



Another EMC resource  
from EMC Standards

## EMC techniques in electronic design Part 3 - Filtering and Suppressing Transients

*Helping you solve your EMC problems*

# Design Techniques for EMC

## Part 3 — Filtering and Suppressing Transients

Originally published in the EMC Compliance Journal in 2006-9,  
and available from <http://www.compliance-club.com/KeithArmstrong.aspx>

Eur Ing Keith Armstrong CEng MIEE MIEEE  
Partner, Cherry Clough Consultants, [www.cherryclough.com](http://www.cherryclough.com), Member EMCIA  
Phone/Fax: +44 (0)1785 660247, Email: [keith.armstrong@cherryclough.com](mailto:keith.armstrong@cherryclough.com)

This is the **third** in a series of six articles on basic good-practice electromagnetic compatibility (EMC) techniques in electronic design, to be published during 2006. It is intended for designers of electronic modules, products and equipment, but to avoid having to write modules/products/equipment throughout – everything that is sold as the result of a design process will be called a ‘product’ here.

This series is an update of the series first published in the UK EMC Journal in 1999 [1], and includes basic good EMC practices relevant for electronic, printed-circuit-board (PCB) and mechanical designers in all applications areas (household, commercial, entertainment, industrial, medical and healthcare, automotive, railway, marine, aerospace, military, etc.). Safety risks caused by electromagnetic interference (EMI) are not covered here; see [2] for more on this issue.

These articles deal with the practical issues of what EMC techniques should generally be used and how they should generally be applied. Why they are needed or why they work is not covered (or, at least, not covered in any theoretical depth) – but they are well understood academically and well proven over decades of practice. A good understanding of the basics of EMC is a great benefit in helping to prevent under- or over-engineering, but goes beyond the scope of these articles.

The techniques covered in these six articles will be:

- 1) Circuit design (digital, analogue, switch-mode, communications), and choosing components
- 2) Cables and connectors
- 3) Filtering and suppressing transients**
- 4) Shielding
- 5) PCB layout (including transmission lines)
- 6) ESD, surge, electromechanical devices, power factor correction, voltage fluctuations, supply dips and dropouts

Many textbooks and articles have been written about all of the above topics, so this magazine article format can do no more than introduce the various issues and point to the most important of the basic good-practice EMC design techniques. References are provided for further study and more in-depth EMC design techniques.

### Table of contents for this article

3.	Filters and transient suppressors .....	2
3.1	Introduction .....	2
3.2	Designing and selecting filters .....	4
3.2.1	How filters work .....	4
3.2.2	Imperfections in the basic filter circuits .....	5
3.2.3	The importance of the RF Reference .....	5
3.2.4	Differential-mode (DM) and Common-mode (CM) .....	6
3.2.5	Maximising impedance discontinuities .....	7
3.2.6	Using soft ferrite cores .....	8
3.2.7	Issues with wound components (inductors, transformers, chokes, etc.) .....	11
3.2.8	Specifying and designing filters .....	14
3.2.9	Problems with real-life supply impedances .....	17
3.2.10	Problems with real-life switch-mode converter input impedances .....	19
3.2.11	Damping filter resonances that cause gain .....	19
3.2.12	Filters and safety .....	19

3.3	Filter installation issues.....	20
3.3.1	Input and output conductors.....	20
3.3.2	Skin effect and the flow of surface currents .....	21
3.3.3	The synergy of filtering and shielding.....	23
3.3.4	Assembly and installation techniques for filters that penetrate shields.....	24
3.3.5	Designing to prevent corrosion.....	29
3.3.6	Filters connected in series.....	29
3.4	Types of overvoltage transients and surges.....	30
3.5	Protecting from surges.....	32
3.5.1	Protecting insulators from surge overvoltages .....	32
3.5.2	Protecting conductors from surge overcurrents .....	32
3.5.3	Protecting electromechanical contacts from surges.....	33
3.5.4	Galvanic isolation is the best defence against surges .....	33
3.5.5	Surge suppression with filters .....	34
3.5.6	Suppression with surge protection devices (SPDs) .....	36
3.5.7	Types of Surge Protection Device (SPD).....	38
3.5.8	Characteristics and comparisons of SPD types .....	40
3.5.9	Minimising the inductance in series with SPDs.....	41
3.5.10	Rating SPDs.....	42
3.5.11	Combining SPDs.....	44
3.5.12	A hierarchy of surge protection .....	44
3.5.13	Protecting SPDs.....	45
3.5.14	Equipment reliability and maintenance issues.....	46
3.5.15	Surge protection products.....	47
3.5.16	'Earth lift' problems in systems .....	48
3.5.17	Data needs error detection/correction .....	49
3.6	References.....	49
3.7	Acknowledgements.....	50

### 3. Filters and transient suppressors

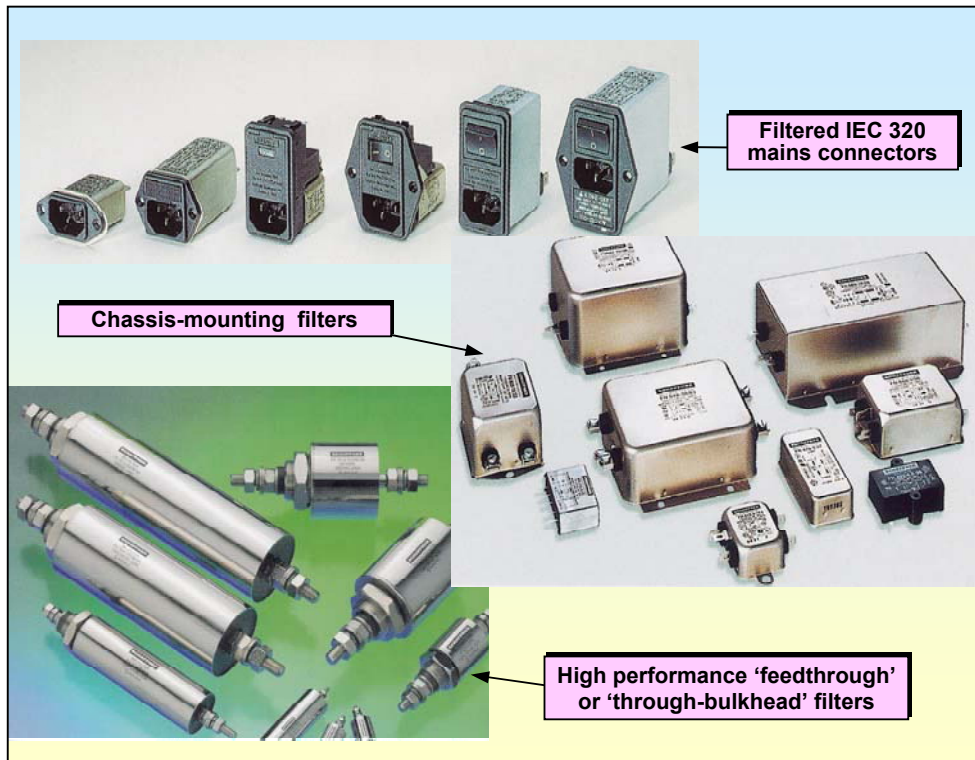
#### 3.1 Introduction

Filters are used to attenuate unwanted frequencies travelling along conductors, and are characterised by attenuation versus frequency curves. Transient suppressors, such as surge protection devices (SPDs, sometimes called surge arrestors), attenuate unwanted voltage surges travelling along conductors, and are characterised by graphs of voltage 'let-through' versus time.

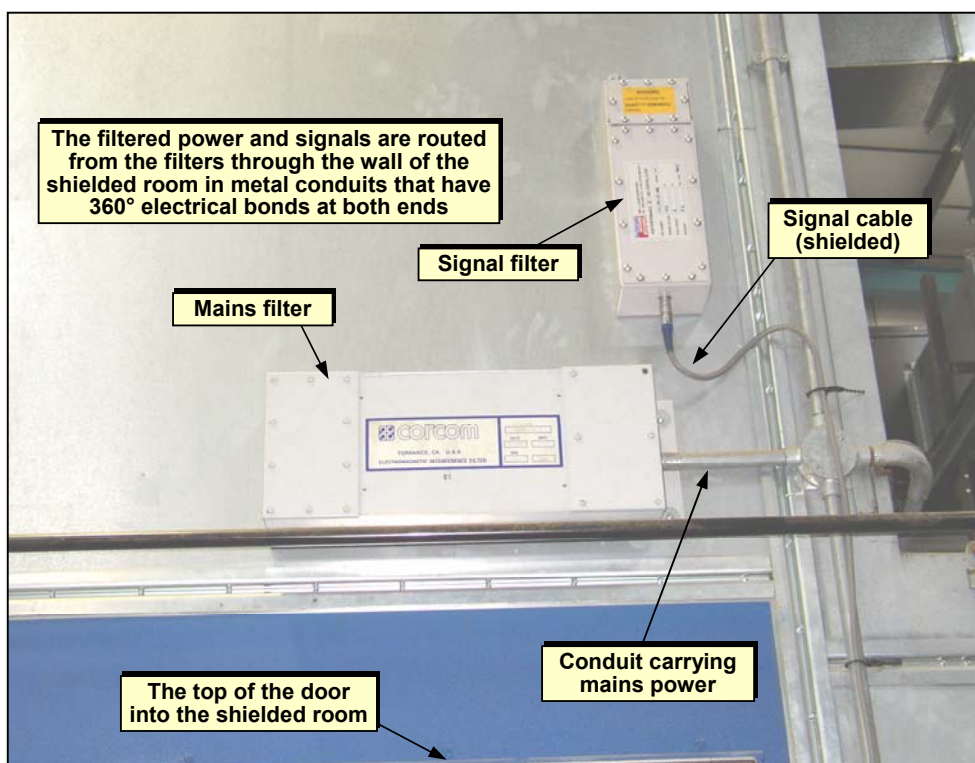
Incorrect use of filters or SPDs can make a product's emissions or immunity worse than if they were not used at all. More expensive filters or SPDs are not necessarily the best. You cannot in general choose a filter or SPD from a distributor's catalogue, by simply checking its ratings, performance and intended application, and expect it to provide the benefits you need for your product.

Many books have been written on filter design, such as Arthur B Williams' [3]. No doubt there is a more modern edition available, but filter design has not changed much over the years. There are also now a number of circuit simulators that run on PCs and can be used to simulate filters. This article will not go into poles and zeroes and that sort of detail – instead it will describe the things which need to be taken into account so that filters designed using textbooks, circuit simulators such as Spice, or chosen from catalogues, stand a chance of performing as required, and avoid unpleasant and/or costly experiences.

Filter design or selection is not a 'black art', but nevertheless it is difficult to predict *exactly* what performance a given filter will achieve when installed in a product, especially at frequencies above 100MHz, so it is often necessary to experiment with different options to find the most cost-effective. Planning and designing for such flexibility from the start of a project is very worthwhile, and an example of what John R Barnes [4] calls "wiggle room" and I call "anti-Murphy design". My approach is based upon the well-known Murphy's Law – I find that designers who try to anticipate the surprises that Murphy might have in store for them reach their design targets and timescales more reliably, and the resulting products have a lower overall cost of manufacture because they have not had to have filters, or larger filters than were hoped for, squeezed in somehow at the end of a project when compliance tests were failed.



**Figure 3A** A wide variety of filter styles is available (such as these mains filters from Schaffner)



**Figure 3B** Examples of high-performance 'room filters' (at BSI's EMC test lab, Hemel Hempstead)

This article considers filters that are fitted at the boundary between a product and its external electromagnetic (EM) environment. Filters used inside an item, for example between a switch-mode power supply and a sensitive analogue circuit, will share most if not all of the same considerations – because filters always separate two areas or zones that should not be allowed to crosstalk or otherwise freely intermingle their signals.

## 3.2 Designing and selecting filters

### 3.2.1 How filters work

Ignoring all the poles and zeroes in the filter textbooks: filters work by creating an intentional discontinuity in the characteristic impedance of a current path, reflecting radio frequency (RF) energy away from a protected circuit, or absorbing the RF energy (converting it to heat) – rather like a shield does, as will be described in Part 4 of this series.

The greater the discontinuity, the greater the attenuation. So if the source impedance of an unwanted signal (noise) is  $100\Omega$  and we put a  $1k\Omega$  impedance in series with it, only about 10% of the signal gets through the high impedance – an attenuation of around 20dB. A similar effect can be created by instead connecting the  $100\Omega$  noise to the 'RF Reference' via an impedance that is much lower than  $100\Omega$ : for example,  $5\Omega$  would provide an attenuation of around 26dB.

Filters use electronic components such as resistors (R), inductors (L), and capacitors (C) to create the desired impedance discontinuities over the ranges of frequencies of concern. R, L, or C can be used as filters on their own, but combining them gives better attenuation. LC types can give better attenuation than RC types, and are often used in power circuits because of their lower losses, but all LC filters are resonators that can produce gain at some frequencies, so they need to be carefully designed, taking their actual source and load impedances into account, to ensure attenuation over the desired range of frequencies. RC types generally provide more reliable filter performance.

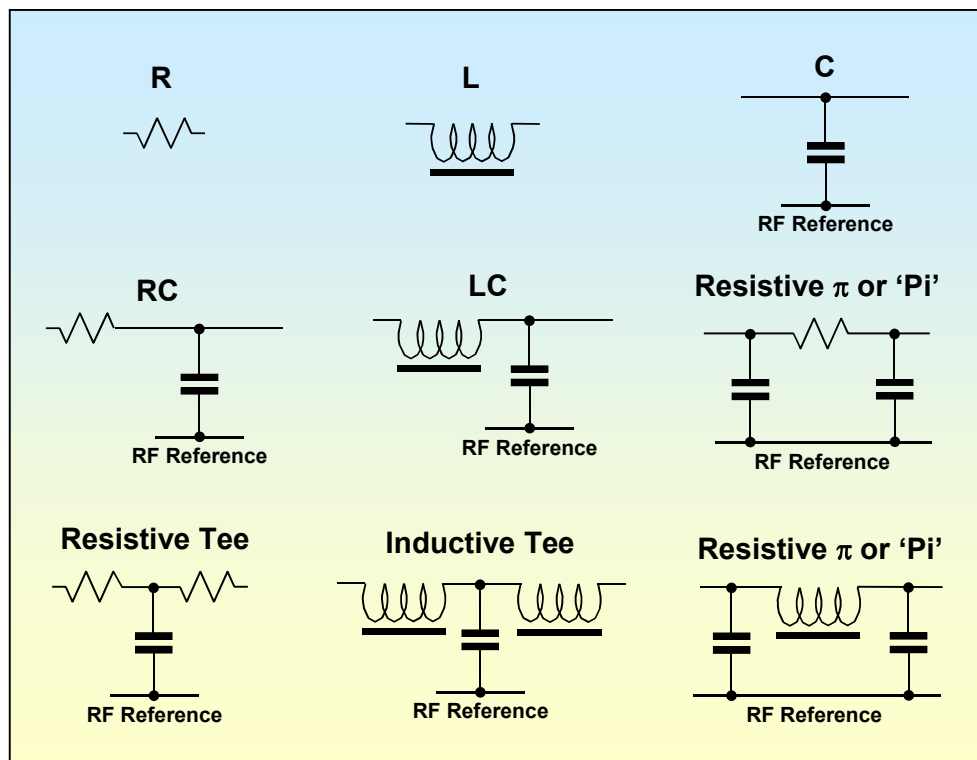


Figure 3C Different types of simple single-line filters

A range of basic schematics exists for low-pass filters based on R, L and C, and is shown in Figure 3A. There are high-pass equivalents, and band-pass or notch filters can also be achieved with passive components like these – but the low-pass filter is the one that is mostly used for EMC so that is the type that is shown in Figure 3A and discussed in this article.

Simple inductive filters (chokes, ferrites, etc.) have no RF Reference connection, so are especially useful where no RF Reference Plane exists, or if it exists but does not have a structure that provides a low enough impedance at the highest frequencies of concern. Unfortunately, such very simple filters are generally unable to achieve very high attenuations – typically between 3 and 20dB, depending on the frequency.

Capacitors can also be used on their own as very simple filters (by creating a 'high-to-low' impedance discontinuity), or as part of a more complex filter circuit that includes inductors and/or resistors. But the effectiveness of a capacitor filter depends upon the impedance of the RF Reference it is using as its 'ground', and also upon the impedance of the interconnection between the capacitor and the RF Reference (e.g. wire leads, PCB traces). As a result, manufacturer's data sheet figures for capacitive filters are rarely achieved in real-life because they were tested with RF Reference Planes that were solid copper sheets covering an entire bench-top, and so had a lower impedance than is usually possible in real life.

Many a well-designed and expensive filter has had its performance wasted by being connected to a poorly performing RF Reference, or by being bonded to an excellent Reference by a short length of wire instead of the direct metal-to-metal contact that was needed.

An example of a common use of RCR filters is to connect computer boards to displays via flexible circuits, to reduce the emissions from the 'flexi'. The resistor values in these filters are often chosen as much for transmission-line matching (see section 2.7 of [6]), as they are for filtering.

Filters must pass the wanted signals/power, while attenuating unwanted 'noise'. So filter specification must begin with knowledge of the full spectrum of the wanted signal or power. It is very common these days for the spectrum of a wanted signal to contain very high frequencies that are not required, caused by the very fast switching edges of modern digital and switch-mode devices. Analogue signals are also polluted with such noise, due to stray coupling from digital and switch-mode circuits nearby. These very high frequencies can be removed by filtering and/or shielding, and it is good EMC practice to remove them at their sources, rather than wait until they have polluted many more conductors, and this was discussed in section 1.1.2 and Figure 1B of Part 1 of this series [7].

Active filters can be designed, based upon operational amplifiers (opamps), using feedback techniques to achieve remarkable attenuations. But the phase-shifts inherent in all opamps converts the attenuation of feedback circuits into amplification, above some frequency. So unless you have the experience and skills to really know what you are doing, and unless you are using op-amps with gain-bandwidth products measured in many GHz – always use passive filters based on Rs, Ls and Cs to control frequencies above 1MHz.

### 3.2.2 Imperfections in the basic filter circuits

All components have imperfections, and these were discussed in section 1.8.1 of Part 1 of this series [7]. These imperfections have a part to play in defeating our attempts to design effective filters quickly and easily. For example: resistors lose attenuation at high frequencies due to their stray parallel capacitance. Inductors lose attenuation when their stray capacitance causes them to self-resonate, and at higher frequencies. Capacitors suffer from self-inductance, causing them to self-resonate and lose attenuation too.

RC filters are the most predictable EMC filters, as they do not resonate strongly. Values of R over the range 1 $\Omega$  to 10k $\Omega$  are commonly used in EMC engineering, with C values typically less than 100nF. RC filters are mostly used where a DC or low-frequency signal from a low source impedance is connected to a high impedance load: the R is connected to the source side, the C connected to the load side, as shown in the lower part of Figure 3E (below), where they provide very high attenuation at low cost.

LC, inductive Tee and inductive  $\pi$  filters can provide higher attenuation with lower losses than filters using resistors, but are resonant circuits and sensitive to their source and load impedances.

### 3.2.3 The importance of the RF Reference

The RF Reference is the node on a circuit's schematic that we define as our reference voltage when designing an RF circuit or measuring its performance. For the most cost-effective EMC, all circuits (digital, analogue, switch-mode, etc.) should now be designed using RF techniques, and this was discussed in Parts 0 and 1 of this series [7].

It is common practice to call the RF Reference 'earth' or 'ground', although it might instead be called 'chassis' or 'frame' in some applications, and in circuits it is usually the same structure as the 0V power supply distribution so it is often called 0V. But all these terms are potentially misleading, because what matters in EMC engineering is the impedance of the conductor structure that is being used as the reference for the RF signals or noises, at the frequencies that you wish to control. The RF Reference is very important indeed, for all filters that are more than simple series impedances. For filters to function as desired, the impedance seen by the return currents as they flow in the RF Reference must be much less than the impedance of any filter elements connected to that Reference.

So, if we are using a 10nF capacitor in an RC filter to shunt the RF noise to 'ground', and we want the RC filter to operate as close to its theoretical performance as possible up to 100MHz, we should realise that the reactive impedance of the capacitor (assuming a self-inductance of 1nH) at 100MHz is approximately 0.65 $\Omega$  (almost all of which, incidentally, is due to its self-inductance). To create a 'ground' structure that has an impedance of much less than 0.65 $\Omega$  at 100MHz is quite difficult, because a 10mm length of 1mm diameter wire or 1mm wide PCB trace has an impedance of about 6.3 $\Omega$  at that frequency. Increasing the diameter of the wire, or the width of the PCB trace, reduces the impedance but not by a great deal – 10mm length of 4mm diameter wire or a 4mm wide trace would still be around 3.2 $\Omega$ .

A great many earths, grounds, chassis, frames, and 0V systems are made of wire or PCB trace conductors, and designers assume that because they are labelled 'earth', 'ground' or '0V' they actually are at earth, ground or 0V potential – but in fact they have such high impedances at RF that they have significantly different

potentials at various points on their structures, depending on the RF currents flowing in them. Above a few tens of MHz, the only conductive structures that can achieve a low enough impedance to be useful as a reference for circuits and especially for filters, are metal areas or planes – which is why RF References are quite often called RF Reference Planes.

The circuits that use a plane as their RF Reference must be located much closer than one-tenth of a wavelength ( $\lambda/10$ ) to it, ideally  $\lambda/100$  or even less – at the highest frequency to be controlled. This helps prevent the connections to the plane from behaving as resonating antennas with impedances possibly in the hundreds of  $\Omega$ , instead of the plain old low-impedance conductors that they look like to our eyes. At 1GHz this would mean a maximum spacing of 30mm, and better EMC would be achieved by being much closer than that, ideally 3Nm or less.

Where a circuit is shielded by placing it in a metal box, it can use one or more walls of the box, and/or the rear, base and top as its RF Reference. Generally, this would still be called an RF Reference Plane, despite that fact that they are different sides of a metal box. An important consideration in the design of the structure of an RF Reference is that surface currents must be able to flow freely where they will, all over the area being used. Surface currents are discussed later in the section on Skin Effect.

Many electronic engineers are familiar with the idea of ‘single point earths/grounds’ – sometimes called ‘star earths’ or ‘star grounds’. In such designs the voltage reference is a single physical point, and everything that needs to use it connects to it by a conductor (a wire or PCB trace). Analysing these conductors in terms of impedance, or in terms of their likelihood of becoming resonating antennas, as discussed in the above paragraphs, quickly shows us that single-point or star conductive structures are no use to us for EMC – their conductors are simply too long.

The continued shrinking of silicon die sizes means that even commonplace digital glue-logic (e.g. HCMOS) now generates significant noise emissions at frequencies up to 1GHz, and modern FPGAs and microprocessors can be very much worse for EMC – both in level and frequency – than such glue logic ICs. To stand any chance of controlling such frequencies requires lengths of wire, PCB traces or via holes, that are no more than a few millimetres long, preferably <1mm. Using flat braid straps instead of round conductors simply raises the useful frequency by a little, but not by enough to control hundreds of MHz. So single-point earthing/grounding techniques are now only of interest to students of the history of technology, regardless of the power or signals involved. All circuits and interconnections now suffer from RF noise that is coupled into them from digital, switch-mode and/or wireless circuits inside the same product, and they also suffer from RF noise coupled from nearby cables and ambient EM fields in their environments. These coupled noises can cause any circuit or interconnection to be source of RF emissions, and/or a victim of interference, and this is true even for DC instrumentation and low-frequency analogue signals such as audio.

Despite the fact that the design of the RF Reference Plane is crucial to cost-effective EMC design, many engineers (me included) still tend to refer to ‘earth’ ‘ground’ or 0V, thereby often leading to confusion and miscommunication with people who think a length of wire can be part of an ‘earth’ structure as long as it has green/yellow insulation. So it is important to look beyond the terms that are being used to identify the physical structure that will be used as the RF Reference Plane, or to create it if it is not yet there.

### 3.2.4 Differential-mode (DM) and Common-mode (CM)

Wanted signals are always DM: they flow along the ‘send’ conductor, and flow back along the ‘return’ conductor(s). In single-ended signalling, all the return currents share a common conducting structure, usually the 0V of the DC power distribution system. In balanced (or ‘differential’) signalling there is a dedicated conductor for the return current path as well as for the send path, and for good signal quality and EMC the two are routed together as a twisted pair.

However, unavoidable imbalances in the physical realisations of interconnections in PCBs and cables cause CM voltages and currents to arise, as shown by Figure 3D. CM currents flow out on both send and return conductors at the same time, and return via another route, often the safety earth structure or the mains power distribution network. CM currents are typically measured in  $\mu\text{A}$ , whereas DM currents are in tens or hundreds of mA, maybe even Amps – but the much larger loop areas associated with CM noise currents and voltages makes them more important for EMC than the DM signals that originated them. Above about 1MHz most unwanted emissions are mostly CM.

A great deal of RF interconnect design is concerned with making cables and PCB traces that have better balance, to reduce the ‘longitudinal conversion loss’ (LCL) that converts the wanted signal energy into unwanted CM noise. The better the LCL, the further the wanted signal will propagate with an acceptable quality, or the higher the frequency that can be sent with acceptable quality over the same distance – hence the computer networking industry’s progress from Cat 5 to Cat 6 and eventually to Cat 7 cables for Ethernet, each increase in Category is accompanied by better balance, resulting in better LCLs at higher frequencies and reduced generation of CM noise for a given type of signal.



Because of the existence of DM and CM signals and noises, we need to be able to apply filtering techniques to both of them. Below 1MHz, we are more likely to be concerned just with filtering DM signals and noise. But at higher frequencies we can use DM filtering to reduce the amounts of RF present in conductors, so as to reduce the amount of CM noise currents and voltages created by the imbalances in the interconnects. We also use CM filtering to reduce the amounts of CM noise present.

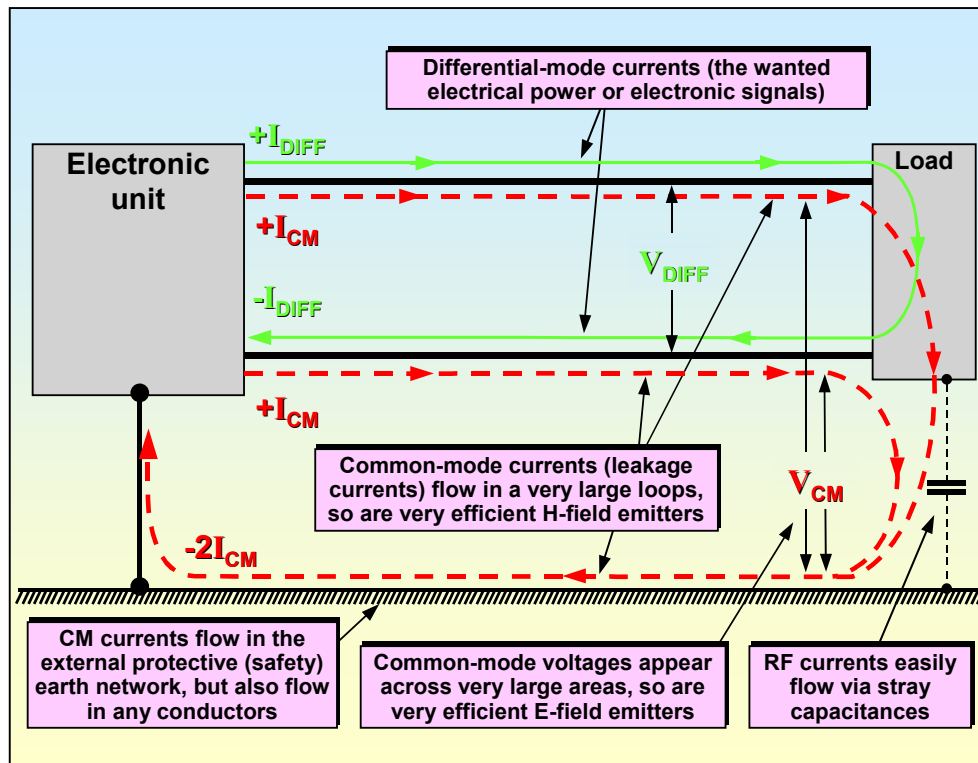


Figure 3D Differential-mode (wanted) signals versus common-mode 'noise leakage'

### 3.2.5 Maximising impedance discontinuities

As mentioned earlier, to design effective filters we must maximise impedance discontinuities, at the frequencies of concern for emissions and/or immunity, and Figure 3E tries to demonstrate this concept for single-ended signals. Capacitors are used in conjunction with the RF Reference Plane (see above) to create low impedances, applied in shunt, whilst resistors or inductors are used to create high impedances, applied in series.

