region of 100 to 400Ω except at resonance, and a figure of 150Ω is commonly taken as a rule of thumb (see also section 11.3.1.1).

14.2.1.2 Common versus differential mode

The distinction between the two modes of interference coupling has been covered in section 11.1.5. Filters to attenuate these modes must be configured appropriately. The filter shown at a) in Figure 14.17 will attenuate interference which appears between terminals 1 and 2, that is, differentially. It will have no effect on interference which appears in common mode between terminals 1 or 2 and ground, since there is no parallel capacitance to ground, and there is a straight-through path via terminal 2.

The common mode filter shown at b) will attenuate interference appearing between terminals 1 + 2 together, and ground. It may also have a lesser effect on differential mode interference; the differential circuit effectively sees a balanced L-filter with capacitance values of 0.5C and an inductance value equal to the leakage inductance of the common mode choke. This may be replaced by two separate chokes in order to increase differential attenuation, at the expense of limiting the bandwidth of the differential circuit.

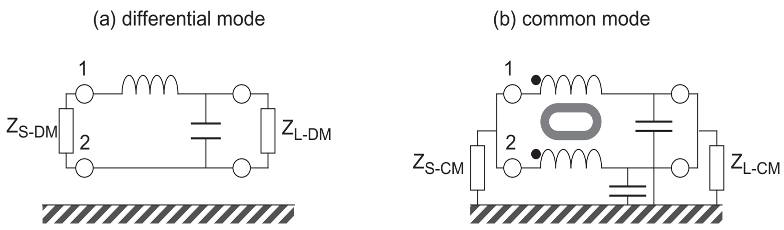


Figure 14.17 Filter modes

14.2.1.3 Parasitics

Filter components, like all others, are imperfect. Inductors have self-capacitance, capacitors have self-inductance. This complicates the equivalent circuit at high frequencies, and means that a typical filter using discrete components will start to lose its performance above a break frequency determined by its parasitics. The larger the components are physically, the lower will be the break frequency. For capacitors, as the frequency increases beyond capacitor self-resonance the impedance of the capacitors in the circuit actually rises, so that the insertion loss begins to fall. This can be countered by using special construction for the capacitors (see the next section). Similarly, inductors have a self-resonant frequency beyond which their impedance starts to fall. Filter circuits using a single choke are normally limited in their performance by the self-resonance of the choke (see Figure 14.18) to 40 or 50dB. Better performance than this requires multiple filter sections.

Capacitors

Multilayer ceramic capacitors (MLCCs) are usually regarded as good for RF purposes, and they are of course easy to source. Small capacitors with short leads (preferably a

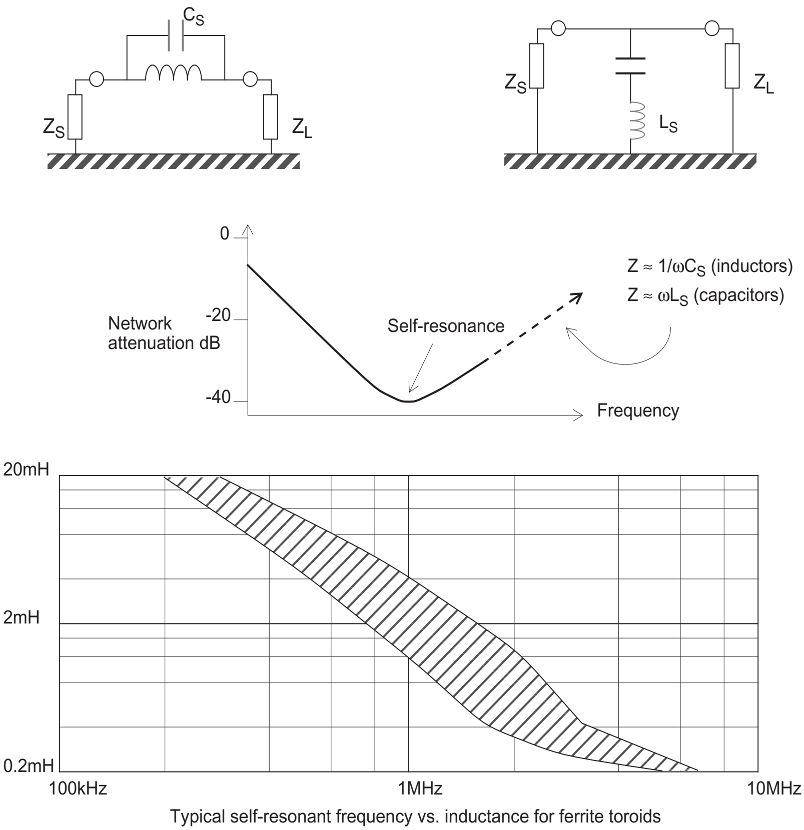


Figure 14.18 Self-resonant effects due to parasitic reactances [55]

chip component) will have the lowest self-inductance. The inductance is made up of a combination of that due to the attached leads and that due to the package; a chip component only exhibits the latter, and their self-resonant frequency can therefore be higher than that of a leaded capacitor of the same value. Some extra inductance is added by the tracking and vias, and tracks to capacitors used for filtering and decoupling should be short and direct, in order not to lose this advantage through additional track inductance. As we have seen in the discussion on decoupling (section 13.1.4.2), smaller packages give better results: for instance widely available chip ceramic data gives a self-resonant frequency for a 10nF component as 65MHz in 0402, 58MHz in 0603, 56MHz in 0805 and 45MHz in 1206. Unless you need a particularly high voltage rating, stay with the smaller parts.

For EMI filtering, lossy dielectrics such as X7R and Y5V are an advantage. Normally, absolute capacitor values, and hence temperature and voltage coefficient of capacitance, are not important. But if you have optimized the capacitor value for a particular performance, be wary of choosing dielectrics such as Y5V or Z5U that show extreme variability with temperature and voltage.

Voltage coefficient of capacitance (VCC)

VCC is a function of the properties of the dielectric material and the voltage stress applied, typically in volts per micron. The effect is negative and non-linear becoming asymptotic towards the limit of dielectric strength. In effect most of the loss occurs long before the part reaches its operational voltage limit; even with de-rating there will be significant capacitance loss. Figure 14.19 shows the effect of voltage stress on X7R dielectric [65]. Manufacturers don't typically quote the actual VCC on their datasheets, but using a conservative approach it would not be unknown for a deterioration of 50% of the marked value when you approach the rated voltage. With worse dielectrics and reducing dielectric thickness, and with voltage de-rating less common since there is always a desire to have the maximum capacitance in the smallest size, greater than 90% loss of capacitance at rated voltage is not unknown.

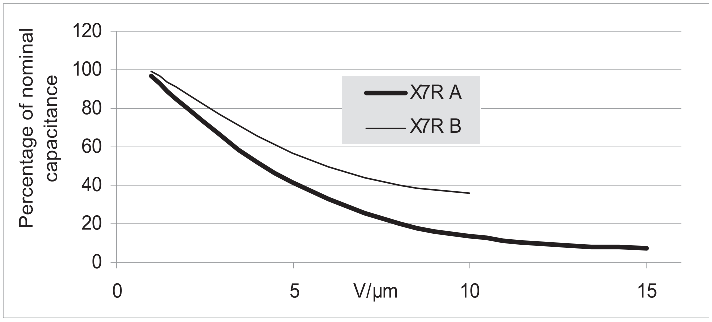


Figure 14.19 Capacitance change versus voltage stress across the dielectric [65]

There are two principal widespread uses for MLCCs that are relevant to EMC: interface filtering, and decoupling. Whether a change in capacitance has a serious effect on either of these applications depends, more than anything, on frequency. If the capacitor is mostly used above self-resonance, as for instance in HF decoupling, then VCC is not so important. But for low frequency filtering, where capacitance is crucial, there is real pressure on the MLCC characteristics, and it will be important to evaluate the effect of VCC on the filter's performance. A necessary consequence is that pre-compliance design EMC testing should be done with worst case DC voltages, low as well as high, present across the relevant components; or, a margin should be incorporated with respect to the LF emissions limits.

Inductors

The more turns N an inductor has, the higher will be its inductance (proportional to N^2) but also the higher its self-capacitance. The number of turns for a given inductance can be reduced by using a high permeability core, but these also exhibit a high dielectric constant which tends to increase the capacitance again, for which reason you should generally use a bobbin on a high-permeability core rather than winding directly onto the core. For minimum self-capacitance the start and finish of a winding should be widely separated; winding in sections on a multi-section bobbin is one way to achieve this. A single layer winding exhibits the lowest self-capacitance. If you have to use more turns than can be accommodated in a single layer, progressive rather than layer winding (see Figure 14.20) will minimize the capacitance.

