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EMC techniques in electronic design Part 6 - ESD, electromechanical devices, power factor correction

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Design Techniques for EMC

Part 6 — ESD, electromechanical devices, power factor correction, voltage fluctuations, immunity to supply dips and dropouts and other power quality issues

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This is the **sixth** and final article in this series on basic good-practice electromagnetic compatibility (EMC) techniques in electronic design, published during 2006-8. 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 (screening)
5) PCB layout (including transmission lines)
6) **ESD, electromechanical devices, power factor correction, voltage fluctuations, immunity to power quality issues**

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.

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6. ESD, electromechanical devices, power factor correction, voltage fluctuations, immunity to power quality issues

6.1 Electrostatic Discharge (ESD)

6.1.1 ESD threats

Normal commercial and industrial ESD tests employ the IEC/EN 61000-4-2 basic test method that attempts to simulate 'personnel discharges' from people's fingers. We have all experienced such discharges when the humidity of the air is low, when touching a metal object such as a door handle. Ordinary people do not generally notice ESD events from their fingers that are less than about ±3kV, and ESD events that make people hop about and complain loudly are generally in excess of ±15kV.

Figure 6A gives some examples of the electrostatic voltages that can be generated on a human body just by moving around in a typical building and doing ordinary things, for various values of the relative humidity of the air, from [3]. The mechanism by which this and most other terrestrial electrostatic charges are generated is called tribocharging, but this is not the article to discuss that phenomenon.

Generation method	Typical electrostatic voltage generated (in kV)	
	10-20% Relative Humidity (RH)	65-90% Relative Humidity (RH)
Walking across carpet	35	1.5
Walking on vinyl floor	12	0.25
Worker moving at non-metal bench	6	0.1
Opening a vinyl envelope	7	0.6
Picking up a polyurethane bag	20	1.2
Sitting on a polyurethane foam padded chair	18	1.5

Figure 6A Examples of personnel electrostatic charging

The IEC/EN 61000-4-2 test method uses an ESD 'gun' that discharges a 150pF capacitor through a 330Ω resistor to create ESD events up to ±8kV at up to ±30A, with risetimes between 0.7 and 1ns. The high dV/dt and dI/dt of these ESD events ensure that they have significant EM energy at frequencies beyond 1GHz. The test method is described in [4] (page 184), Chapter 43 of [5], Part 3 of [6] (which also describes some low-cost alternatives), and in the guide to EN 61000-4-2 in [7].

Figure 6B sketches the basic circuit elements of an IEC/EN 61000-4-2 ESD gun, which can be fitted with two types of discharge tip: a round one that simulates a human finger and is used for creating discharges in the air, and a pointed tip used for discharging by direct contact with conductive surfaces or objects.

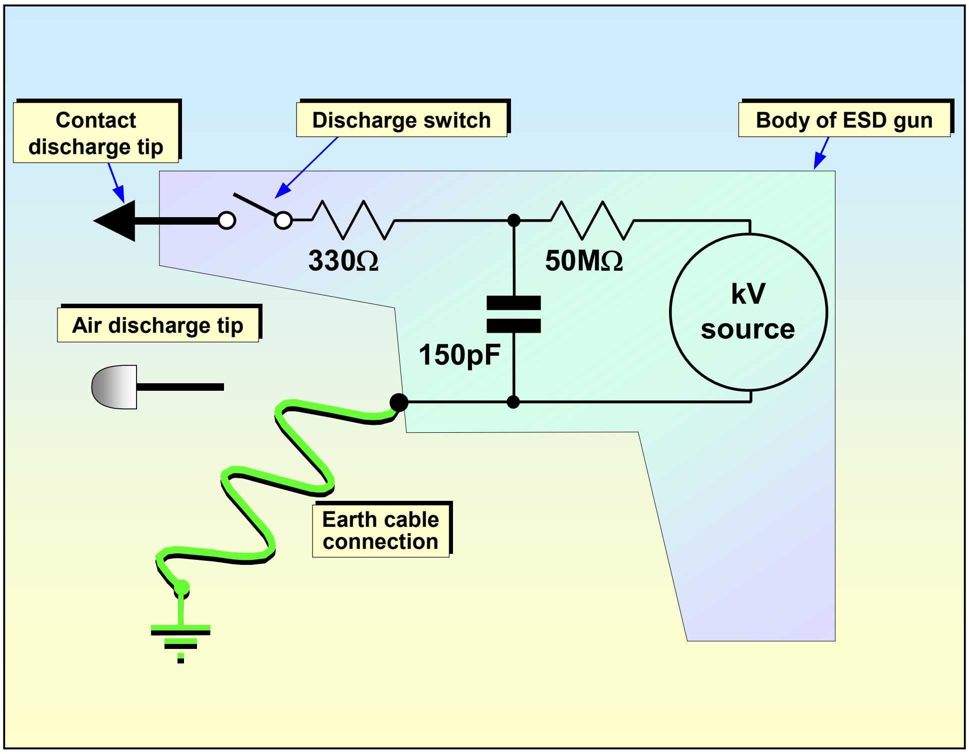


Figure 6B Overview of an IEC/EN 61000-4-2 'ESD Gun'

Figure 6C shows an example of a commercially available ESD gun, that has plug-in modules for various discharge waveshapes, including that specified in IEC/EN 61000-4-2 (see Figure 6F).

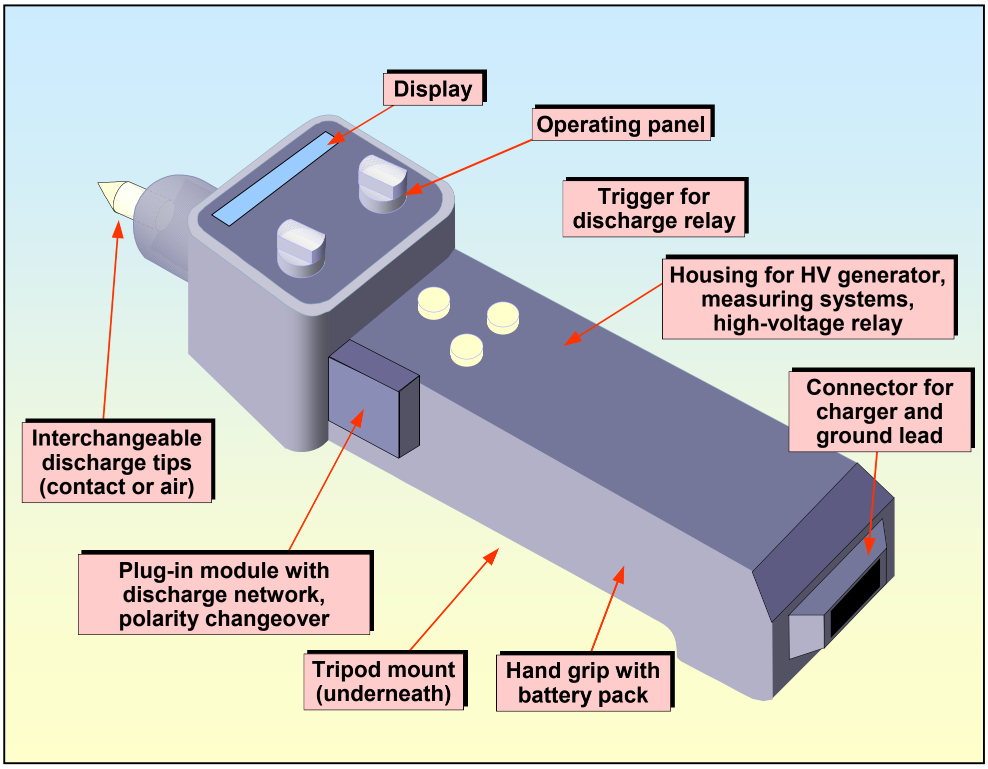


Figure 6C Example of a KeyTek MiniZap®

Figure 6D shows that testing to IEC/EN 61000-4-2 simulates three different kinds of 'personnel ESD' events, and can also cause 'secondary arcing' to occur – effectively ESD events within the product's structure. The 'near-fields' from an ESD test can be kV/m at 1m from a discharge, and kA/m within 100mm of a discharge – these are very intense fields indeed.

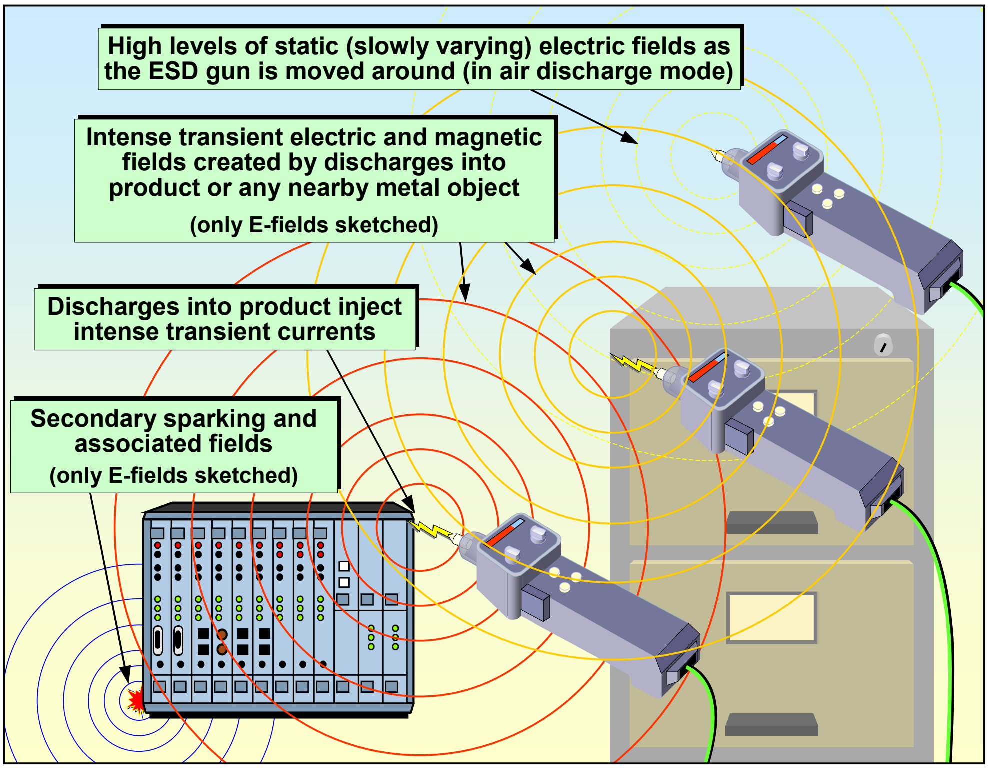


Figure 6D Various EM phenomena associated with ESD testing

The automotive industry uses the ESD test method ISO 10605 instead of IEC/EN 61000-4-2, and tests up to ±25kV with products powered, and when unpowered to simulate handling during shipping and installation [8].