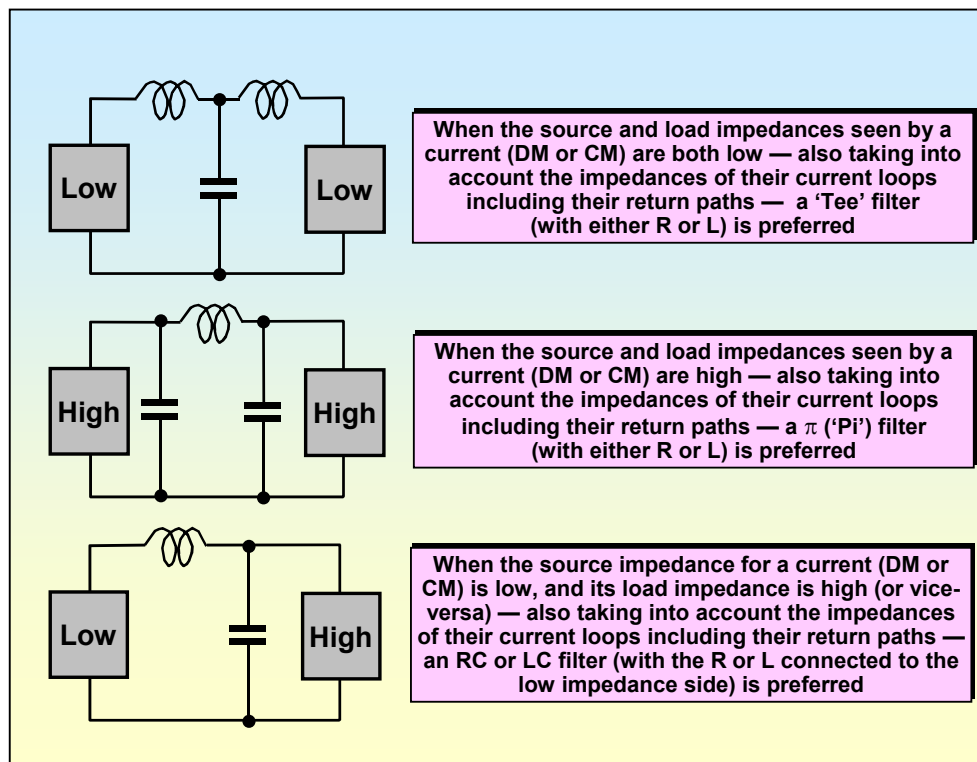
When the source and load impedances seen by a current (DM or CM) are both low – also taking into account the impedances of their current loops including their return paths – a 'Tee' filter (with either R or L) is preferred. When the source and load impedances seen by a current (DM or CM) are high – also taking into account the impedances of their current loops including their return paths – a  $\pi$  ('Pi') filter (with either R or L) is preferred. When the source impedance for a current (DM or CM) is low, and its load impedance is high (or vice-versa) – also taking into account the impedances of their current loops including their return paths – an RC or LC filter (with the R or L connected to the low impedance side) is preferred.

For low-power circuits with low-frequency wanted signals and high impedance loads, it is often possible to replace the inductors in these simple circuits with resistors of between 100 $\Omega$  and 10k $\Omega$  to save cost and even sometimes achieve higher attenuations over wider frequency ranges.

When the source impedance of the noisy current (DM or CM) to be filtered is low, fitting a simple C filter will increase the noise currents flowing in the circuit, increasing H field emissions, and will also increase the noise voltages across the circuit's 0V plane, increasing its CM emissions. Preventing DM or CM noise currents from increasing is another reason why we always follow a low impedance source with a series resistor or soft ferrite choke (to create an RC, LC or Tee filter).

For balanced (differential) signals the RF Reference in Figure 3E is replaced by the return conductor for the conductor pair – but only for DM filtering. For CM filtering we need two circuits as shown in Figure 3E – one for the send conductor and one for the return conductor, both of them connecting their capacitors to the RF Reference Plane. CM filtering can also benefit greatly from the use of 'CM chokes' – described later and shown in Figures 3M, 3N and 3P.





**Figure 3E Maximising impedance discontinuities to improve attenuation Using soft ferrite cores**

All inductors (L) suffer from RF resonances, and are only effective in filters at frequencies not far above their first (parallel) resonance (see section 1.8.1 of [7]). But so-called 'soft ferrites' behave resistively at RF, and the resulting lack of RF resonances helps make filters that use them have better and more predictable performance at RF. For example, a typical small 'soft ferrite' bead a few millimetres in diameter will have around  $1\mu\text{H}$  of inductance and  $0.1\Omega$  of resistance at DC, but around  $80\Omega$  of real resistance (*not* inductive reactance) at frequencies from 30MHz to 1GHz or more. Some leaded soft ferrites are available with resistances of over  $1\text{k}\Omega$  at 100MHz, but a much wider range of surface mounted device (SMD) soft ferrites is available with resistances up to  $1\text{k}\Omega$  or more at selectable frequencies from 30MHz to 2GHz.

Soft ferrite components are known by a variety of names, including 'RF suppressers', 'Interference suppressers', 'Suppression chokes', and 'Shield Beads'. Figures 3F and 3G show some of the cable-mounted soft ferrite parts available. A very wide range of PCB-mounted soft ferrite components is also available, but not shown in these figures. Figure 3G includes a standard VGA cable, showing the standard soft-ferrite CM choke that all VGA cables are required to have at each end, for the products they interconnect to meet emissions regulations in Europe and the USA (at least).

The bottom right-hand-side of Figure 3G shows a toroidal soft ferrite core used as a CM choke, in this case with '4½ turns' of cable wrapped around it. As will be described below, the attenuation of a filter at the highest frequencies is governed by the stray coupling between its input and output conductors — so it is important that the input and output conductors of a CM choke, such as the one in this photograph, are as far apart from each other as possible. This results in the winding format that can clearly be seen in the Figure — only half the core is wound and the input and output cables are on opposite sides from each other, the perhaps odd description of it as having '4½ turns' is a common way of making clear that input and output conductors are on opposite sides.

A useful soft ferrite component is a cylinder split lengthways and held in a plastic clip-on housing, and some examples of this are also included in Figure 3G, for round cables as well as for flat cable styles. Such split ferrites are very easy to apply to cables (and to remove if found to be ineffective), and EMC engineers tend to carry many of these around with them, using them for the diagnosis, isolation and curing of EMC problems, both DM and CM. A ferrite cylinder clipped around an entire cable or cable bundle, including all the send and return conductors, is a CM choke, but if clipped over just a send or return conductor it is a DM choke.

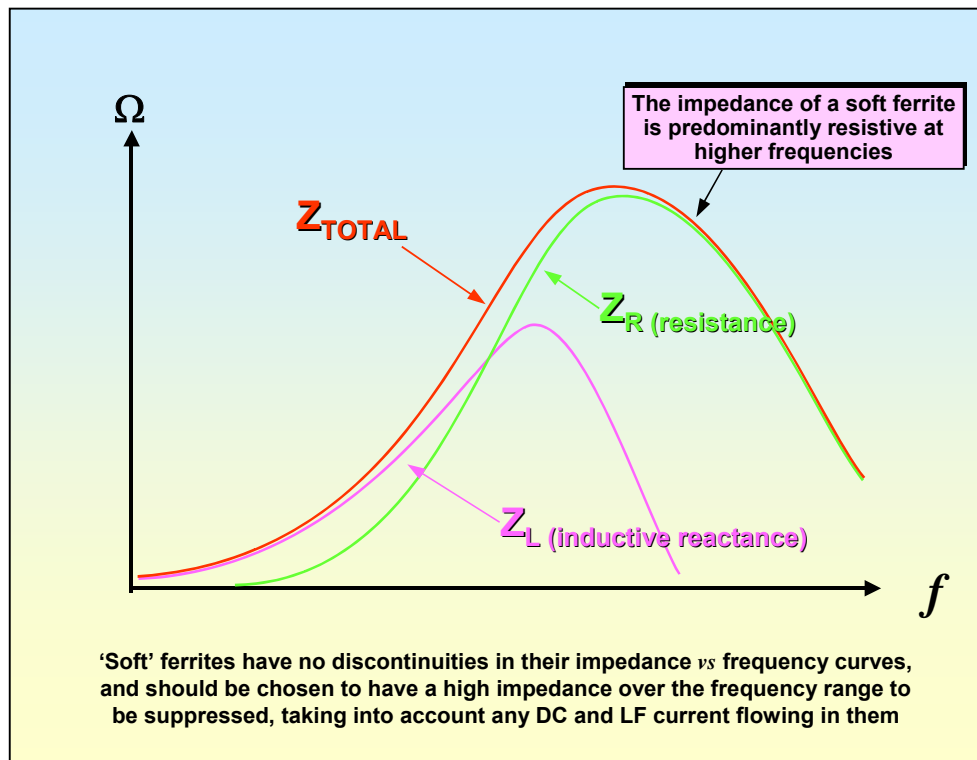


**Figure 3F** A wide variety of soft ferrite cores is available  
(these examples of toroidal and cylindrical types are from Philips)



**Figure 3G** Some more examples of soft ferrite cores

Choosing soft ferrites involves checking that their impedance is as high as required over the frequency range for which significant attenuation is required. Soft ferrite components always have impedance versus frequency curves that are smooth and not discontinuous, whereas the curves for inductors will show one or more discontinuities (changes in slope from positive to negative, or vice-versa, that occur at a point) that reveal the presence of resonances.



**Figure 3H Choosing soft ferrites**

Some data sheets only provide impedance data for a portion of the frequency range you are concerned with. But it would be a mistake to assume anything about the impedance they will achieve in the frequency range for which no data is provided – *always* make sure you have manufacturers data on the impedance over the entire frequency range that you wish to control.

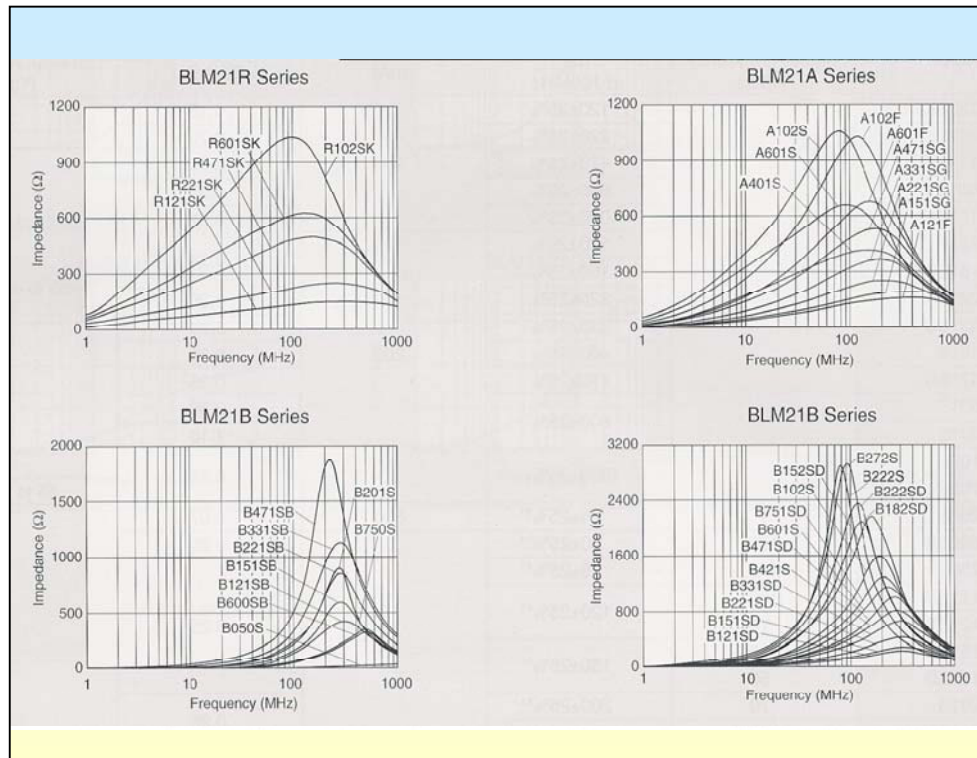
An aspect of choosing soft ferrites that is often overlooked, is that their impedance versus frequency curves vary with their DC and/or LF current. Typical data sheet curves assume zero current in the device, but as the current increases the frequency at which the peak impedance occurs will also increase, and may not be as high as it was with no current. Often, when an emissions or immunity test is failing at some frequency (e.g. a clock harmonic at 228MHz), a soft ferrite will be chosen that has a very high impedance close to this frequency, and it will be added in series with traces on the PCB that are thought to be the cause of the problem.

But the DC or LF current in those traces could make the frequency of the peak impedance increase by enough that the actual impedance achieved at the problem frequency is not high enough to provide significant attenuation and pass the test. Instead, the currents in the traces and the frequency/current variation of the type of devices to be used should have been taken into account, to select a device that would have its peak impedance at the problem frequency *when the trace current is passing through it*.

Several manufacturers offer wide ranges of soft ferrite RF suppression components, and are continually adding to them. Recent additions include SMD parts rated at 3A continuous current at DC and low frequencies; yet provide 1kΩ or more around 100MHz. Other recent additions include parts that provide impedances of 1kΩ or more over the range 100MHz to 2GHz.

Curves such as those in the two top graphs in Figure 3J are most suitable for filtering low-frequency signals, whereas the two bottom curves show devices that have been tailored for filtering unwanted harmonics from digital waveforms whilst leaving sufficient lower-frequency harmonics to create reasonable digital waveforms that have rise and fall-times fast enough to reliably meet the maximum skew requirements of the circuit. CM ferrites tend to have impedance versus frequency curves similar to the top left-hand graph, since there is no need for any CM currents at any frequencies.

When simulating filters or other circuits using soft ferrite components, a simple device model cannot be used – the parameters are frequency-dependant and current-dependant, and may also be temperature dependant, and should be modelled as such to achieve any accuracy in the simulation over a range of frequencies. Some circuit simulators may be unable to handle models with such complex parameters.



**Figure 3J A wide range of soft ferrite devices is available**  
(these examples from the Murata BLM21 series, 0805, typically rated 100 - 200mA)

### 3.2.7 Issues with wound components (inductors, transformers, chokes, etc.)

Three main issues of concern for wound (inductive) components are the control of their stray magnetic fields, the variability of their parameters with current and temperature, and their resistivity.

#### Closed magnetic circuits are preferred, to reduce stray fields

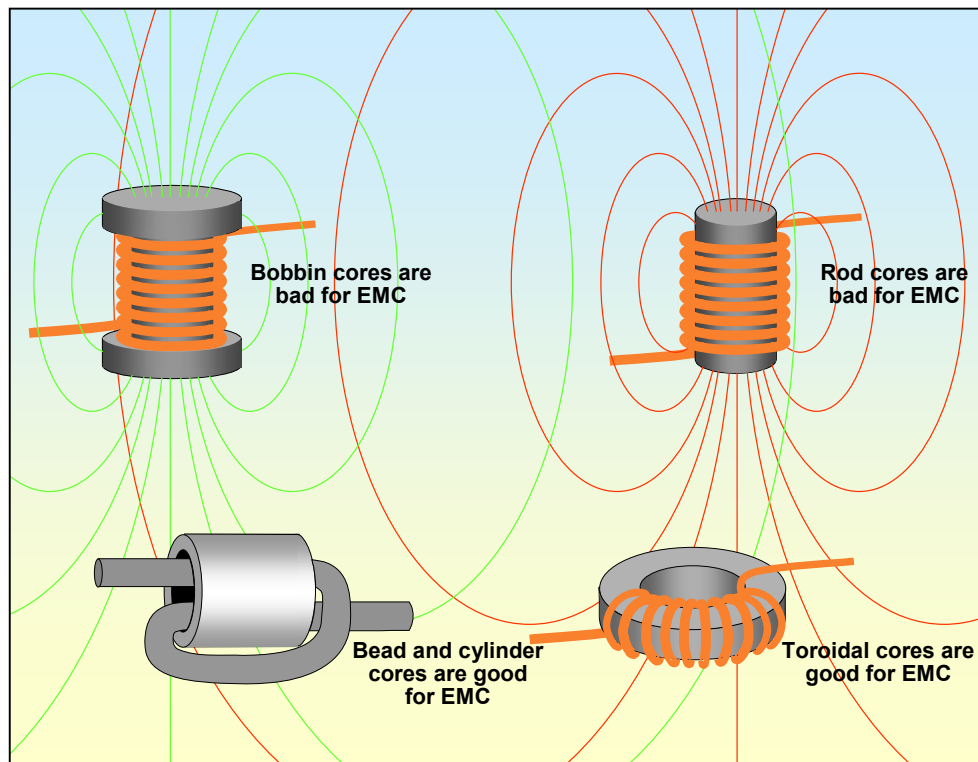
If the path of the magnetic field includes air, the component can be a significant cause of emissions, and can also pick up ambient magnetic fields and cause noise in the circuit. As a result, the EMC of wound components benefits from having a closed magnetic circuit, such as a ferrite bead, cylinder or toroid. This is important for filters, and also for chokes and transformers in switch-mode power converters, whether AC-DC, DC-AC (e.g. inverters) or DC-DC.

For example, an inductor wound on a rod core is essentially the same as a ferrite rod antenna in a typical AM radio. Many types of low-current inductors (up to about 1mH) or ferrite suppression components are available as axial components, essentially ferrite rods with a winding on them. While they can be perfectly acceptable in some applications, in others they can pick up noise from ambient fields and cause or suffer from crosstalk and/or emissions or immunity problems – so it is generally more reliable to use types based on ferrite cores that are beads, cylinders or toroids, which have closed magnetic circuits. Figure 3K illustrates this issue.

This can be a significant problem for DM chokes carrying higher currents, because they have large magnetic fluxes and rely on air gaps to prevent their cores from saturating. Where an air gap is unavoidable, solutions include applying a shield over the component to contain its stray fields, or else use a core made of iron or iron oxide powder in an epoxy binder, or a similar distributed-air-gap material.

Close proximity of a shield reduces the component's inductance – and the shield might become saturated from the high field strength and no longer provide shielding – so the shield should not be too close to the core or the windings. Iron powder or similar cores have no discrete air gaps, relying instead on hundreds of millions of microscopic air gaps between the magnetic particles in their cores, with much smaller stray magnetic fields as a result.





**Figure 3K Stray fields from air gaps in the cores of wound components are bad for EMC**

#### **Allowing for the variation of parameters due to DC and/or LF currents, and temperature**

As the DC and/or low-frequency current in a filter increases, the inductance value of all the series inductors falls, affecting the filter performance and resonant frequencies. Also – for all soft ferrites – the frequency at which the peak impedance occurs increases. Sufficient levels of current will saturate the magnetic circuit, causing the inductance/impedance to fall to practically zero. This is a common cause of the differences between simple calculations or simulations and real-life filter performance.

Variations in temperature have a similar effect. Above about 25°C, the inductance of a magnetic core reduces, falling more rapidly as the 'Curie Point' is approached, and zero above that point. The Curie Point depends upon the type of material, but is generally between 100 and 200°C. An experiment that showed how the performance of a mains filter could be reduced by as much as 20dB, by variations in current loading and ambient temperature that remained within the operational ratings of the filter, is described in [8].

So if designing a filter, get all the appropriate graphs of current and temperature dependency from the core suppliers and take this effect into account during the filter design for the full range of currents and temperatures they are required to operate over. And if choosing a filter from a supplier, make sure you understand what its minimum performance will be over the range of currents and temperatures (simultaneously) that it will experience in operation.

A particular problem is AC-DC mains power supplies that do not meet EN/IEC 61000-3-2 Class D. They draw their mains currents as peaks that are many times higher than their rated or measured RMS supply current. These peaks will degrade the attenuation 100 times per second (120 times/second in 60Hz countries) and might even cause momentary saturation of filter inductors and seriously compromise the filtering performance achieved.

CM chokes aim to balance the send and return currents so there is no net magnetisation of their cores by the wanted DM currents. But because no windings can be perfect, there is always some imbalance, which manifests itself as a DM choke in series with the CM choke. The resulting imbalance currents can saturate the cores of CM chokes, especially because their cores are made as small as possible to save space and reduce cost.

If designed carefully, CM chokes always run cool. Saturated inductors run warm, and if used on power frequencies they may be heard to hum or buzz, or felt to vibrate, both of which are clues to possible errors in design.

#### **Resistivity of ferrite cores**

All ferrites are ceramics, but they are not insulators – different types of material have differing resistivities. So if the insulation on the winding wire is inadequate for the stresses and strains imposed by winding it on the core, the core can 'short-out' the windings, or at least provide a parallel resistance that affects performance. If the

winding is carrying a hazardous voltage, failed insulation can result in a 'live' core that can be a safety hazard to service personnel, or even to a user.

There is always stray capacitance between a winding and its core, and because ferrites are ceramics their dielectric constant is high so the stray C is increased. RF voltages on the windings will therefore induce RF currents in the core, and because ferrite is conductive, these currents can flow out of the core into other conductors via stray capacitance or resistive contact.

Service personnel could get shocks and RF burns from the cores of wound filter components or transformers that are handling large amounts of RF. These stray RF core currents are CM, and so can cause significant problems for EMC. It is important either to insulate the cores to help inhibit the flow of stray CM currents, or else to provide a connection to the core that returns the stray current to the appropriate part of the circuit.

The most appropriate part of the circuit to return the core current to is not necessarily the 'earth', 'ground', 'chassis', 'frame' or '0V'. All currents flow in loops, and for good EMC the loop areas must be minimised – so the correct technique is to figure out where the stray RF core currents originally came from, and return them back to their source by the path that encloses the smallest area. For the filters in 'typical' electronic products there is usually no need to worry about the RF currents and voltages associated with ferrite cores. But control of core currents can be important for achieving EMC in the ferrites associated with switch-mode power conversion (flyback chokes, transformers, etc.) because switch-mode waveforms contain a great deal of energy in the RF spectrum. But this article is about filters so will not go any further into switch-mode EMC design issues.

### **Stray capacitance and its effect on high frequency impedance**

Stray capacitances between the input and output of a choke limits its high-frequency impedance, by acting as a 'bypass' in parallel with the choke. This effect can be seen in the impedance versus frequency curves of all of the ferrite chokes shown in Figure 3J – as frequency increases their impedance increases, but eventually a point is reached when their impedance starts to decrease as frequency continues to increase. In this region the impedance of a choke is dominated by the stray capacitances between its windings, and also between its input and output terminals and/or leads.

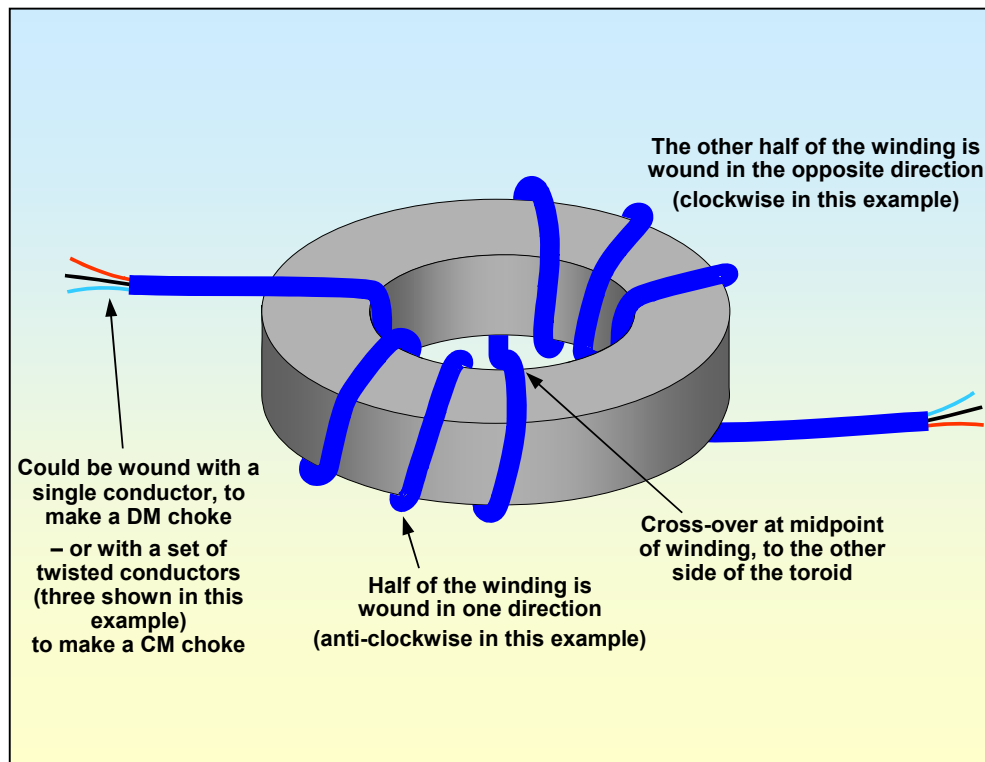
For this reason it is very important for high-frequency performance to keep the input and output terminals of series filtering elements (such as chokes or resistors) – and any leads, circuits or PCB traces attached to those terminals – as far apart from each other as practical. For good performance at 100MHz and above it can even be important to shield the input circuit from the output circuit – a topic discussed in 3.3.3 below.

Surface mounted ferrite beads can achieve high impedances at very high frequencies, because their parasitic capacitances are so small. But it is easy to ruin their performance above 100MHz by routing their input and output traces near to each other, increasing the stray input-output capacitance, so PCB layout is very important (see Part 2 of [9] for more on PCB design and layout techniques for EMC filters).

To obtain increased attenuation, it is tempting to wind the conductor(s) several times around the same magnetic core to increase its impedance. But this might not be as effective as required at higher frequencies, because of the increase in the stray capacitances created by the extra windings. The higher the number of windings on a core, the closer their proximity to each other, and the higher the stray capacitance between them. This effect is very strong with multilayer windings, especially if they are pile-wound – as this winding technique does not control where the windings lie with respect to each other, from one unit to another.

Chokes with multiple winding layers, including pile-winding, can achieve very high impedances at lower frequencies, and in some applications this is all that is required, for example when suppressing the RF noises emitted by a phase-angle controlled triac in a lighting dimmer, where significant levels of attenuation may only be needed up to a few MHz. Interestingly, at least one manufacturer of such lighting dimmers has found that chokes with layered or piled windings do not always provide the same attenuation performance when the chokes are reversed in the circuit. This is because of the complex nature of the stray capacitances in such a choke.

To improve the impedance at higher frequencies with highly-wound or multilayer chokes, it is better to use sectional winding techniques to reduce the overall input-output stray capacitance. One example of this type of choke is the 'super-toroid' winding shown in Figure 3L. The aim of this technique is to split the winding in half – each half being wound on different portions of the toroid to reduce the stray capacitance coupling between them and increase the impedance of the choke at higher frequencies. Notice that each half of the winding is wound in the opposite direction to the other half – but because these halves are wound in different directions their fluxes do not cancel out, they add together to maximise impedance.



**Figure 3L Winding a ‘super-toroid’**

A super-toroid wound with a single conductor is a DM choke, whereas one wound with a multi-filar twisted cable, as shown in Figure 3L, is a CM choke (see later).

Another way to improve the impedance at higher frequencies is to string a number of ferrite tubes or toroids in series along a cable. However, it is possible that their impedances and stray capacitances could interact to create resonances that will defeat the aim of this technique. So when using this technique it is best to simulate all such designs, using reasonably accurate estimations for the stray ‘components’ (or extract them from a 3-dimensional field simulation of the physical structure), or simply to build them and test their transfer function with instruments.

One way to reliably achieve a choke that has a very low stray interwinding capacitance and hence the very highest series impedance that the ferrite material is capable of, is to string a number of ferrite tubes or toroids along a conductor, with all of the ferrites touching each other. Instead of multiple windings around each ferrite core, the conductor is only passed once through the centre of each core. Of course this makes a very large or long device overall.

### 3.2.8 Specifying and designing filters

When specifying a filter prior to design or selection, it is necessary to know the spectrum of the wanted signals, so that the filter’s response can be tailored to pass the wanted signals whilst impeding unwanted noises (interference). It is easy to specify 50 or 60Hz mains filters, because the RF noises to be filtered are at a much higher frequency than the wanted 50 or 60Hz sinewaves. But it is not so easy to specify a filter when the signal and noise spectra overlap, as they do for most digital signals and interconnections.

However, most emissions and immunity problems are caused by CM noises, whereas wanted signals are DM – so we can use CM filtering on noise that is within the signal’s spectrum without attenuating or distorting the signal. Of course, nothing is perfect, so CM filtering will have some effect on the wanted DM signals, and it is part of the design to make sure that the attenuation and/or distortion of the DM signals are within acceptable limits.