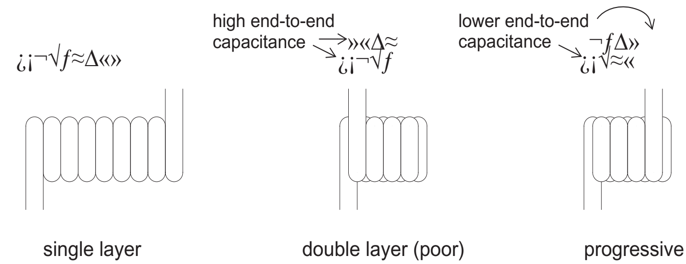


Figure 14.20 Inductor winding techniques

14.2.1.4 Component layout

Lead inductance and stray capacitance degrade filter performance markedly at high frequency. Two common faults in filter applications are not to provide a low-inductance ground connection, and to wire the input and output leads in the same loom or at least close to or passing each other. Low frequency differential mode attenuation will still exist but high frequency common mode attenuation will be minimal.

A poor ground offers a common impedance which rises with frequency (section D.5.2) and couples through HF interference via the filter's local ground path. Common input-output wiring does the same thing through mutual capacitance or inductance, and it is also possible for the "clean" wiring to couple with the unfiltered side through inappropriate routing. The cures (Figure 14.21) are to directly couple the filter's ground terminal to the lowest inductance ground of the equipment, preferably the chassis, and to keep the I/O leads separate, preferably screened from each other. It is best to position the filter so that it straddles the equipment shielding, where this exists.

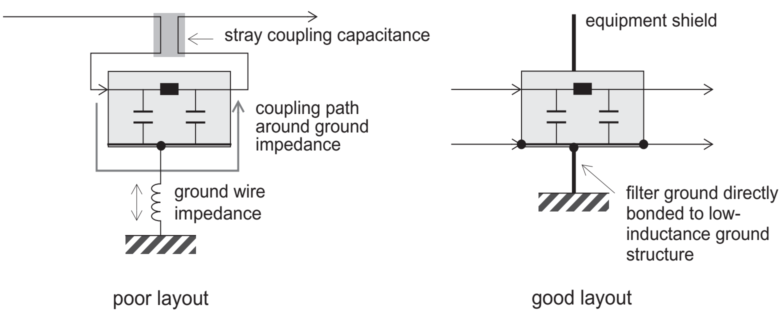


Figure 14.21 The effect of filter layout

Component layout within the filter itself is also important. Input and output components should be well separated from each other for minimum coupling capacitance, while all tracks and in particular the ground track should be short and substantial. It is best to lay out the filter components exactly as they are drawn on the circuit diagram. If there are several inductive components, these should be designed and positioned so that magnetic coupling between them (through leakage flux) is minimized: toroidal cores are helpful for this. Electric field coupling between individual components in different stages of a multi-stage filter should also be minimized, and where this can't be achieved by separation, you may need to implement electric field screens between the stages.

14.2.2 Components

There are a number of specialized components which are intended for EMI filtering applications.

14.2.2.1 Ferrites

One of the most common techniques for reducing both incoming and outgoing RF interference is the application of ferrite sleeves to cables and at interfaces. The attractiveness of the ferrite choke is that it involves no circuit redesign, and often no mechanical redesign either. It is therefore very popular for retro-fit applications. Several manufacturers offer kits which include halved ferrites, which can be applied to cable looms immediately to check for improvement.

As described in section 11.1.1.2, current flowing through a conductor creates a magnetic field around it. Transfer of energy between the current and the magnetic field is effected through the "inductance" of the conductor – for a straight wire the self-inductance is typically 10nH per cm. Placing a magnetically permeable material around the conductor increases the flux density for a given field strength and therefore increases the inductance.

Ferrite is such a material; its permeability is controlled by the exact composition of the different oxides that make it up (ferric, with typically nickel and zinc) and depends heavily on frequency. Also the permeability is complex and has both real and imaginary parts, which translate into both inductive and resistive components of the impedance "inserted" into the line passed through the ferrite (Figure 14.22). The ratio of these components varies with frequency – at the higher frequencies the resistive part dominates (the ferrite can be viewed as a frequency-dependent resistor) and the assembly becomes lossy, so that RF energy is dissipated in the bulk of the material and resonances with stray capacitances are avoided or damped.

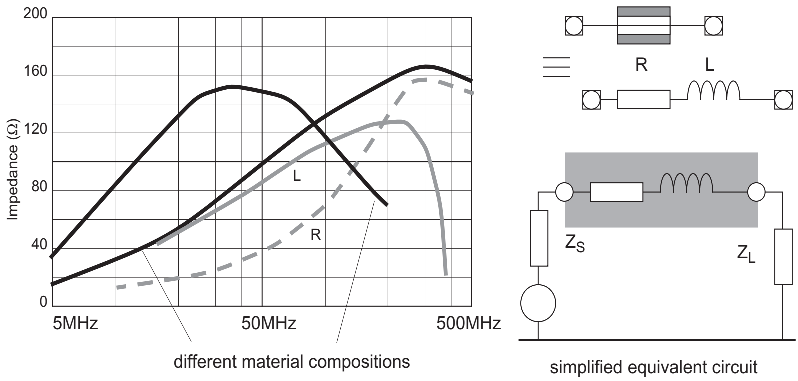


Figure 14.22 Ferrite impedance versus frequency and equivalent circuit

Cable currents

Cables will normally carry signal and return, and/or power and return, conductor pairs. Multiway cables may carry several such pairs. The magnetic field produced by the intended "go" current in each circuit pair is cancelled by the field produced by its equal and opposite "return" current, provided that the two conductors are contained within the cable. Therefore any magnetic material, such as a ferrite sleeve, placed around the whole cable will be invisible to these differential mode currents. This will be true however many pairs there are, as long as the total sum of differential mode currents in the cable harness is zero.

Placing a ferrite around a cable, then, has no effect on the differential mode signals carried within it. On the other hand, common mode currents on a cable *do* generate a net magnetic field around the cable, since by definition these currents are flowing in the same direction in each conductor in the cable. Therefore, a ferrite placed around the cable will affect the resulting magnetic field and will increase the cable's local impedance to these currents. This action is highly desirable, since it provides a means of discriminating between unwanted common mode interference, which should be attenuated, and wanted differential mode signals, which should be unaffected, when these are both within a similar frequency range.

The effect of impedance

As with any other component, when a ferrite is placed in circuit it operates between source and load impedances. A quick glance at the equivalent circuit in Figure 14.22 shows that maximum attenuation due to the simple impedance divider will occur when Z_S and Z_L are low. For example, if Z_S and Z_L are 10 ohms and the ferrite impedance at a given frequency is 100 ohms, the total attenuation (with versus without ferrite) is:

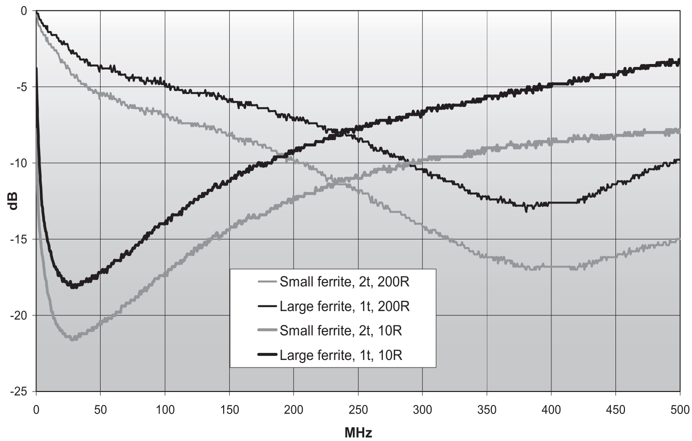
$$A = 20\log_{10}[(10+10)/(10+100+10)] = -15.6\text{dB} \qquad (14.7)$$

but if the circuit impedance is 150 ohms, the attenuation becomes:

$$A = 20\log_{10}[(150+150)/(150+100+150)] = -2.5\text{dB} \qquad (14.8)$$

For cable interfaces, low source impedance means that the ferrite should be applied adjacent to a capacitive filter to ground or to a good screen ground connection. For electrically long cables, the RF common mode load impedance varies with frequency and cable length and termination (see section 11.3.1.1): a quarter wavelength from an open circuit, the impedance is low, a few ohms or tens of ohms; a quarter wavelength from a short circuit, the impedance is high, a few hundred ohms. Since you do not normally know the length and layout of any cable that will be attached to a particular interface, and since the impedance is frequency dependent anyway, it is usual to take an average value for the cable common mode impedance, and 150 ohms has become the norm.

Clip-on ferrite impedances rarely exceed 200–300 ohms, and consequently the attenuation that can be expected from placing a ferrite on an open cable is typically no more than 6–10dB, with 20 dB being achievable at certain frequencies and positions where the cable shows a low impedance. The plots in Figure 14.23 show the actual attenuation for two core sizes at two different resistive circuit impedances.



Small ferrite is Steward 28B0375-100 (9.5mm OD x 14.5mm long) with 2 turns, large ferrite is Steward 28B0562-200 (14mm OD x 28.5mm long) with one turn

Figure 14.23 Ferrite attenuation in different circuit impedances

14.2.2.2 Choosing and using ferrite components

Size and shape

There are two rules of thumb in selecting a ferrite sleeve for highest impedance:

- where you have a choice of shape, longer is better than fatter;
- get the maximum amount of material into your chosen volume that you can afford.