These ESD tests inject sufficient voltage and current into products to permanently damage ICs and transistors, and even some passive components, see Figure 6E. And the intense electric (E) and magnetic (H) fields they create can couple transient noises into nearby circuits and disrupt signals, causing errors and often creating big problems for software.

Type of Device (typical 2002 technology)	Typical level at which damage occurs (kV)
MR heads, RF FETs	0.01 - 0.1
Power MOSFET transistors	0.1 - 0.3
VLSI (e.g. microprocessors, FPGAs, memory)	1 - 3
Film resistor	1 - 5
HC and similar CMOS glue logic	1.5 - 5
Small-signal bipolar transistor	2 - 8
Power bipolar transistor	7 - 25

Figure 6E Examples of ESD damage levels for devices

The damage levels in Figure 6E are based on tests in 2002 on unassembled devices (so ignores any protection provided by their circuits and enclosures) using the semiconductor manufacturing industry's 'human body model' – which discharges a 100pF capacitor through a 1.5kΩ resistor, generating a peak current of ±2A with a risetime of between 5 and 20ns. Modern (2007) microprocessors, memory devices and glue logic use insulating layers that are a great deal smaller than their 2002 ancestors, and probably have damage levels around 100V – reducing all the time as silicon feature size continues to reduces according to Moore's law [26].

Figure 6F shows the waveform of the discharge current that is specified when calibrating an ESD gun to IEC/EN 61000-4-2, and it also shows the sort of waveform that can obtain in real life due to radio-frequency (RF) resonances in the product being tested. Resonances can extend the ns transient of an ESD spark into a complex electromagnetic event that lasts for tens of μs.

There are concerns [9] that the IEC/EN 61000-4-2 tests do not simulate real-life personnel ESD events well-enough to prove that products incorporating modern microprocessors and memory chips will be reliable in real life. [9] also claims that the standard does not specify the design of the ESD gun well enough to prevent significant differences when testing a given real product with different manufacturers' guns. This is also discussed in the guide to EN 61000-4-2 in [7].

But in real-life applications, ESD events can originate from a wide variety of sources other than people's fingers, as sketched in Figure 6G. These sources can have much higher values of capacitance than 150pF, and/or much higher voltages (up to ±40kV has been seen) or risetimes as low as 10ps.

It may seem odd that, as indicated in Figure 6G, fluid flow can cause ESD – but many serious incidents and accidents in the petroleum and other industries have occurred due to this very problem (see No. 458 in [12]), and it is also implicated in the crash of TWA 800 from an explosion caused by sparking in one of its fuel tanks. Spacecraft can suffer from very high levels of ESD due to charging of insulated parts by the solar wind, and by charged particles from outer space. And aircraft (fixed and rotary wing) can become charged up to very high levels due their passage through the air, especially during certain weather conditions (see No's 22, 23, 294, 295 and 431 in [12]). Motor vehicles can also become highly charged (see No. 366 in [12]).

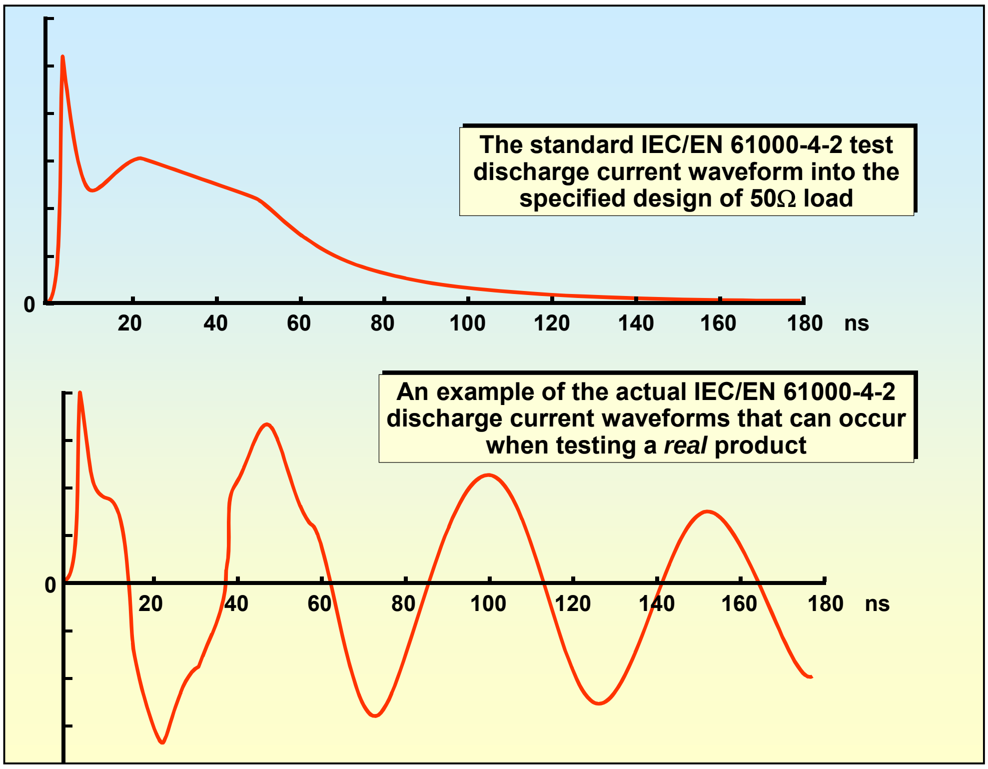


Figure 6F Examples of ESD waveforms

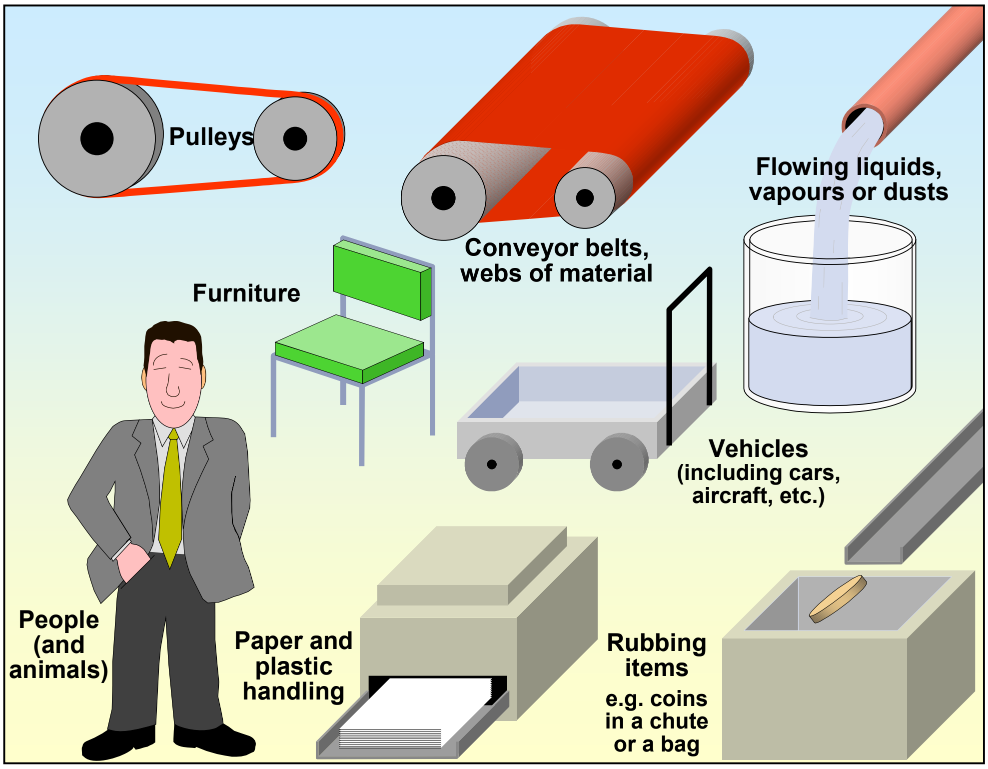


Figure 6G Examples of some ESD sources

Products that pass ESD tests in a laboratory can fail in the field due to more aggressive ESD events in their operational environments. Very high voltages and very low risetimes do not generally go together. High-voltage events tend to have a risetime of 1ns or longer, whereas low-voltage events (such as caused by jingling coins in a plastic bag, see [10]) can have risetimes as low as 10ps – with a spectrum of energy that extends well beyond 10GHz. The measurement of ESD risetimes is limited by the availability of suitable instrumentation, and it seems that as oscilloscopes get faster, we discover that real-life ESD events can be faster than we previously thought.

For more background on ESD and the forms it can take visit [10] and [11]. 21 examples of real-life ESD problems are described in [12]. An interesting example is the ESD caused by the rotors of AC motors running in nylon or other insulating bearings. Few designers would expect the little motor embedded within their product to be an ESD source, but the E- and H-fields created when its rotor discharges across its bearing to its frame can easily upset microprocessors and cause software to malfunction or crash.

6.1.2 Prevent ESD by preventing electrostatic charge from building up

This is generally a system or installation design technique, but it is used in all semiconductor manufacturing areas, and widely used in electronic assembly areas, so products intended for use in such environments can benefit and may not need to pass any ESD tests.

There are two basic methods: one is to make sure that all the materials used are dissipative (i.e. have a resistance between 10^6 and 10^9 Ω/square), and are connected to the ground reference, so that electrostatic charges decay quicker than they are generated and high voltages cannot build up. Some materials are made dissipative by coating them with appropriate materials (e.g. antistatic spray for carpets and furnishings). But many coatings only function as intended in atmospheres with a certain minimum relative humidity – so humidity control becomes a necessary feature of the heating, ventilating and air-conditioning system of the area.

The other method is to make the air itself conductive by using high-voltage needles to create alternating batches of negative and positive ions in a fan-blown air-stream. Ionised air is conductive, and alternating negative and positive ionisation results in air that is neutrally charged on average and so does not cause electrostatic charges to accumulate. Blowing the ionised, conductive air around the area to be protected causes any static charges on products, furniture or people to dissipate. In fact, a neutrally ionised air stream is the one sure way to remove charge from the surface of an insulator without having to wipe all over it with a grounded conductive brush or cloth.