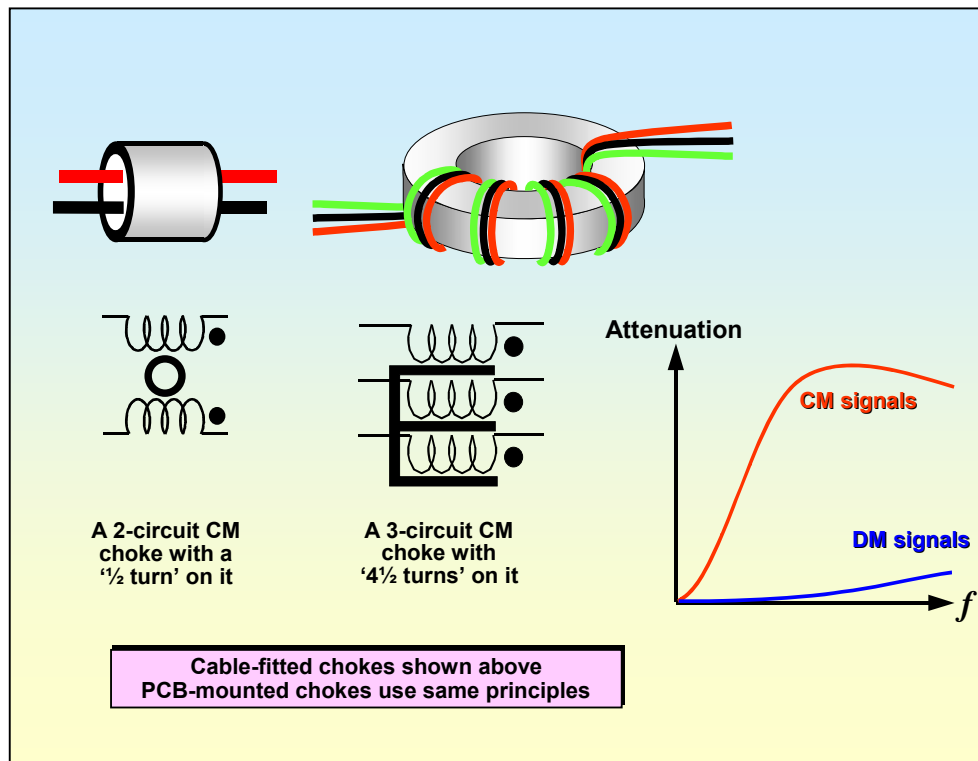
The key component for CM filtering is the CM ‘choke’, which generally uses a soft ferrite core with the send and return conductors for the DM signal wound together on the choke’s ferrite core, usually wound bi-filar, tri-filar, etc. with the input and output leads/pins on opposite sides of the core, to reduce the stray input-output capacitance that limits the performance at the higher frequencies.

When a magnetic circuit is wrapped around both (or all) of the send and return conductors associated with a signal or power circuit, it will only attenuate CM currents. The magnetic fluxes created by the send and return current paths of the wanted DM signals cancel out, so they experience no effect from the magnetic circuit. In practice there is always some leakage inductance, caused by imbalances in the send/return windings, hence there is always some DM attenuation. This inevitable leakage can be turned into a benefit by providing both CM



and DM filtering in one component. In ordinary CM chokes, the DM leakage is not controlled, and can vary considerably, but some EMC filter component manufacturers (e.g. Murata) offer ranges of CM chokes with specified DM impedances.

Some aspects of CM choke filtering are shown in Figure 3M.



**Figure 3M Common-mode (CM) filtering with CM chokes**

The cancellation of DM flux in CM chokes (sometimes called 'current-balanced' chokes) allows large inductance values (e.g. milliHenries) to be achieved with small components, whereas DM chokes of the same physical size achieve much lower inductances (e.g. microHenries) and must become physically larger as DM currents increase.

Of course we want to design so as to have confidence in passing EMC tests the first time – to avoid delays caused by iterations and retesting – but for volume-manufactured products we also do not want to add too much cost by over-engineering. Designing CM filters creates some difficulties here, because CM voltages and currents are mostly unknown in both frequency and amplitude until the product is EMC tested.

Computer-aided EMC simulation packages are now available that allow CM effects to be predicted. To give sufficiently accurate results the simulator must model all of the stray (parasitic) parameters of the components (which can be done in any circuit simulator), but it must also solve the fields associated with the components, conductors and other physical structures, in three dimensions. These simulators require powerful computers with a great deal of memory, such as modern 64-bit PCs, and the software packages can cost as much as a top-of-the range luxury car, but they can be very effective if sufficient effort is put into learning how to use them, modelling the components and digitising the actual physical design. Computer simulation offers the possibility of designing EMC right first time with the lowest component cost – or at least getting much closer to the optimum EMC design more quickly.

In the absence of an accurate computer simulation, it is best to design prototypes for a range of filter component or packaged filter options, and keep a stock of all the filter components or packages that could be needed, from the lowest-cost to the highest-specification (often available as free samples from their manufacturers). In the case of PCB-mounted filters, 'universal filter pad patterns' can be developed that can accommodate from zero-ohm links through individual resistors or ferrite beads to CM chokes, plus two or three-terminal capacitors, to create a wide range of filter types including RC, LC, Tee, and  $\pi$  filters.

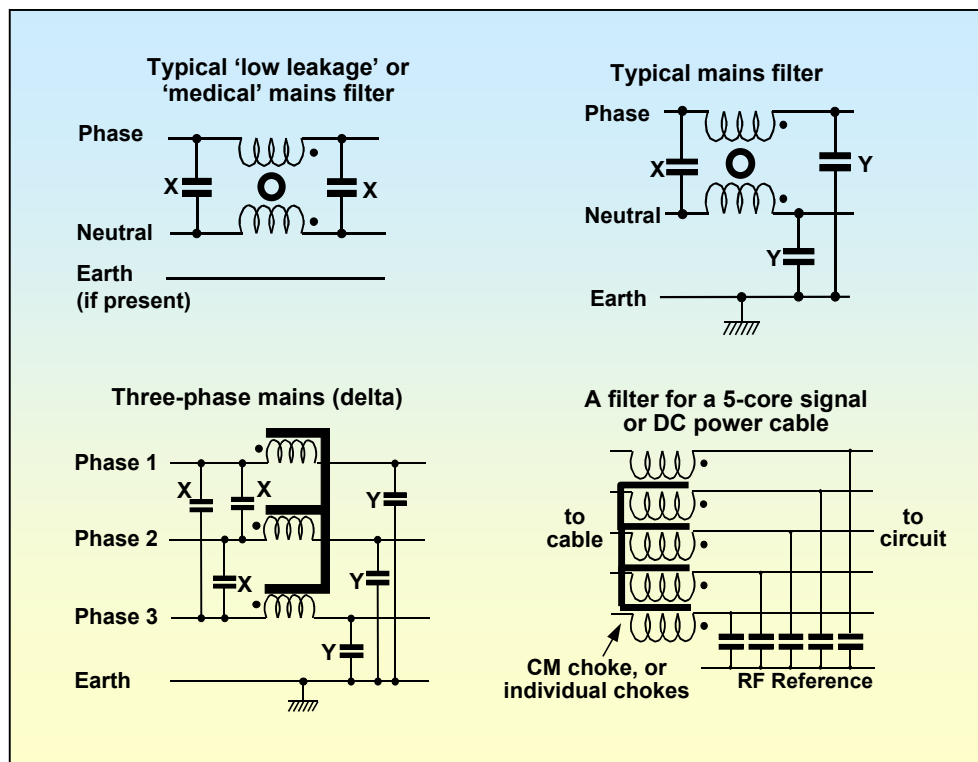
EMC pre-compliance testing begins with the lowest-cost filter options fitted, and if these turn out to be inadequate different components or higher-specification filters are fitted until the tests are passed. An alternative and arguably better approach starts off with the highest-specification filtering, to make sure no other problems exist – and when the tests are passed the filters are then reduced in cost to what works well enough. For products containing digital and switch-mode electronics, it is usually enough to use pre-compliance

emissions tests – low emissions usually being a sign of adequate immunity, but only the full suite of emissions and immunity tests will prove whether adequate filters have been employed.

Such approaches help avoid the ‘no room for the filter’ problem, which could require a complete redesign of the product. Most engineers soon gain experience with their own technologies and applications, permitting smaller-sized filter pad patterns – but the resulting design rules must be considered afresh whenever construction, ICs, or technology changes (e.g. due to a die-shrunk microprocessor, see Part 1 of [7]).

It is best if filter performance provides an ‘engineering margin’ of at least 6dB beyond the emissions or immunity requirements of the test standards, to allow for device and assembly tolerances, measurement accuracy, etc. If the EMC tests are less accurate than those achieved by nationally accredited laboratories (e.g. by [10]) it is best to allow even greater margins for error, remembering that EMC test repeatability at a given test laboratory is usually no better than  $\pm 4\text{dB}$ , and repeatability between test laboratories that have been accredited by the same accreditation body is often no better than  $\pm 10\text{dB}$ .

Some types of interconnections, such as ribbon cables using single-ended signalling, suffer from high levels of DM emissions and immunity, and it might prove impossible to achieve adequate EMC by using filtering alone – because the necessary filtering attenuates the wanted signals by too much. In such cases it may be necessary to use shielded interconnections (see Part 1 of [7]) to attenuate frequencies that can’t be filtered sufficiently without damaging the wanted signals. Sometimes it is most cost-effective or necessary to use filtering and shielding at the same time.



**Figure 3N** Typical single-stage filters, using CM and DM filtering techniques

Figure 3N shows how CM and DM filtering techniques are combined in three examples of simple mains filters. These are called single-stage filters, because they only use one inductive element. Because mains-powered products must isolate the hazardous mains voltages from any touchable parts of its body for safety reasons, their CM emissions will generally have a high source impedance, so they are firstly attenuated with capacitors between phase and the RF Reference, and subsequently with a CM choke, to maximise the impedance discontinuities for the CM noises (see Figure 3E).

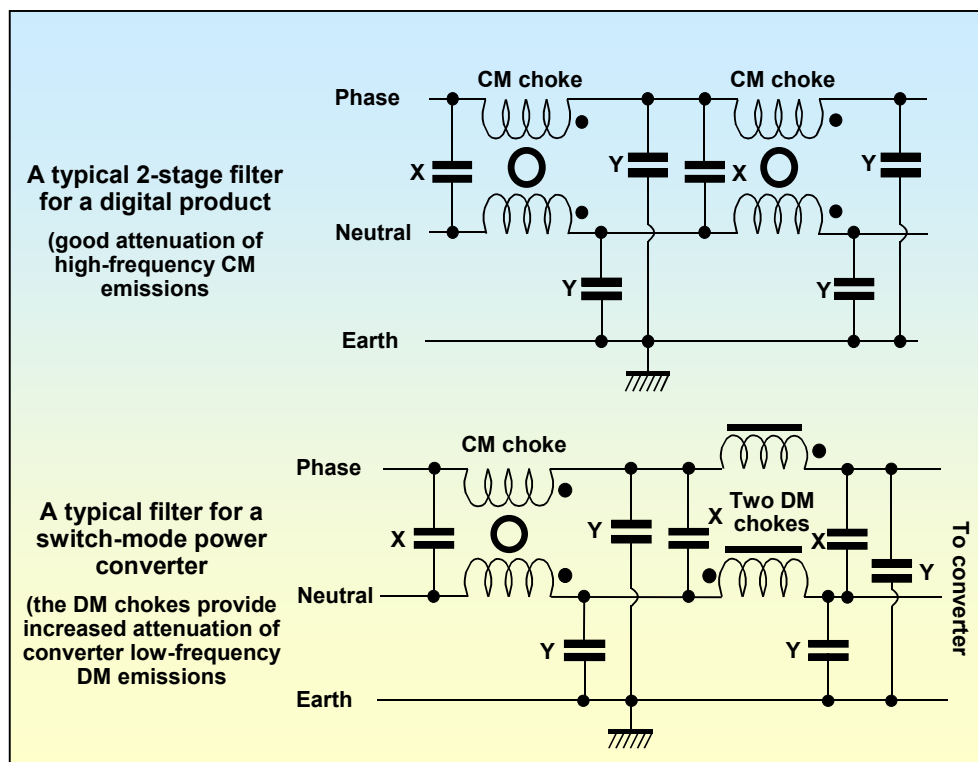
Because products draw power from their phase and neutral conductors, their DM emissions tend to have low impedance. So they are firstly attenuated by a DM choke (maybe the leakage inductance of a CM choke), and subsequently by a capacitor between the phases, to maximise the impedance discontinuities for the DM noises.

The low leakage (often called ‘medical’) filter shown in Figure 3N has no capacitors between phase and earth, so that it may be used in medical applications where very low earth leakage currents are required to protect patients (maximum 50 or 60Hz leakages for some medical products can be as low as  $10\mu\text{A}$ ). This type of filter tends to rely on larger, higher-impedance CM chokes, and its lack of ‘earthy’ capacitors can sometimes make it useful in applications other than medical.

Figure 3N also shows an example signal filter using a CM choke in an LC filter type. Military signal cable filters tend to rely on C-only and  $\pi$  types, probably because most traditional military equipment has a substantial and well-engineered RF Reference Plane between items of equipment (die-cast metal boxes electrically bonded by multiple mounting bolts directly onto metal surfaces in metal-bodied vehicles) and between items of equipment and their electrical power sources (generators and/or batteries).

Some civilian applications (such as digital telecommunications exchanges, computer rooms, and semiconductor manufacturing) achieve or approach such high-performance RF References in systems and installations, but most domestic, commercial, and industrial products are used in systems and installations that rely on networks of green/yellow insulated 'safety earth' wires for their reference, which of course have very high impedances and resonances at RF frequencies. The most predictable signal filters in such applications tend to be R, L, RC, LC, or Tee types (using soft ferrites for any Ls). These types of filters impose lower levels of RF currents on the RF Reference than C-only or  $\pi$  filters, reducing the RF potential differences between different parts of the RF Reference and helping to reduce CM emissions as a result. As military vehicles use new materials such as carbon fibre, their RF Reference Planes suffer higher impedances, and they may find R, L, RC, LC, or Tee filters more cost-effective than C or  $\pi$ .

The use of a CM choke as shown in Figure 3N, instead of a series of individual ferrite beads, can allow substantial CM filtering to be achieved at frequencies as low as 150kHz, whilst allowing wanted (DM) signals of 15MHz or more to pass through unattenuated. There are inevitable tolerances between individual components of the same type, so (all else being equal) the CM attenuation of a filter that uses a single CM choke for all the conductors in a cable will give superior CM attenuation than one using a row of individual ferrites. However, CM chokes are often more costly than the equivalent number of single ferrites, and careful PCB layout and component choice can achieve a design that can be fitted with a CM choke if the single ferrites aren't adequate, with just a few minutes with a soldering iron, instead of a PCB design iteration.



**Figure 3P Some typical 2-stage mains filters using CM and DM filtering techniques**

Mains filters with more than two stages are frequently used, and Figure 3P shows two typical designs. The multiple filter stages provide more impedance discontinuities for DM and CM noises, hence more attenuation. Standard filter products are available with up to four stages, and with ratings of up to 100 Amps, single or three-phase. These are the types of filters that are often required for use with high-power variable-speed AC motor drives, uninterruptible power supplies (UPSs), or other powerful switch-mode power converters.

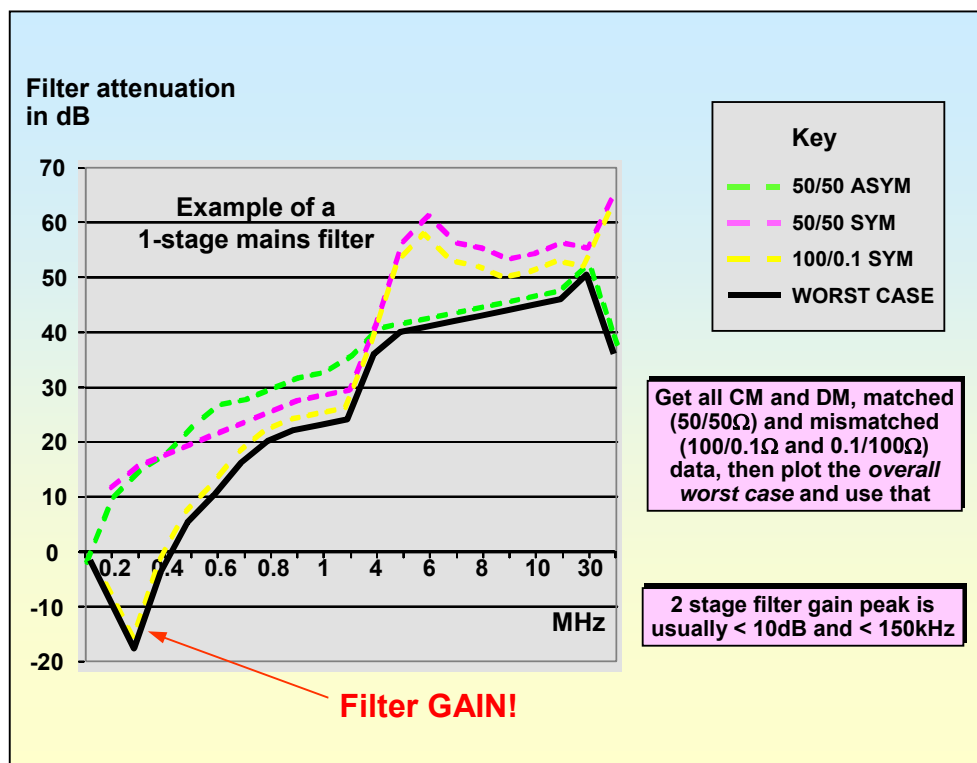
### 3.2.9 Problems with real-life supply impedances

The CM and DM impedances of the public AC mains supply can vary from about  $2\Omega$  to  $2,000\Omega$  depending on the frequency and time of day. All filters that use inductors and capacitors are resonant circuits, with their resonant frequencies and hence their attenuations depending critically on their source and load impedances.

But most filter datasheets are based upon CISPR17 measurements taken with  $50\Omega$  source and load impedances, which means that their specifications are always better than their real-life performance.

Single-stage filters are very sensitive to source and load impedances, and have a resonant peak that provides *gain*, rather than attenuation, when operated with certain source and load impedances. Single-stage filter gain usually pops up in the 150kHz to 10MHz region and can be as bad as 20dB, so it is possible that fitting a mains filter with a good ( $50\Omega/50\Omega$ ) specification can actually *increase* emissions and/or *worsen* susceptibility. It is not uncommon to have a switch-mode power supply or inverter motor drive with excessive emissions between 200kHz and 1MHz, and fit a low-cost filter with a data sheet that shows sufficient attenuation to pass the tests – only to find that the filter increases the emissions!

To avoid this situation, only consider filters whose manufacturers specify both CM (sometimes called ‘asymmetrical’) and DM (sometimes called ‘symmetrical’) performance, for both matched  $50\Omega/50\Omega$  and mismatched sources and loads. Mismatched figures are taken with  $0.1\Omega$  source and  $100\Omega$  load, and vice versa, using the CISPR17 test standard that is also used for  $50\Omega/50\Omega$  tests. Drawing a line that represents the worst-cases of all the different datasheet curves results in a filter specification that can generally be relied upon – providing the filter is not overloaded or overheated (see earlier) and is installed correctly (see later). Figure 3Q shows an example of estimating a filter’s worst-case attenuation curve.



**Figure 3Q Deriving reliable estimates of real-life mains filter attenuation from manufacturer's data**

Mains filters with two or more stages (some examples are shown in Figure 3P) have at least one internal circuit node, and these have impedances that do not depend as much on the source or load impedances. As a result, when installed correctly and not overloaded or overheated, they are more likely to provide real-life attenuation that approaches their  $50\Omega/50\Omega$  datasheet specifications. Of course, they are larger and cost more than simpler single-stage filters, so if they might be required – the design should allow sufficient room.

Multi-stage filters still suffer from gain with some combinations of input and output impedances, but when designed correctly adding more stages reduces the maximum gain and the frequency at which it occurs. For example, some manufacturers' data sheets show that their two-stage mains filters have gains of up to 10dB at frequencies between 10 and 100kHz.

Figure 3Q and the above discussion concerned mains filters, but exactly the same resonant gain issues arise for signal filters that use  $L$ s and  $C$ s, at frequencies below about 30MHz. Signal filters almost always use soft ferrites, which are resistive in the upper part of their frequency range and so do not cause resonances there – but they are inductive in the lower part so can cause resonances at those frequencies. Unfortunately, few if any signal filter manufacturers provide attenuation figures for source or load impedances other than  $50\Omega$  – so if RC, resistive Tee or  $\pi$  filters are not good enough and inductive components must be used, it might be best to experiment.

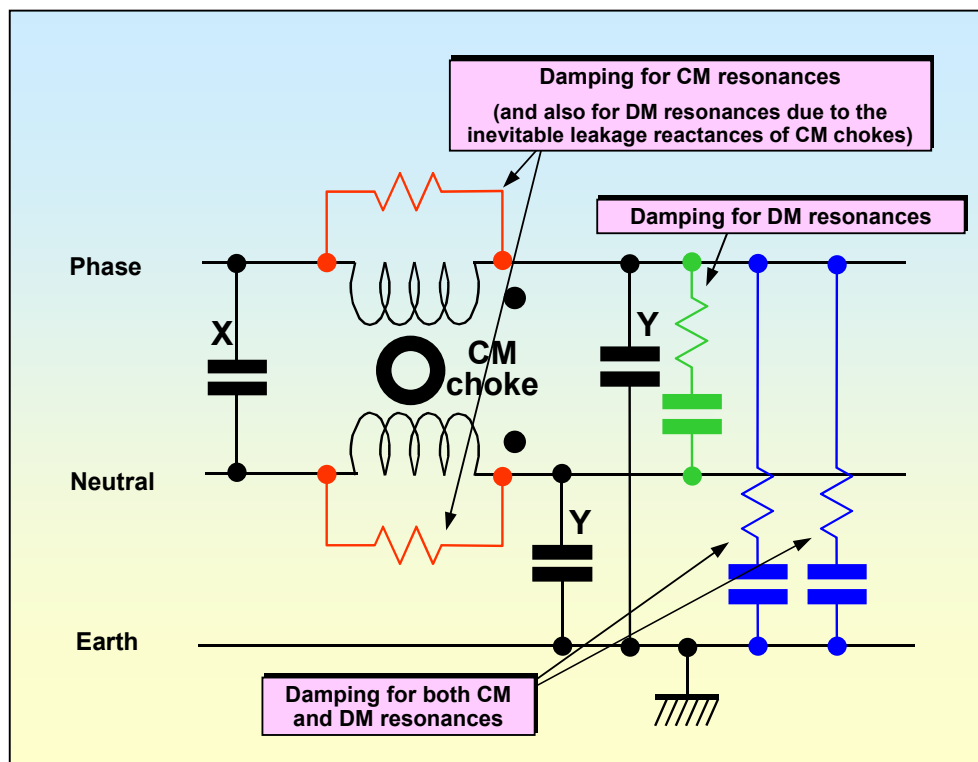
Experiments should take place at an early design stage, using sample filters and an RF signal generator and oscilloscope or spectrum analyser (if you don't have a network analyser), to see what attenuation can realistically be expected when used with the actual source and load impedances in your application. The differences between data sheet figures and real-life attenuation can be dramatic, and you do not want to discover such interesting effects during the EMC testing of a supposedly finished design!

### 3.2.10 Problems with real-life switch-mode converter input impedances

Switch-mode power converters (e.g. switch-mode regulators, switch-mode amplifiers, DC/DC converters or DC/AC inverters) can have a negative dynamic input resistance at their power input terminals. This can interact with the impedance of an input filter, resulting in instability and even oscillation that can destroy the switch-mode converter and/or other equipment connected to the same power source. The problem is mainly caused by the series inductance in the filter, which can be compensated by using a larger value of capacitance connected across the input terminals, and also by damping the inductance (see 3.2.11 below).

[5], written in 1973, describes these issues, and defines the conditions required for oscillation to take place. It also describes methods for preventing the instability and/or oscillation. Many more references can be found on the Internet, by searching with appropriate terms.

### 3.2.11 Damping filter resonances that cause gain



**Figure 3R Damping the filter resonances that cause gain**

All LC filter circuits resonate, and at resonance they have a gain that depends upon the amount of loss in the circuit. 3.2.9 discussed resonances that can be caused in power supply filters by variations in their input (power source) impedances, whereas 3.2.10 discussed resonances that can be caused by the negative input impedance characteristics of switch-mode power converters. Filter resonances can amplify surge overvoltages, as discussed in 3.5.5, increasing their potential to cause damage.

Loss is caused by resistance, but most L and C filter components are designed to have low internal resistances to minimise their internal heating – so their resonant gain can be high. Figure 3R shows how resistors can be added to an example mains supply filter circuit to provide damping that reduces the resonant gain. Such circuits cannot eliminate resonance entirely in any practical filter, but in some applications they can tame the resonances sufficiently well.

Although a simple single-stage mains filter is shown in Figure 3R, the resistive damping techniques it shows can also be used for other types of power and signal filters containing inductors.

### 3.2.12 Filters and safety

Class 1 products have a protective (safety) earth conductor connected to their metal structures, which are also their RF References, and mains filter capacitors connected to the filter's earth/ground cause leakage currents in

the safety earthing/grounding system that can be dangerous. The maximum limits for these currents should be no larger than those specified by the appropriate safety standard(s) for the type of product concerned, e.g. IEC/EN: 60950, 61010-1, 60601-1, 60335-1, 60204-1, etc. Typically: double insulated products (no protective earth connection) must have  $<0.25\text{mA}$ , Class 1 protectively-earthed *portable* products must have  $<0.75\text{mA}$  and *fixed* ones must have  $<3.5\text{mA}$ . Class 1 protectively-earthed industrial fixed products might be permitted to have earth-leakage currents up to 5% of the rated phase current – when specified warning labels are fitted – but at the other extreme patient-connected medical products can be limited to  $<0.01\text{mA}$ .

In systems and installations the earth leakages from numbers of filters can build up to create large earth currents that can be very dangerous indeed. Tens of amps of leakage current is not unusual in the main protective earth terminal of a modern office building, due to its very many PCs and PC monitors, each with their mains filters leaking up to  $3.5\text{mA}$ .

Mains filtering is an area where EMC requirements can often come into conflict with safety needs, and of course safety must always come first. So always take the relevant safety standards into account when designing or selecting mains filters, remembering that most X and Y capacitors have tolerances of  $\pm 20\%$ .

Mains filters sold for 50Hz use may generally be used on supplies from DC to 400Hz with the same performance (but check with their manufacturers). Also remember that the earth-leakage currents caused by filter capacitors connected to the earth/ground will increase as the supply frequency increases, so filters that meet the relevant safety standards at 50Hz might not comply at 60Hz, and may be decidedly dangerous on 400Hz.

Capacitors connected between the phases and RF Reference should always be approved to all relevant safety standards for both the application and the voltage. They will usually be Y1 (for double insulated products) or Y2 (for Class 1 products with an earthed protective bonding network). Capacitors between phases should also be safety approved, e.g. types X1 and X2, more to prevent fire hazards than to prevent shocks.

It is always best to use mains filters (or components) for which third-party safety approval certificates have been obtained and checked for their authenticity, filter model and variant, temperature range, voltage and current ratings, and the application of the correct safety standard. Forged safety approval certificates are not unknown, even from manufacturers who might be expected not to run such risks, so I always recommend that certificates are checked with their issuing Approvals Bodies, who in my experience are always happy to help, to make sure they are not forgeries.

### 3.3 Filter installation issues

Real-life filter performance is totally dependant on how they are installed, especially on the impedance of the RF Reference and the impedance of the method used to electrically bond the filter to its RF Reference. Not only should these impedances be much lower than that of the shunt capacitors in the filters, they should also allow the internal and external CM surface currents to find their optimum return paths. This section discusses these issues, and the practical installation guidance that results.

#### 3.3.1 Input and output conductors

Stray RF coupling between the conductors associated with their unfiltered and filtered sides easily degrades filter attenuation. This problem is generally worse at higher frequencies, because the impedances of stray capacitances and stray mutual inductances reduce as frequencies increase, increasing the amount of stray coupling bypassing the filter. Many engineers have been very surprised by the ease with which high frequencies will bypass ('leak around') a filter, given half a chance.

In an unshielded enclosure, filters should be positioned as near to the point of entry of the cable as possible. The maximum possible separation distances should be maintained between the filter's external and internal cables, and between all of the conductors associated with the circuits on either side of the filter. Conductors in air should be spaced *at least* 100mm apart, more if they are routed in parallel for more than a few centimetres. Closer spacings might be acceptable for PCB traces and components – but only if they are much closer to the PCB's RF Reference Plane than the spacing between them. Filter input and output conductors should never, *ever*, be bundled together, or share the same cable or cable route, unless they are each well shielded. See [6] for how to shield conductors effectively.