The impedance for a given core material is proportional to the log of the ratio of outside to inside diameter but directly proportional to length. This means that for a certain volume (and weight) of ferrite, best performance will be obtained if the inside diameter fits the cable sheath snugly, and if the sleeve is made as long as possible. A string of sleeves is perfectly acceptable and will increase the impedance pro rata, though the law of diminishing returns sets in with respect to the attenuation.

Number of turns

Inductance can be increased by winding the cable more than one turn around a core; theoretically the inductance is increased proportional to the square of the number of turns, and at the low frequencies this does indeed increase the attenuation. But it is usual to want broadband performance from a ferrite suppressor and at higher frequencies other factors come into play. These are:

- the core geometry already referred to; the optimum shape is long and snugly fitting on the cable, and this does not lend itself to multiple turns;
- inter-turn capacitance, which appears as a parasitic component across the ferrite impedance and reduces the self-resonant frequency of the assembly.

The normal effect of multiple turns is to shift the frequency of maximum attenuation downwards. It will also increase the value of maximum attenuation achieved but not by as much as hoped. The source and load impedances are critical in determining the effect: the lower the impedances, the less the effect of parasitic capacitance.

Capacitance

Because a ferrite material is in fact a ceramic, it has a high permittivity as well as permeability, and hence will increase the capacitance to nearby objects of the cable on which it is placed. This property can be used to advantage especially within equipment. If the ferrite is placed next to a grounded metal surface, such as the chassis, an L-C filter is formed which uses the ferrite both as an inductor and as the dielectric in a distributed capacitor. This will improve the filtering properties compared to using the ferrite in free space. For best effect the cable should be against the ferrite inner surface and the ferrite itself should be flat against the chassis so that no air gaps exist; this can work well with ribbon cable assemblies.

Resistance

A ferrite material is also slightly conductive. This is rarely a disadvantage unless you intend to place the ferrite over a bare conductor, in which case you should be aware of the possible hazards it might bring, such as leakage in high-impedance circuits. Volume resistivities of 10^5 to 10^8 ohm-cm are typical with 10^9 achievable using special materials. Alternatively, specify a ferrite core with an enamel coating.

Saturation

As with other types of ferrite, suppression cores can saturate if a high level of flux is passed through them. At saturation, the magnetic material no longer supports an increase in flux density and the effective permeability drops towards unity, so the attenuation effect of the core disappears. The great virtue of the common mode configuration (Figure 14.28) is that to a first approximation, differential currents cancel and the core is not subjected to the magnetic field they induce, but this only happens if the core is placed around a cable carrying both go and return currents. If you must place a core around a single conductor (such as a power supply lead) or a cable carrying a net low frequency current, be sure that the current flowing does not create a flux which exceeds the core's capability; it is usually necessary to derive this from the generic material curves for a particular core geometry.

14.2.2.3 Ferrite chip components

Surface mount ferrite parts are widely available which translate the high frequency impedance aspects of a ferrite sleeve into a small chip inductor that can easily be used in circuit design. As a generalization, in signal lines it is preferable to use series resistors wherever possible for such impedance; they are cheaper and they include no inductive component which might induce unwanted circuit resonances. But you can't do this with power supply circuits or other circuits which pass appreciable current, since the loss or voltage drop introduced even by a low-value resistor would be unacceptable. In these applications the ferrite chip reigns supreme and experiences wide use.

There are few rules for choosing an exact value; usually the criterion is to achieve an adequately high RF impedance over an adequately wide bandwidth, while using the smallest package size with the highest current rating. In power supply circuits, voltage drop for a given maximum current tends to be the determining factor. Ferrite chips are generally specified for an impedance at 100MHz, with values from a few ohms to well over 1000 ohms being easily available, and 600 ohms as a general purpose value which can be used in many applications. But a glance through the suppliers' data will show that the impedance versus frequency curve can be tailored to give substantial peaks at other frequencies, and anywhere between 100 and 500MHz is typical. Below 100MHz the performance declines but useable impedance can still be obtained down to 30MHz. For this reason the part is used almost as a matter of course in interface circuits for low-speed signals, since it attenuates the HF noise developed by circuit operation elsewhere on the board without any serious effects on the wanted signal. If you use it in conjunction with a properly grounded three-terminal capacitor you can expect the interface to be able to cope with most commercial specifications.

14.2.2.4 Three-terminal capacitors

Any low-pass filter configuration except for the simple inductor uses a capacitor in parallel with the signal path. A perfect capacitor would give an attenuation increasing at a constant 20dB per decade as the frequency increased, but a practical wire-ended capacitor has some inherent lead inductance which in the conventional configuration puts a limit to its high frequency performance as a filter (Figure 14.18). The impedance characteristics show a minimum at some frequency and rise with frequency above this minimum.

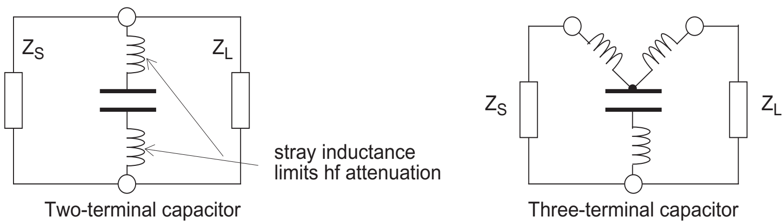


Figure 14.24 The three-terminal capacitor

This lead inductance can be put to some use if the capacitor is given a three-terminal construction (Figure 14.24), separating the input and output connections. The lead inductance now forms a T-filter with the capacitor, greatly improving its high frequency performance. A ferrite bead on each of the upper leads will further enhance the lead inductance and increase the effectiveness of the filter when it is used with a relatively low-impedance source or load. The three-terminal configuration can extend the range of a small ceramic capacitor from below 50MHz to beyond 200MHz, which is particularly useful for interference in the vhf band. To fully benefit from this approach, you must terminate the middle (ground) lead directly to a low-inductance ground such as a ground plane, otherwise the inductance remaining in this connection will defeat the capacitor's purpose. A constraint with the three-terminal construction is that the component now has a current rating as well as a voltage rating.

Surface mount capacitors are also available in a quasi-feedthrough or three-terminal form, as well as with π-, T- or L- configuration including embedded ferrite. In these the device has an elongated shape, and the end terminals carry the signal through the capacitor while the middle body terminal is bonded directly to a ground plane on the PCB. The proper PCB layout for this is shown in Figure 14.25.

14.2.2.5 Feedthrough capacitors

Any leaded capacitor is still limited in effectiveness by the inductance of the connection to the ground point. For the ultimate performance, and especially where penetration of

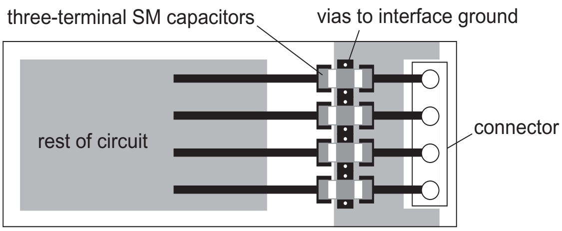


Figure 14.25 Using three-terminal surface mount capacitors

a screened enclosure must be protected at UHF and above then a feedthrough (or leadthrough) construction (Figure 14.26) is essential. Here, the ground connection is made by screwing or soldering the outer body of the capacitor directly to the metal screening or bulkhead. Because the current to ground can spread out for 360° around the central conductor, there is effectively no inductance associated with this terminal and the capacitor performance is maintained well into the GHz region. This performance is compromised if a 360° connection is not made or if the bulkhead is limited in extent. To create a π-section filter, the inductance of the through lead can be increased by separating the ceramic metallization into two parts and incorporating a ferrite bead within the construction. Feedthrough capacitors are available in a wide range of voltage and capacitance ratings but their cost increases with size.

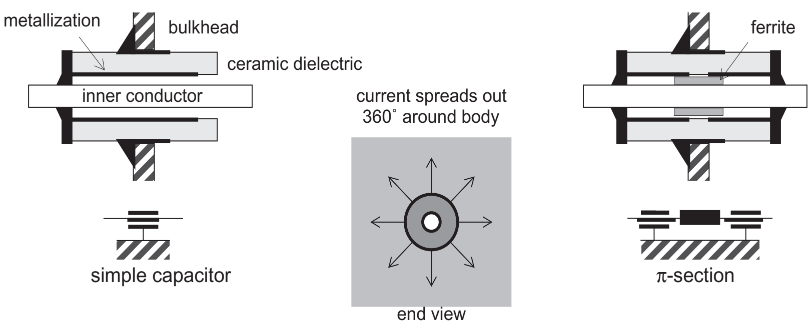


Figure 14.26 The feedthrough capacitor

14.2.3 Mains filters

RFI filters for mains supply inputs have developed as a separate species and are available in many physical and electrical forms from several specialist manufacturers. Some of the reasons for the development and use of block mains filters are:

- mandatory conducted emission standards concentrate on the mains port, hence there is an established market for filter units;
- add-on "fit and forget" filters can be retro-fitted;
- safety approvals for the filter have already been achieved;
- many equipment designers are not familiar with RF filter design.