The above techniques can also be used *within* products, to improve their reliability by discharging rotating belts, pulleys, motors with nylon bearings and the like so that they don't give rise to internal sparks that could upset their electronics. Dissipative materials can be used, such as conductive rubber (instead of insulating rubber) for drive belts, conductive plastics for wheels and pulleys, etc.

Insulating parts that move, including consumables such as the paper in a photocopier or printer, can also be discharged with grounded conductive brushes, often made of stainless steel or carbon fibre for longevity. Also, neutrally ionised air streams can be blown inside equipment to prevent the build-up of static charges.

6.1.3 Prevent the discharge from happening with insulation

When a product has to cope with external ESD from people or other sources, a very powerful design technique is insulation. We use plastic enclosures, membrane keyboards, plastic knobs, switch caps and control shafts, etc. to prevent the injection of the intense discharge currents into the product – in effect we simply do not permit the charged person or object to discharge into our product.

Figure 6D shows that this technique still leaves the product exposed to the slowly varying electrostatic fields and the intense E and H-fields from nearby discharges. Slowly varying fields are generally only of concern for very high impedance circuits (typically >1MΩ), and both these and the intense fields can be dealt with by techniques described in 6.1.8: for example a product might use all the techniques described in the earlier parts of this series [13] [14] [15] [16] and [17] – and then have insulation applied all over as well, to prevent direct ESD.

Typical plastics have a breakdown voltage through their thickness of about 40kV/mm, which can be a problem for membrane type control panels in extreme ESD environments if we want to use 'clicky' tactile buttons. To get a good button-clicking experience we need to a top plastic layer of about 0.5mm that will insulate up to about 20kV, if we have to use thicker layers it will ruin the feel of the button.

Membrane panels can employ a metal shielding layer immediately below their top layer of insulation, as described in 4.3.13 in [16]. This should intercept any discharges that manage to penetrate the top insulating layer, but to be effective it must be connected to the product's RF Reference, which in a shielded enclosure will be the enclosure shield itself [16], and in an unshielded enclosure will be the PCB's 0V plane [17] [18].

4.3.13 in [6] says that the shield layer should RF-bond to the metal enclosure all around its perimeter. This not the normal method used by membrane panels manufacturers, who generally use a 'shield grounding' trace in the flexi-ribbon cable that connects the panel's switch traces to the PCB. This is effectively a 'pigtail', like the bad-practice method of terminating cable shields discussed in 2.6.6 of [14], and it allows stray RF coupling into the membrane panel's conductors. If we do not use a metal shield within our insulating enclosure, for example as discussed in 4.7.7 of [16], we might need to filter the membrane panel's interconnections as described in [15] or ESD-suppress them as described 6.1.5 below.

Capacitive sensing techniques will work through almost any practical thickness of plastic, glass or ceramics and so can be made to withstand any ESD voltages, but they provide no tactile feedback at all. Using a remote control, such as a wireless remote, allows us to locate the human interface in a more benign ESD environment.

Air and vacuum are the biggest problems when using insulation to prevent actual discharges from occurring to the product. Enclosures must have seams and joints to make it possible to assemble them, and these create gaps in the insulation, and the gaps contain air. Air has a breakdown voltage of only about 1kV/mm, less if humidity is high. In space the gaps are filled with vacuum, which also has a breakdown voltage of about 1kV/mm but does not suffer from variations due to weather.

The resulting problem is shown in Figure 6H – we need very large air gaps between conductors and places that could be touched by people or other ESD sources, to be sure they don't break down and allow a discharge into the product.

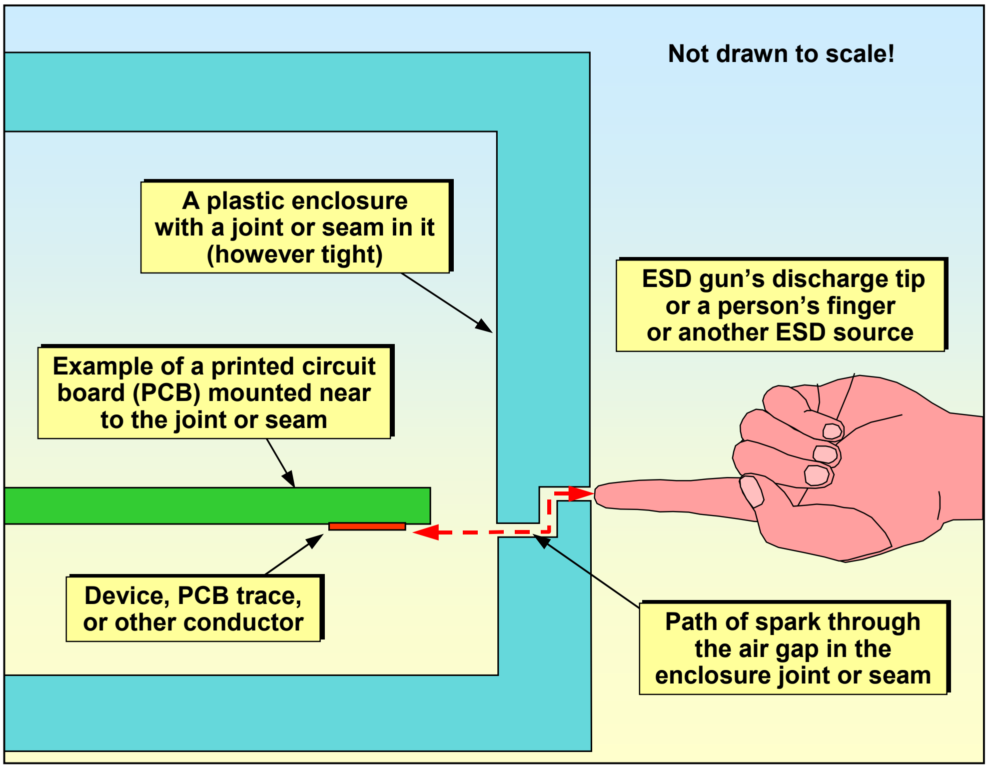


Figure 6H ESD penetrating a seam in a plastic enclosure

To withstand ±8kV we need *at least* an 8mm air-gap, and *at least* 25mm for ±25kV, taking into account the reduced breakdown voltage with increased humidity (unless the product is intended to work in a vacuum instead). Another issue is that all practical insulating surfaces are coated with dirt, damp, greasy fingerprints, possibly even mould, so discharges will find the surface of an insulator to be an easier path than even the air. It is not uncommon during ESD testing to see a spark from an 8kV discharge wriggle around on a painted metal surface for several tens of millimetres, tracking through the dirt and other contamination before finding a microscopic pinhole that allows it to reach the metal surface underneath.

So the best approach to insulating surfaces is to assume they are conductors and not take them into account at all in the total length of the air gap. Figure 6J shows how the air gap in Figure 6H can be increased, and also introduces the 'guard ring' PCB technique and the possible need to use shielding for very sensitive devices or traces.

People have been using perimeter guard rings on PCBs for decades, but because of the prevalence of the 'single-point grounding' myth, they thought it best to use a long trace around the PCB perimeter, connected to the 'chassis ground' – or whatever – at one point. As [14] shows, all they were really doing was creating resonant structures that were very effective antennas at certain frequencies. These take the broadband energy in the ESD discharge and re-radiate it as a very intense field at their resonant frequency, possibly replacing one type of ESD failure with another.

Another possibility is that because the inductance of such a guard ring was so high, when it received a discharge its voltage could rise so high that it then caused a secondary discharge to the devices and traces it was supposed to be protecting.

The *only way* to implement an effective guard trace for ESD (or for any RF purpose above a few MHz) is to start with an RF Reference plane layer that maintains a very low impedance up to the highest frequency of concern for ESD (in excess of 1GHz) – and then connect the guard trace to the plane with via holes whose spacing is much smaller than the wavelength of the highest frequency. The effect of the dielectric constant of the PCB on the wavelength must be taken into account [17] [18].

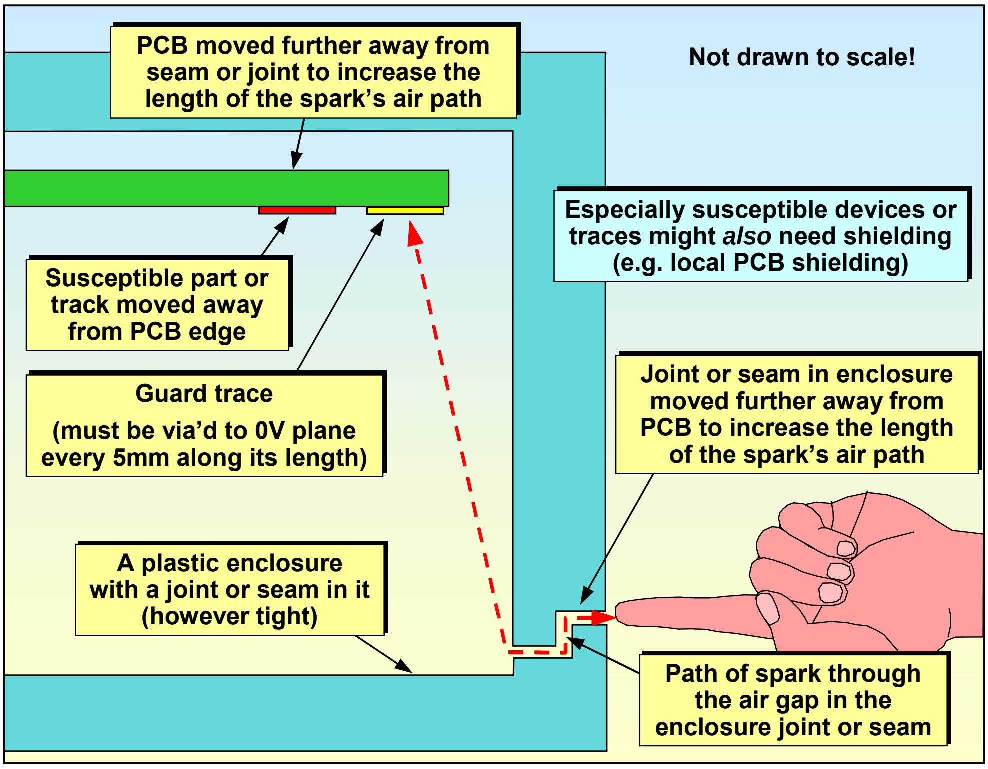


Figure 6J Solutions to the problem of ESD penetrating seams in plastic enclosures

The previous part of this series [17] only describes basic EMC techniques for PCBs, so for the details of implementing perimeter (or other) guard traces that are effective for RF and/or ESD, see [18].