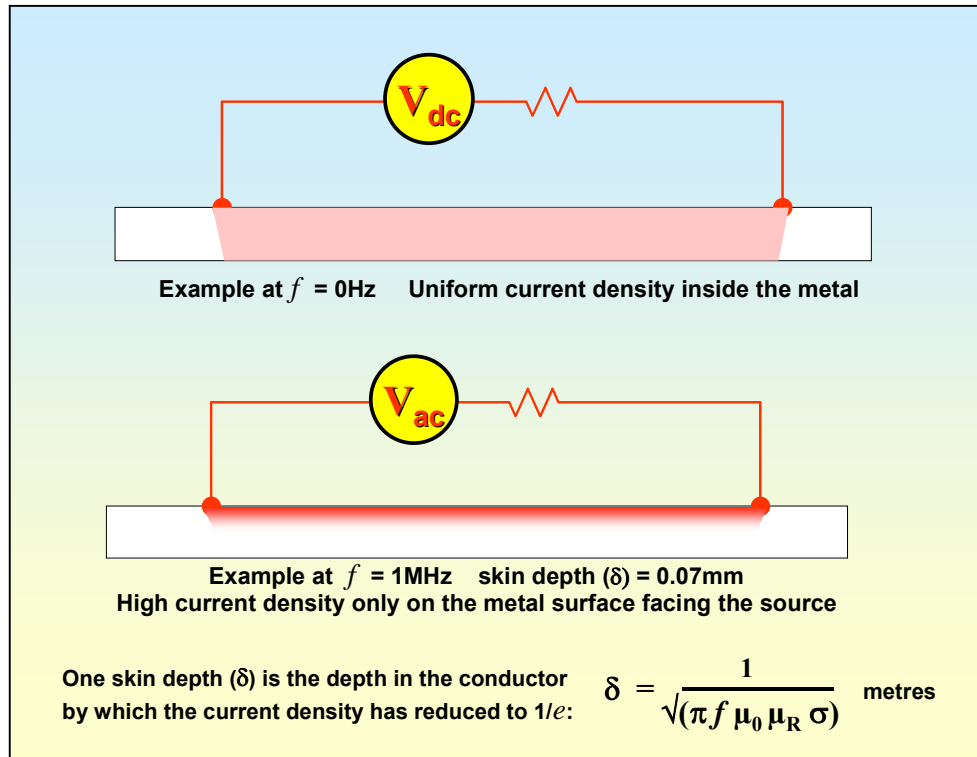
Where the enclosure is shielded, it is essential to mount the filter in the wall of the enclosure, with the filter's body electrically bonded directly to the shielding surface of the wall, otherwise both the filtering and shielding performances will be degraded by stray coupling around the filter. The type of filter required is often called a bulkhead-mounting, or through-bulkhead filter, because it fits through the metal wall (bulkhead) that it is mounted upon. The shielded enclosure considerably reduces the stray coupling between the filter's input and output. It may even be necessary to fit a conductive EMC gasket around the aperture in the shield where the filter is mounted, for the maximum possible filtering and shielding. Issues of filtering with shielded enclosures are covered in more detail below.



### 3.3.2 Skin effect and the flow of surface currents

Where the frequencies to be attenuated are not very high, it could be acceptable to use a few direct bonds, or a few millimetres of wire or braid to provide the electrical bonding to the RF Reference, providing the impedance of the bonding method is much less than that of the filter's shunt capacitors at the highest frequency of concern. But to understand how to assemble/install filters correctly for good RF performance at high frequencies, we need to understand 'skin effect'.

All RF currents travel as surface currents, because all conductors have a skin effect that effectively causes them to shield their inner depths from RF currents. Figure 3S illustrates the general principle, and shows that as the frequency increases, the current is constrained to flow closer to the surface, increasing the current density at the surface of the conductor. One skin-depth is the depth into the conductor by which the current density has decreased to  $1/e$  of what it was – about 0.368. By two skin-depths into a conductor the current density has reduced to  $(1/e)^2$ , or 0.135, by three skin-depths it has reduced to  $(1/e)^3$ , or 0.05, and so on.



**Figure 3S Skin effect: examples of cross-sectional current density in a metal sheet**

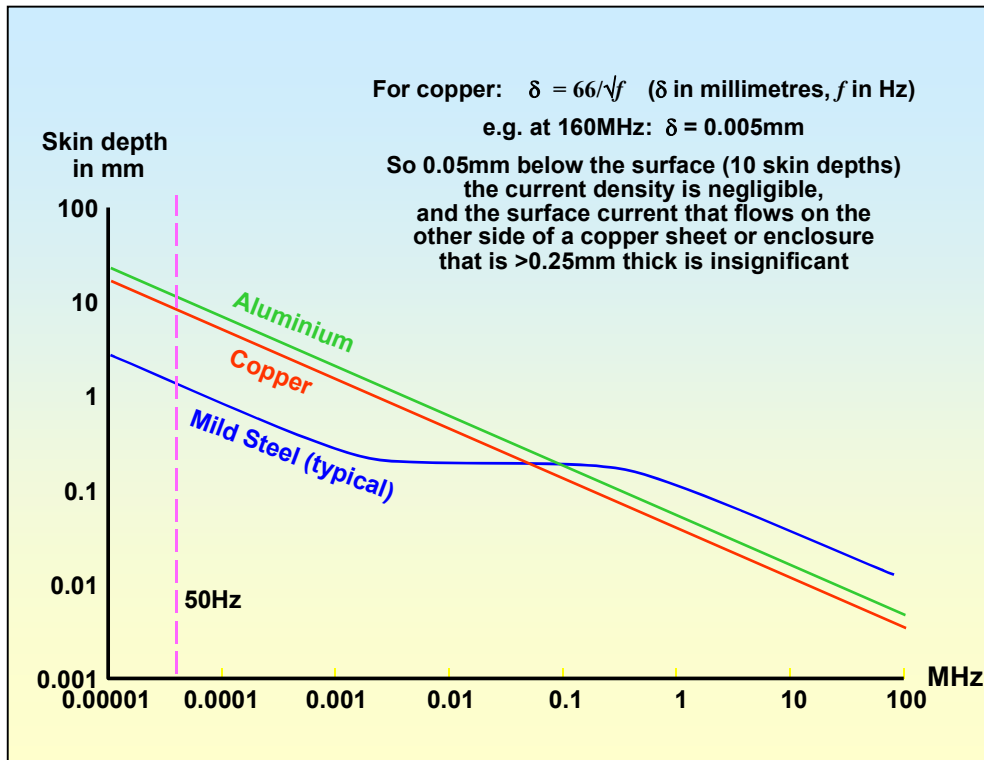
Figure 3S gives the formula for calculating one skin-depth  $\delta$ , where  $\mu_0$  is the permeability of free space ( $4\pi \cdot 10^{-7}$  Henrys per metre);  $\mu_R$  is the (dimensionless) relative permeability of the conductor material (most common conductors, such as copper, aluminium and tin, have a  $\mu_R$  of 1.0) and  $\sigma$  is the conductivity of the conductor material in mho/metre. Copper has a nominal volume resistivity  $\rho_v$  of  $1.72 \cdot 10^{-8} \Omega\text{-m}$ , giving it a nominal conductivity of  $58 \cdot 10^6$ , so one skin-depth in nominal copper is given by  $\delta = 66/\sqrt{f}$  ( $\delta$  is given in millimetres when  $f$  is in Hz). For example, at 160MHz: one skin-depth is 0.005mm, so 0.05mm below the surface of a copper conductor, the RF current density is 0.0025 of the density at the surface, an attenuation of 52dB.

Figure 3T shows graphs of skin-depth versus frequency for some common materials, to save having to find out the values of their conductivity and calculate  $\delta$ . Mild steel is shown as an example of a ferromagnetic material (nickel is another), and to show that their high values of  $\mu_R$  result in smaller skin-depths, but also that their permeability is frequency-sensitive and disappears above some critical frequency.

[11] contains information on the material properties of a wide range of conductors, for calculating skin depth, and also a great deal of other useful information for designers. [12] is a useful source for information on skin-depth.

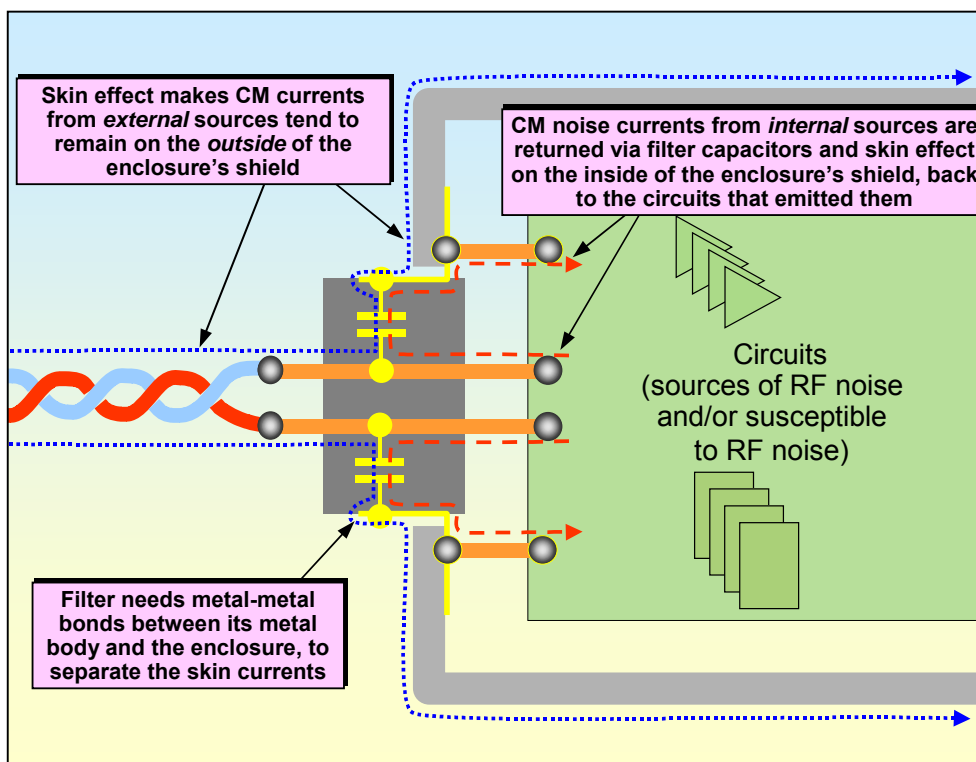
Above a few tens of MHz most conductors and metal items (such as the cases of filters) are several skin-depths thick, so RF currents travel as surface currents in them. Taking this phenomenon into account in the design of a filter's assembly/installation is essential for the achievement of good emissions and/or immunity performance.





**Figure 3T** Graph of skin-depth ( $\delta$ ) for copper, aluminium, and mild steel

Figure 3U shows how providing a continuous metal bond between a filter and the shielding enclosure of a product ensures that the external CM noise currents do not enter the enclosure and cause interference and immunity problems, and the internal CM noise currents remain inside the enclosure and do not escape to cause emissions problems. Figure 3U shows a simple capacitor filter, but the principle applies to all types of filters.



**Figure 3U** Filter RF bonding and the skin effect

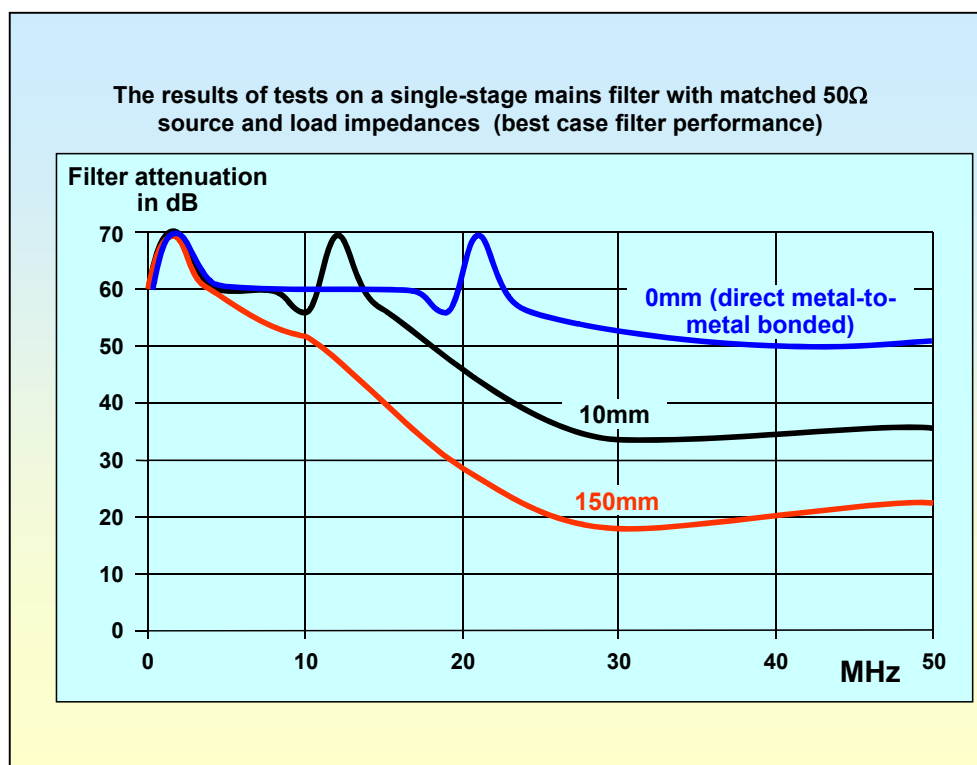
As a result, the optimum way to bond a filter to its RF Reference Plane, for the best performance at the highest frequencies, is what is often called '360° direct metal-metal contact' – meaning that the filter's metalwork and the RF Reference Plane are in direct contact with each other *all around the periphery* of the filter (hence the term 360°).

Commercial and industrial conducted emissions standards generally only measure up to 30MHz, and at such low frequencies it is often sufficient to bond a filter to an enclosure with a single direct metal-to-metal connection between the filter's case and the enclosure. Where the filter is only required for low frequencies, e.g. below 1MHz, it may even be possible to use a very short length of wire or braid to connect its metal case to the enclosure metalwork, plus of course the enclosure will not need to be a proper shield either (enclosure shielding will be covered by Part 4 of this series). But there is a synergistic relationship between filtering and shielding, discussed in more detail in the following section.

Filters that employ capacitors connected between power or signal conductors and the RF Reference depend upon the RF Reference – and their connection to it – having a much lower impedance than the filter capacitors, at all of the frequencies to be attenuated. The connection between the capacitors and RF Reference should be very short and direct, less than one-hundredth of a wavelength long at the highest frequency to be attenuated, and should also have a very low inductance. This usually means that wires or even braid straps cannot be used to electrically bond filters to the RF Reference Plane, except for low frequencies (say, below 1MHz).

Figure 3V, which is taken from [13], shows the sorts of bad effects that even a short length of interconnecting wire can have on a standard single-stage mains filter even when measured with 50Ω/50Ω source and load impedances – its best possible case. If the 10mm wire were replaced with at least one direct metal-to-metal bond, performance at 30MHz and above would improve dramatically.

It is acceptable to fit green/yellow wires of any length to mains filters, for safety reasons, as long as there is also at least one direct metal-to-metal electrical bond between the filter's metal case and the product's RF Reference. When a mains filter's metal-to-metal bonds have been designed to maintain a very low impedance over the lifecycle of the product, there is no need for a green/yellow 'safety earth' wire as well – but safety inspectors are generally much more reassured when they can see a green/yellow bonding wire with anti-vibration anti-corrosion connections at both ends. (But, as discussed above, it would be a mistake to assume that the green/yellow safety wire was adequate for achieving the filter's EMC performance.)



**Figure 3V Comparison of two lengths of filter earth-bonding wire**

### 3.3.3 The synergy of filtering and shielding

Some mains filter manufacturers only design and specify their filters to provide attenuation over the frequency range of the conducted emissions tests (typically up to 30MHz for commercial and industrial products), to keep costs low. Unfortunately, if such filters have poor attenuation above 30MHz, they will degrade the shielding effectiveness (SE) of a shielded enclosure above that frequency by permitting RF signals to leak out via the filtered cables – resulting in problems for both emissions and immunity.

It does not matter what is the ostensible purpose of a conductor, e.g. mains or DC power, audio, whatever – if its filtering and/or shielding provides less attenuation than is required for the shielded enclosure, it will degrade the SE of the enclosure. The filtering and/or shielding of cables used for audio, mice or keyboards are often

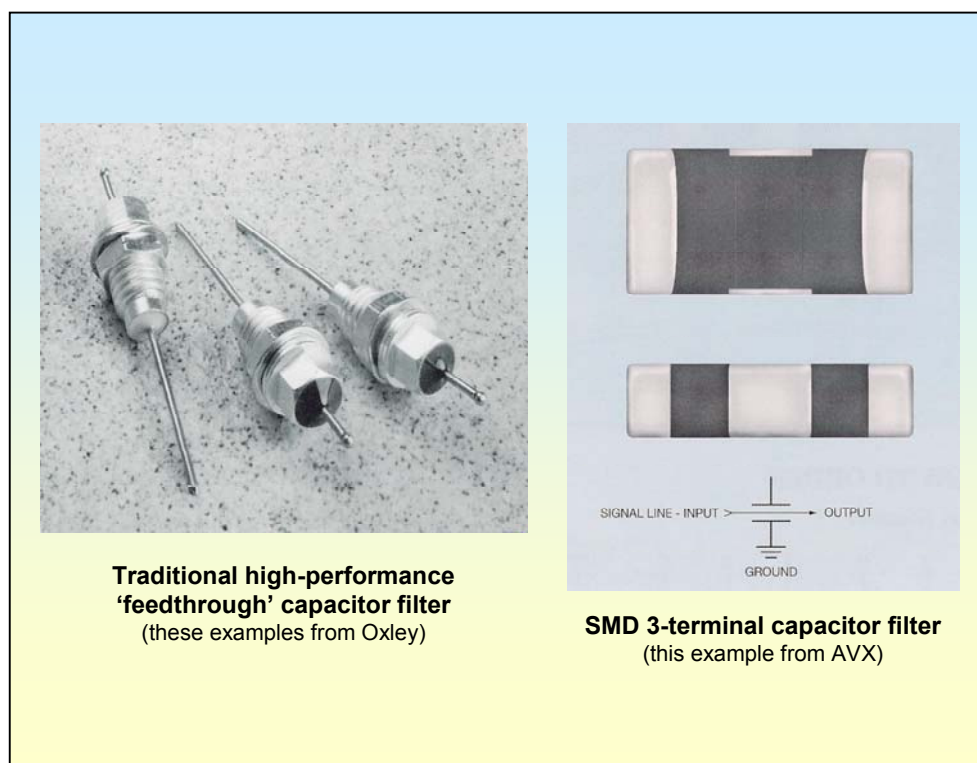
ignored when they exit a shielded enclosure. The assumption is usually that the signals they carry will not cause a problem for EMC. But this overlooks the fact that *all* conductors or *whatever* type or signal designation always behave as 'accidental antennas' (see [6]), very readily picking-up EM noises on either side of a shielded barrier and retransmitting them on the other side – unless specifically prevented from doing so by the application of shielding and/or filtering.

If good high frequency shielding is required, all unshielded cables that enter the enclosure (including mains) must be filtered with good attenuation at the highest frequency of concern for shielding purposes. So where shielding is required up to 1GHz (for example), only employ filters with data showing good attenuation up to at least 1GHz. Few mains filters intended for commercial and industrial equipment specify attenuation above 100MHz, so additional high-frequency filtering might be needed. However, some filter manufacturers (e.g. EMC Solutions) specify their filters up to 1GHz.

### 3.3.4 Assembly and installation techniques for filters that penetrate shields

As discussed above, the performance of shielded enclosures can easily be degraded by RF noise that 'leaks' out along the cables that enter and exit the enclosure. The shielding/filtering synergy issues discussed above are vital considerations when high levels of shielding or filtering are required (e.g. >40dB) at frequencies >100MHz.

The design of shielded cables was covered in [6], and the design of shielded enclosures will be covered in Part 4 of this series. This section discusses how filters should be installed in shielded enclosures so that they do not permit RF noises to pass through them that could compromise the SE of the enclosure.



**Figure 3W Three-electrode ('three-terminal') and feedthrough capacitors**

Figure 3W shows an example of a 'feedthrough' capacitor specifically designed for use where unshielded conductors penetrate a shielding enclosure, which could be a product's enclosure or an internal shielded volume. Another, higher current style of feedthrough capacitor was shown in Figure 3A. Feedthrough capacitors have three terminals, for input, output and 'ground'. The signal to be filtered enters at one side of its electrodes and exits at the other, having to pass the ground electrode as it does so. The middle 'ground' terminal connects directly to the shield, using a 360° electrical bond so that the internal and external surface currents stay separated on either side of the shield, as shown in see Figure 3U, allowing the shield to function correctly. If designed correctly, the shielded enclosure prevents stray coupling between the capacitor's input and output terminals, and also provides the filter with an RF Reference Plane with negligible impedance at the highest frequency of concern, all of which helps the filter employing the feedthrough capacitor to achieve the best performance it is capable of.

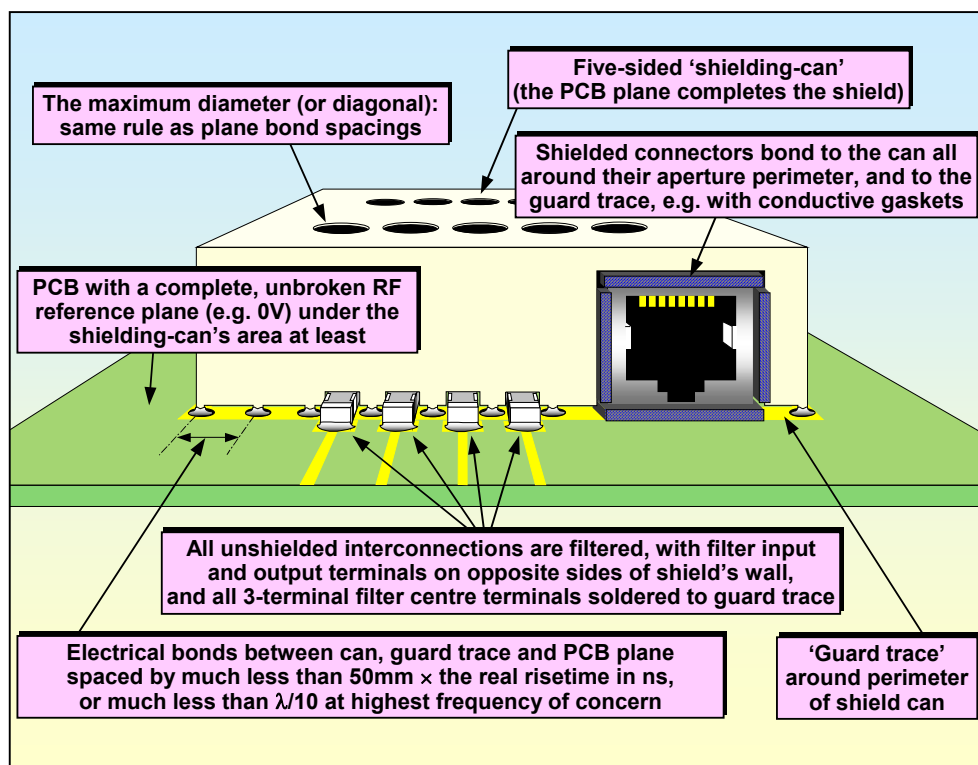
When used as (or in) filters, traditional feedthrough capacitors such as the ones shown in Figures 3W and 3A provide much better attenuation, at much higher frequencies, than is possible by using ordinary two-terminal capacitors. Traditional feedthrough capacitors are soldered or screwed into a shield wall and connected to the

circuits on either side by wire conductors. They are often used between shielded compartments within RF equipment, e.g. to filter the DC power that passes between the RF, IF and digital sections of an RF receiver or spectrum analyser.

Traditional feedthrough filters, such as those shown in Figure 3W, are also available as 'filter pins' in some standard connectors, such as some D-types and military circular connectors. (Note that not all connectors with built-in filters use feedthrough filter pins, some use discrete components on miniature internal PCBs, which will not achieve as good an attenuation at the highest frequencies.)

Traditional feedthrough filters are not favoured for modern volume-manufactured products because of their high component cost, and the high cost of their manual assembly and the assembly of the wires they connect to. Volume-manufactured products prefer to use SMD components automatically assembled on PCBs – but since a true feedthrough capacitor cannot be automatically assembled, three-terminal capacitors have been developed to fulfil this purpose.

Figure 3W includes an example of a three-terminal capacitor intended for SMD assembly processes, and Figure 3X shows an example of how it is used in conjunction with PCB shielding. The capacitor is aligned with the shield wall so that its input and output terminals are shielded from each other by the PCB-mounted shielding-can, and the capacitor's centre 'ground' terminal is soldered directly to a guard trace that follows the wall of the shield-can and connects it to a PCB plane (almost always 0V) with a wall of via holes.



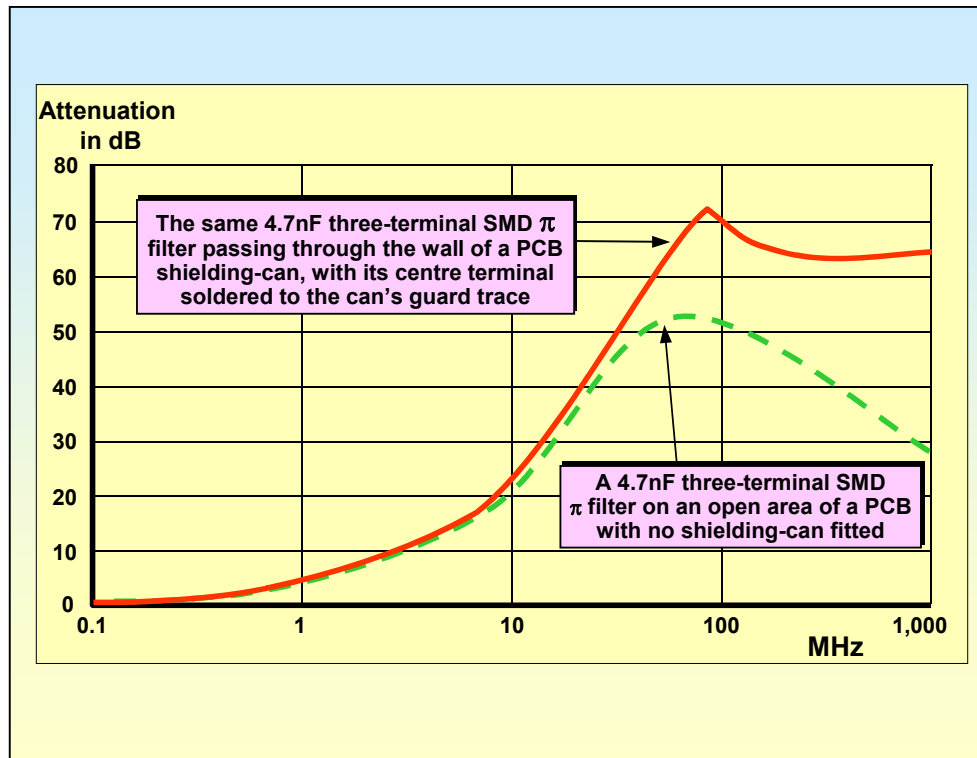
**Figure 3X Overview of PCB shielding and filtering**

The gaps that are cut out of the shield-can's wall for the bodies of the filters are known as 'mouseholes' (for reasons that should be obvious to anyone who enjoys 'Tom and Jerry' cartoons). Three-terminal capacitors and the filters that use them cannot be as good as proper 360° shield-bonded feedthrough types, because there will always be some stray coupling through the mouseholes in the shield. But careful control of the maximum dimensions of the mouseholes, and of the spacing between the via-holes connecting the shield wall to the PCB plane that provides the shield's sixth side, can nevertheless achieve excellent performance. For more details on this, see Part 2 of [9].

Figure 3Y shows attenuation of a three-terminal SMD  $\pi$  filter assembled on a PCB, and the effect of adding a PCB-mounted shielding-can in the manner shown in Figure 3X. Without the PCB's shield-can fitted, the filter performance is quite respectable at about 50dB at 100MHz, but above that frequency it falls off at 20dB per decade, so that it is only about 30dB at 1GHz, and it would presumably be about 10dB at 10GHz.

But the addition of the shielding-can reduces the stray coupling bypassing the filter very considerably, and also allows correct separation of internal and external surface currents and provides an RF Reference Plane that has a much lower impedance over the frequencies measured. The result is an attenuation of around 70dB at 100MHz, and a more-or-less flat attenuation that maintains about 65dB up to 1GHz – easily 35dB more attenuation than was achieved without the shield. It is not clear what the performance with the shield-can would

be above 1GHz, but as there is no sign of any roll-off even at 1GHz it is likely that an attenuation of at least 45dB would be achieved at 10GHz.



**Figure 3Y The synergy of filtering and shielding**

As described in section 2.6 of [6], shielded cables exiting a shielded PCB region require shielded connectors or glands that are electrically bonded to the shield-can's wall by mechanical fixings, soldering or gasketing that makes multiple connections around its periphery – preferably full 360° bonding.