In fact, the market for mains filters really took off with the introduction of regulations on conducted mains emissions, compounded by the rising popularity of the switchmode power supply. With a switching supply, a mains filter is essential to meet these regulations. A real benefit is that safety approvals needed for all components on the mains side of the equipment have been already dealt with by the filter manufacturer if a single-unit filter is used.

14.2.3.1 Application of mains filters

Merely adding a block filter to a mains input will improve low frequency emissions such as the low harmonics of a switching power supply. But HF emissions (above 1MHz) require attention to the layout of the circuitry around the filter (see section 14.2.1.4). Treating it like any other power supply component will not give good HF attenuation and may actually worsen the coupling, through the addition of spurious resonances and coupling paths. Combined filter and CEE22 inlet connector modules are a good method of ensuring correct layout, providing they are properly bonded to a specified clean ground.

A common layout fault is to wire the mains switch in before the filter, and then to bring the switch wiring all the way across the circuit to the front panel and back. This ensures that the filter components are only exposed to the mains supply while the equipment is switched on, but it also provides a ready-made coupling path via stray induction to the unfiltered wiring. The filter should be the first thing the mains input encounters. If this is impossible, then mount switches, fuses etc. immediately next to the inlet so that unfiltered wiring lengths are minimal, or use a combined inlet/switch/fuse/filter component. Wiring on either side of the filter should be well separated and extend straight out from the connections. If this also is impossible, try to maintain the two sections of wiring at 90° to each other to minimize coupling.

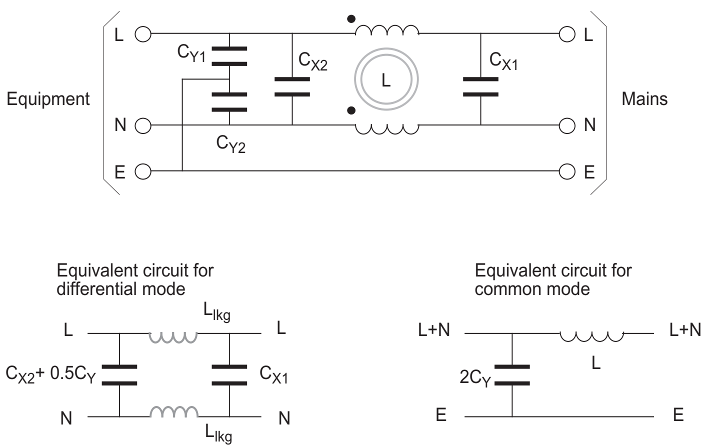


Figure 14.27 Typical mains filter and its equivalent circuit

14.2.3.2 Typical mains filter

A typical filter (Figure 14.27) includes components to block both common mode and differential mode components. The common mode choke L consists of two identical windings on a single high-permeability core, usually toroidal but sometimes of E-I construction, configured so that differential (line-to-neutral) currents cancel each other. This allows high inductance values, typically 1–10mH, in a small volume without fear of choke saturation caused by the mains frequency supply current.

The sense of the windings (Figure 14.28) is such that differential currents, in which the "go" current in one wire is equal and opposite to the "return" current in the other, each create a magnetic flux in the core, but because they are equal and opposite the two fluxes cancel, leaving no net magnetic flux. Thus since the core is invisible the differential mode inductance is very small, being dominated by the residual difference between the windings, known as the leakage inductance.

By contrast the flux from common mode currents in the wires adds in the core, and therefore the full inductance of the choke is presented to common mode signals. To put it another way, the magnetic permeability of the core has maximum effect for common mode currents and negligible effect for differential mode currents.

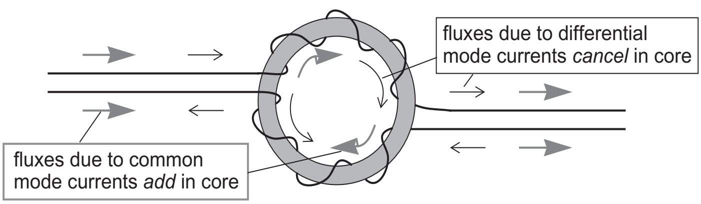


Figure 14.28 The common mode choke

Chokes used in this way are sometimes known as "current-compensated" chokes, since as well as being invisible to differential signals, they can carry large values of low frequency or DC current without fear of core saturation and loss of inductance. Alternatively, by designing in a suitable amount of imbalance, the differential mode inductance can be tailored to provide some DM attenuation as well as CM, at the cost of a reduced current rating.

Common mode capacitors

Capacitors C_{Y1} and C_{Y2} attenuate common mode interference and if C_{X2} is large, have no significant effect on differential mode. The effectiveness of the C_Y capacitors depends very much on the common mode source impedance of the equipment (Figure 14.29). This is usually a function of stray capacitance coupling to earth, which depends critically on the mechanical layout of the circuit and the primary-to-secondary capacitance of the mains isolating transformer, which for large equipment can easily exceed 1000pF. The attenuation offered by the potential divider effect of C_Y may be no more than 15–20dB. The common mode choke is the more effective component, and in cases where C_Y is very severely limited more than one common mode choke may be needed. The LISN impedance shown in Figure 14.29 is for the standard CISPR 16 unit discussed in section 7.2.2.1. Calculation of appropriate component values is covered in Appendix D (section D.7).

Differential mode capacitors

Capacitors C_{X1} and C_{X2} attenuate differential mode only but can have fairly high values, 0.1 to 0.47μF being typical. The main limitation is their size and their interaction with the safety bleed resistance (see next section). Either may be omitted depending on the detailed performance required, remembering that the source and load impedances may be too low for the capacitor to be useful. For example a 0.1µF capacitor has an impedance of about 10Ω at 150kHz, and the differential mode source impedance seen by C_{X2} may be considerably less than this, because of the reservoir capacitor on the DC link for a switching power supply in the hundreds of watts range, so that a C_{X2} of this value would have no effect at the lower end of the frequency range where it may be most needed.

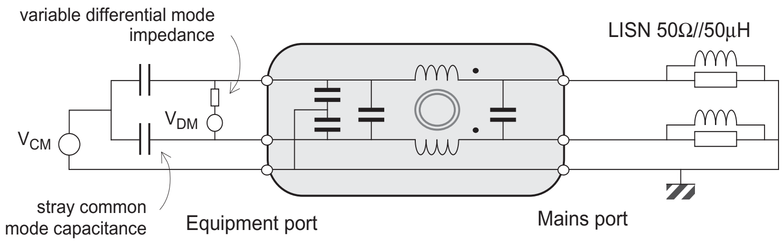


Figure 14.29 Impedances seen by the mains filter

14.2.3.3 Safety considerations

C_{Y1} and C_{Y2} are limited in value by the permissible continuous current which may flow in the safety earth, due to the mains operating voltage impressed across C_{Y1} (or C_{Y2} under certain fault conditions). Leakage current for these capacitors only can easily be calculated from

$$I_{LKG} = V \cdot 2\pi F \cdot C \cdot 1.2 \quad \mu A \tag{14.9}$$

with C in µF and V the maximum supply voltage, and where the factor 1.2 allows for the maximum 20% capacitor tolerance

Values for this current range from 0.25 to 3.5mA depending on the applicable standard, safety class and use of the apparatus (Table 14.2). Special installation conditions apply where the leakage exceeds 3.5mA. Medical equipment has an even lower leakage requirement, typically 0.1mA. Note that this is the *total* leakage current due to the apparatus; if there are other components (such as transient suppressors) which also form a leakage path to earth, the current due to them must be added to that due to C_Y, putting a further constraint on the value of C_Y.

Table 14.2 Allowed earth leakage limits in common safety standards

Standard	Class I portable	Class I stationary	Class II
EN 60335-1, EN 60950-1	0.75mA	3.5mA	0.25mA
EN 61010-1	**Sinusoidal**	**Non-sinusoidal**	**DC**
	0.5mA	0.7mA	2mA
EN 60601-1-1 (medical)	**Type B**	**Type BF**	**Type CF**
	0.5mA (whole equipment)		
Patient leakage	0.1mA	0.1mA	0.01mA

The frequently specified value of 0.75mA leakage current gives a maximum capacitance of around 4nF on each phase for a voltage of 250V at 50Hz, so something less than this value is typical in general-purpose filter units.

Component ratings

Both C_X and C_Y carry mains voltages continuously and must be specifically rated to do this. Failure of C_X will result in a fire hazard, while failure of C_Y will result in both a fire hazard and a potential shock hazard. “X” and “Y” class components to EN 132400 (similar to IEC 60384-14) are designed and marketed specifically for these positions; safety standards mandate their use. EN 132400 has various requirements (Table 14.3) including peak impulse voltage, voltage endurance and flammability.