An alternative approach to the problem of joints and seams in insulating enclosures is simply to fill them up with insulator, such as a (gas-tight) rubber gasket, silicone or epoxy sealant. The sealant approach is very acceptable for joints and seams that should never need to be opened during the life of the product (e.g. around the edges of a display), and the rubber gasket approach can be practical where access is required.

Beware of the temptation to try to make a totally sealed product. It is more difficult than it seems, and there have been many attempts that ended up with excessive amounts of condensation sloshing around inside, causing rapid corrosion.

Controls and displays are weak points in any ESD scheme, because they must somehow connect between the protected circuitry and the world inhabited by charged-up people and other ESD threats. Plus, of course, the charged-up people keep insisting on pointing at things on displays and touching the controls. (Cable and antenna connections are also weak points, and these are discussed in 6.1.5.)

LED and filament lamp indicators can use surface-mounted devices with plastic light guides to communicate their light to the human interface, as shown in Figure 6K. The light guides can be cheaply made from injection mouldings that snap into the enclosure and align with the LEDs or lamps on the PCB. This method is very low-cost, but it is important that any seams or joints between the light guides and the enclosure's insulating surface are friction-welded, glued or sealed so that sparks cannot track along contamination on the light guide's surface and get into the PCB.

Another technique is to present the LEDs or lamps at apertures in the enclosure, but cover them with a glued-on plastic overlay, generally the one carrying the control panel markings, that has transparent areas over the displays.

A problem with glued plastic layers, that also afflicts membrane panels, is the uniformity of the glue layer. Any missing glue, or imperfect bonding with the insulating enclosure surface, will create an air-gap that will allow sparks to slip under the overlay or between the laminated layers in the membrane panel and inject discharge currents into indicator devices, or into printed traces in membrane panels.

Where glue uniformity and quality cannot be guaranteed, make sure the edges of the plastic layer extend a very long way beyond the vulnerable components or traces. For 8kV ESD, 20mm would not be excessive. Or else seal them with silicone or other insulation as shown in Figure 6K.

For rotary shafts for switches, potentiometers and encoders, toggle switches, and similar manual controls, plastic knobs, shafts and toggles are recommended. For many years now, equipment has been so miniaturised that their control knobs are so small that discharges from operator's fingers can easily track across their surfaces and into any metal shafts they are mounted on – thereby penetrating the insulating enclosure and damaging some vital device or scrambling its software. So plastic shafts should be used, as shown in Figure 6K.

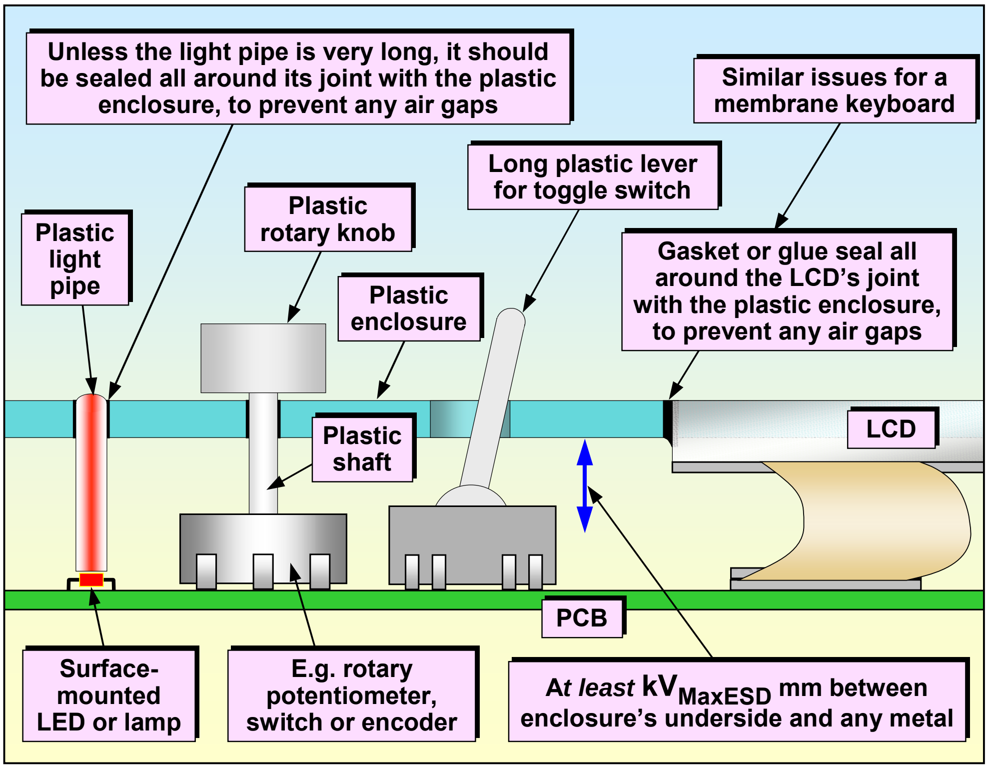


Figure 6K Indicators, displays and controls penetrating plastic enclosures

Figure 6K also shows that the assembly of LCD panels and graphics displays should avoid exposing their edges to ESD events. This is where a good mounting bezel with a rubber gasket, or a silicone or other type of sealant (that could also be used to hold the display in place, simplifying assembly) can be very useful. The fixed windscreens in almost all models of motorcars introduced since 1990 are an example of appropriate assembly techniques. Older vehicles used to have all sorts of bezel contraptions to hold them in place, and it was not unusual for them to leak (allow water to penetrate the enclosure) – but these days they simply glue them in, and incidences of water penetration are rare.

6.1.4 Control the discharge with shielding

Shielding is an alternative ESD suppression method to insulation (see 6.1.3). It allows the discharge to occur to the product, but then seeks to control it so that it doesn't upset any of the product's electronics. Shielding techniques were discussed in Part 4 of this series [16], and at first sight it might seem that all we need to do is design our shielding to be effective enough at a high-enough frequency.

For normal ESD testing, with risetimes close of 0.7ns or longer, we can assume the highest frequency of concern ($1/\pi t_r$) is about 500MHz, but real ESD events are much faster than this, at 0.3ns or less [9] and so we should assume 1GHz or more instead.

But the very high intensities associated with ESD events (tens of amps with sub-ns risetime, E-fields of kV/m, H-fields of kA/m) significantly increase the demands on our shielding, In fact, designing shielding for ESD is rather like designing it for military or aerospace purposes, where we can be dealing with kV/m E-fields at 1GHz or more from nearby radars, so in this section we need to discuss how to apply the techniques described in [16] to the ESD situation.

Clearly, with such high levels of E and H near-fields, the shielding effectiveness (SE) required at the highest frequency of concern will be higher than what is usually required to cope with the normal domestic, commercial and industrial environments (typically tested at 3V/m or 10V/m, although achieving immunity to the close proximity of cellphones, walkie-talkies or GPRS-enabled computing devices can require testing at 60V/m or more).

Because of the very high levels of ESD current flowing around the outer skin of a shielded enclosure, any gaps or apertures that make these surface currents divert from their natural paths become very intense sources of secondary E and H-fields. So it is very important to locate sensitive devices, PCB traces and conductors *very far away* from even tiny gaps or joints in the shield. Figure 4R in [16] shows the general principle, but much more than its 40dB of SE might be required.

The voltages developed across a gap or aperture in a shield, due to their diversion of the flow of the ESD currents, can be so high that they break down the air (or vacuum, in the case of spacecraft) at that point and spark across the gap. This is known as *secondary arcing* and, as might be expected, where it occurs it can cause very great problems. It can even occur *inside* products whose external shielding provisions are not as good as they should be, generally playing havoc with their electronics.

Secondary arcs are often small faint blue things that are hard to see even when right in front of your eyes, but more often than not they are hidden within a metal seam, or inside or on the bottom surface of the product being tested and so even less visible, as sketched in Figure 6L. When secondary arcing is suspected, for example when the ESD gun is applied to the top of the product, but the microprocessor that resets (and its reset lines) are located near the bottom, a powerful diagnostic technique is to do ESD testing in the dark.

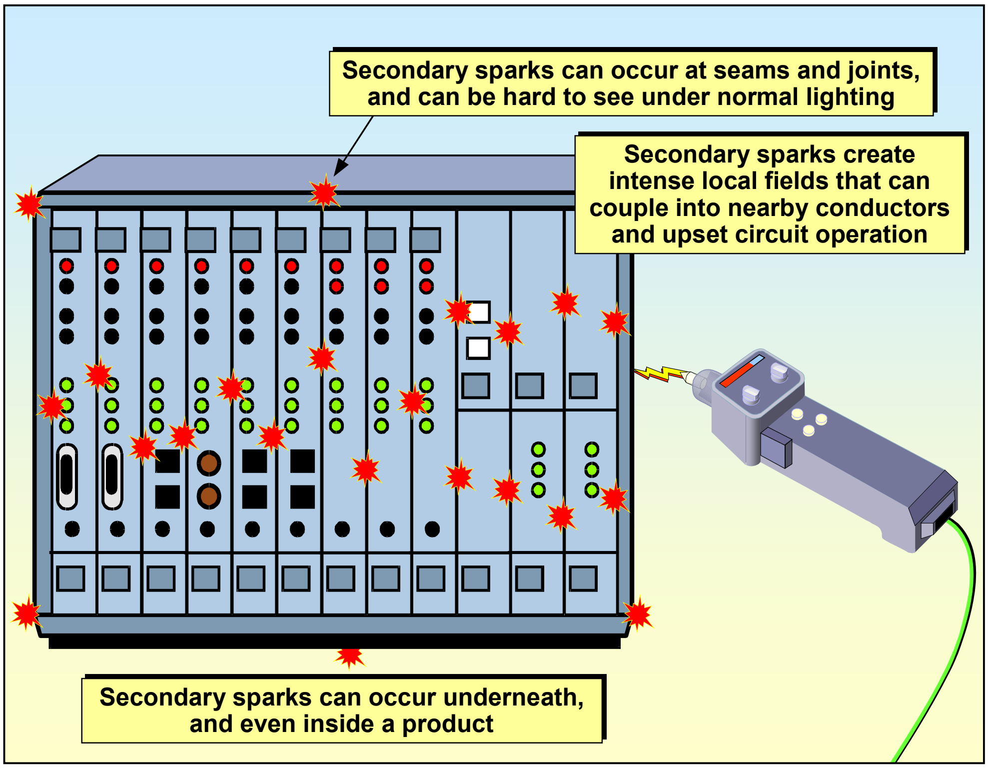


Figure 6L Sketch of some secondary arcing possibilities

It is of course inadvisable to do ESD testing in total darkness, because we need to see what we are doing well enough to:

a) Apply the ESD gun to the correct point on the product, and in the specified manner

b) Not accidentally discharge the gun to ourselves (painful, but not damaging to people. Where one's health depends on implanted or portable electronic medical devices such as pacemakers or defibrillators, you should not be anywhere near an ESD test anyway.)