The experiment whose results are shown in Figure 3Y reveals two important things:

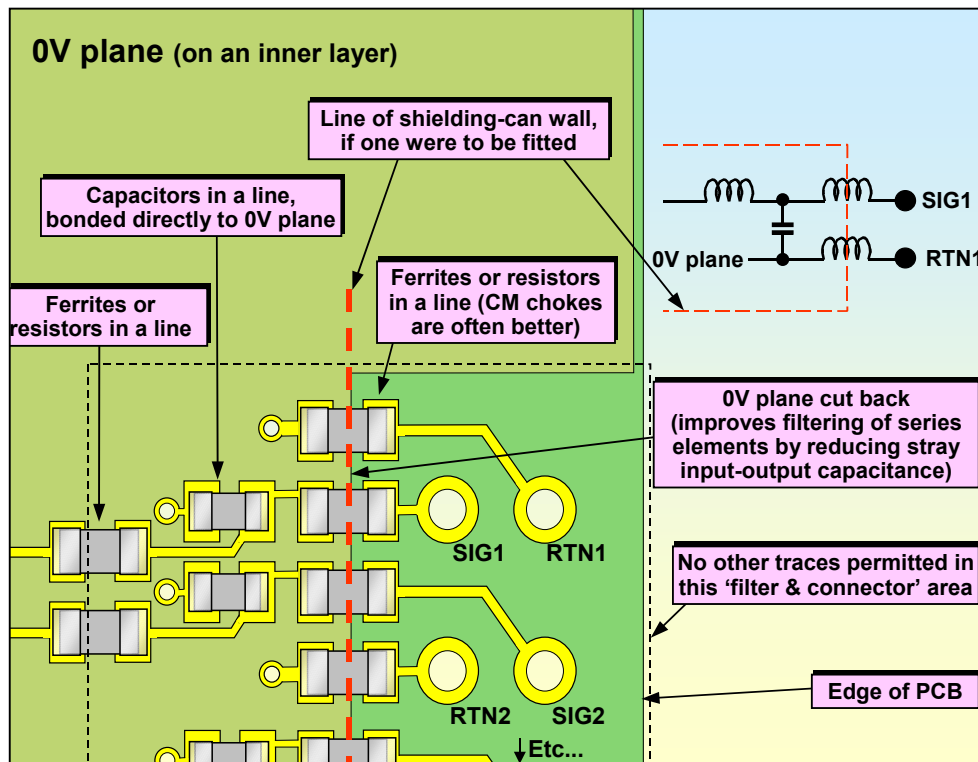
- Filters that must provide significant levels of attenuation at frequencies above 100MHz, must employ shielding techniques as well. They will not be able to achieve the required performance otherwise.
- Modern digital ICs produce large amounts of CM and DM noise at frequencies above 1GHz, and products supplied to the USA already have to comply with FCC emissions limits above this frequency. The EN standards used to achieve a presumption of conformity with the EMC Directive for products supplied to Europe will soon be changing to include emissions and immunity requirements above 1GHz – at least to 2.7GHz and maybe higher. To comply with these requirements using low-cost SMD PCB assemblies will require the use of shielding wherever GHz frequencies need to be filtered.

There are now many suppliers of PCB-mounted shielding-cans that can be used with three-terminal filters, and they have many types that can be automatically assembled like any other SMD component. Part 2 of [9] has more details on these shield-cans, and also describes a number of different PCB layouts appropriate for filtering off-board connectors. An example of one of these layouts is shown in Figure 3Z.

Where filters must penetrate the shield of a product's overall enclosure, and PCB-mounted components are not suitable, more traditional feedthrough or 'bulkhead-mounted' filters in metal cases are the best. A point to watch out for is whether the metal cases of such filters are seamless – good filters are enclosed in what are actually well-shielded enclosures themselves. Filters that have metal cases with apertures, seams or gaps in them give poor attenuation at high frequencies regardless of what their data sheet says, because they compromise the attenuation of the shielded enclosure they are assembled/installed onto.

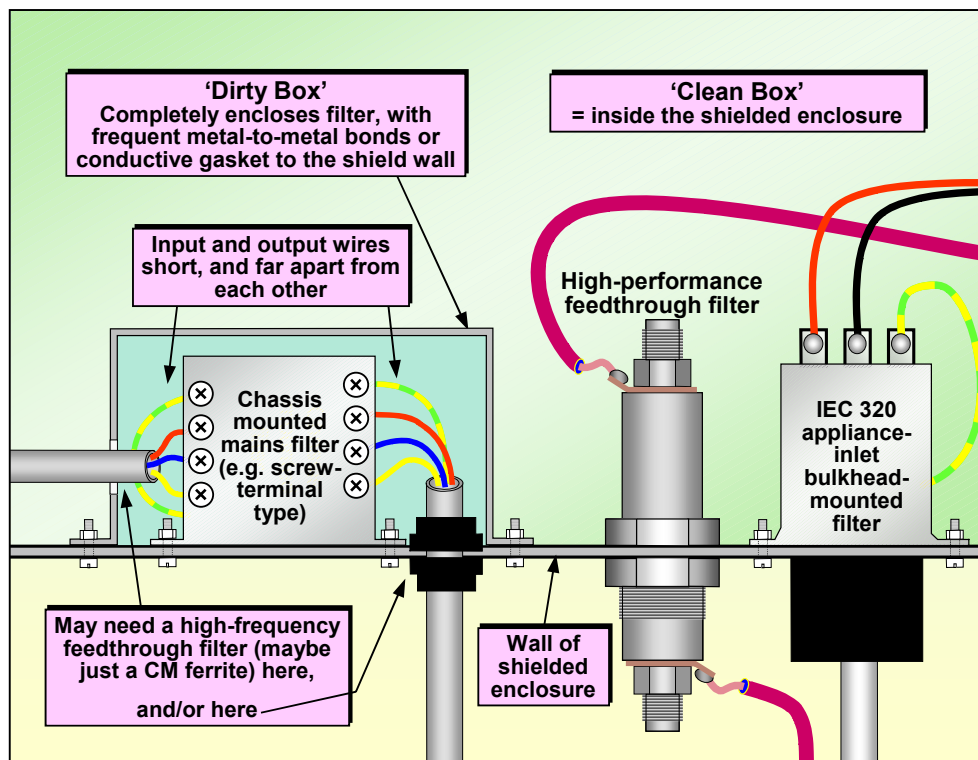
'Chassis mounted' filters include types with screw terminals, spade or blade terminals, or flying leads (Figure 3A shows some examples of chassis mounted filters with spade terminals) and cost less than proper bulkhead or feedthrough types, but cannot be assembled to shields so as to reduce stray coupling between their inputs and outputs. The result is that they are not as effective as feedthrough or bulkhead mounting types at higher frequencies, especially above about 10MHz. Their performance can be maximised by mounting them with multiple direct metal-to-metal bonds to an RF Reference Plane that is a shielded enclosure wall, or at least a very large metal plate, plus routing their input and output cables very close to the RF Reference Plane and keeping them and any circuits or components they connect to very far apart.





**Figure 3Z** Example layout for an unshielded off-board connector, using Tee filtering

However, their performance can be significantly improved by the use of what is known as the 'dirty box' shielding technique illustrated in Figure 3AA. This figure shows a shielded enclosure, and an example of the correct installation of a traditional high-performance feedthrough filter. It also shows an example of an IEC 320 appliance mains inlet connector with an internal filter. The important issue with such inlet filters is that they should have seamless metal bodies that make a direct metal-to-metal connection to the wall of the shielded enclosure.



**Figure 3AA** Mounting filters in the walls of shielded enclosures (examples shown are all power filters)

Many manufacturers have fitted mains connectors with built-in filters, relying on their mounting screws and green/yellow safety earth wire to make the necessary electrical bonds, and have found the EMC performance to be almost useless. As discussed above, the length of the green/yellow safety wire is simply too long, and a

problem with most built-in filter connectors is that their mounting screws bear onto plastic mouldings, so they don't provide any metal-to-metal connections. The correct way to install such filters is to ensure that an area of the enclosure's shield wall is free from paint or anodising, and has a highly conductive surface that will be pressed firmly against the filter's metal body when it is assembled. Sometimes it may even be necessary to bond the bodies of such filters 360° to the shield wall all around the perimeter of the filter's metal case, requiring high surface conductivity for the metalwork on both sides of the gasket, and protection from corrosion (see below).

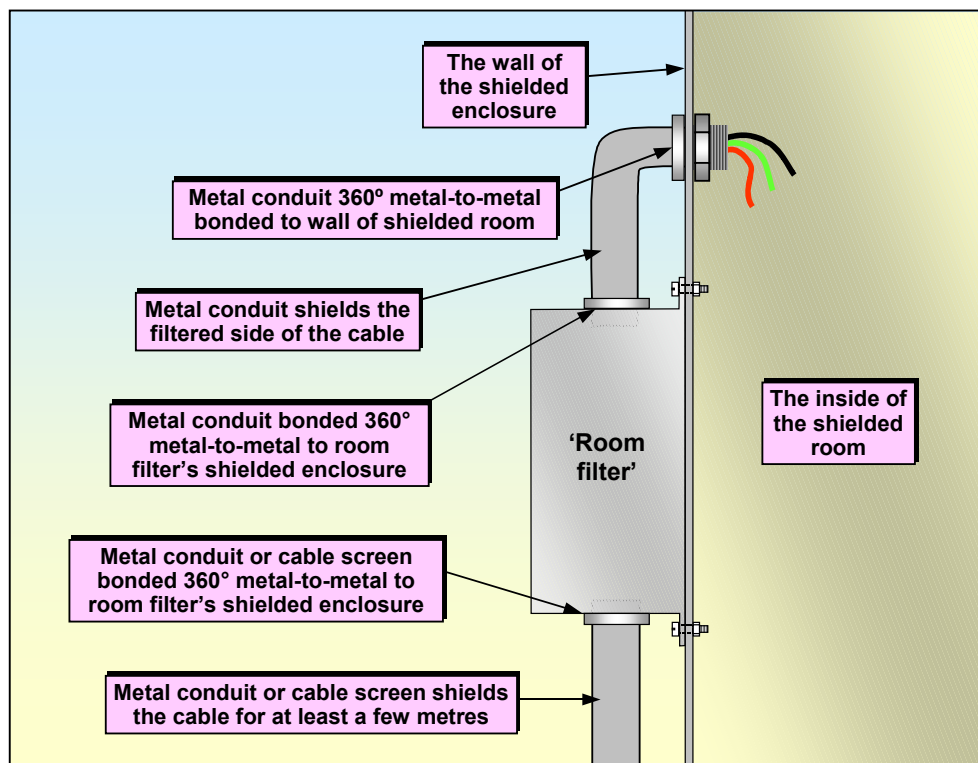
When chassis-mounted filters are applied to cables entering or exiting a shielded enclosure, the portion of the cable that enters the enclosure to connect to the filter degrades the attenuation of the filter by causing stray coupling to its other terminals. This portion of cable also degrades the SE of the enclosure by acting as an accidental antenna (see [6]), especially at higher frequencies.

To maximise the high-frequency performance of such filters and prevent degradation of the enclosure shielding, such filters should be installed using the 'dirty-box' method illustrated in Figure 3AA. The Dirty Box is a five-sided shielded cover that fits over the filter and the external cable entry, within the overall shielded enclosure. It must have metal-to-metal bonds at multiple points between its walls and the wall of the shielded enclosure, spaced apart by much less than  $\lambda/10$  at the highest frequency to be controlled, and covering the entire perimeter of the Dirty Box's walls. Conductive gaskets might help reduce assembly time by reducing the number of fixing screws, or might even be necessary to achieve sufficiently good bonding to the enclosure wall.

The filter is mounted inside the Dirty Box, with its input and output conductors kept as short and as far apart as possible, to reduce their stray coupling – but even so the higher frequencies will still couple between them. If the resulting high-frequency stray coupling is problematic and cannot be reduced by careful cable routing within the Dirty Box, soft-ferrite CM chokes and/or high-frequency feedthrough filters may be needed on either (or both) the input and output cables, fitted at the point where they enter or exit the Dirty Box.

'Shielded room' filters are also available, and although intended for EMC test chambers (as shown in Figure 3B) they can be used for shielded equipment cabinets as well. These are essentially screw, spade or blade terminal filters with two Dirty Boxes, one over the input terminals and their conductors, and one over the output terminals and their conductors, to minimise the stray coupling between input and output.

Conduit fittings are usually provided for the filtered side of room filters, to provide shielding for their conductors whilst they enter the shielded room or enclosure. Where the conduit enters the shielded room or enclosure it must electrically bond 360° at the shield wall, as illustrated in Figure 3AB. Shielded cables may be used instead of conduits, as long as they bond 360° at both ends, to the filter's case and the shielded room or enclosure wall using appropriate glands or connectors.



**Figure 3AB Mounting 'room filters' to the walls of shielded rooms**



Figure 3AC shows an overview of shielding and filtering at the level of the final system or installation. Where an electrical/electronic product has an overall shielded enclosure, all of the conductors that enter or exit that enclosure must be shielded, and/or filtered, at the point where they enter/exit the enclosure. There are no exceptions to this rule, whatever the purpose of the conductors, including safety earth wires: metal armour or draw-wires for cables, fibre-optics, or hydraulic hoses; metal pipes for gasses or liquids; metal ductwork for cables, air-conditioning, etc. Conductors permitted to be connected directly to the shield wall should be so connected, using 360° bonding techniques just as if they were cable shields (see [6]).

Unshielded conductors that are not directly bonded to the enclosure at point of entry/exit must be filtered, taking into account all of the techniques discussed above concerning the synergy of filtering and shielding.

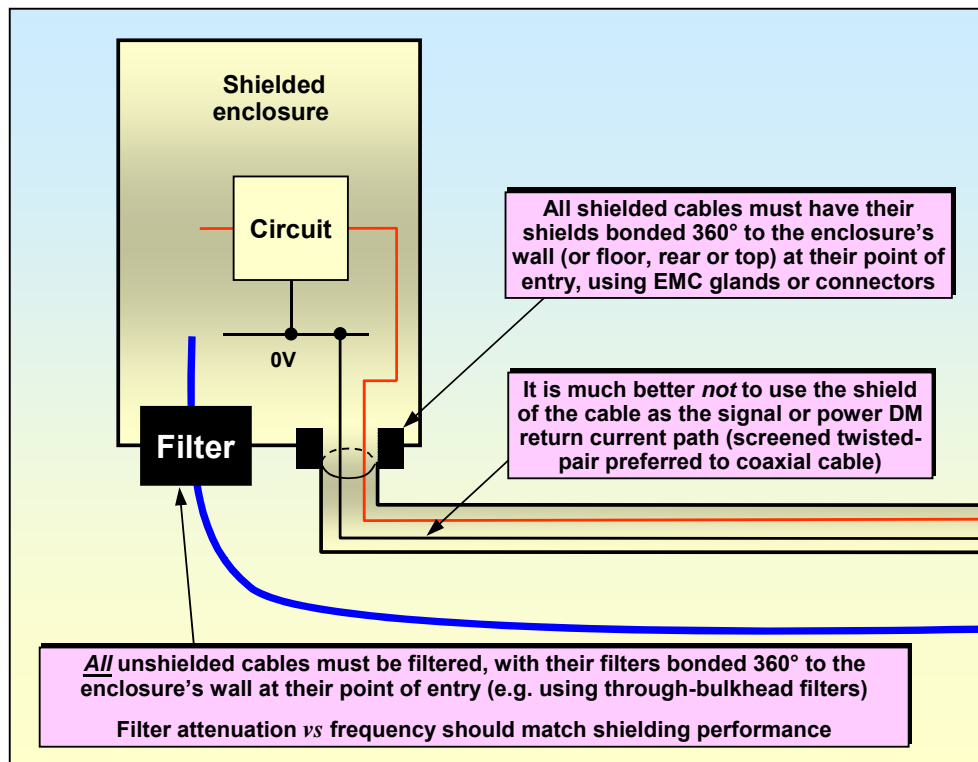


Figure 3AC Shielding and filtering at installation level

### 3.3.5 Designing to prevent corrosion

All metal-to-metal bonds associated with filters (and shielding), and all conductive gaskets, must be designed to provide low impedance for the anticipated lifecycle of the product, despite the mechanical, climatic, biological, chemical and other physical environments the product is exposed to. This generally means choosing metals, platings and gasket materials that resist oxidation, and it also means ensuring that the materials in contact are sufficiently close in the galvanic series so that they don't suffer unduly from galvanic corrosion. IEC 60950 is a safety standard but provides some useful guidance on these issues, and there is also a lot of information available freely on the Internet.

Effective 'vapour-phase corrosion inhibition techniques' are claimed to have been developed in recent years, by Cortec Corporation (<http://www.cortecVpCI.com>), and should be investigated, especially where corrosion is a significant problem.

### 3.3.6 Filters connected in series

It sometimes happens that a product is supplied with mains filtering, but its RF emissions are too high (or its immunity too low) for the equipment, system or installation it is used in. This is often a problem where a large number of identical or similar devices are used in one product or system, for example a number of low-power inverter motor drives in one industrial cabinet. Each product may meet the relevant emissions limits individually, but when a number are all operating at once the aggregate of their emissions might exceed the permitted limits.

In such situations it is tempting to simply add another mains filter, which would then appear in series with the mains filters already fitted in the products. Often a single-stage filter is chosen because the filtering requirements are only modest. The gain problems that can occur with filters with 'mismatched' source/load impedances, especially single-stage types, were discussed earlier – but sometimes connecting filters in series can result in resonances that are not present in any of the filters when they are tested individually. So adding the extra filter can sometimes create worse emissions or immunity than before.

Solutions include replacing the original filters in the products with ones that achieve higher performance, or experimenting with different types of additional filters to find ones that work well when connected in series with the filters in the products. If the circuits of the filters involved (product and additional) are known, circuit simulators such as Spice should be able to predict resonance problems in advance, and guide the choice of appropriate devices.

### 3.4 Types of overvoltage transients and surges

There are many types of transient overvoltage phenomena that can cause damage or interference due to overvoltages and/or overcurrents, for example:

**Electro-static discharge (ESD).** ‘Personnel ESD’ has risetimes under 1ns, peak voltages typically in the region  $\pm 8$  to  $\pm 115$  kV when humidity is 20% or more (but maybe up to  $\pm 35$  kV in some circumstances). The initial ESD charge-transfer impulse is often a unidirectional impulse waveform lasting a few tens of nanoseconds, but resonances mean that conductors generally experience a damped oscillatory waveform that can last for milliseconds. The energy content of an ESD event is low, although sufficient to damage sensitive semiconductors. There are other kinds of ESD phenomenon, including machine ESD, and furniture ESD. As well as conducted transients, ESD events generate very intense pulsed electric and magnetic fields, up to kV/m and kA/m in close proximity. Design techniques specifically for suppressing ESD will be discussed in Part 6 of this series.

**Fast Transient Bursts.** Caused by arcs and sparks at electromechanical contacts or degraded insulation, their conducted noise is characterised by a broadband spectrum with an upper frequency and risetime that depends on the length of conductor between the spark and the observation point. Within a metre or two of a spark, the upper frequency of the conducted noise can exceed 300MHz and the risetime be less than 1ns.

Typical peak voltages on mains cables can be up to  $\pm 2$  kV, maybe  $\pm 4$  kV in industries where high-powers are switched, but in exceptional circumstances they can be much higher. The energy content of a burst is quite low, although sufficient to damage sensitive semiconductors. This phenomenon is suppressed by the application of the techniques described in all six parts of this series, taking the characteristics of the fast transient bursts into account.

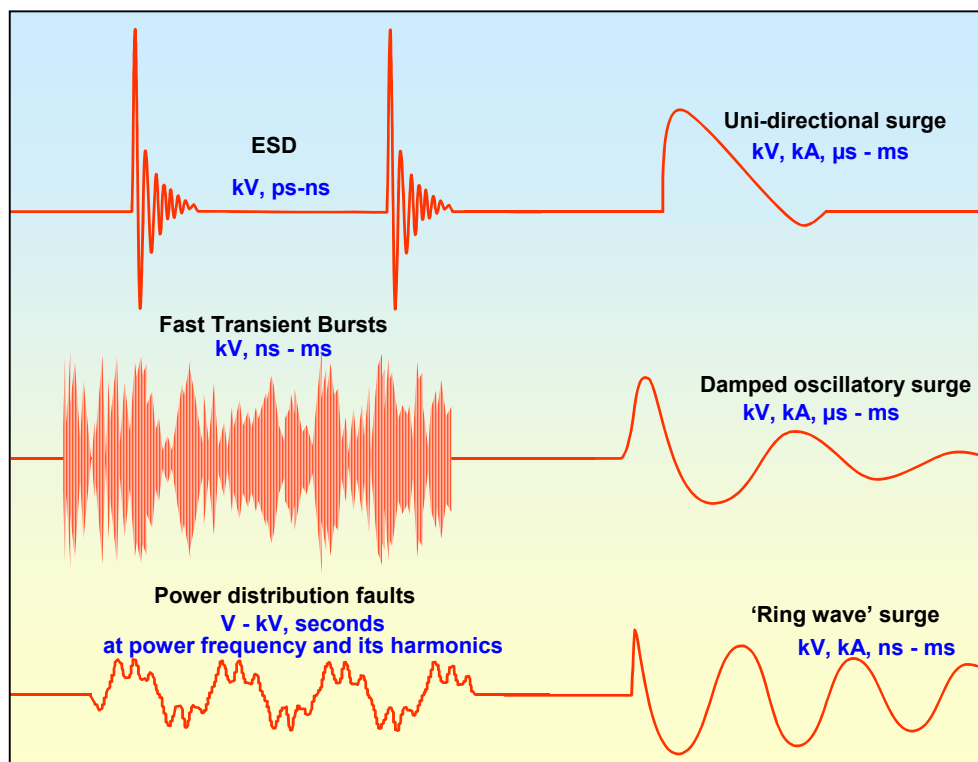


Figure 3AD Some types of overvoltage transients and surges

**Nuclear and High-Altitude Electromagnetic Pulse (NEMP and HEMP).** Originally only a concern for military and telecommunication infrastructure organisations, the possibility of ‘electronic warfare’ or ‘electronic terrorism’ in the commercial and industrial sectors is making it a more mainstream issue. A number of standards in the IEC 61000-5 series are being written to cover these phenomena, how to test for them, and how to design to protect from them. Essentially, suppression requires the application of the various techniques described in all six parts of this series, taking the characteristics of the EM threats into account.

**Surges.** Surges are generally fairly slow but powerful EM phenomena, with risetimes typically in the range 50ns -10 $\mu$ s, caused by lightning and also by the sudden release of energy stored in reactors (e.g. the energy in the rotor of a motor, released as a surge when the motor is switched off). They are generally measured in  $\pm$ kV, and can last from milliseconds to as long as seconds in some unusual cases. There are many surge voltage and current waveshapes associated with different situations, including: 'unidirectional', 'damped oscillatory' and 'ring wave'.

The surge test standard used most commonly in commercial/industrial applications for surge testing is EN/IEC 61000-4-5, which uses unidirectional surge waveshapes and assumes (presumably on the basis of real information) that line-to-line surges in AC power cables have a source impedance of 2 $\Omega$ , and line-to-ground surges have 12 $\Omega$ . It also assumes that surge currents flowing in power and earth/ground conductors induce CM surges into any long conductors (whether power, signal, data or control) with a source impedance of 42 $\Omega$ , and earth-lift surges due to currents flowing in the common earth/ground structure have an impedance of 2 $\Omega$ . Induction and earth-lift surges are usually only considered significant where cables are longer than 10 metres, but in some circumstances much shorter cables could be vulnerable.

The low surge impedances mean that, for example, a  $\pm$ 2kV mains voltage surge line-to-line has an associated surge current of around  $\pm$ 1kA. The voltage and current risetimes differ due to the inductance of the circuits, but the resulting peak powers are typically measured in MW, and the total energy associated with a surge is typically measured in tens of Joules (enough to vaporise the iron wire in metal-clad resistors, and blow the terminals off their ends). As a result, the main problems caused by surges is actual damage to devices, and even to conductors (e.g. PCB traces) by overvoltage, overcurrent or over-dissipation – but it should not be forgotten that they also cause errors in signals and data.

**Power frequency overvoltages.** There are various types of faults in power distribution systems, and they can cause the 'earth' or 'ground' voltage of an item of equipment to suffer various levels of voltage at the powerline frequency (and its harmonics, in the case of distorted AC power waveforms). In the (usual) case of earthed/grounded neutrals, the neutral experiences the same overvoltage. The overvoltages are typically up to 50% of the power systems phase-neutral voltage, but in some cases can approach 100%. Their durations are measured in fractions of a second, or even seconds, so although their voltages are not generally as high as the other overvoltages listed above, the total energy delivered can be very large indeed. Faults in high-voltage power systems, can be a very significant indeed, so it is lucky that they are quite rare.

These overvoltages can also be caused when one of the other types of overvoltage listed above cause power current to flow to earth/ground, e.g. via a spark or the operation of a line-to-ground SPD. This is often called a follow-on SPD current. Overvoltages can also be caused if a power cable is accidentally shorted to a conductor associated with the circuit to be protected, sometimes called a 'power-cross', which can last for minutes, hours, or even be continuous.

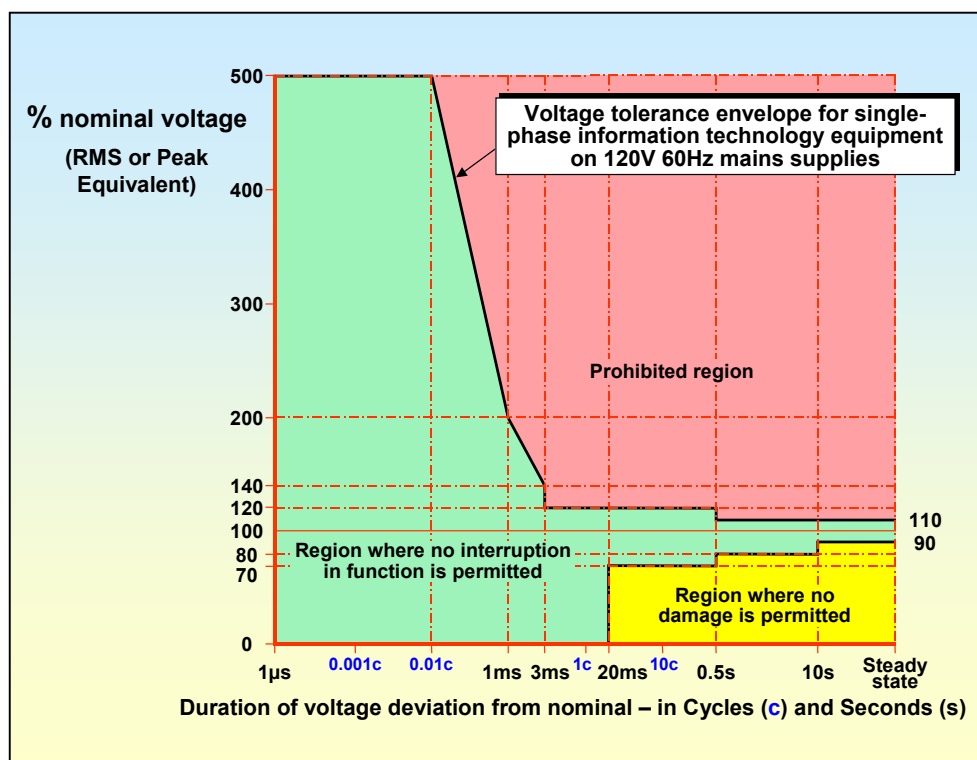


Figure 3AE The ITI (CBEMA) Curve

**Other types of overvoltage.** Figure 3AE shows the “ITI (CBEMA) Curve”. This curve and its Application Note [14] describe an AC input voltage envelope for 120V 60Hz power systems, which typically can be tolerated (no interruption in function) by most Information Technology Equipment (ITE). The ITI Curve and its Application Note are not intended to serve as a design specification for products or AC distribution systems but this is often how it is employed.

## 3.5 Protecting from surges

### 3.5.1 Protecting insulators from surge overvoltages

Conductors in cables, connectors, transformers, etc., are protected by insulation, which can be overcome by surge voltages, causing interference and even possible safety problems. Solid insulation can also degrade over time, due to repeated surges. Even where a surge causes a spark to an earthed chassis and does not cause any *direct* interference, the RF noise caused by the spark can be a significant source of interference within a product.

So insulation (whether air or solid materials) must at least satisfy the ‘creepage’, ‘clearance’ and ‘voltage withstand’ requirements of the relevant safety standards (e.g. IEC/EN: 60950, 60335-1, 61010-1, 60204-1, 60601-1, etc.) – which usually provide sufficient performance for EMC purposes. However, environments suffering from especially high levels and/or high rates of surge overvoltages could require increased insulation voltage withstand, unless the surge voltages are limited by other means, such as galvanic isolation or SPDs (see below).