Table 14.3 EN132400 impulse voltage and endurance ratings

Class	Application	Peak impulse 1.2/50μs before endurance		Endurance, 1000 hr
X1	High pulse, 2.5kV < V_P ≤ 4kV	C ≤ 1μF: 4kV, C > 1μF: (4/√C)kV		1.25 x rated voltage with 1kV AC for 0.1 seconds each hour
X2	General purpose, V_P ≤ 2.5kV	C ≤ 1μF: 2.5kV, C > 1μF: (2.5/√C)kV		
X3	General purpose, V_P ≤ 1.2kV	None		
	Insulation bridged	**Rated voltage**		
Y1	Double or reinforced	≤ 500V	8kV	1.7 x rated voltage with 1kV AC for 0.1 seconds each hour
Y2	Basic or supplementary	≥ 150V ≤ 250V	5kV	
Y3		≥ 150V ≤ 250V	None	
Y4		< 150V	2.5kV	

Bleed resistance

Large values of C_X should be protected with a bleeder resistor in parallel, to prevent a hazardous charge remaining between L and N when the power is removed, if the mains switch is placed after the filter on portable apparatus. The value of resistance is determined by its time constant with the total value of differential capacitance, such that the charge decays adequately soon after the unit is unplugged from its supply (detailed requirements can be found in safety specifications such as IEC 60335/EN 60335). One consequence of this for apparatus which must now comply with energy efficiency legislation is that the bleed resistor’s minimum value may be limited by standby power dissipation requirements; and this in turn can place a limit on the maximum value of X capacitance in the filter.

14.2.3.4 Insertion loss versus impedance

Ready-made filters are almost universally specified between 50Ω source and load impedances. The typical filter configuration outlined above is capable of 40–50dB attenuation up to 30MHz in both common and differential modes. Above 30MHz stray component reactances limit the achievable loss and also make it more difficult to predict behaviour. Below 1MHz the attenuation falls off substantially as the effectiveness of the components reduces.

The 50Ω termination does not reflect the real situation. The mains port HF impedance can be generalized for both common and differential mode by a 50Ω//50µH network as provided by a CISPR 16 LISN (section 7.2.2.1); when the product is tested for compliance, this network will be used anyway. (At high frequencies, it's also necessary to account for the impedance of the mains lead itself.) The equipment port impedance will depend on power rating and on the HF characteristics of the input components such as the rectifier diodes and reservoir capacitor. Differential mode impedance is typically a few ohms for small electronic products, while common mode impedance as discussed above can normally be approximated by a capacitive reactance of 100–1000pF.

Damping

While accurate design of a filter's overall stop- and pass-band characteristic is not normally done, it's often necessary to optimize a design at the low frequency end. This is because the need for attenuation typically extends down into the kHz region, especially for military and aerospace applications, but as the frequency gets lower so the needed capacitance and inductance values get larger, as do the components themselves. The simple L-C filter can easily be characterized at the low frequency end where parasitic components have little effect.

The basic parameter of a single stage series L – parallel C filter is its "cutoff" frequency F_0, determined by $1/(2\pi\sqrt{LC})$ (shown as ω_n in section D.7). The second most important parameter is the critical damping resistance R_0, determined by $\sqrt{(L/C)}$. At frequencies well above F_0, the attenuation of the filter increases (until it is limited by parasitics) at a rate of 40dB per decade. So for an example theoretical attenuation of 80dB at 1MHz, the cutoff frequency needs to be 10kHz. But this frequency can be achieved by any combination of L and C; for instance 1mH and 250nF, but also 0.25mH and 1µF, or 2.5mH and 0.1µF, or 50µH and 5µF, and so on. The actual values chosen will have to reflect the practical and circuit limitations.

But once these are chosen, you have also set R_0. Then the performance near to F_0 becomes a factor to consider, and this is determined by the ratio of R_0 to the source and load impedances. With resistive impedances, the filter will be overdamped if R_0 is higher than the external impedance, so that the attenuation starts at a lower frequency than F_0; but if it is lower, then the filter is underdamped, and this gives a degree of *gain* around F_0 which is almost certainly undesirable. It is therefore necessary to have at least an estimate of the external impedances to get a true picture of the behaviour of the filter across the required frequency range.

One alternative method for measuring filter insertion loss is to use terminating impedances of 0.1Ω/100Ω and reverse. This method is defined in CISPR publication 17, and results in more realistic performance data under some circuit conditions, but not all filter manufacturers provide figures using this test method. If you can assume pure resistive terminating impedances, you can derive the expected insertion loss performance from the published 50Ω data using the equations in section D.7.

The circuit of Figure 14.29 is easy to set up with a circuit analysis program. Provided that you remember to include estimated parasitic reactances (capacitance across the choke coils, inductance in series with capacitors) you can make a fair attempt at designing the mains filter circuit for the optimum required attenuation from known emission levels.

14.2.3.5 Core saturation

Filters are specified for a maximum working RMS current, which is mainly determined by allowable heating in the common mode choke. Capacitor input power supplies have a distinctly non-sinusoidal input current waveform (see the discussion on mains harmonics in section 11.4, and Figure 14.30(a)), with a peak current of between three and ten times the RMS, which represents a high "crest factor". But, for reasons of space efficiency, the ferrite core of a mains filter is typically designed so that the common mode inductance has dropped to 80% of its nominal value at the filter's rated current. To protect the common mode inductance against saturation, this 80% point should occur at the maximum value of the current waveform: that is, the filter should be rated not for the RMS but for the peak current. The magnetic field due to the supply current is compensated inside a common mode ferrite core, but the stray inductance (around 1% of nominal) still leads to saturation at the peak current. The effect of saturation is mainly felt at the bottom end of the spectrum, where a loss of attenuation of more than 10dB can occur.

The problem is made worse when the choke is passing its rated current, because winding losses lead to heating. At higher temperatures, saturation is reached at lower values of flux density, translating to lower attenuation values.

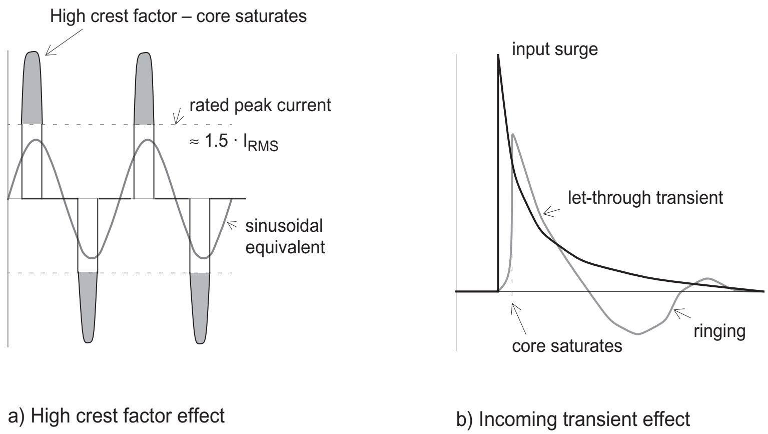


Figure 14.30 Core saturation effects

The core will also saturate when it is presented with a high-voltage, high energy common mode surge, such as a lightning transient on the mains (Figure 14.30(b)). The surge voltage will be let through delayed and with a slower risetime but only slightly attenuated with attendant ringing on the trailing edge. Standard mains filters designed only for attenuating frequency-domain emissions are inadequate to cope with large incoming common mode transients, though some are better than others. Differential mode transients require considerably more energy to saturate the core and these are more satisfactorily suppressed.

14.2.3.6 Extended performance

In some cases the insertion loss offered by the typical configuration won't be adequate. This may be the case when for example a high-power switching supply must meet the most stringent emission limits, or there is excessive coupling of common mode interference, or greater incoming transient immunity is needed. The basic filter design can be extended in a number of ways (Figure 14.31):

- *extra differential line chokes*: these are separate chokes in L and N lines which are not cross-coupled and therefore present a higher impedance to differential mode signals, giving better attenuation in conjunction with C_X. Because they must not saturate at the full AC line current they are much larger and heavier for a given inductance.
- *an earth line choke*: this increases the impedance to common mode currents flowing in the safety earth and may be the only way of dealing with common mode interference, both incoming and outgoing, when C_Y is already at its maximum limit and nothing can be done about the interference at source. Because it is in series with the safety earth its fault current carrying capability must satisfy safety standards. Ensure that it is not short-circuited by any extra earth connection: this is most often provided by a cable to another item of equipment, and makes the earth line choke unusable except in apparatus with only a mains supply connection.
- *transient suppressors*: a device such as a voltage-dependent resistor (VDR) across L and N will clip incoming differential mode surges (see also section 14.2.5). If it is placed at the mains port then it must be rated for the full expected transient energy, but it will prevent the choke from saturating and protect the filter's C_X; if it is placed on the equipment side then it can be substantially downrated since it is protected by the impedance of the filter. Of course, it has no effect on common mode transients.

In addition to these extra techniques the basic filter π-section can be cascaded with further similar sections, perhaps with inter-section screens and feedthroughs to obtain much higher insertion loss. For these levels of performance the filter must be used in conjunction with a well-screened enclosure to prevent high frequency coupling around it.