So close the blinds and/or turn the lights down quite a lot, and wait a few minutes for your eyes to get accustomed to the gloom. People can see quite well by moonlight, which has one-millionth the luminous intensity of sunlight, so given time our eyes adapt to gloom very well.

The tester has to watch where the ESD gun is applied and has a limited ability to monitor other areas of the product, so spotting any secondary arcing can be made much easier if someone else looks closely at different parts of the product during the tests. It can also help to reorient the product, for example lying it on its side to see its underside. Detecting internal secondaries can require more radical techniques.

Indicators, controls and displays are weak points for ESD when relying on shielding, just as they are for the insulation techniques discussed in 6.1.3. The insulation-based techniques sketched in Figure 6K (plastic light guides, knobs, shafts and toggles, etc.) are effective with metal enclosures too, but the apertures they create in the shield might cause problems for nearby sensitive devices when discharges occur to the metal surface, or when shielding for frequencies above 300MHz. Figure 6M shows some alternative techniques that prevent the creation of apertures in the shield.

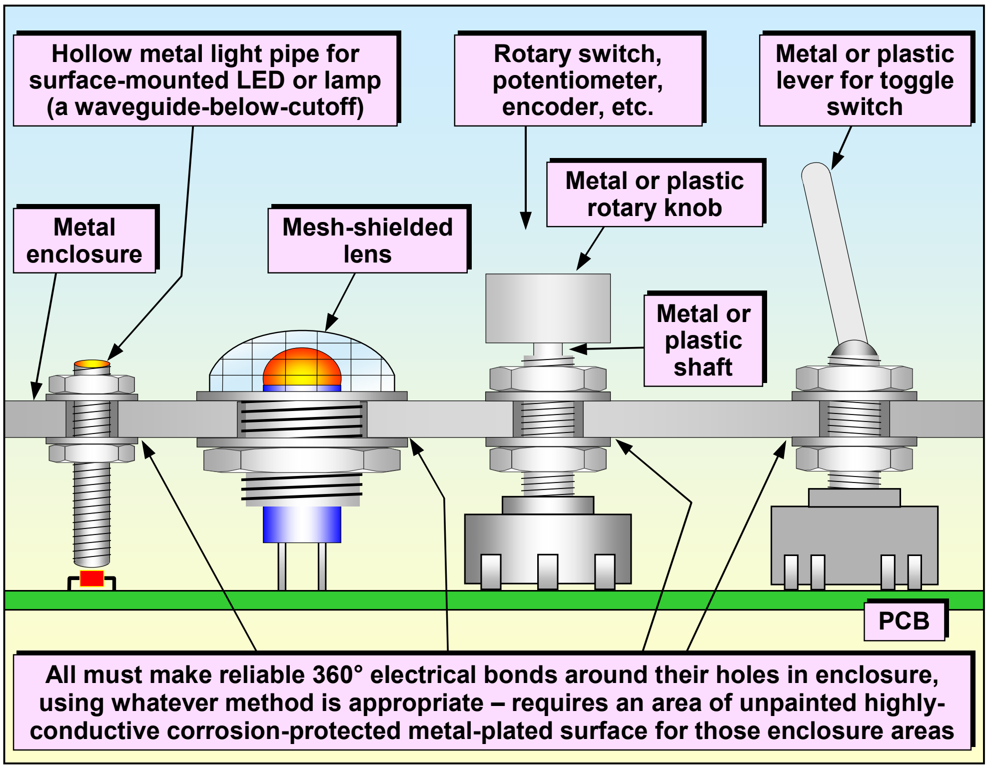


Figure 6M Indicators, displays and controls penetrating shielded enclosures

Figure 6M does not show how to deal with panel-type displays or membrane keyboards. Displays need to be treated using one of the techniques described in [16] (e.g. Figure 4AH). Membrane keyboards should RF-bond their metal backing plates, and/or any internal shielding layers to the shield all around their perimeter, using conductive gaskets, see 4.3.13 in [16]. Care should be taken to prevent discharges into their edges or backs.

It is often assumed that a Faraday Cage (i.e. an effective shielded enclosure) always prevents any external voltages from creating voltage differences within it. Whilst this is true for established DC voltages (as Michael Faraday found, when sitting inside his eponymous cage holding a gold-leaf electroscope) – and also true for continuous RF for enclosures with no apertures made of a metal with at least 10 skin-depths at the frequency concerned. But it is not true for transient voltage fluctuations such as ESD.

When a discharge is applied to a shielded enclosure, at first the transferred charge spreads all over the outer skin of the metal shield, and in the short-term whilst current is still flowing, it is confined to the outer surfaces by skin effect. Once the currents have equalised the voltage all over the outer skin of the shield (which they do at the speed of light, so it would only take about 1ns for a small enclosure) they stop flowing, and the charge then becomes static – an electrostatic voltage on the outside of the metal enclosure.

Over the next few ns the charge diffuses through the thickness of the metal shield material until it appears on its inner surface. The rate of diffusion depends on the relative permeability of the metal – the higher it is, the smaller the skin depth and the slower the rate of diffusion.

Inside any metal enclosure there are hundreds or thousands of stray capacitances between the shield material and each device (in fact, each pin of each device), PCB traces and other conductors – and they are all different. When the charge appears on the inner surfaces of the shield, it charges up these stray capacitances, and during this process they carry charging currents (sometimes called displacement currents). These transient charging currents will of course be injected into the devices, PCB traces and other conductors that they are 'strays' to. Figure 6N shows the general idea, but really needs an animated sequence to better show the sequence of events.

Eventually the 0V-chassis connection (if there is one) will carry currents that equalise the static voltages throughout the interior of the product. If there is no intentional connection, equalisation will happen more slowly due to ionisation of the air in the product. The transient currents in the stray capacitances are different for each device pin, PCB trace or other conductor, and so they cause transient voltage differences between different parts of the circuits. These transient differential-mode voltages can upset the operation of circuits, and reset or crashed microprocessors are a common consequence.

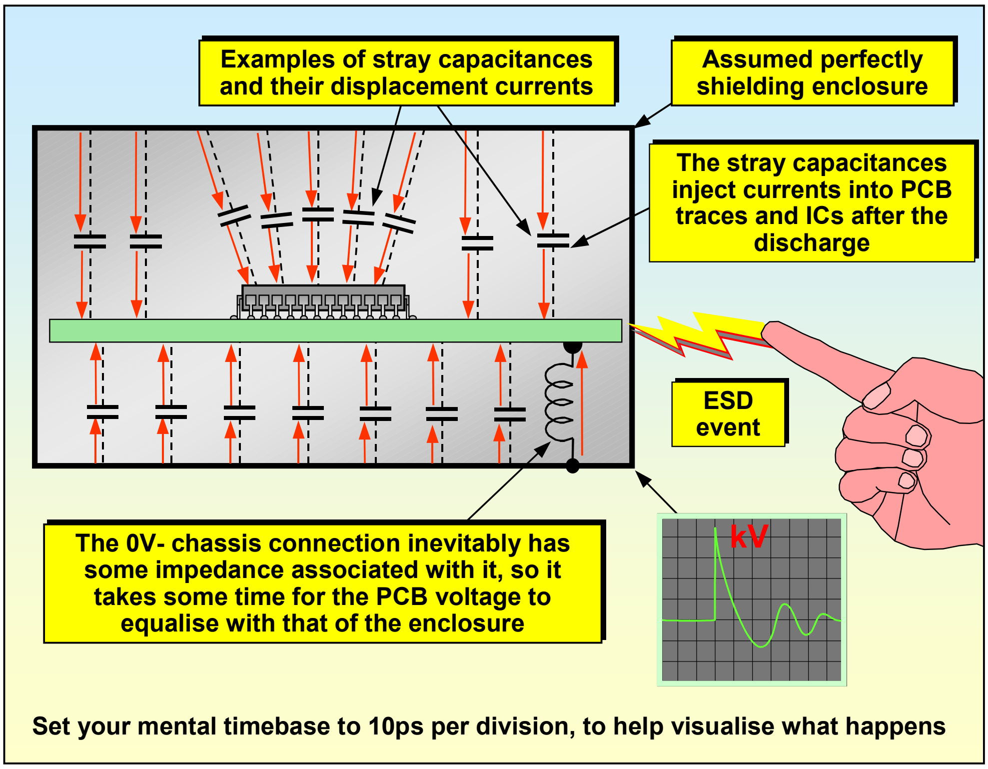


Figure 6N Effects of a discharge to a shielded enclosure

Note that a wired connection between the enclosure shield and a safety earth or other external 'ground' has no effect over the process described above. The length (and hence inductance) of the earthing/grounding wire or strap is simply too great for it to carry the charge away from the outer surface of the shield before it has time to diffuse inside. However, direct metal-to-metal bonding at multiple points around the perimeter of a metal enclosure, to a large metal surface (e.g. the metal hull of a ship or metal fuselage of an aircraft) should allow the surface charge to 'drain away' fast enough to have *some* effect.

One solution to this ESD problem is to create a high-quality 0V plane on the PCB, as described briefly in [17] and in detail in Chapter 4 of [18]. Then 'RF bond' this plane to the shield with multiple low-impedance (at 1GHz) bonds – described briefly in [17] and in detail in Chapter 3 of [18].