### 3.5.2 Protecting conductors from surge overcurrents

The high levels of transient currents flowing during a surge can damage conductors themselves, so we need to provide appropriate protection. If conductors have sufficient CSA (Cross Sectional Area) they can handle momentary surges that are many times higher than their continuous current ratings. For copper conductors, the following guide can be used to calculate its maximum current/time before it fuses (melts):  $I = 290 \times \text{CSA} / \sqrt{t}$ , where  $I$  is the peak current in amps, CSA is given in square millimetres, and  $t$  is the duration of the current in seconds. This guide is only valid for  $t$  up to 5 seconds, because thermal convection and conduction effects start to become significant at longer times, requiring a very much more complex equation.

On a PCB, ½oz (finished) copper foil has a thickness of 17.5µm; 1oz copper 35µm; 2oz copper 70µm and 3oz copper 105µm. Multiplying the finished copper thickness by trace width gives its CSA, which can be used in the above guide. For example, a trace that is 0.18mm (7 thousandths of an inch) wide in 1oz (finished) copper (35µm), subjected to a rectangular current surge lasting 40µs, would have a maximum transient or surge current (before melting) of around 290A. Of course real surges do not have rectangular current waveforms, so there is some error in using the above guide to estimate real surge capability of conductors – a large ‘engineering margin’ is recommended!

Underwriters Laboratories (UL) carried out some tests of their own [15][16], resulting in some guidance on the maximum current handling that would not result in an open-circuited PCB trace. The guidance given by Figures 7 and 8 in [15][16] seems to correspond pretty well with the above formula (but note that the horizontal axes of the graphs in its Figures 7 and 8 say they are trace thickness, when in fact they are trace width).

But the above guide only considers the copper conductor’s transient or surge current carrying capacity before it melts, and such high temperatures will damage wire insulation and PCB dielectrics, possibly causing delamination with consequent reliability problems.

Some suppliers publish transient current data for wires, typically for current surges with a 1s duration. A typical 1s rating for wire with a 0.5mm<sup>2</sup> CSA and standard PVC insulation, with a normal operating temperature of 25°C, is 55A. This current will not raise the copper temperature above the standard PVC insulation’s 160°C short-term temperature rating (standard PVC is rated 70°C for continuous use). If the short-term temperature ratings of the PCB dielectric used is not known, the same guidance should probably be applied for FR4 and similar materials, since their continuous and maximum operating temperatures are similar to those of PVC.

The transient current rating (for  $t < 5s$ ) is proportional to the conductor’s CSA and inversely proportional to the square root of the duration, i.e. the current rating for duration  $t$  is the 1s rating divided by the square root of  $t$ .

Applying this guide to our example trace above – 0.18mm (7 thousandths of an inch) wide in 1oz (finished) copper (35µm), subjected to a rectangular current surge lasting 40µs – assuming the normal ambient temperature is 25°C suggests that repetitive transient or surge currents of up to 100A should not cause damage to FR4 dielectrics. Of course, this assumes that the transients or surges occur at a rate that allows the trace to cool down to 25°C after each event.

To protect a trace or PCB from an overcurrent that exceeds the values given by the above guides, we might want to use a fuse. But even the fastest fuses cannot respond in less than 10ms (which is a very slow transient

or surge in EMC terms), and all fuses have tolerances on their time-to-opening values, so it is important to design for the slowest of the fuse type that will be used.

When calculating the maximum permissible current for normal operating temperatures above 25°C, remember that the current allowed when the trace is at the maximum temperature of the insulation (e.g. 160°C) is zero, and that the heating effect of a current is proportional to the square of its value in Amps. So, in the above example, if the operating temperature was 60°C instead of 25°C, the maximum current that could be permitted before the trace temperature exceeded 160°C would fall from around 100A to around 86A.

High-temperature cable insulation materials, and grades of FR4 and other PCB dielectrics are available, and using them allows conductors to carry higher transient or surge currents without damage. On the other hand, some types of PCB dielectrics (especially some low-loss and microwave types, increasingly used in modern PCBs) can have lower short-term temperature ratings than FR4, so they could only withstand lower transient or surge currents than the above guide.

It is best to use 'UL Recognised' or 'UL Approved' PCB materials, especially to prevent safety hazards from fire, smoke and toxic fumes, so always obtain and check the UL Approval certificate for the basic PCB material used (or other evidence). Some suppliers don't always deliver what they said they would, so check that all delivered PCBs have the appropriate UL logo stamped all over them.

### **3.5.3 Protecting electromechanical contacts from surges**

Overvoltages can cause electromechanical contacts to spark-over, and since the resistance of an arc-channel is low, this can apply power to circuits that should be off. The high currents during a surge can also cause electromechanical contacts to weld together, so they might not open when supposed to. Either of these effects can cause erroneous operation or malfunction, even safety risks in some cases.

Spark-over is prevented by using large contact gaps and or SPDs (see below) to limit the maximum voltage. Some types of switches and electrical contactors are available with contact gaps up to 8mm, which will generally withstand nearly 8kV. Most switches and relays will change their mechanical state even when one or more of their contacts are welded, and the solution here is either to divert the surge current away from the contacts using SPDs (see below) so they do not weld, or to use devices with positively guided contacts, which will only change their mechanical state if all the contacts also change state. Automatically operated switches, relays and contactors with positively-guided contacts must be combined with position sensing to detect welded contacts and prevent problems from resulting.

### **3.5.4 Galvanic isolation is the best defence against surges**

Various techniques are available for achieving galvanic isolation that completely prevents surges from entering equipment, including...

- High-voltage opto-isolators or opto-couplers, fibre-optic links
- Wireless, infra-red, free-space microwave or laser voice or data communications
- Motor-generator sets
- On-line continuous-conversion double-conversion uninterruptible power supplies (UPSs) employing isolation transformers

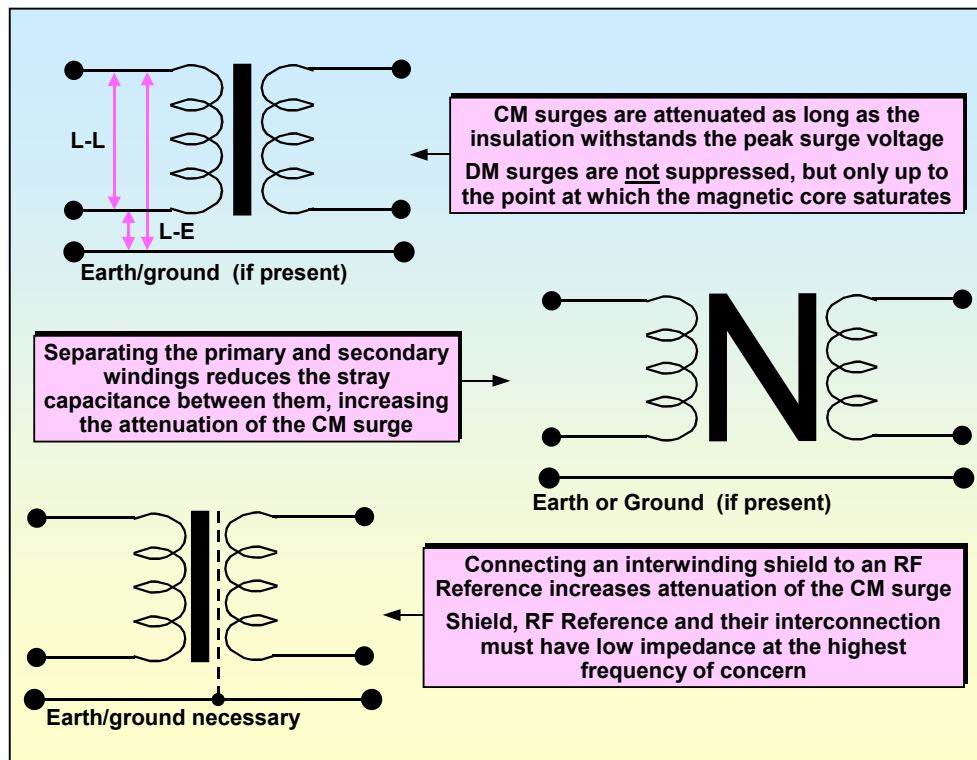
Of course, the part of the galvanic isolation device that is connected to the surge-exposed circuits must themselves be resistant to the surges they will experience (e.g. the mains input circuitry of a UPS).

The above provide protection against CM and DM surges. Isolating transformers are commonly used for mains power (e.g. in linear or switch-mode DC power supplies) or signals (e.g. Ethernet), but they only provide galvanic isolation for CM surges, as shown in Figure 3AF. Even so, because normal mains transformers are not wound for good balance at RF, there is conversion from the CM surge on the primary to DM surge on the secondary [17].

DM surges are rare on most types of signal (e.g. analogue signals or digital data carried by twisted-pair cables) but are common on DC or AC power, and can pass straight through an isolating transformer to the protected circuits, although if the transformer is a step-down type, the overvoltage will generally be stepped down too.

However, if the DM surge voltage exceeds a level that saturates the magnetic core, this excess will not be transformed to the secondary. 'Constant voltage transformers' (CVTs) run their cores in saturation (and run hot as a result) so they are effective at suppressing DM surges.

Increased attenuation of CM surges, especially their higher-frequency spectral content, can be achieved through the use of high-isolation transformers, for example using separated primary and secondary windings to reduce the stray capacitance and stray mutual inductance in the air between them, even placing them on separate limbs of the transformer core.



**Figure 3AF CM surge suppression with isolating transformers**

Another technique is to use an interwinding shield, but this needs to be connected to a RF Reference (often misleadingly called earth or ground), with the shield, Reference and their interconnection all having a very low impedance at the frequencies to be suppressed. The RF Reference for an item of equipment is usually an appropriately-designed chassis or metal enclosure (see 3.2.3 and 3.3 above), but if the equipment is a PCB in an unmetallised plastic enclosure it could be the PCB's 0V plane (see Part 5 of this series).

[17] gives more details on the use of transformers to suppress surges, and states that ordinary isolating transformers are not very good at surge suppression, even if fitted with electrostatic shields.

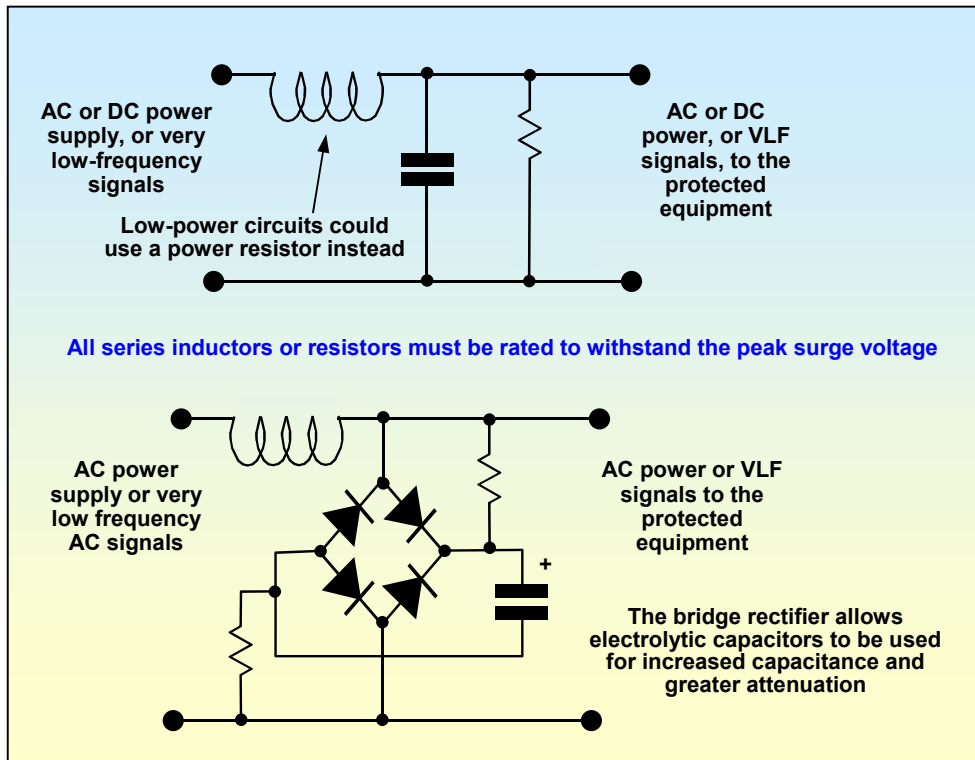
### 3.5.5 Surge suppression with filters

Filters are often found following the SPDs described below, and are not often recommended for suppressing surges on their own. However, they can be quite effective in either application providing issues of resonance and voltage/current/energy handling are taken into account. Figure 3AG shows some examples of low-pass filters used for suppressing surges. The spectra of many surges have high amplitude at low frequencies (e.g. a few kHz), so simple filters like the single-pole types shown in this figure need to have low corner frequencies, often just a few tens or hundreds of Hz, to provide sufficient attenuation of the surge.

Filters do not dissipate the energy of a surge – they aim to convert the waveforms from spikes with high peak values, to gentle 'bumps' with lower peaks that are more easily survived by the protected circuits. The areas under the voltage/time and current/time curves stay much the same, as a surge passes through a filter.

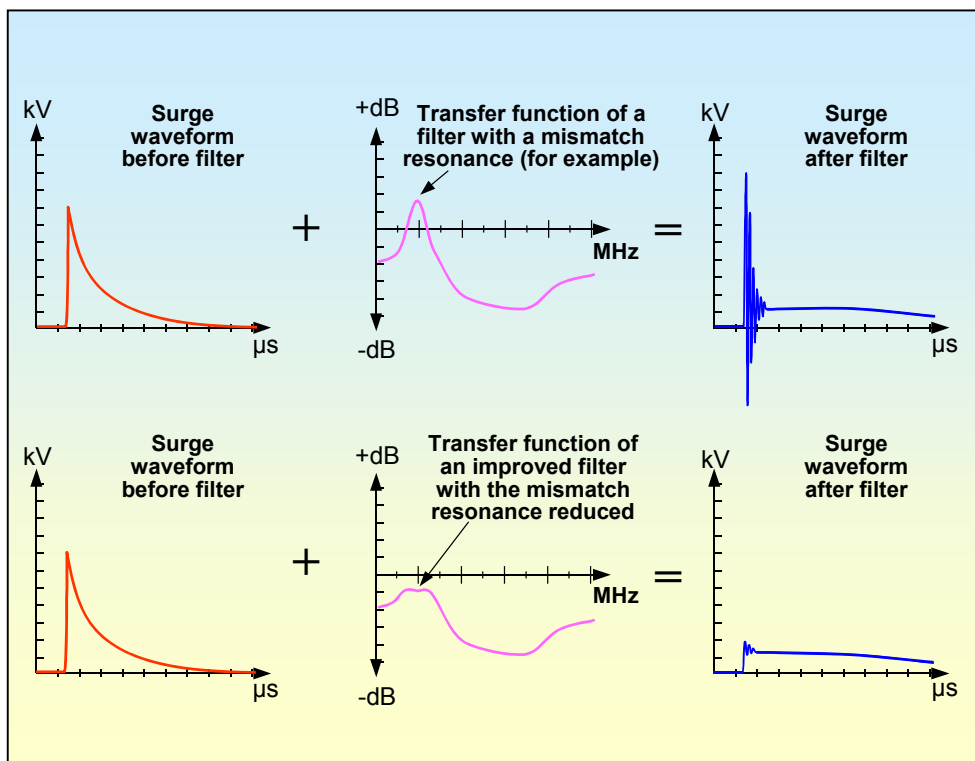
But all LC filters are resonant circuits, and there is the possibility that most of the energy in a surge might emerge from a filter as a burst of damped sinusoidal waves at the resonant frequency. The peak amplitude of such a burst could be much higher, even as much as ten times higher, than the peak of the original surge voltage, and the author has seen examples of this effect in real life. Even where filters are not supposed to be providing surge suppression, if they connect to external cables (especially mains cables) they will always be exposed to surges, and it is important that at least they not make a surge overvoltage worse.

So all filters that could be exposed to surges, or are intended for surge suppression, should be carefully designed not to exhibit significant resonances in the frequency range of concern (the spectrum of the energy in the surges), and/or be damped (see 3.2.9 and Figure 3Q, and 3.2.10 and Figure 3R) so that at least they don't increase the overvoltage. RC filters do not suffer from resonances, and can easily be used for surge suppression in low-power circuits. When soft ferrites are used as the L element in filters this dampens down resonances at higher frequencies, but at the lower frequencies they have a significant inductive reactance, so cause resonances.



**Figure 3AG Surge suppression with filters**

Like all filters, filters used for surge suppression need RF References, and interconnections to them, that have very low impedances at the frequencies to be suppressed (see 3.2.3, 3.3 and 3.5.5 above).

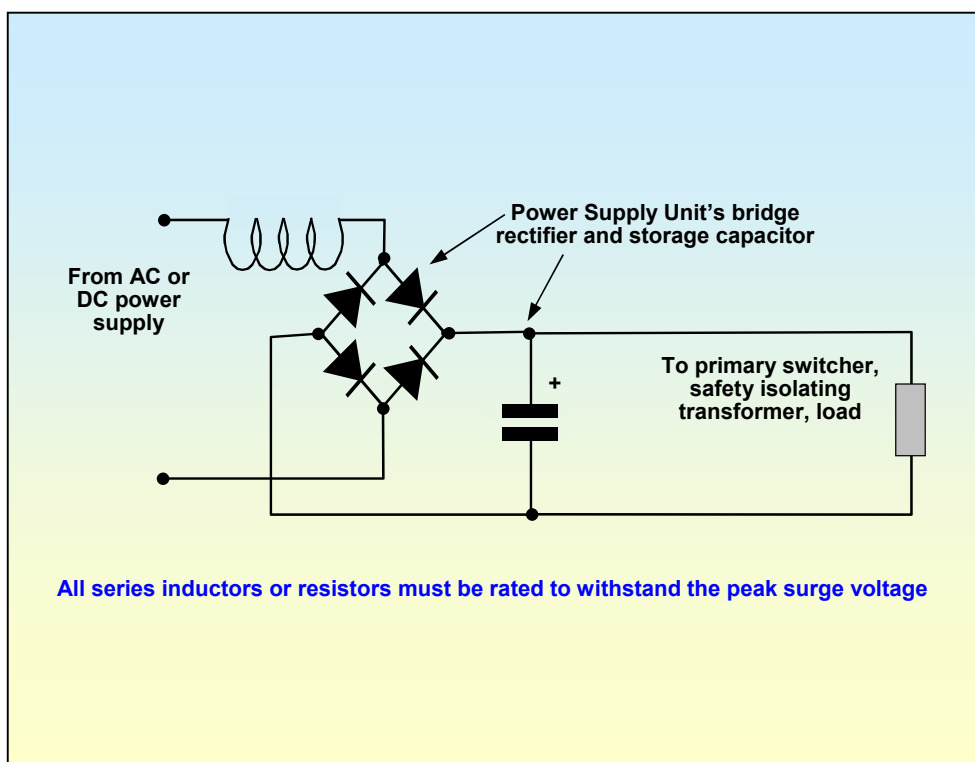


**Figure 3AH Filter resonances can increase peak voltages**

When a filter is used to suppress surges on AC power conductors, its cut-off frequency should be chosen to avoid excessive levels of power-frequency currents in the filter capacitor, that could cause it to heat up and reduce its reliability. If used to suppress line-to-earth surges the filter capacitor powerline currents should not cause the equipment's protective conductor or touch current to exceed the levels set by the relevant safety standards. These issues compromise the design of a filter, making it less effective on certain kinds of surges on AC power conductors.



But these corner frequency compromises do not apply when filtering DC conductors, or where AC power is converted to DC in the rectifier of an off-line mains power supply unit (PSU). The storage capacitor that follows a rectifier will provide some DM surge suppression – providing it can handle the surge voltages and currents without reducing its life by too much. But adding an inductor on the AC side of the rectifier, as shown in Figure 3AJ, adds a filtering function that can be designed to improve the attenuation of the surges, whilst also reducing the voltage and current stresses on the storage capacitor.



**Figure 3AJ Using the input circuit of a PSU as surge protection, by adding a series inductor**

It is quite common in industrial applications to fit such external inductors in the AC power inputs to electronic equipment such as variable-speed motor drives, to improve their reliability by increasing their surge protection as shown in Figure 3AJ. These inductors are usually called 'line inductors' or 'line reactors', and they also help reduce the emissions of mains harmonics from the electronic units' AC power rectifiers (sometimes this is the primary reason for their use, and the surge suppression is a bonus).

When designing a filter for surge suppression, in circuits similar to those in Figures 3AG and 3AJ, the filter components need to be rated appropriately for their high-voltages and high-currents, including...

- Choose the value of capacitor so that it can absorb the total charge of the highest-energy surge without its voltage rising so much that the following circuit is damaged. A useful guide is  $C\Delta V = I\Delta t$ , where  $\Delta V$  is the change in capacitor voltage caused by the current  $I$  flowing for time  $\Delta t$ . Because surges never have rectangular waveforms, some assumptions and/or additional calculations are necessary when using this guide.
- Choose the capacitor voltage rating so it is not damaged itself by its voltage rise during the highest-energy surge
- Design the inductor and capacitor as a low-pass filter that attenuates even the slowest surge rising edges by enough to protect the following circuits from damage
- The capacitor's ESR and ESL should be low enough to maintain good attenuation whilst absorbing the highest surge currents
- Use a type of capacitor that will handle the highest current and fastest surges expected, it may be necessary to use specially 'pulse-rated' capacitors, rather than types intended for storage, filtering or decoupling
- Filter series elements (inductors or resistors) must not spark-over due to the high-voltage (this makes it difficult to use surface-mounted components)

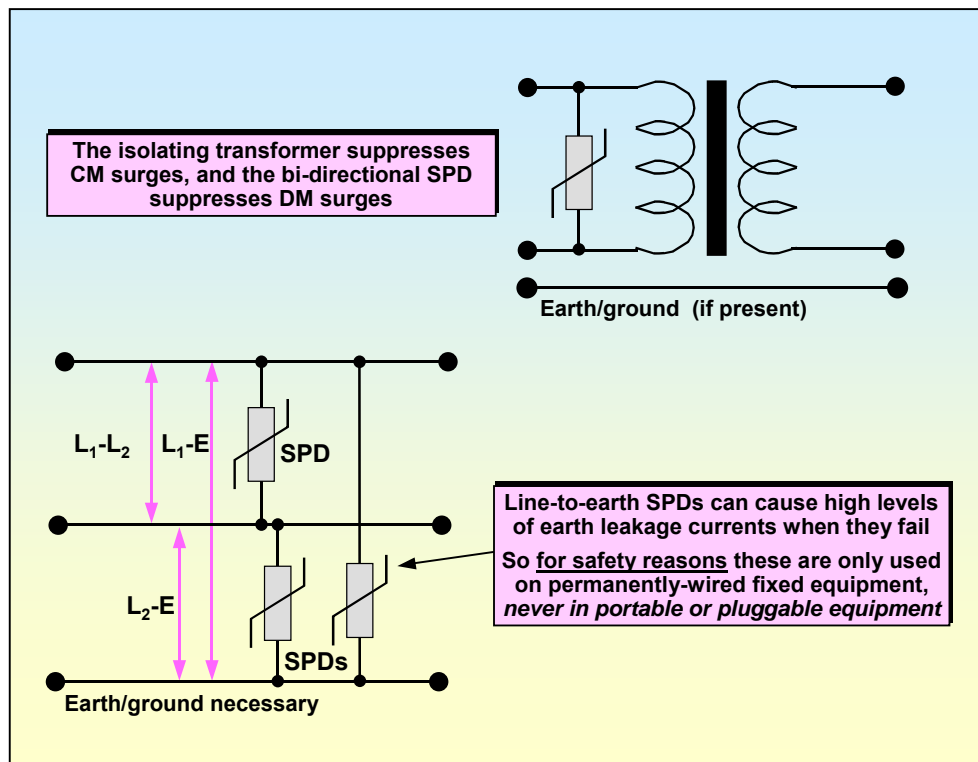
### 3.5.6 Suppression with surge protection devices (SPDs)

SPDs are high-resistance devices that switch to a low-resistance state above a certain voltage. They are connected in parallel (shunt) with the circuit to be protected to limit the overvoltage it is exposed to, and to divert

the surge currents away via a different route. SPD techniques can be used as alternatives to the methods described above, or in conjunction with them.

The choice of SPDs and design of the circuits that use them is far from trivial, with a number of issues to be dealt with to achieve reliable protection at a reasonable cost. For example, SPDs are non-linear devices, so adding them to a circuit can add EMC problems. Some may need filtering to prevent them from demodulating RF, and gas-discharge or spark-gap types create RF noise when they operate, which may need filtering to prevent interference.

Figure 3AK shows some typical examples of the use of SPDs. When used in conjunction with an isolating transformer, the CM surges are suppressed by the primary-secondary insulation (see 3.5.4), and the DM surges are suppressed by the SPD connected between the lines (phases). Three-phase power systems would need three SPDs to suppress their DM surges, connected either in star or delta.



**Figure 3AK** Some surge protection circuits using SPDs

The lower circuit in Figure 3AK shows SPDs used for both CM and DM suppression of a single-phase AC power supply. An important safety consideration is that all SPDs fail eventually, due to wear-out, and when they do those connected between line and earth/ground (CM suppression) can cause hazardous levels of leakage currents (usually called 'touch currents' or 'protective conductor currents' in product safety standards).

So, for safety reasons, SPDs between line and earth/ground are only used (when used at all) in permanently-wired fixed (stationary) equipment, such as mains distribution cabinets, and never in portable or pluggable equipment.

Of course, a surge that is energetic enough to cause an SPD to explode will result in an open-circuit failure, and this can be a consequence of not choosing the right device ratings for their EM environments. Other types of devices can also explode when exposed to surges they cannot handle, and this is why appropriate safety precautions should always be taken when testing with surges – always placing a blast-proof screen around the tested unit so that no-one can be injured by ejecta from the equipment being tested, or their ricochets. It is also a good idea to have a power isolation switch in the blast-protected area, and ready access to fire extinguishers suitable for electrical fires.

It is possible for surges that are below the threshold of activation of the SPDs to cause more damage than surges with higher voltages. Surges below the threshold voltage can source all of their current into to the circuitry being protected, possibly causing damage. Low-voltage surges are much more common than high-voltage ones, so if the SPDs are not activating at low enough voltages the consequences for reliability can be severe.

Most people assume that testing with the most severe surge is the worst-case, and avoids the need to test with lower levels, but this is not the case, and to check that a design is going to be reliable enough it is always recommended to test with surge voltages that are just too small to activate the SPDs.

### 3.5.7 Types of Surge Protection Device (SPD)

SPDs have a voltage-dependant resistance. When the voltage exceeds their rating, their resistance falls rapidly so they carry some (or all) of the surge current whilst maintaining a low level of voltage between their terminals. They are highly stressed components, and degrade (wear out) and fail after a number of surges, depending on the type of device and the energy and number of the surges – so it is very important to choose appropriate SPD types and ratings for the surges expected in their EM environment.

There are four basic types of SPD: Metal oxide varistor (MOV); Avalanche Diode; Spark gap / Gas Discharge Tube (GDT); and silicon controlled rectifier (SCR), as shown in Figure 3AL along with some of their common schematic symbols.

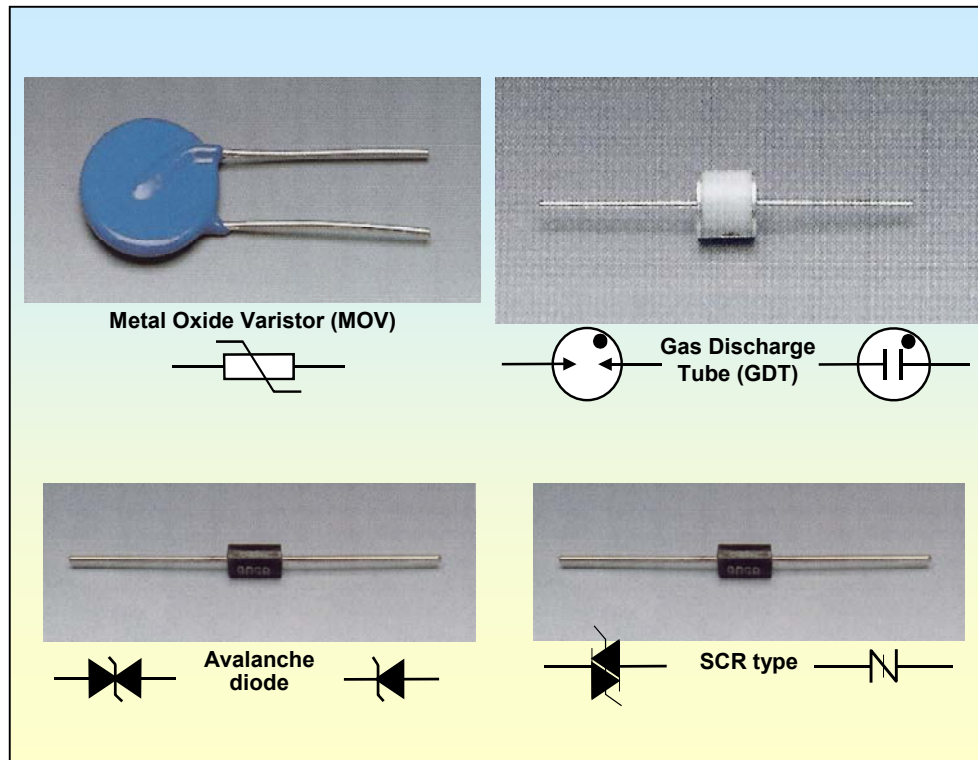


Figure 3AL Types of SPDs

#### Metal oxide varistors (MOV)

These are low cost, inherently bi-directional devices that operate in about 1ns, available in a very wide range of energy ratings, from small devices for PCB mounting to very large devices for primary lightning protection. MOVs are available in a very wide range of voltage ratings, from a few tens to hundreds of volts, and as discrete devices (see Figure 3AL) or arrays in IC style packages.