14.2.4 I/O filtering

If I/O connections carry only low bandwidth signals and low current it is possible to filter them using simple RC low-pass networks (Figure 14.32(a)). The decoupling capacitor must be connected to the clean I/O ground (see section 12.2.4) which may not be the same as circuit 0V.

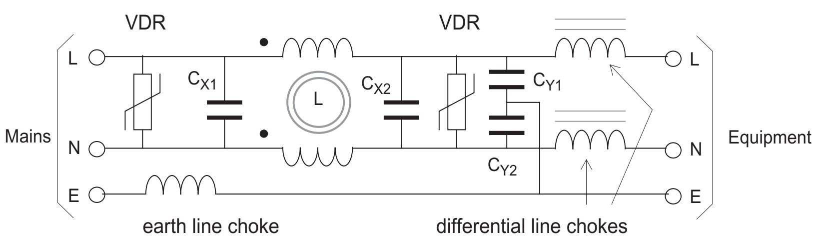


Figure 14.31 Higher-performance mains filter

14.2.4.1 Common mode choke

Adequately high values of R and C may not be possible with high-speed data links or interfaces carrying significant current, but it is possible to attenuate common mode currents entering or leaving the equipment without affecting the signal frequencies by using a discrete common mode choke arrangement. The choke has several identical windings on the same core such that the fields from differential currents cancel each other whereas common mode currents add, in the same fashion as the mains common mode choke described in section 14.2.3.2. Such units are available commercially (sometimes described as "data line filters") or can be custom designed. It is important when designing in a signal line common mode choke that you make sure the differential currents really do cancel in the core: this means that all lines handled by the interface, *including the 0V line* if it is used as a return, must pass through individual windings on a single choke, as shown in Figure 14.32(b).

The exact form of construction will depend on the desired frequency range. Above 100MHz, small multilayer ferrite chip parts are usually adequate. For lower frequencies, multi-turn windings are needed, and stray capacitance across each winding will degrade high frequency attenuation.

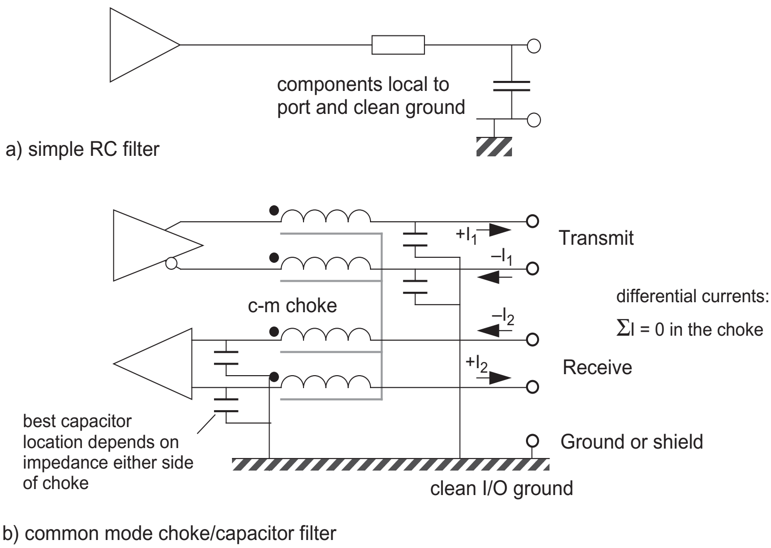


Figure 14.32 I/O filtering techniques

14.2.4.2 Filtered connectors

A convenient way to incorporate both the capacitors and to a lesser extent the inductive components of Figure 14.32 is within the external connector itself. Each pin can be configured as a feedthrough capacitor with a ceramic jacket, its outside metallization connected to a matrix which is grounded directly to the connector shell (Figure 14.33(a)). Thus the inductance of the ground connection is minimized, provided that the connector shell itself is correctly bonded to a clean ground, normally the metal backplate or chassis of the unit. Any series impedance in the ground path not only

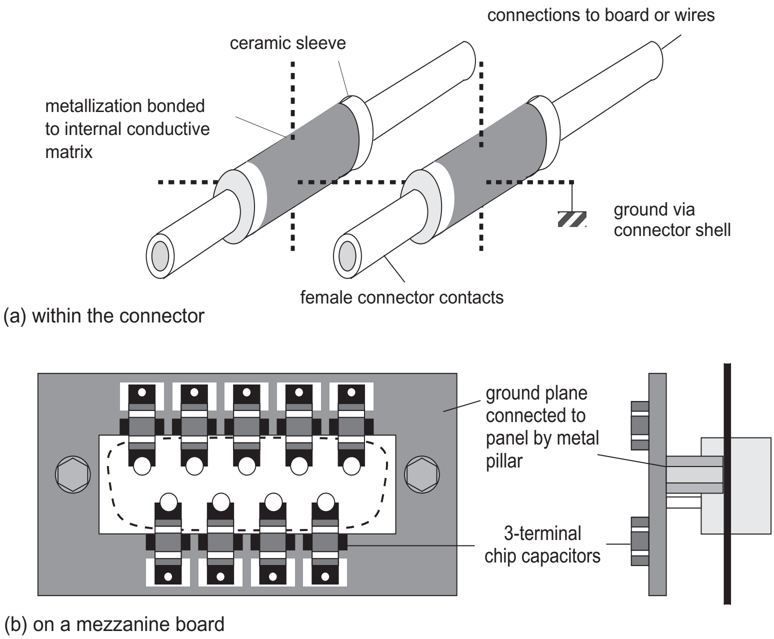


Figure 14.33 Filtered connector pins

degrades the filtering but will also couple signals from one line into another, leading to designed-in crosstalk.

The advantage of this construction is that the insertion loss can extend to over 1GHz, the low frequency loss depending entirely on the actual capacitance (typically 50–2000pF) inserted in parallel with each contact. With some ferrite incorporated as part of the construction, a π-filter can be formed as with the conventional feedthrough (section 14.2.2.5). No extra space for filtering needs to be provided. The filtered connector has obvious attractions for retro-fit purposes, and may frequently solve interface problems at a stroke. You can also obtain ferrite blocks tailored to the pinout dimensions of common multiway connectors, which effectively offer individual choking for each line with a single component; and the "EESeal®" silicone rubber insert for retrofit to various connector types, which has a matrix of embedded multilayer capacitors or transient suppressors which make contact to each pin.

Disadvantages are the significant extra cost over an unfiltered connector; if not all contacts are filtered, or different contacts need different capacitor values, you will need a custom part. Voltage ratings may be barely adequate and reliability may be worsened. A small "piggy-back" or "mezzanine" board of chip capacitors, or preferably three-terminal components mounted immediately next to the connector (Figure 14.33(b)), with their ground connection made via the lowest possible inductance, can be equally effective up to hundreds of MHz, is cheaper, and is inherently customized. Because it has a larger ground loop area than the filtered pin, it may be more vulnerable to high frequency magnetic pickup, a potential issue for military applications.

14.2.4.3 Circuit effects of filtering

Imbalance

When you use any form of capacitive filtering, the circuit must be able to handle the extra capacitance to ground, particularly when filtering an isolated circuit at radio frequencies. Apart from reducing the available circuit signal bandwidth, the RF filter capacitance provides a ready-made AC path to ground for the signal circuit and will seriously degrade the AC isolation, to such an extent that an RF filter may actually increase susceptibility to lower frequency common mode interference. This is a result of the capacitance imbalance between the isolated signal and return lines. It is impossible to assure perfect matching between individual capacitors, and typical tolerances of up to 10% mean that the circuit common mode rejection would be no more than 20dB – far less than could be achieved without filter capacitors.

This may restrict the allowable RF filter capacitance to a few tens of picofarads, and in the extreme with wideband balanced circuits, even a few tens of pF will be too much. In these cases, only a common mode choke and/or isolation is acceptable at the interface. The Ethernet interface is the most commonly encountered example of this.

Voltage rating

Any filter capacitor will have to be chosen for a particular maximum voltage rating. This is easy enough to decide for differential mode capacitors across the circuit terminals, since you will know the peak voltage to be expected in any given circuit. But what about common mode capacitors to chassis? To choose a rating for these you have to know the maximum voltage that is to be expected between the operating circuit and the chassis. This is not always obvious, because it is determined more by system factors which may not be known to the product designer. If the circuit is deliberately isolated from earth then the actual maximum voltage may not be specified anywhere, but it needs to be considered; often it's the peak transient surge amplitude that determines the voltage rating here, which can be more than 1kV. You can get multilayer ceramic capacitors at these levels, but they are large, and don't have high capacitance values. Or, in many cases it is acceptable to bridge circuit and chassis with a transient suppressor, in which case the capacitor voltage rating is limited to the highest clamping voltage.

Circuit stability

As a further hazard, capacitive loading of low frequency analogue amplifier outputs may also push the output stage into instability (see section 13.1.5.3).