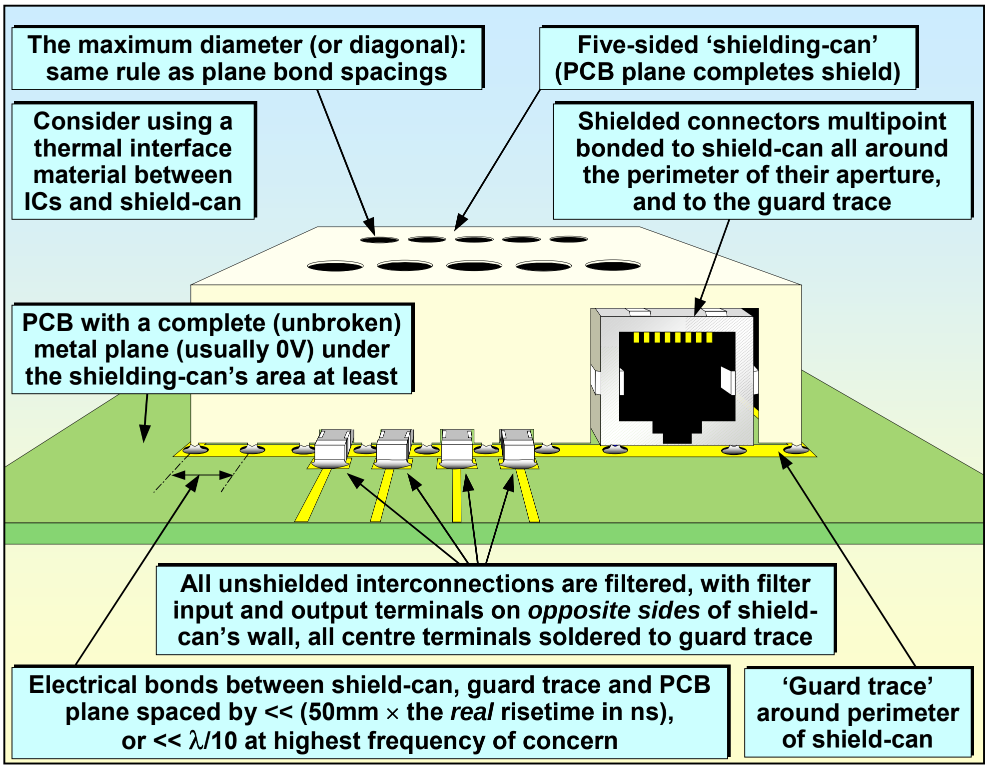


Figure 6P Overview of PCB-level filtering and shielding techniques

Another solution is to use filtering and shielding techniques on the PCB, at least over the most sensitive components. These techniques are described briefly in section 5.3 of [17] and in detail in Chapter 2 of [18], from which Figure 6P is taken, and PCB shielding-cans can be quite low-cost. It may be necessary to apply the 0V planes and RF bonding at the same time as the PCB-level filtering and shielding.

6.1.5 Protecting signal, data, control or power conductors

Discharges into semiconductors can be fatal for them, so if it is not possible to protect conductors with the methods described in 6.1.2 - 6.1.4 above, and if we really have no choice but to expose conductors to ESD discharges, for example the antennas on portable radio receivers, we need to apply appropriate suppression or filtering techniques to them.

This is especially a problem for the pins of connectors, which are exposed to ESD discharges from:

- Personnel discharges (e.g. people fingers)
- Plugging in other equipment (equipment with two-core mains leads are not earthed, so could be charged up to kV)
- Charged-up cables (dragging a cable over the floor can cause all of its conductors to take on an electrostatic charge at several kV)
- Discharges from a variety of other sources

A solution for passing tests to IEC/EN 61000-4-2, which attempts to simulate personnel ESD, is to use small metal-shrouded connectors with their shroud directly connected to the product's RF Reference (its shielding, or PCB 0V plane if unshielded). Appropriate connectors include D-types, USB, Firewire, RJ45, etc. If the 8mm diameter 'air discharge' tip was used for such tests it would most likely discharge to the metal shroud, sparing the connector pins. But in any case there is a clause in IEC/EN 61000-4-2 that mandates using the pointed 'contact discharge' tip, and only applying it to the metal shrouds of such connectors.

Although this is an appropriate technique for personnel discharge, it doesn't deal with the remaining three bullet points above. It allows a product to pass the ESD tests as part of declaring compliance with the EMC Directive, but it doesn't necessarily mean that the product is protected against all the ESD events it will experience in real life.

(Some manufacturers place all their connectors on the rear of their product, so they can state that they are not "accessible to persons during normal use" to take advantage of a clause in IEC/EN 61000-4-2 that removes the requirement to do *any ESD tests at all* on those connectors. I'm sure I don't need to say why I don't recommend that approach!)

For signal conductors that could be exposed to any types of ESD discharges, current-limiting, transient suppression, or filtering techniques will almost always be needed to protect their circuits from upset and damage to their devices. Some DC power conductors may also need similar protection, although if they are well decoupled (see [17] and Chapter 5 of [18]) this should be sufficient.

As far as I am aware, almost all ICs are fitted with ESD protection diodes that shunt overvoltages and undervoltages to their DC power rails, and those that are not have bold warnings of this fact on their data sheets. But because of the commercial pressures to make devices cheaper, hence use smaller silicon die, these diodes have never been very large or powerful and they are becoming progressively smaller and less powerful. We can help these diodes do their job by putting impedances in series with the conductors that are suffering from the ESD event (e.g. connector pins), as shown in Figure 6Q.

As Figure 6Q shows, the resistors or soft-ferrite chokes used must be rated to withstand the full ESD voltage. If ordinary 0805 or similar types are used, *at least* ten (possibly twenty) will be required in series, otherwise their terminals will spark-over – defeating their purpose of limiting discharge currents. When many resistors are used in series, they must not be placed close to each other on the PCB, otherwise the ESD might flash-over between different resistors or track across the inevitable surface contamination on the PCB.

The values of the resistors or chokes are chosen to limit the worst-case discharge current to one that the IC's own protection diodes can handle, the data for which should be provided on the data sheet. All such designs should be proven by assiduous testing, not just a few discharges. And some ESD testing should also be done in the near-dark, to reveal any 'sneak' discharges on the PCB. Where contamination by dust or condensation is likely, test in the dark with foreseeable contamination simulated.

Unfortunately, the values of impedance required may be so high that high-data-rate signals suffer from degraded signal quality (e.g. collapsed eye-pattern). Also, many types of individual semiconductors are unprotected against overvoltages and some are especially susceptible, so just adding series impedance isn't going to work for them.

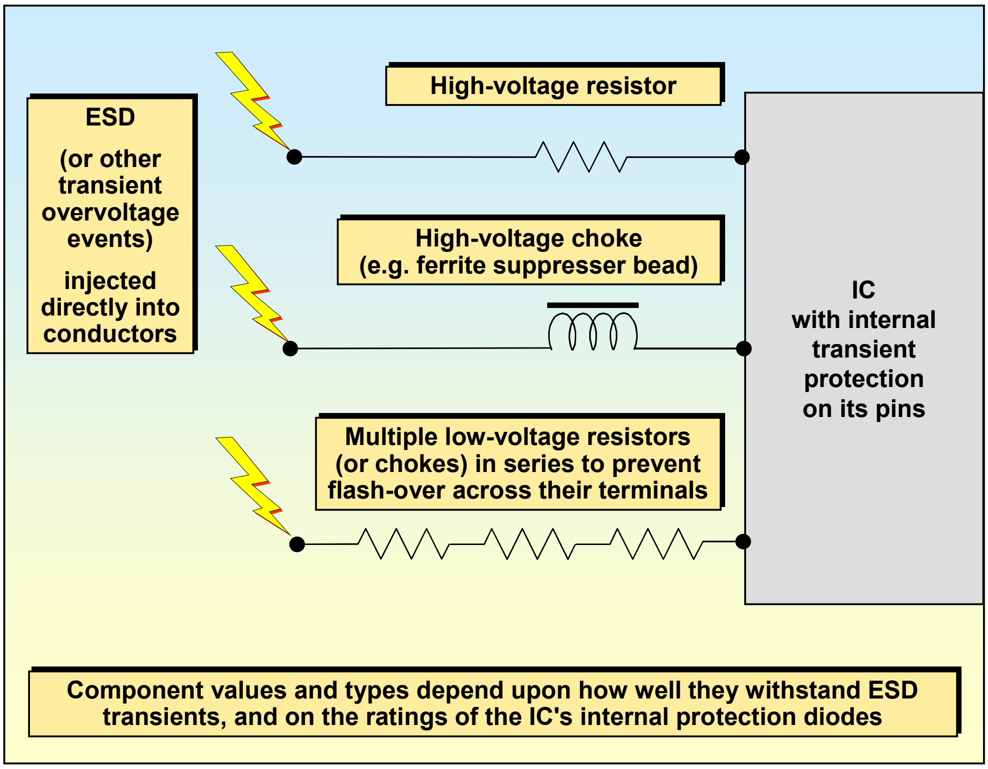


Figure 6Q Adding impedance in signal lines to limit discharge current

Filtering or suppression techniques must provide at least 40dB of attenuation (e.g. reduce 8kV to 80V) and possibly as much as 70dB (e.g. from 24kV to 8V) for transients with risetimes of 0.7ns (ideally 0.2ns) – equivalent to a frequency of 460MHz (ideally 1.6GHz). They will not be able to achieve this very high performance without an RF Reference that provides a very low impedance up to the highest frequency of concern. Suitable RF References will either be a metal plane in the PCB (see [17] and Chapter 4 of [18]) or the wall of a shielded enclosure that has a good SE at the highest frequency of concern (see [16]). Figure 6R shows two general techniques: a TVS device, and transient-rated diodes.