Their natural wear-out mode is to degrade gradually to a low-resistance, even short-circuit, and they can degrade very quickly indeed if not rated correctly. MOVs naturally have a high capacitance, generally measured in nF, making it possible to combine filtering and overvoltage protection in one device, but making them unsuitable for use on RF and high-data-rate signals.

#### Avalanche diodes

These are like zener diodes but with very high transient current and power ratings. They are available only with low voltage ratings, from a few volts to a few tens of volts. Uni- or bi-directional types exist, so it is important to choose the correct type. They are available as arrays of devices in IC package styles, as well as discrete items such as the type shown in Figure 3AL.

They are the fastest type of SPD, operating in well under 1ns so can also be used for ESD protection, but are relatively expensive and tend to have low current and energy ratings. Like all semiconductors they can fail short or open, and like all SPDs they do wear out eventually. Proprietary types include 'Transil', 'Transorb' and 'TVS'. (TVS is often used as a generic name for these devices, but is short for 'Transient Voltage Suppressor' – so generic a term that it could be used to mean any kind of SPD.)

Avalanche diodes historically have had quite high capacitances, measured in 100s of pFs, but because of the need to protect the recently developed USB2.0, Firewire and similar high-speed datacommunication interfaces, versions with capacitances of 5pF or are now available.

## Gas Discharge Tubes (GDT)

These are low cost, inherently bi-directional devices that are available in a very wide range of energy ratings, from small devices for PCB mounting to very large devices for primary lightning protection. They are available with voltage ratings (trigger voltages) from around 100V to kV.

They have very low capacitances, usually just a few pF, which makes them very useful for protecting RF and high-speed signals, and they have been used to protect radio transmitters and receivers from overvoltages on their antennas for many years. They can eventually degrade to a short-circuit, as their electrodes vaporise and plate the inside of their packages, and this can happen very quickly if they are not rated correctly.

GDTs have to reach a trigger voltage before they start to conduct significantly, but once the discharge is 'ignited' it creates an arc, which has a very low resistance and then the GDT has a very low terminal voltage at any level of current (SPDs are typically measured with pulses of 1kA, or more).

The arc created inside a GDT when it 'strikes' makes them glow a very pretty violet colour, but arcs are hot and take a little time to cool down enough to restore the high-resistance ('off') state. This can mean that when used to protect AC power, there might not be enough time for them to cool down between mains cycles – so once turned on, they might never turn off. However, some manufacturers offer GDTs designed for rapid arc quenching, so that they do not restrike after the mains zero-crossing. Most distributors' catalogues and some manufacturers' data sheets do not make it clear whether their GDTs are suitable for use on AC mains supplies, so always check.

Another consequence of the arc in a GDT is that when used to protect AC mains supplies, an SPD that is in a low-resistance state carries mains current as well as surge current. In the case of GDTs and spark gaps these currents can continue for up to 10ms (half a cycle of the mains) after the surge has finished, until the arc is quenched by the zero-crossing of the AC mains waveform (but see earlier comments on choosing GDTs suitable for use on AC mains). These are known as 'follow-on' currents, and the SPD's conductors need to be sized to cope with them safely and reliably. It is possible for power frequency overvoltages (see 3.4 above, and Figure 3AD) to be caused by the earth-lift (ground-lift) resulting from follow-on currents in line-to-ground connected GDTs.

GDT trigger voltage is imprecise, so special three-terminal types are used for protecting differential signal lines (e.g. telephone lines). If two ordinary two-terminal GDTs were used, one would trigger before the other, causing the CM surge on the line to be converted into a DM surge – potentially more damaging to the following circuitry than the CM surge would have been. The three-terminal GDTs are designed so that both sides of the device trigger at the same instant.

## Spark gaps

These are simply a point where two conductors come close together, spaced apart by a suitable distance so that a surge voltage above the gap's breakdown voltage will cause it to spark over, thereby creating a low-impedance (the arc channel) that diverts the surge energy. The breakdown voltage of the air at sea level and 50% humidity is approximately 1kV/mm, but variations in humidity and pressure have a significant effect so it is impossible to design spark gaps so that they operate with any precision.

Spark gaps created between two traces on PCBs, or two conductors on a plastic or ceramic substrate, suffer additional variations in trigger voltage due to contaminants on the surfaces of the PCB or substrate. Another problem with spark gaps is that their electrodes tend to get further apart due to melting when they operate, increasing their trigger voltage over time; they can splash molten metal onto other conductors causing short-circuits; and if mounted on organic substrates (PCBs, plastics, etc.) their arcs can cause carbonisation, and they can become resistive and start to dissipate power, possibly leading to malfunctions, even overheating and fire.

Some parts of the world suffer from very intense thunderstorms, and the author knows of some Australian manufacturers who, having had rapid wear-out problems with GDTs, construct their own spark gap protection devices out of plates of metal.

A GDT is essentially a spark-gap in a controlled atmosphere, so that it has a repeatable trigger voltage. A spark gap behaves just like a GDT, except that it has an unreliable and unpredictable trigger voltage.

## Silicon Controlled Rectifier (SCR)

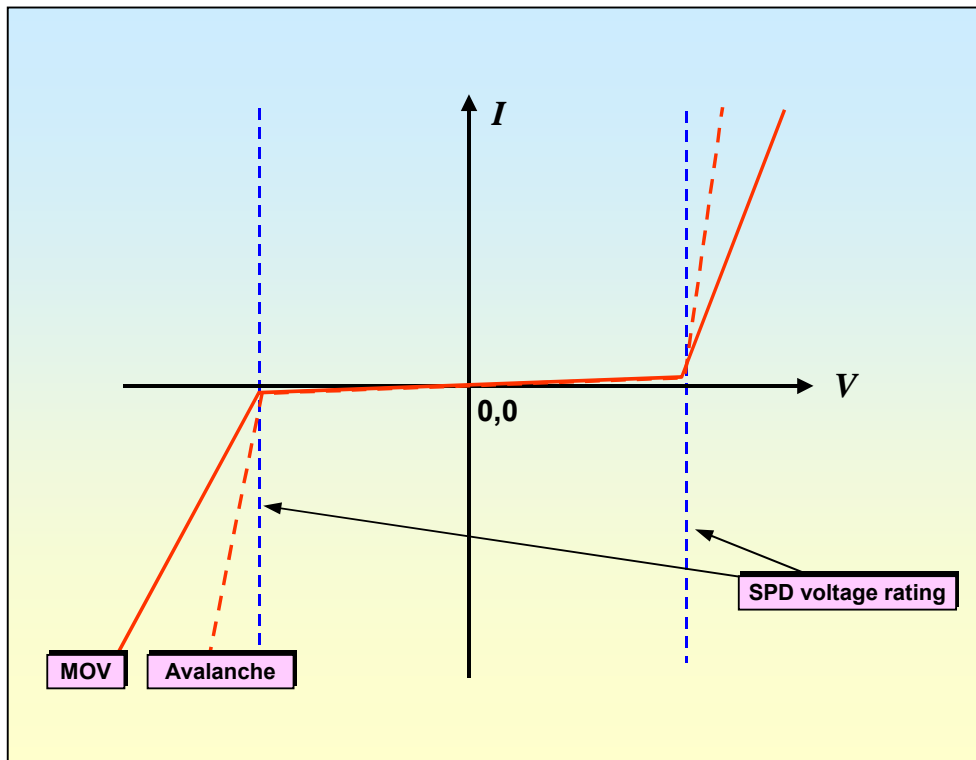
These are available as uni-directional (based on thyristors) or bi-directional (based on triacs), so it is important to choose the correct type. They tend to have high current and energy ratings for their package size and cost, and are available as arrays of devices in an IC package as well as discrete items such as the type shown in Figure 3AL. Like all semiconductors they can fail short or open, and like all SPDs they do wear out eventually. Proprietary devices include 'Surgector', 'Sidac', 'Sibar' and 'Trisil'.

SCR types have to reach a trigger voltage before they start to conduct significantly, but once they start to conduct they switch to a low resistance state which persists until the current through the device drops below some 'holding' value.

### 3.5.8 Characteristics and comparisons of SPD types

Figure 3AM shows the voltage/current characteristics of MOV and Avalanche Diode devices, which are often called 'voltage-limiting' types, because their terminal voltages increase much more slowly as current increases, once they pass their threshold voltage.

The figure is drawn assuming the threshold voltage for the MOV is the same as that of the Avalanche device, to show that the Avalanche type generally has a lower on-resistance, so for a given surge current it generally has a lower terminal voltage and provides better protection (in this regard, at least).



**Figure 3AM MOV and Avalanche Diode characteristics**

Figure 3AN shows the voltage/current characteristics of GDT and SCR devices, which are called 'voltage-switching', 'foldback' or 'crowbar' devices, because once they reach their triggering voltage their terminal voltage suddenly decreases, and stays very low even with very high currents in the device.

An important issue for these types of SPDs is the system's maximum DC voltage. If the clamping voltage of the device is the same or lower than the system's DC voltage, the SPDs will never switch off – a condition that should generally be avoided, not least because without appropriate protection the SPD could overheat, be damaged, emit toxic fumes or even catch fire.

Figure 3AP compares the generic performance of the various types of SPDs against a time axis, as they are activated by an example surge voltage. MOV and Avalanche devices act like zener diodes, with a threshold or knee voltage where they start to draw current. As their current increases their clamping voltage rises, sometimes to more than double their knee voltage. The figure also shows that GDT and SCR devices have to reach a trigger voltage and during this time they could let through surge voltages that could be high enough to cause damage. Then they foldback (or 'crowbar') and the voltage across them drops to well below the voltage they previously would have blocked. The figure also shows the minimum system DC voltage that was discussed in Figure 3AN.

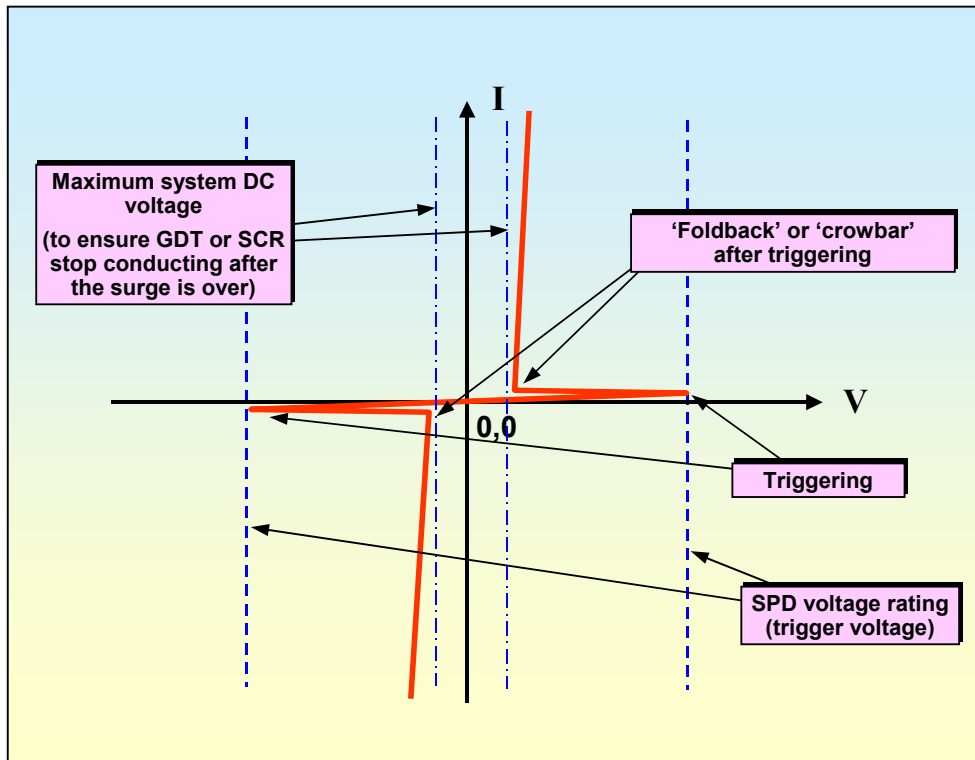


Figure 3AN Spark-gap / GDT and SCR characteristics

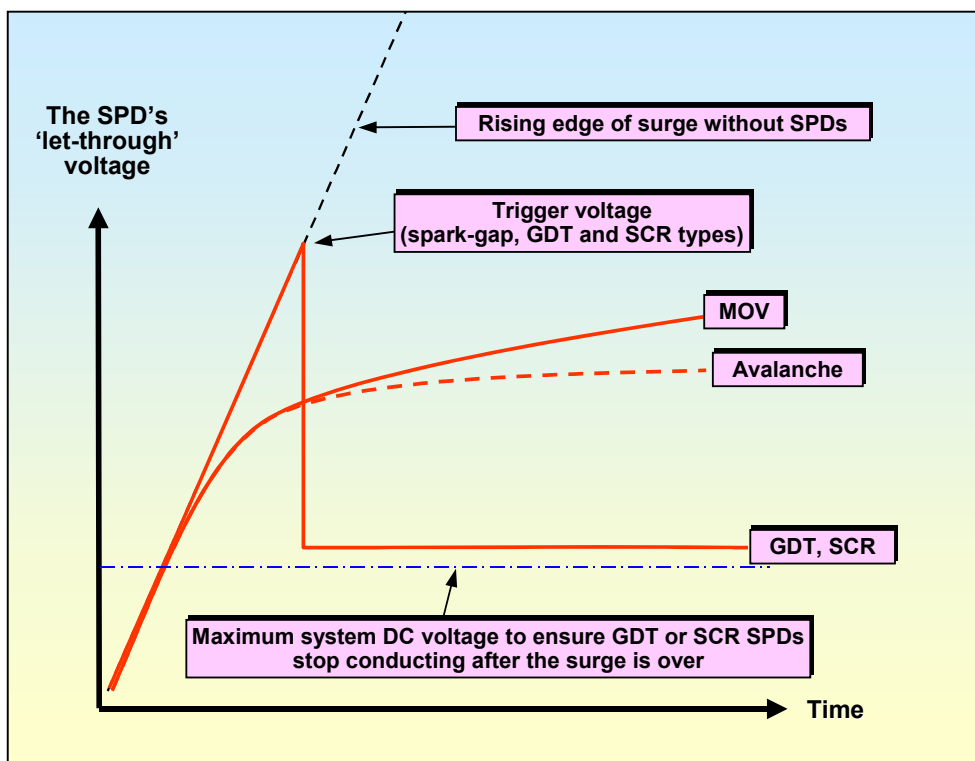


Figure 3AP Comparing the time-domain performance of SPD types

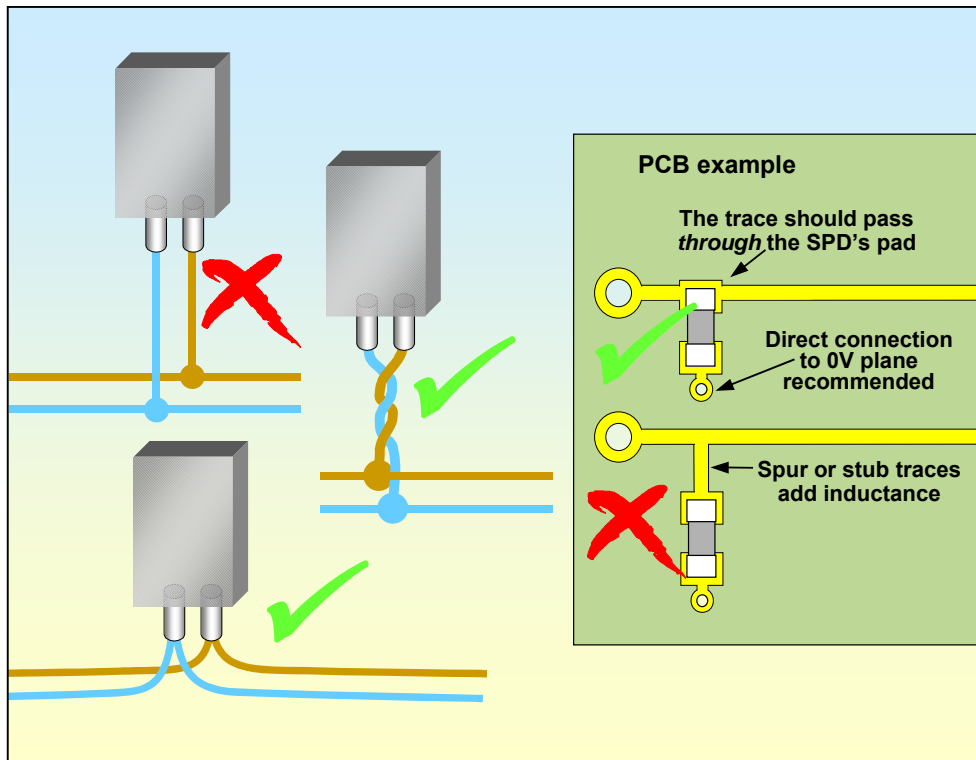
### 3.5.9 Minimising the inductance in series with SPDs

When surge currents flow in the inductance inherent in any conductors that are in series with an SPD, they increase the 'let-through' voltage. It is best to connect the incoming power (or signal) directly to the terminals of its SPD, and then connect the protected circuitry to the SPD terminals too.

Just as for filter capacitors, it is important that SPDs have wires or PCBs trace layouts that add just the minimum inductance in series with the device. It is best to take the conductors (wires or PCB traces) for the circuit to be protected directly to the SPD's terminals, and then route the conductors from the SPD's terminals to the protected circuit.



Where wires are used to connect to a GDT, and it is not possible to avoid using a long 'spur' connection to the protected circuit, the wires should be as short as possible and twisted as shown in Figure 3AQ.



**Figure 3AQ Reducing SPD let-through voltage by minimising series inductance**

### 3.5.10 Rating SPDs

The lightning protection systems fitted to most sites are not intended to protect electronics. On such sites, and on those without any lightning protection at all (e.g. most residential properties that are not high-rise apartments), real-life AC mains supply surges can be expected to reach at least  $\pm 6\text{kV}$  several times each year. At sites fed by overhead mains cables, the number of such events can be as high as several hundred a year, according to figures collected in the UK (which is hardly a high lightning incidence area).

Figures 3AR and 3AS show some statistics that have been collected for peak voltages on AC supplies in the USA. Similar statistics apply in most of the developed countries, but in less well-developed countries the incidence of surges can be higher.

Where equipment could be operated on a private mains supply (e.g. hospital equipment expected to work on local generators during power cuts, and tested every week), it might have a quite different surge exposure. For example, there are standards, reports and conference papers describing the types of surges that should be expected in the AC and DC power supplies used in land, sea, rail, air and space vehicles (these are not listed under the EMC Directive or tested by the IEC/EN 61000-4 series of standards).

6kV "or higher" surges are permitted by EN 50160, the European standard for the power quality of public mains supplies, and the author's experience is that such high voltages reliably occur over most of Europe. In fact, a peak surge voltage of 6kV is typical of single-phase public mains supplies worldwide, and is this value because it is the typical spark-over voltage at the terminals of typical single-phase mains sockets. The single-phase mains sockets are acting as accidental spark gaps, protecting the equipment connected to the mains supply from overvoltages higher than approximately 6kV.

In dedicated three-phase mains distribution systems, where there are no single-phase plugs or sockets, the clearance between terminals is greater (e.g. between the terminals in IEC 309 style mains plugs and sockets) and peak voltages of at least 12kV should be expected according to certain manufacturers who have had to design to this level to solve real problems in the field. Maybe as much as 20kV could be possible.

The venerable lightning protection standard BS6651 Appendix C (which will be replaced by EN 62305-4:2006 in August 2008) deals with this issue, and specifies the SPD ratings for equipment fitted in different parts of a site. This standard, or others which deal with the lightning protection of electronic equipment (e.g. IEEE C62.41-1991, IEC 61312-1) should be used where the EMC test standards applied are lacking in surge requirements, or where their surge requirements are incomplete, or where there is concern that the use of these ordinary EMC test standards might not give sufficient protection for the desired level of reliability in the intended environment.

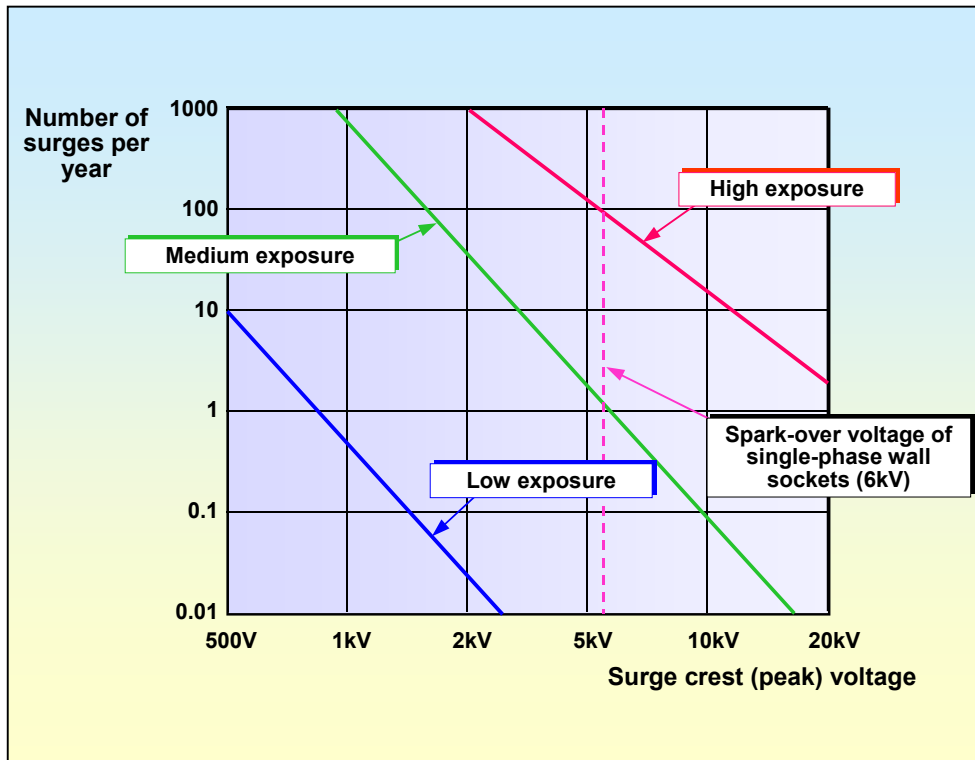


Figure 3AR Power line surge exposure chart (from IEEE C62-41:1991)

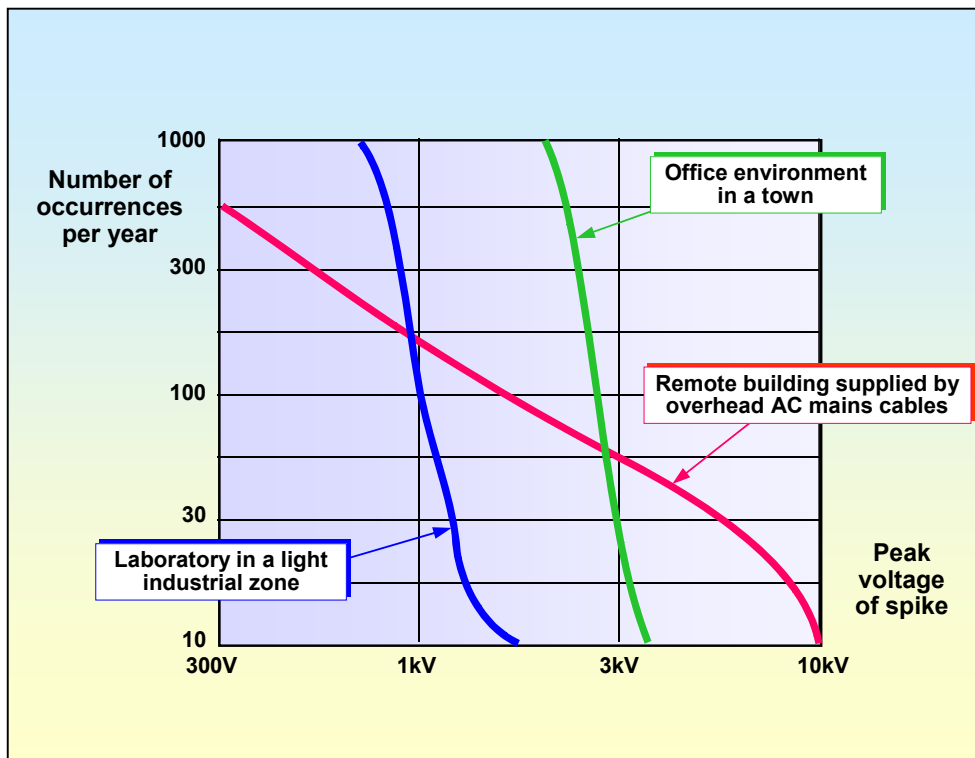


Figure 3AS "How many times will the mains attack your data?"

When a product is adequately protected against lightning surges, it is generally protected well enough against common surges generated by other means, such as switchgear. But some industrial or medical environments can suffer from significantly high levels of mains surges that are not caused by lightning, and these could possibly also have a higher rate of occurrence. Such sites include superconducting magnet or power generation applications, and high-power switched reactive loads such as very large motors or transformers.

To achieve adequate reliability of equipment, hence low warranty costs, satisfied and loyal customers, increased levels of repeat sales, and a virtuous circle leading to greater profitability, it is important to rate SPDs to handle the current and energy related to the likely surge exposure at the intended operational sites. It will not

generally be sufficient for reliability to rely on testing at the 1 or 2kV levels specified by the EMC test standards listed under the EMC Directive.

Taking the above into account, the maximum ratings of SPDs should be carefully chosen for:

- The maximum clamping voltage when activated, when carrying the maximum current
- Surge energy handling, in Joules
- Peak current handling, in kA
- The number of peak current or peak energy events that can be handled before the device degrades too much
- Continuous power rating in Watts, especially where surges are frequent, or where power-frequency overvoltages or power-cross incidents can occur (see 3.4 and 3.5.12)

### 3.5.11 Combining SPDs

It is sometimes hard to obtain all the necessary characteristics in one device, and to maximise surge protection performance and optimise cost, it might be necessary to combine different types of SPD. In the example shown in Figure 3AT the high power handling of a GDT is combined with the fast clamping action of an MOV. The GDT absorbs most of the surge energy – but only after it has triggered. The series inductor allows the GDT to be triggered by overvoltage, while the MOV protects the load against the GDT's trigger voltage.