14.2.5 Transient suppression

Incoming transients on either mains or signal lines are reduced by non-linear devices: the most common are metal oxide varistors (voltage-dependent resistors, or VDRs), zeners and spark gaps (gas discharge tubes). The device is placed in parallel with the line to be protected (Figure 14.34) and to normal signal or power levels it appears as a high impedance – essentially determined by its self-capacitance and leakage specifications. When a transient which exceeds its breakdown voltage appears, the device changes to a low impedance which diverts current from the transient source away from the protected circuit, limiting the transient voltage (Figure 14.35). It must be sized to withstand the continuous operating voltage of the circuit, with a safety margin, and to be able to absorb the energy from any expected transient.

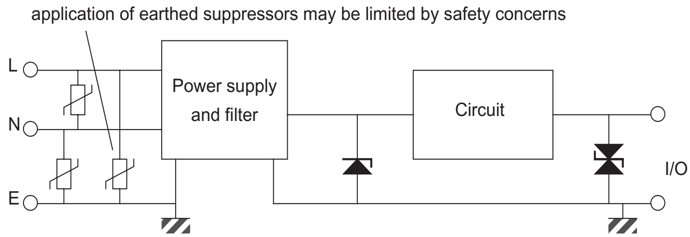


Figure 14.34 Typical locations for transient suppressors

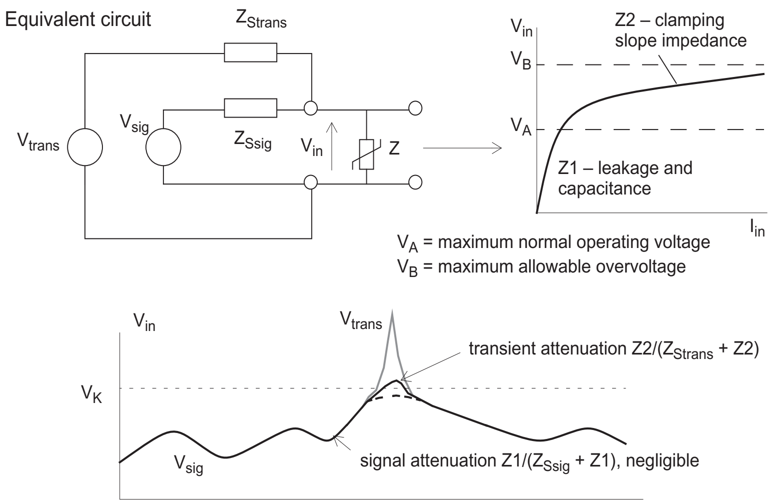


Figure 14.35 The operation of a transient suppressor

The first requirement is fairly simple to design to, although it means that the peak transient clamping voltage (V_B) is usually at least twice the continuous voltage (V_A), and circuits that are protected by the suppressor must be able to withstand this. The second requirement calls for a knowledge of the source impedance Z_{Strans} and probable amplitude of the transients, which is often difficult to predict accurately especially for external connections. This determines the amount of energy which the suppressor will have to absorb. Reference [48] gives details of how to determine the required suppressor characteristics from a knowledge of the circuit parameters, and also suggests design values for the energy requirement for suppressors on AC power supplies. These are summarized in Table 14.4. IEEE C62.41 [229] gives further details of expected transient sources and amplitudes, as well as detailing the test methods recommended to check transient immunity.

Table 14.5 compares the characteristics of the most common varieties of transient suppressor. Variations on the three basic types are available; for instance the ZnO varistor is available in monolithic multilayer form which allows clamp voltages down

Table 14.4 Suggested transient suppressor design parameters for mains supplies [48]

Type of location	Waveform	Amplitude	Energy deposited in a suppressor with clamping voltage of 500V (120V system)	1000V (240V system)
Long branch circuits and outlets	0.5µs/100kHz oscillatory	6kV/200A	0.8J	1.6J
Major feeders and short branch circuits	0.5µs/100kHz oscillatory	6kV/500A	2J	4J
	8/20µs surge	6kV/3kA	40J	80J

Table 14.5 Comparison of transient suppressor types

Device	Leakage	Follow-on current	Clamp voltage	Energy capability	Capacitance	Response time	Cost
ZnO varistor	Moderate	No	Medium	High	High	Medium	Low
Zener	Low	No	Low to medium	Low	Low	Fast	Moderate
Spark gap GDT	Zero	Yes	High ignition, low clamp	High	Very low	Slow	Moderate to high

to 5V, and can also be fabricated with a specified capacitance to offer a varistor-cum-capacitor in a single component.

14.2.5.1 Clamping characteristics and destruction

Some experiments have been performed on various types of varistor and zener suppressors [151] to try and compare their characteristics when stressed with a standard IEC 61000-4-5 surge waveform. A couple of sample waveforms for low-voltage devices are shown in Figure 14.36. The conclusions reached from these experiments were:

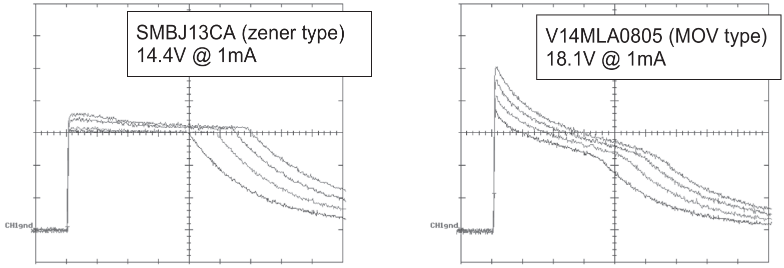


Figure 14.36 Transient suppressor clamping waveforms

- MOV devices have a high slope resistance and a greater peak clamping voltage, but are very robust and can take high energy repeatedly. Even small parts (1206 and smaller) are able to withstand severe surges without destruction, although they may in consequence experience up to three times their rated stand off voltage.
- TVS (Zener avalanche) devices have a low slope resistance and a very flat clamping profile at low energies, but are more easily destroyed by higher energy surges. These would generally be the preferred part if the source impedance is high but the downstream circuit withstand voltage is not, as in signal circuits.
- Ordinary zeners will work but aren't characterized for surges; again, for signal circuits with some series impedance, they may well be adequate.

Degradation

One of the often-quoted disadvantages of varistor devices is that their characteristics are said to degrade with each transient they capture. It’s hard to find authoritative manufacturers’ data on this; to find out if there is anything in this view, one sample of a 1206 device was subjected to 100 surges at its rated current. The device was a V33MLA1206 which has a rated maximum current of 180A; to achieve this a voltage of 450V was applied through a 2.5 ohm source impedance. The surges were applied at a rate of one every 10 seconds; no significant heating effect was observed on the device. Between the first and 100th application there was a roughly 5% increase in the peak clamping voltage. If a particularly stressful environment is expected, you may want to derate the device in anticipation of this.

Destruction

When a device is overstressed to failure, it can fail in any of three ways: open circuit, short circuit (less than a few ohms) or with a degraded clamping voltage [146]. Depending on the circuit to be protected, you may have to anticipate any or all of these mechanisms. In high integrity applications it can be necessary to include manual or automatic checking mechanisms to confirm that the protection is still present (see also section 6.3.3.2), since if the device fails open then the system will carry on operating happily – until the next surge comes along, when it will be unprotected. In very high integrity systems you may well find that this aspect in fact prevents you from using transient suppressors at all.

Follow-on current after a device has failed short can create problems, especially on power supplies, for which reason it is good practice to include a fuse in series with the protected line just before the suppressor. The fuse has to be selected with an I^2t characteristic less than the device’s own fusing capacity and with a rating dependent on the expected follow-on current.

14.2.5.2 Cascading types

You may sometimes have to parallel different types of suppressor in order to achieve a performance which could not be given by one type alone. For example, telecoms applications require signal line protection not just from straightforward transients, but also from local lightning strikes and shorting between the signal line and AC power lines [37]. These can only be dealt with by providing both primary and secondary protection; primary protection, offered by gas discharge tubes (GDTs) or carbon spark gaps, will remove the major part of the incoming energy but leaves an initial spike due to its slow response time. Secondary protection, provided by a semiconductor device, is faster and needs only to deal with the residual energy.

For a really serious degree of protection, you will be looking at a combination of all three types (Figure 14.37). Notice that series impedances are needed between suppressors to ensure that each type handles only the part of the surge that it is intended to. The design of this type of protection really needs to use a circuit simulation package with all non-linear and time-dependent parameters included (most suppliers now offer Spice models for their devices), if you are going to have any confidence that the final circuit will protect as intended. Notice also that the series impedances, particularly any that precede the first suppressor, will face the full might of the applied surge, perhaps several kV. You will need to use appropriately surge-qualified components here – an 0603 resistor will definitely not do!