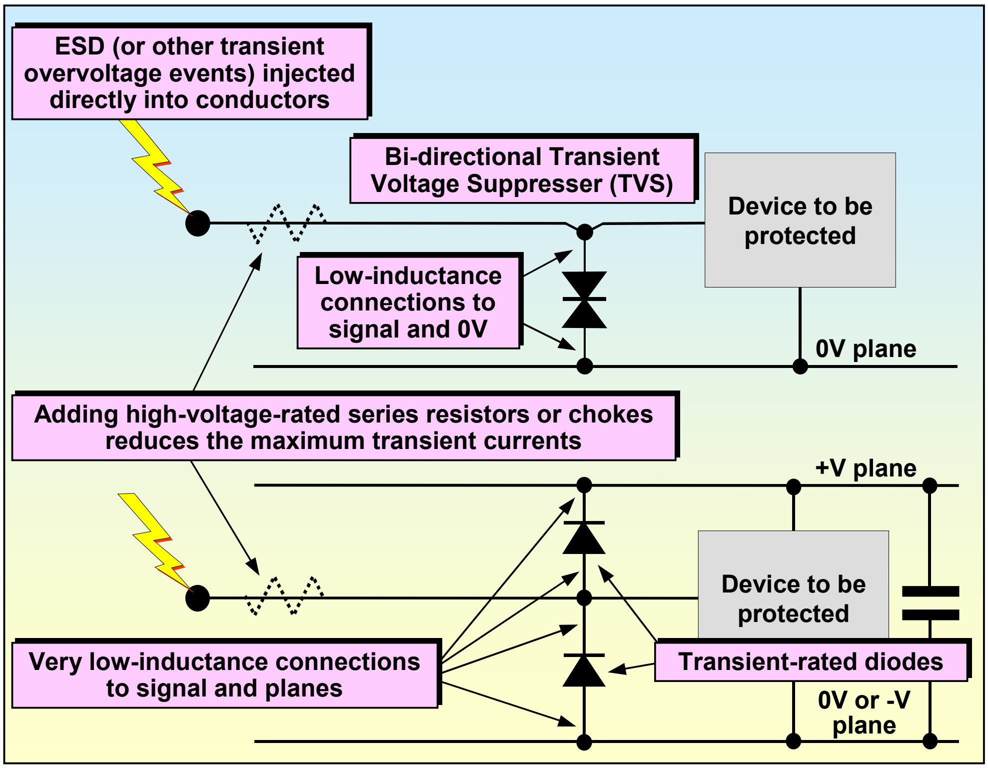


Figure 6R Adding transient voltage protection

The basic principles of transient overvoltage protection were covered in the final sections of [15], for low-frequency surges. TVS devices are generally avalanche diodes, which seem to be increasingly referred to as

SADs (silicon avalanche diodes) – the fastest-operating type of transient protection device. Surface-mounted metal-oxide-varistors (MOVs, sometimes called VDRs) should be fast enough to suppress ESD with 0.7ns risetimes, but might not be quick enough where risetimes of 0.5ns or less could occur.

Transient-rated diodes usually come in pairs in SOT-23 packages or similar, and are specified exclusively for transient suppression applications, not for use as ordinary diodes. Sometimes the level of ESD exposure makes it necessary to use high-voltage-rated impedances in series with TVSs or transient diodes, in which case all the high-voltage issues discussed earlier apply to the resistors or chokes.

Many TVS devices have too large a self-capacitance for high-frequency or high-data-rate signals, although low-capacitance versions, some as low as 1pF, are becoming increasingly available – spurred by the rapid increase in high-speed interconnections such as USB2 and Firewire. Transient diodes are reverse-biased in normal operation and so have a low capacitance, which makes them suitable for high-speed signals, as long as the discharge currents are not too high. When using transient diodes, the DC power rail they connect to should be a plane with very low impedance at the highest frequency of concern, just like the RF Reference plane.

A problem with reverse-biased transient diodes is that their leakage currents double with every 10°C rise in temperature, making them difficult to use in high-temperature applications, or on sensitive high-impedance DC-coupled circuits. Many other exotic solutions are possible, for example using an isolated-gate FET (IGFET) arranged so that an incoming overvoltage turns it on and temporarily shorts the trace to the RF Reference.

The placement of the components on a PCB, and the routing of their traces, is vital if the required transient attenuation is to be achieved, and Figure 6S sketches the details, for a TVS. The same layout rules apply to transient-rated diodes as well.

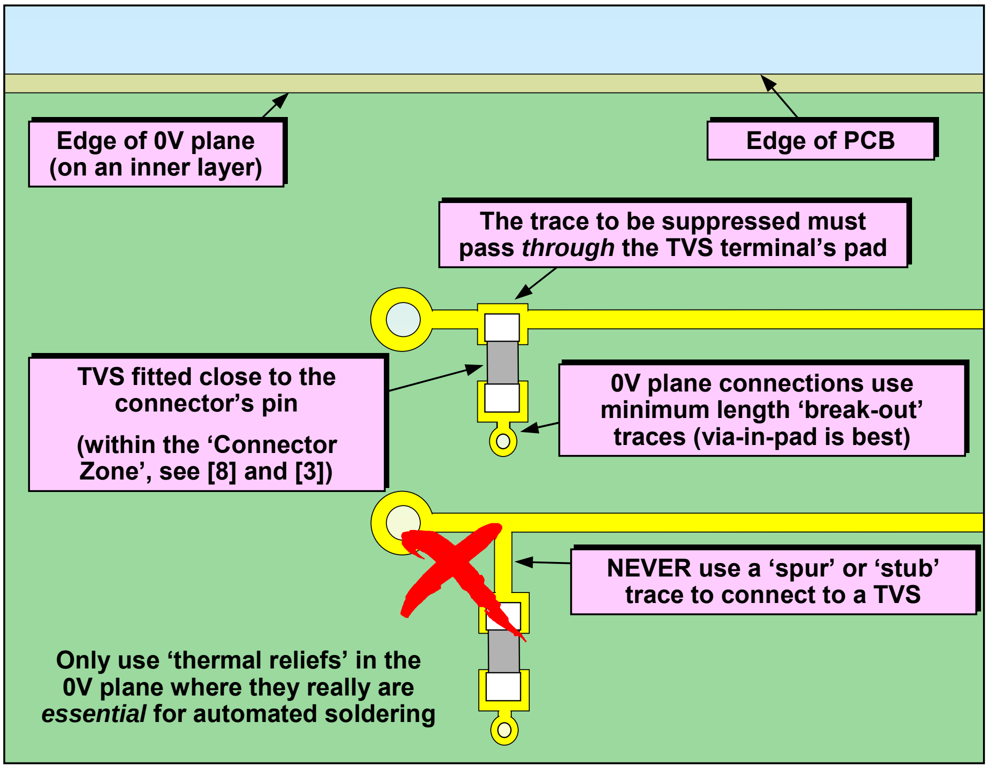


Figure 6S PCB layout issues for transient voltage protection

An IEC/EN 61000-4-2 ESD test at 8kV can generate a dI/dt in excess of 43A/ns. A 1mm wide trace just 1mm long, routed over an RF Reference plane layer, will have a self-inductance somewhere between 0.3 and 0.6nH (depending on its height above the plane) [18]. At 43A/ns the peak voltage drop along the 1mm trace will therefore be between 13 and 26V. So the short trace shown connecting the TVS to the via hole in Figure 6S could limit the efficacy of the transient suppression.

The via hole to the reference plane shown on Figure 6S also has self-inductance. On a two-layer 1.6mm thick board the length of the via hole carrying TVS current will be 1.6mm, giving it a self-inductance of 1.6nH. 43A/ns in such a via hole would drop 70V peak. The TVS device itself will also have internal series resistance and self-inductance, which will also add more peak volts.

The faster risetimes or higher voltages possible with some types of real-world ESD events could double or even triple the above estimates. Clearly it is possible for the PCB layout itself to degrade the performance of the TVS or transient diodes by so much that even if they had zero clamping voltage (which of course they do not) the

ICs would still be exposed to quite high peak voltages, just from the self inductances of very short traces, via holes and the transient suppression devices themselves.

However, as long as these voltages are not *too* high, the internal transient protection devices in the ICs themselves should cope with them.

The very best suppression devices are three-terminal types, like the three terminal filter components discussed in [15]. To get the best performance from them, they should be used with at least two parallel vias to their Reference plane, arranged symmetrically around the device and very close to it. Also, the PCB dielectric between the Reference plane layer and the layer on which the suppression device is mounted should be as thin as is practical, say 0.15mm or less. Such precautions are not yet generally necessary, but perhaps they will become more common in future as devices explore silicon processes at 45nm and smaller.

[15] covered the basic principles of filtering, and the above descriptions of the issues associated with ESD suppression using a TVS also apply to the high-voltage-rated series resistors (or chokes) and shunt capacitors when using filtering instead. Where the signals are very slow, some manufacturers just use a large capacitor on its own, with no series impedance, to act as a capacitive voltage divider with the capacitance of the ESD gun, by charge redistribution.

For example, if the ESD gun had a 150pF capacitor charged to 15kV, using a 150pF shunt capacitor to protect an IC's pin would result in 7.5kV at the IC's pin, 1.5nF would give about 1.5kV, 15nF would give 150V and 150nF would give 15V. These are idealised calculations – as shown above the self inductances of even short PCB traces and via holes could easily add tens of peak volts to these values, and there is also the issue of the behaviour of the capacitor with such transient charging currents.

Ceramic capacitors are the only suitable types, with COG or NPO being the best. A typical surface-mounted capacitor might have an internal series resistance of 10mΩ and a series inductance of 1nH, generating an additional peak voltage of 43V with a current risetime of 43A/ns.

The voltage ratings for any series resistors or chokes are the peak ESD voltage itself. Because kV can leap large distances through air or vacuum, their location on the PCB and proximity of them and their traces to other devices and their traces is very important. The voltage rating for the capacitors in any voltage dividers or filters is set by charge redistribution. The value of capacitor or 'clamping voltage' of a TVS should, of course, be less than the level that damages the IC it protects, taking into account the additional transient voltages caused by the self-inductances of shunt components, traces and via holes, and itself. The TVS's capacitance is set by the circuit impedance and data rate; and its peak current rating by the magnitude of ESD event (taking into account any current limiting by high-voltage-rated series resistors or chokes).

6.1.6 'Earth lift' problems for interconnected items of equipment

The above discussions have only considered the ESD protection of a single product, on its own, but when two or more products are interconnected (e.g. by power, signal, control or data cables), 'earth lift' adds a new type of ESD problem.

As Figure 6T shows, when an ESD event injects current into a chassis or enclosure (either directly or via a shunt suppresser like a TVS, transient diode or capacitor), the chassis (etc.) suffers an 'earth lift' transient as the discharge current flows in the very high inductance of the earth-bonding network.

As mentioned before, self-inductance of ordinary conductors is so high that earthing using wires or even braid straps has little/no effect on the peak transient voltage attained by the chassis of the product suffering the discharge. In fact, the peak voltage attained will be almost the same as it would be for an unearthed product, for example one that was battery powered, or 'double-insulated' from the mains power and so powered by a two-core mains lead with no safety earthing conductor.