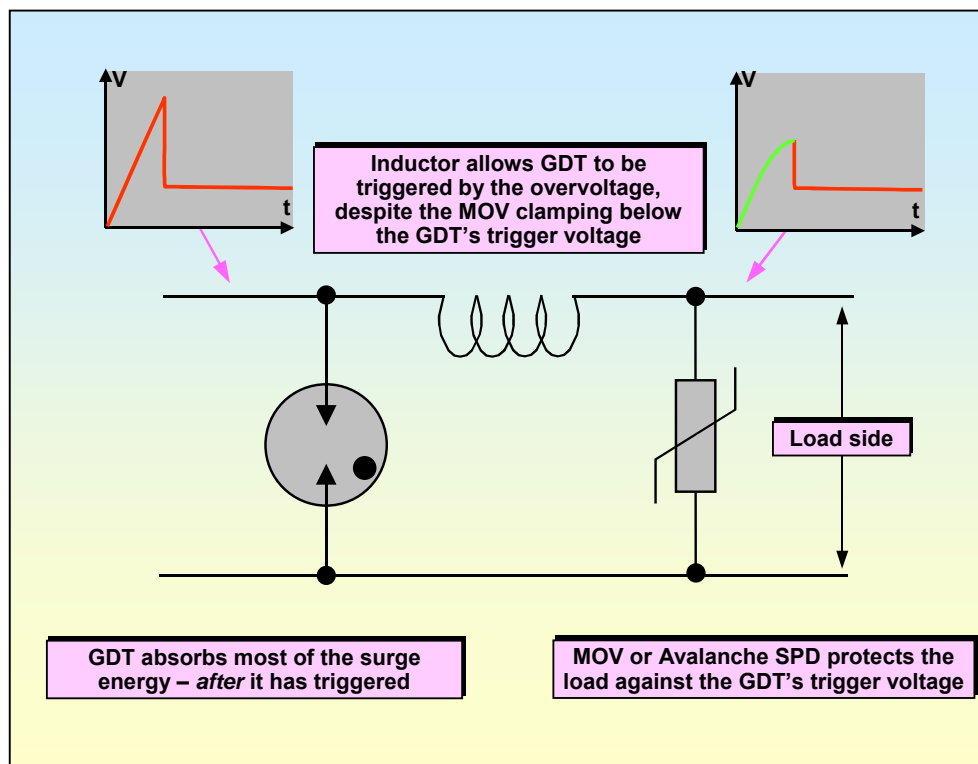


Figure 3AT Example of combining a GDT with an MOV

Simply paralleling a voltage-switching (crowbar) type of SPD with a voltage-limiting type such as an MOV usually prevents the crowbar device from triggering. There are a wide range of surge protection units available from lightning protection companies, and their proprietary designs often include series inductors where they combine voltage-limiting with voltage-switching SPDs.

Devices are available (e.g. TISP from Texas Instruments) that behave in Avalanche Diode mode up to the point where the voltage reaches a trigger level, at which point they change to SCR mode and crowbar the voltage to a lower level.

### 3.5.12 A hierarchy of surge protection

Full protection from surges when using SPDs usually requires a hierarchy of devices. High-energy SPDs such as large GDTs or very large MOVs are fitted at the incoming mains supply to the building or other structure. Medium-energy SPDs, such as MOVs, are installed in local mains distribution cabinets or at the mains inputs of equipment enclosures. Finally, low-energy (but fast) SPDs are fitted where needed to the PCBs inside equipment to protect electronic devices that interface with external cables. The PCB-mounted SPDs are often designed to protect against ESD as well, during (unpowered) assembly, as well as during operation. See Part 6

of this original series [1] for more on designing this type of protection. Part 6 of this revised series will be published during 2007.

For more on using a hierarchy of SPDs on a site, refer to Chapter 9 of “EMC for Systems and Installations” [13].

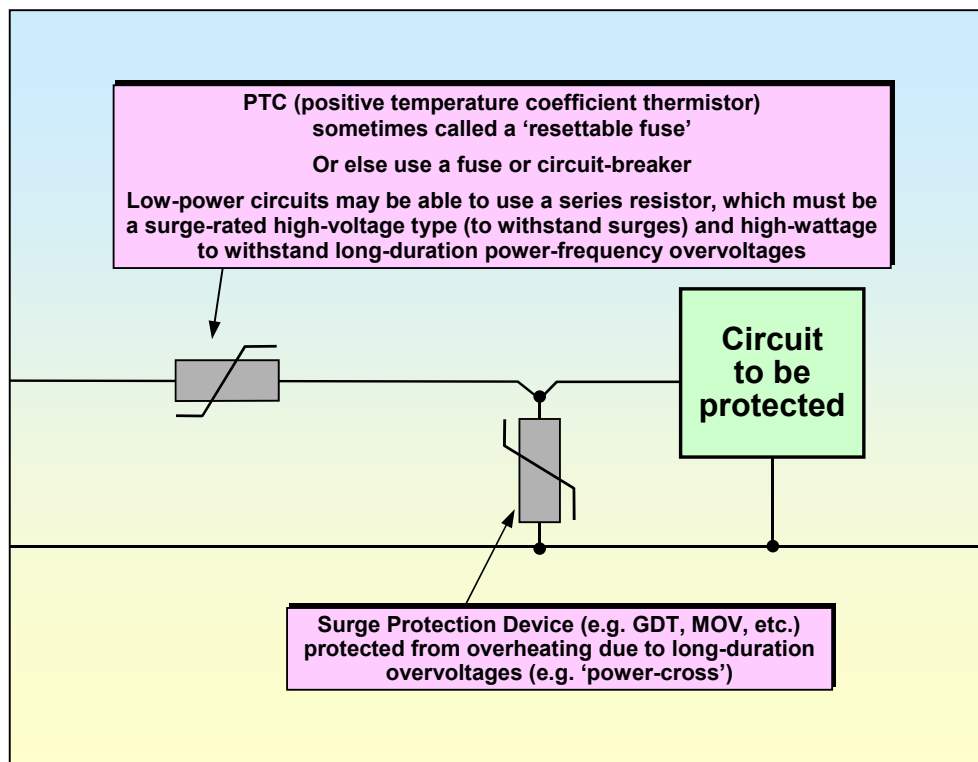
### 3.5.13 Protecting SPDs

Some faults in power networks can cause the supply voltage to increase significantly (even nearly double) for a few seconds. In some cable applications mechanical damage can short cables together, applying mains power to signal circuits, which is why telecommunications equipment has to withstand ‘power cross’ tests that apply 230V AC for several *minutes* onto their signal cables. Where SPDs are used, and the increased voltages are sufficient to trigger them, the very long durations of these overvoltages will cause them to dissipate large amounts of power, causing overheating, damage, possibly even toxic fumes, fire or explosion.

Although SPDs can handle kA and kW, they can only do so for very short periods of time. Their total energy ratings are quite small – for example a surface mounted MOV in an 0603 package might be rated at 0.1J; a 7mm diameter radial wire-led MOV might be 3J, and a 60mm square block-type MOV with spade terminals might be 500J. But these ratings are achieved with test pulses that last no more than 1ms, when even a 3J device can handle 3kW. But even a 500J device can handle no more than 10W for 50 seconds, so it is clear that SPDs need to be protected from overheating.

Figure 3AU shows the principle of using a (surge-rated, high-voltage, high-wattage) resistor, PTC thermistor, fuse or circuit-breaker in series with the input to an SPD, to prevent unreliability or damage caused by overheating.

Surge-rated resistors are special types that can handle very high levels of overcurrent (for a short time). They are manufactured by several companies, and may also be known as pulse or transient rated resistors. Some types, known as fusible resistors, are designed to open-circuit to prevent their thermal rating from being exceeded, helping to avoid smoke and fire hazards.



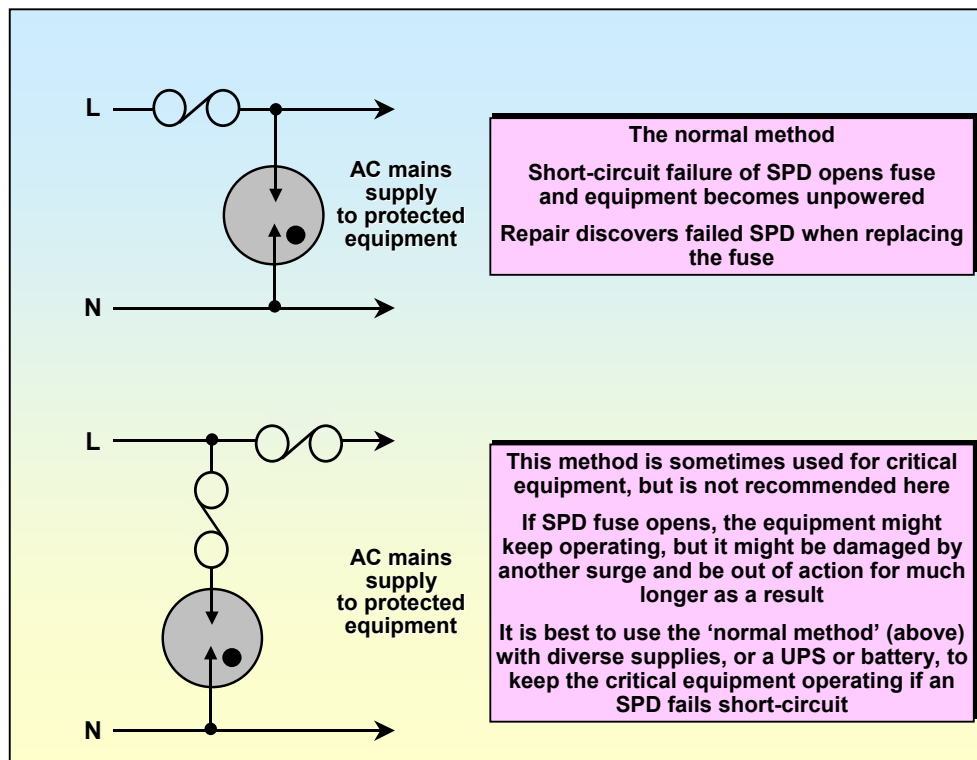
**Figure 3AU Protecting SPDs from overheating due to power-frequency overvoltages**

As already mentioned, SPDs are highly stressed components and wear out eventually (quite quickly, if not rated adequately for the surges in their EM environment). So another reason for using resistors, PTCs, fuses or circuit breakers in series as shown in Figure 3AU, is to prevent fire hazards arising if the SPD fails low-resistance (as MOVs always do, unless they are removed from the circuit by ‘explosive disassembly’). Some manufacturers offer MOVs that include thermal protection in their packages (e.g. TMOV and TPMOV products, see [18]).

SPDs can carry very large currents whilst suppressing surges, even kA, and it might seem that passing this current through a PTC, fuse or circuit-breaker is bound to cause it to operate. Of course, it would not be good if an equipment’s mains fuse (for example) opened every time its SPDs suppressed a surge. But it is not *current*

alone that operates PTCs, fuses or circuit breakers – it is current  $\times$  time, and co-ordinating of the current/time ratings of the fuses (etc.) with the SPD ratings (for example, using time-delay fuses) allows SPDs to suppress surges and be protected by the fuses, without suffering fuse reliability problems. More detail on how to coordinate fuses and SPDs is given in [18], which also describes a type of fuse specifically designed for use with SPDs.

Figure 3AV shows the normal method of fusing an SPD, and the principle applies equally well to SPDs protected by PTCs or circuit-breakers, see Figure 3AU. The fuse that protects the SPD is in series with the conductor that provides power or signal to the protected circuit. SPD failure or overheating protection will open the fuse, disconnecting the equipment from the input, making it easy for the user to discover that the fuse has opened and needs replacement. Replacing the fuse when an SPD has failed just makes the fuse open again, so the user knows to have the SPD replaced.



**Figure 3AV Examples of fusing SPDs**

Figure 3AV also shows another method, in which the SPD is fused separately from the equipment. This method is sometimes used for critical equipment, on the (mainly erroneous) assumption that if SPD fuse opens the equipment will keep operating, although it will not be protected against another surge. But this is not a generally recommended method for ensuring that critical equipment keeps functioning.

An SPD is most likely to fail whilst it is carrying a heavy surge current, and when the fuse opens this will cause a large flyback voltage that could well damage the equipment, especially if it adds to the surge voltage. Also, during a thunderstorm it is quite likely that several surges will occur, so if one of them has caused the SPD to fail there is quite a good chance that another surge will be along in a few seconds to damage the equipment.

Instead, it is generally much better to use the normal method of fusing, and if the operation of the equipment is critical, to use diverse power sources or add a UPS or other power back-up facility (such as a battery) to keep the equipment running when a mains fuse has opened.

Some proprietary surge-protection units use many SPDs of the same type, connected in parallel and individually fused, so that any that fail short-circuit do not prevent the unit from continuing to protect the equipment, and do not cause the mains fuse that powers the protected equipment to open. This is an application of the second method of fusing that is generally quite acceptable, providing the failed SPDs are replaced before the unit ceases to provide surge protection. Some units of this type incorporate condition monitors and indicators, to encourage users to get them refurbished before all protection is lost.

### 3.5.14 Equipment reliability and maintenance issues

When protecting AC power inputs (e.g. to off-line DC power converters) from surges using voltage-limiting SPDs, if a resistor is used in series with the input and a long-duration power-frequency overvoltage occurs – the voltage peaks will be clamped but the protected circuit will still have an AC power input (although very distorted)

and will still function. However, using a crowbar SPD or PTCs, fuses or circuit breakers will remove the AC power and so the protect circuit will no longer function.

As long as power-frequency overvoltages that activate the SPDs are rare enough, losing the functionality of the circuit is generally not a problem (assuming this does not increase safety risks). But some AC supplies suffer from quite common 'swells' or voltage fluctuations, and it would generally be unacceptable for equipment to frequently stop working, especially if it meant replacing a fuse or resetting a circuit breaker to make it work again.

When designing a surge protection it is tempting to choose SPDs that have as low a threshold/trigger voltage as possible, to provide maximum protection to the following circuit. But this increases the likelihood that commonplace swells and fluctuations will activate the SPD overheating protection devices and cause the circuit to stop working.

A particular problem is equipment designed in countries where they are used to a 220V or 230V mains supply, but used in the UK where the nominal mains voltage is still 240V despite being *called* 230V. In rural parts of the UK it is not uncommon to have 245V as the normal mains voltage. An overheat protection circuit for an SPD that has quite a low threshold/trigger voltage might have an acceptable rate of activation in mainland Europe, but cause reliability problems in the UK.

Designers should also be aware that where local generation is used, in many third-world countries, and even in some areas in more developed countries (even parts of the USA and Australia, for example) the mains supply voltage and waveform can be very poorly controlled indeed. Designers who assume a mains supply of 230V  $\pm 10\%$  with a nice sine waveshape can find their products failing frequently when used in parts of the world that don't have such well-controlled mains supplies.

So, for example, when designing an off-line switching power converter for use on 230V mains supply, instead of using 600V rated PowerFETs or IGBTs and trying to protect them with GDTs that trigger at 350V (the peak of a 240V pure sinewave is 339.4V) – a more reliable and robust design can be achieved, with less effort, using 900V transistors and GDTs that trigger at 600V.

SPDs eventually fail, and if they fail short-circuit they will generally be detected through their effect on the operation of the protected equipment, for example, by opening a mains fuse as discussed in 3.5.13. But they might also fail open-circuit, in which case the protected equipment might keep functioning as normal but no longer be protected from surges.

So maintenance is always important where SPDs are employed. Some types of proprietary surge protection units are available with condition indicators that inform the user if they are healthy, need repair but are still protecting, or are not providing surge protection any more. If using such units, it is important to have a reliable procedure that ensures the condition indicators are checked often enough, and that any units that need it are repaired.

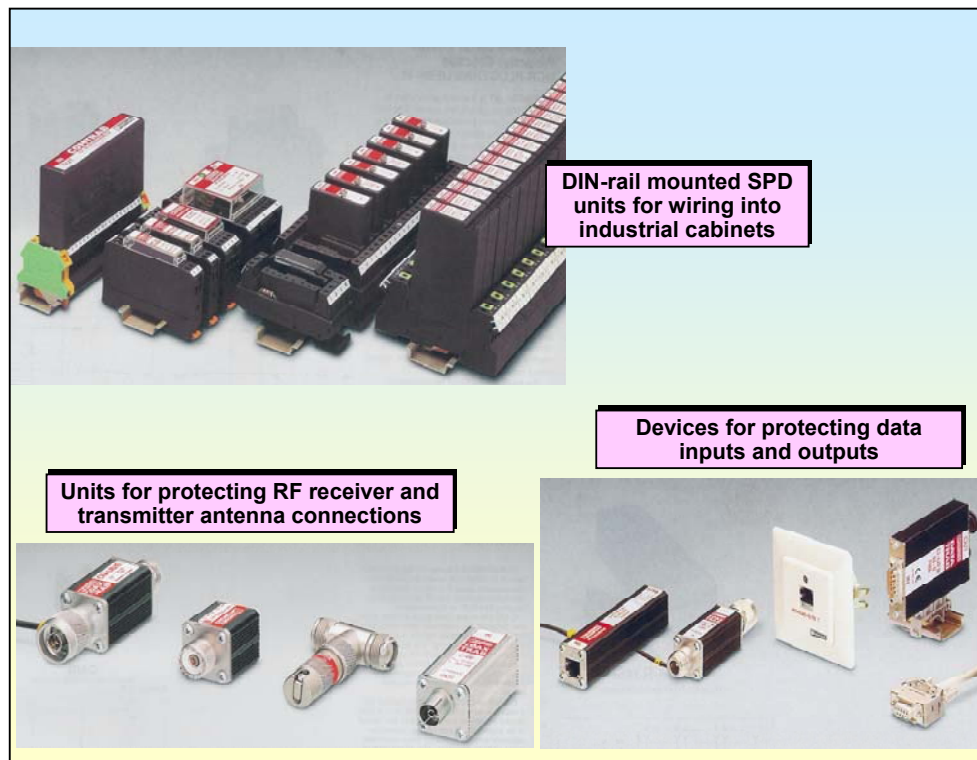
But where ordinary SPDs are used, a procedure will be needed for checking the state of its surge protection at regular intervals, to be sure of replacing failing SPDs before their protection is lost completely. The designer of the equipment needs to design so that this can be easily done, and make sure that all corresponding user maintenance instructions are written and communicated to the users.

### 3.5.15 Surge protection products

Figure 3AL showed some examples of small SPD devices. There are a very large number of surge protection products available that use SPDs to provide protection for power, signals and data in every different application, from protecting the three-phase supplies entering large buildings from direct lightning strike, to protecting telephone, radio and Ethernet circuits.

Figure 3AW shows some examples from a very wide range of proprietary surge products offered by numerous manufacturers. Industrial cabinets often use DIN-rail mounted protection units for protection AC supplies, for their ease of wiring. But such units are unsuitable for protecting RF or high-speed signals, so a range of surge protection units is available in different styles for these applications, as shown in the figure.





**Figure 3AW Some examples of SPD products (from Phoenix Contact)**

### 3.5.16 'Earth lift' problems in systems

So far, the above has discussed surge protection as it applies to an individual 'port' (e.g. a connection to a cable) on an item of equipment. But additional issues arise in systems, especially the phenomenon of 'earth lift' or 'ground lift' – caused by surge currents flowing in an earth/ground structure that is shared between several items of equipment.

A related issue occurs when surge currents flow in the neutral lead of an AC supply, or the return lead of a DC supply, that is shared between several items of equipment.

Figure 3AX illustrates the problem – all conductors and conductive structures have an impedance, that is predominantly inductive above a few kHz. When kA surge currents with fast rise-times are allowed to flow in them (either from SPD operation, or a spark-over to a chassis due to inadequate insulation) significant potential drops arise. These potentials arising in the common earth/ground structure then expose the electronic devices associated with interconnecting cables to surge voltages.

In a typical building with wired earth/ground structures, the resulting differences in earth/ground potential between two items of equipment can even approach the voltage of the initial surge. These voltages are called 'earth-lift' or 'ground-lift', and values up to 10kV are not unknown in large buildings. They are CM voltages, and can cause damage or interference problems for circuits connected to any power, signal, control or data cables that interconnect different items of equipment.

For example, a single straight wire in air (such as a green/yellow insulated earth/ground wire) has an inductance of about 1µH per metre, and a typical test that aims to simulate commonplace uni-directional surges has a peak current of 1000A with a current risetime of 10µs. Since  $V = -LdI/dt$  the earth/ground wire develops a longitudinal voltage (along the length of the wire) of 100V for every meter of its length.

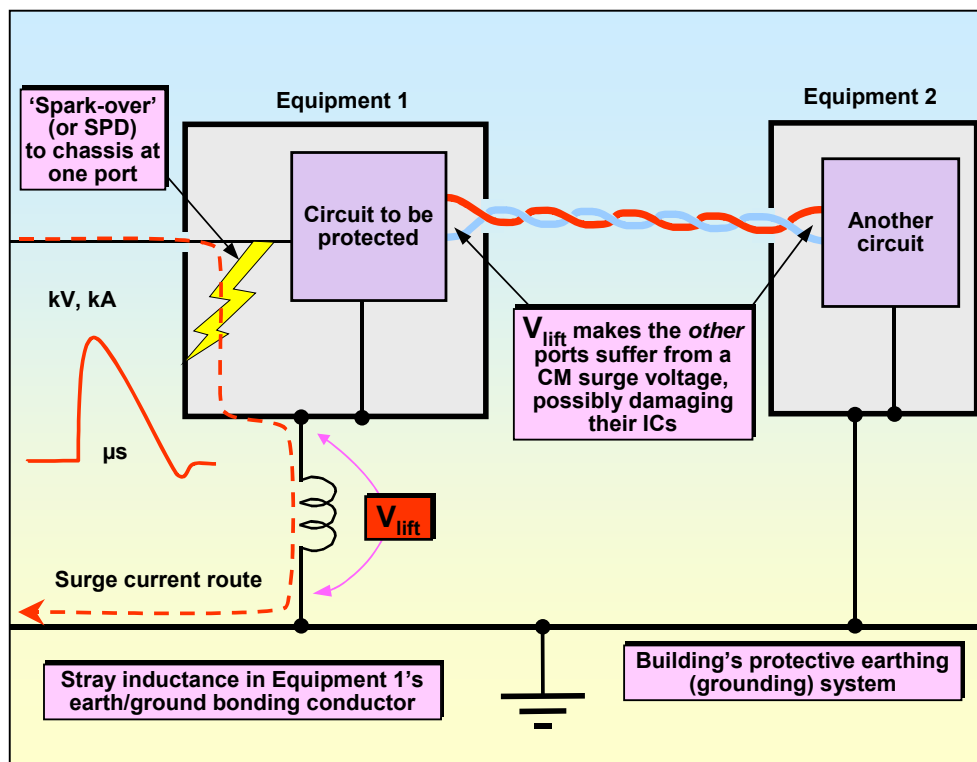
Most systems and installations still use the outdated single-point earthing/grounding principle, that could have been designed to cause maximum surge voltage exposure to equipment [19], so lengths of earth/ground cable 10m long are not unusual, leading to 1kV earth-lifts with the above surge current waveform.

So a consequence of using SPDs in equipment – rather than relying on galvanic isolation (see 3.5.1 and 3.5.4) – is that when that equipment is used in systems and interconnected by signal, control or data cables to other equipment some distance away, it will often be necessary to provide surge protection at all of its ports.

This issue is often overlooked because of the tendency to assume earths and grounds are perfect conductors, with no impedance. Solutions include:

- Not connecting any SPDs to the equipment chassis, frame or earth/ground. Instead, use galvanic isolation (e.g. isolating mains transformers, fibre-optics, etc.), plus adequate creepage and clearance distances between the incoming/primary circuits and the earth/chassis/frame/etc.

- Insulating sufficiently to prevent spark-over to the chassis, frame or earth/ground
- Reducing the impedance of the protective earthing/grounding system (sometimes known as the common bonding network: CBN), e.g. by connecting the chassis of the interconnected items of equipment together using short lengths of earth wire, or metal structures
- Protecting signal data and control inputs and outputs from damage, using galvanic isolation, filters and/or SPDs



**Figure 3AX 'Earth lift' surge issues in systems**

The ideal earthing/grounding system for controlling earth-lift is a mesh ('MESH-CBN') as described in IEC 61000-5-2. Such structures are commonly used in large computer/telecom installations, and are recommended by their relevant IEC standards and ITU Recommendations. Ships and offshore oil/gas platforms are generally made completely of welded or riveted steel sheets, which can be used as an ideal earth/ground by bonding all the equipment directly to it.

### 3.5.17 Data needs error detection/correction

SPDs on data lines only protect the devices from damage, they do not prevent false data from occurring during a surge. So data lines exposed to surges also need to use a good error-detecting, or (better still) error-correcting protocol. The very best bus may well be the military '1553' bus, versions of which are now available in civilian guise.

It is not recommended that designers try to create their own error-correcting protocols. Surges and similar transient phenomena occurring in real life are usually not as 'clean' as the waveforms used in the EMC tests supposed to simulate them, and typical protocols that are well proven to be robust in real life often have as much as 50 man-years of experience in robust communications behind them. The best advice is to purchase the devices and/or software that implement a proven robust communications protocol, rather than try to design one.

## 3.6 References

- [1] Keith Armstrong, "Design Techniques for EMC", UK EMC Journal, a 6-part series published bimonthly over the period February – December 1999. An improved version of this original series is available via the "Publications & Downloads" page at <http://www.cherryclough.com>
- [2] "EMC for Functional Safety", The IET, 2008, a 180-page practical guide suitable for all industries and applications, available for free download from [www.theiet.org/factfiles/emc/index.cfm](http://www.theiet.org/factfiles/emc/index.cfm) or as a colour printed book for £27 from: <http://www.emcacademy.org/books.asp>
- [3] Arthur B Williams, "Electronic Filter Design Handbook", McGraw Hill, 1981, ISBN 0-07-070430-9

- [4] John R Barnes, *“Robust Electronic Design Reference Book, Volume I”*, Kluwer Academic Publishers, 2004, ISBN: 1-4020-7737-8
- [5] Sokal, N. O., *System Oscillations From Negative Input Resistance at Power Input Port of Switching-Mode Regulator, Amplifier, DC/DC Converter, or DC/AC Inverter*, IEEE Power Electronics Specialists Conference (PESC) 1973 Record, pp. 138-140.
- [6] Keith Armstrong, *“Design Techniques for EMC, Part 2 – Cables and Connectors”*, The EMC Journal, May and July 2006, available from <http://www.compliance-club.com>.
- [7] Keith Armstrong, *“Design Techniques for EMC, Part 0 – Introduction and Part 1 – Circuit Design and Choice of Components”*, The EMC Journal, January 2006 pp 29-41, plus March 2006 pp 30-37, available from <http://www.compliance-club.com>.
- [8] F Beck and J Sroka, *“EMC Performance of Drive Application Under Real Load Condition”*, presented at the Industrial Forums in EMC Zurich 2001, and also a Schaffner EMV AG application note dated 11th March 1999. It was also presented by W L Klampfer at the 8th International Conference on Electromagnetic Interference and Compatibility, INCEMIC 2003, ISBN: 81-900652-1-1, publication date: 18-19 Dec. 2003.
- [9] Keith Armstrong, *“EMC for Printed Circuit Boards – Basic and Advanced Design and Layout Techniques”*, February 2007, [www.emcacademy.org/books.asp](http://www.emcacademy.org/books.asp), £47 plus p&p, perfect bound: ISBN 978-0-9555118-1-3, spiral bound (lays flat): ISBN 978-0-9555118-0-6
- [10] The United Kingdom Accreditation Service, <http://www.ukas.com>
- [11] John R Barnes, *Robust Electronic Design Reference Book, Volume II, Appendices*, Kluwer Academic Publishers, 2004, ISBN 1-4020-7738-6
- [12] RF Café, *Skin Depth*, [http://www.rfcafe.com/references/electrical/skin\\_depth.htm](http://www.rfcafe.com/references/electrical/skin_depth.htm)
- [13] Tim Williams and Keith Armstrong, *“EMC for Systems and Installations”*, Newnes 2000, ISBN 0 7506 4167 3, especially chapter 8, [www.newnespress.com](http://www.newnespress.com), RS Components Part No. 377-6463
- [14] ITI (CBEMA) Curve and Application Note: <http://www.itic.org/archives/iticurv.pdf>
- [15] *“PCB Layout: The Impact of Lighting and Power-Cross Transients”*, Milton Hilliard, Compliance Engineering, January/February 2003 pp 25-30, available from the archives at <http://www.ce-mag.com>
- [16] MIL-STD-275 *“Printed Wiring for Electronic Equipment, Revision: E, Dated: 31 December 1984”*, available via: <http://www.dscc.dla.mil/Programs/MilSpec/ListDocs.asp?BasicDoc=MIL-STD-275>
- [17] Akihiko Yagasaki, *Characteristics of a Special-Isolation Transformer Capable of Protecting from High-Voltage Surges and Its Performance*, IEEE Trans. EMC, Vol. 43, No. 3, August 2001, pp 340-347
- [18] Daniel Dunlap, *Protection of SPD Products*, ITEM 2000, pp 148-157, visit [www.interference-technology.com](http://www.interference-technology.com) and search by 'Daniel Dunlap'.
- [19] Keith Armstrong, *The Benefits of Applying IEC 61000-5-2 to Cable Screen Bonding and Earthing*, IEE Seminar entitled: “EMC – its nearly all about the cabling”, The IEE, Savoy Place, London, January 22<sup>nd</sup> 2003.

### 3.7 Acknowledgements

I am very grateful to the following people for suggesting a number of corrections, modifications and additions to the first series published in 1999 [1]: Feng Chen, Kevin Ellis, Neil Helsby, Alan Keenan, Mike Langrish, Tom Liszka, Tom Sato, and John Woodgate. I am also indebted to more recent input from Richard Marshall.