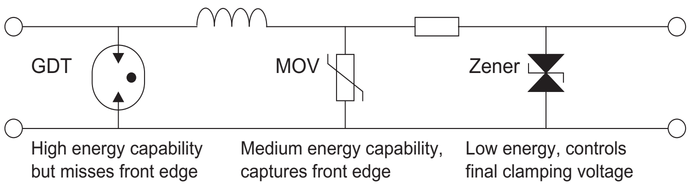


Figure 14.37 Cascading suppressors

Foldback suppressors

The disadvantages of straightforward zener suppressors, that their energy handling capability is limited because they must dissipate the full transient current at their breakdown voltage, are overcome by a family of related suppressors which integrate a thyristor with a zener. When the overvoltage breaks down the zener, the thyristor conducts and limits the applied voltage to a low value ("folds back"), so that the power dissipated is low and a given package can handle about ten times the current of a zener on its own. Provided that the operating circuit current is less than the thyristor holding current, the thyristor stops conducting once the transient has passed.

14.2.5.3 Layout of transient suppressors

Short and direct connections to the suppressor (including the ground return path) are vital to avoid compromising the high-speed performance by undesired extra inductance. When the suppressor clamps it will of course pass a high current, and transient edges have very fast risetimes (a few nanoseconds for switching-induced interference down to sub-nanosecond for ESD) so that any inductance in the clamping circuit will generate a high differential voltage during the transient plus ringing after it, which will defeat the purpose of the suppressor.

The component leads must be short (suppressors are available in SM chip form) and they must be connected locally to the circuit that is to be clamped (Figure 14.38). Any common impedance coupling, via ground or otherwise, must be avoided. ESD spike suppression can be improved by raising the RF impedance of the input circuit with a lossy ferrite component – although this is less successful for lightning-level surges since the ferrite will most probably saturate. Where suppressors are to be combined with I/O filtering you may be able to use the three-terminal varistor/capacitor devices that are now available. You will also need to consider thermal issues if the suppressor is expected to dissipate significant power.

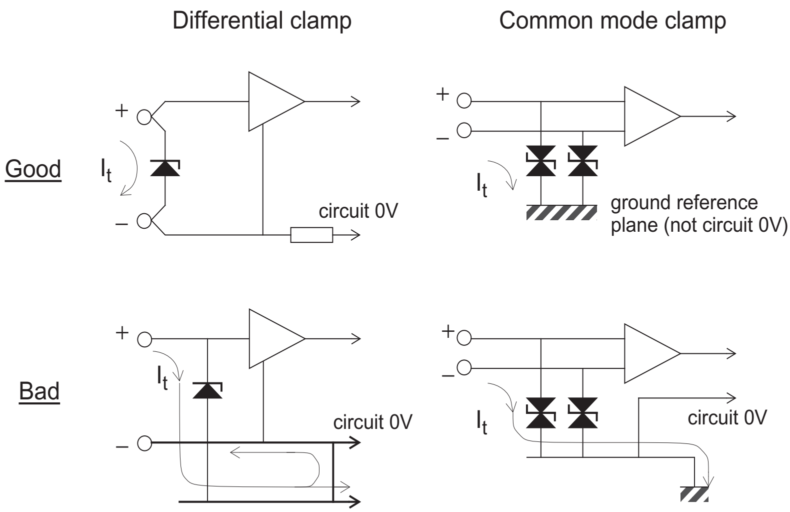


Figure 14.38 Layout and configuration of I/O transient suppressors

14.2.6 Contact suppression

An opening contact which interrupts a flow of current – typically a switch or relay – will initiate an arc across the contact gap. The arc will continue until the available current is not enough to sustain a voltage across the gap (Figure 14.39). The stray capacitance and inductance associated with the contacts and their circuit will in practice cause a repetitive discharge until their energy is exhausted, and this is responsible for considerable broadband interference [10][86]. A closing contact can also cause interference because of contact bounce.

Any spark-capable contact should be suppressed. The criteria for spark capability are a voltage across the contacts of greater than 320V, and/or a circuit impedance which allows a dV/dt of greater than typically 1V/µs – this latter criterion being met by many low-voltage circuits. The conventional suppression circuit is an RC snubber network connected directly across the contacts. The capacitor is sized to limit the rate-of-rise of voltage across the gap to below that which initiates an arc. The resistor limits the capacitor discharge current on contact closure; its value is a compromise between maximum rated contact current and limiting the effectiveness of the capacitor. A parallel diode can be added in DC circuits if this compromise cannot be met.

14.2.6.1 Suppression of inductive loads

When current through an inductance is interrupted a large transient voltage is generated, governed by $V = -L \cdot di/dt$. Theoretically if di/dt is infinite then the voltage is infinite too; in practice it is limited by stray capacitance if no other measures are taken, and the voltage waveform is a damped sinusoid (if no breakdown occurs) whose frequency is determined by the values of inductance and stray capacitance. Typical examples of switched inductive loads are motors, solenoid or relay coils and transformers, but even a long cable can have enough distributed inductance to generate

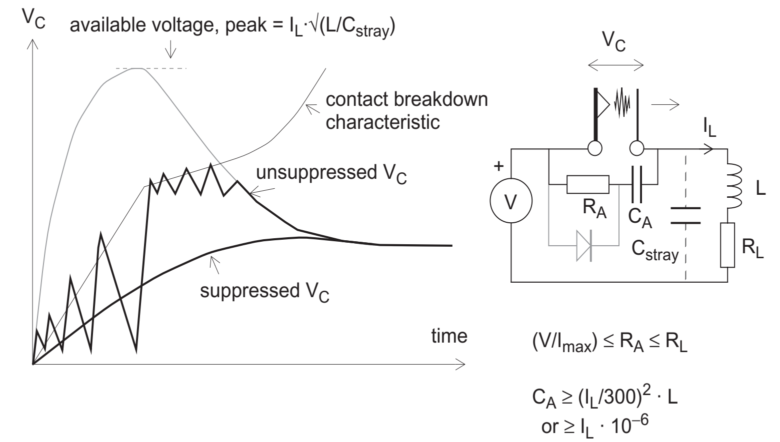


Figure 14.39 Contact noise generation and suppression

a significant transient amplitude. Switching can either be via an electromechanical contact as in the previous section, or a semiconductor, and the latter can easily suffer avalanche breakdown due to the overvoltage if the transient is unsuppressed. RF interference is generated in both cases at frequencies determined by stray circuit resonances, and is usually radiated from the wiring between switch and load.

The RC snubber circuit can be used in some cases to damp an inductive transient. Other circuits use diode, Zener or varistor clamps as shown in Figure 14.40. In all cases the suppression components must be mounted immediately next to the load terminals, otherwise a radiating current loop is formed by the intervening wiring. Protection of a driver transistor mounted remotely must be considered as a separate function from RF suppression.

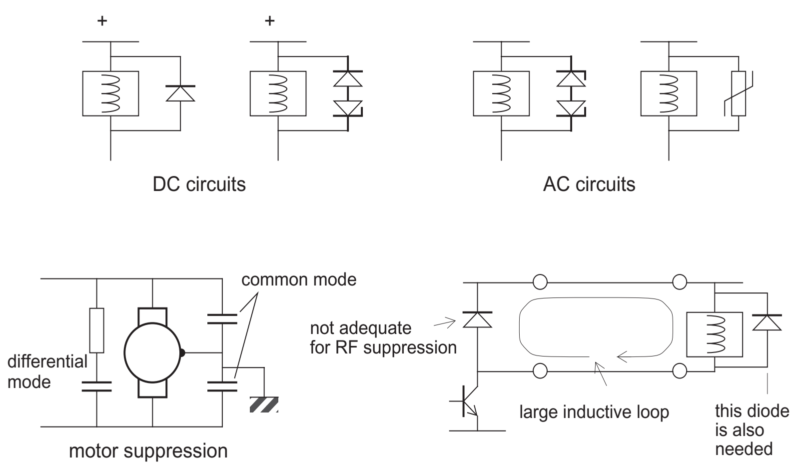


Figure 14.40 Inductive load suppression

14.2.6.2 Motor suppression

DC brushed motor noise is particularly aggressive, since it consists of impulsive and hence wideband transients repeated at a rate determined by the commutation speed – in other words, several hundred to several thousand times a second. The spectral composition of this noise may extend up to several hundred MHz. This appears both as differential mode noise across the terminals and common mode with respect to the housing, coupled through stray capacitance.

The best method of suppression is to prevent the motor from generating impulsive voltages across the commutator segments. This can only be achieved by the motor manufacturer, by incorporating varistor or RC components between each commutator segment, but ensures that the motor is quiet without further suppression being necessary. Motors are available with this addition, but of course it's an extra expense and they aren't usually standard items. Otherwise, a capacitor or snubber across the terminals (for differential mode) and capacitors from the terminals to the local earth (for common mode) are required; if the local earth is not available, then you can only use a choke at the motor terminals. Motor interference often appears in common mode with respect to the housing as a result of the high stray capacitance between the housing and the windings, hence you will often need to use common mode techniques.

Note that the above comments don't apply to induction motors or other types which don't have brushed contacts. They do not develop the same kind of impulsive interference – but if they are fed by a variable speed drive with a high voltage pulse-width-modulated waveform, there is plenty of interference from this source instead (see section 13.2.3).