The peak transient voltage of a product can be determined by charge redistribution between the ESD source's capacitance and the space-charge capacitance of the product. We can calculate the capacitance of the product very approximately as:

$$C = 4\cdot\pi\cdot 8.85\left(\frac{1}{a} - \frac{1}{b}\right) \quad \text{pF}$$

where: a = the radius of the sphere representing the product, and…
b = the distance of the product from the nearest floor or wall or ceiling that is either made of masonry or has substantial metal in it (e.g. a suspended ceiling)

For the value of a I suggest using half of the average of the two longest product dimensions, e.g. width and length, solely on the basis that it feels about right. Obviously, we do not expect to get a very accurate assessment using the above formula, and this carries across into the accuracy we can expect of the peak transient voltage we would calculate by charge redistribution. Accurate calculations of peak transient voltage can be achieved using modern computer simulation techniques, which can also determine the capacitance of the ESD source.

The earth-lift voltage is common to all of the conductors in an interconnecting cable, so it is a common-mode (CM) ESD transient, which just means that it can damage a number of input and output devices simultaneously.

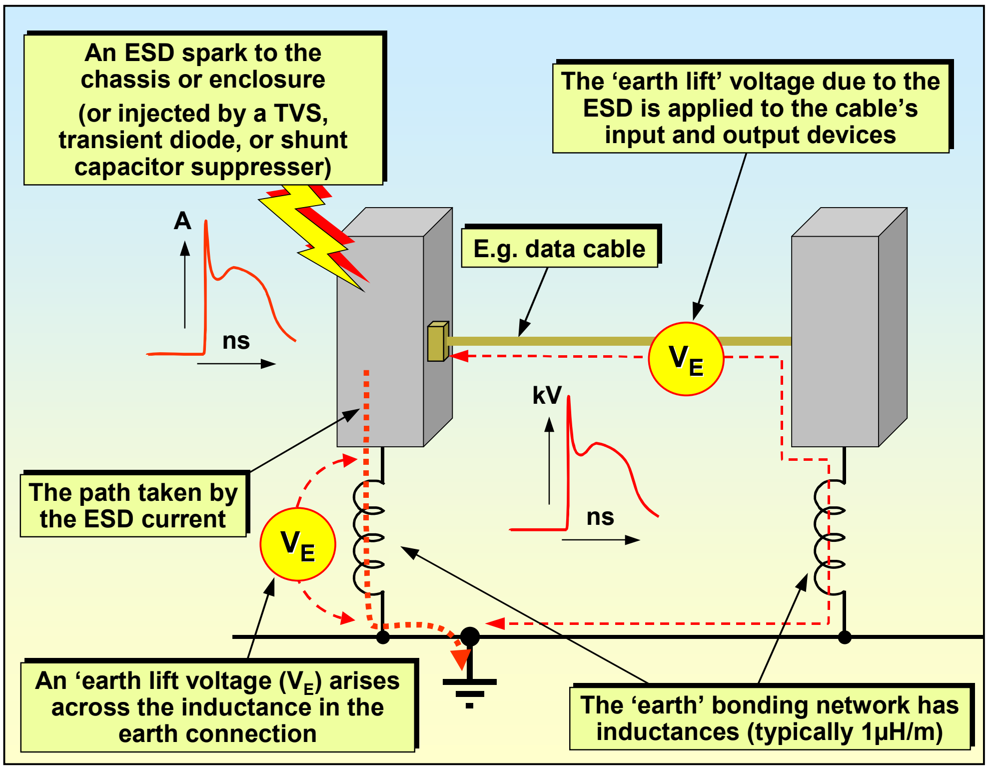


Figure 6T `Earth lift' problems for interconnected items of equipment

For protection, either prevent the ESD from happening in the first place using the techniques described in 6.1.2 or 6.1.3, or reduce the impedance of the products' local earthing network to negligible amounts by direct metal-to-metal bonding each of the interconnected products to the same sheet of metal, at multiple points around their perimeters. Such brute-force 'earth bonding' can be quite straightforward in a ship, aircraft, offshore oil platform or other structure made solely from large metal sheets.

Also, using well-shielded cables and connectors to interconnect the products, each cable shield 360° bonded to the frame/chassis/shield etc. of the product at *both* ends as described in [14], will reduce the earth-lift voltage; although I am not sure whether this method on its own always guarantees freedom from earth-lift problems.

If all of the above methods are impractical, or inadequate, we are left with applying circuit techniques to the input and output devices: either galvanic isolation or suppression with TVSs, transient diodes, filters or just capacitors, as discussed in 6.1.5.

Since earth-lift is a CM phenomenon, a CM choke may be just as effective, if not more so, than individual chokes in series with each of the conductors in the interconnecting cables. PCB-mounting CM choke components are available, but are probably not rated to withstand ESD voltages.

Figure 6U shows a few of the very wide variety of cable-mounted CM chokes that are available for round and flat cables. Whilst these do not generally offer as high values of impedance as the board-mounted types, they have very good high-voltage performance, limited only by the insulation of the cable they are used on.

Galvanic isolation is by far the most robust technique, and can use transformers (e.g. pulse transformers in Ethernet, microphone transformers in professional audio), optical isolators, fibre-optics, wireless, infra-red, free-space modulated lasers, and other techniques. But the vast majority of transformers and optical isolators are not rated for kV isolation and will spark-over when subjected to ESD test voltages.

Transformers can be designed and made with appropriate ratings, usually to special order. 10kV-rated optoisolators have been available from some suppliers for many years, essentially just an emitter and receiver spaced 20-30mm apart by a light guide. It is difficult to obtain such devices that will also handle high-rate digital data, for which fibre-optics and free-space lasers will generally be required. Fibre-optics are generally preferred for EMC reasons anyway, see [14].

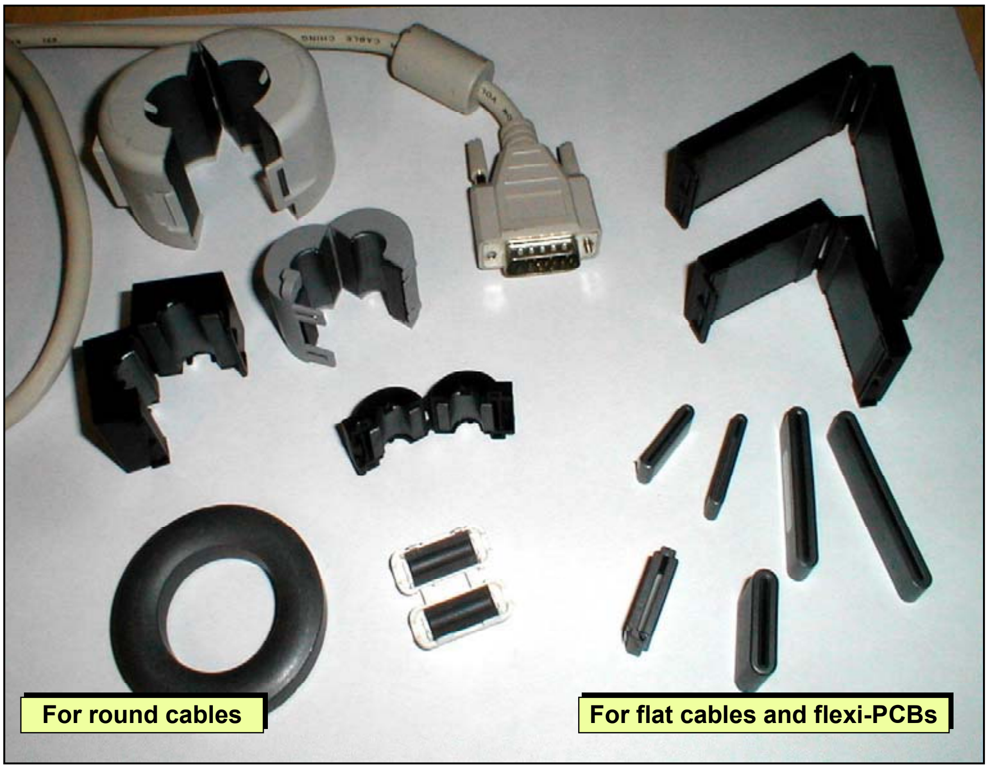


Figure 6U Examples of cable-mounted CM chokes

6.1.7 Protecting data and signals from errors

Transient suppression devices such as TVSs, transient diodes and shunt capacitors only prevent actual damage to devices, they don't prevent signal corruption. But after a typical ESD test, if the operating state of the product has altered, or any data has been lost, the result is a failure. So it is not sufficient to simply prevent device damage, we have to maintain signal integrity too.

Hardware and/or software design is generally needed to discriminate between ESD events and valid signals. Keyboard strokes, button presses and slow signals, control or data are all easy to distinguish from ESD events using simple techniques, because the ESD events are so brief.

For example, a very quick 'jab' at a momentary contact switch might last as little as 25ms, which is at least 10,000 times longer than almost any ESD event, including the decay of any product resonances it excites. So a simple resistor-capacitor low-pass filter, or a couple of lines of keyboard polling software that checks whether the data is still valid after a few ms, is often perfectly adequate.

But high-speed data uses signals with risetimes and/or durations that might not be so very different from ESD events, making simple discrimination schemes unreliable. High-speed analogue signals should use high-quality shielding (see [14]) and digital data can too. Alternatively, convert all signals into digital data and employ error-detecting or error-correcting protocols.

Any digital engineer can design error detecting/correcting communications protocols, but the temptation to do so should be *resisted at all costs*! It is not at all easy to get a robust product unless you are an expert *in this type of design*. As Figure 6F shows, even a single very short ESD event can cause surprising EM phenomena whose amplitudes, frequencies and durations are hard to predict, and some ESD events are neither single nor short.

But ESD is not the only type of transient that a data communication link needs to be protected from. Fast transient bursts have hardly been mentioned in this series, because they are generally dealt with quite adequately by techniques already described (making allowances for their frequency range and amplitude), but our error detecting or correcting protocol needs to cope with these long bursts of noise too. In real life, fast transient bursts can last for several hundred ms, sometimes for several seconds, especially in high-power industries or near high-voltage distribution switchyards.

We can easily purchase ICs and/or software that have enjoyed the benefit of experts with aggregate experiences of hundreds of man-years *solely* in protecting data in communications links. So we should always buy these, as they will be much more cost-effective than anything we might think we can do ourselves, no matter how clever we are.

Ethernet and CAN bus are but two examples of robust datacommunications, but they are not perfect – in extreme EM environments the data rate of Ethernet can drop to zero, and so can the CAN bus, due to a small oversight in the CAN bus standard [19]. More sophisticated protocols exist, one highly respected example being that used by the real-time MIL-STD-1553 bus, of which commercialised versions are now available.

6.1.8 Use all the other EM design techniques too...

The EM engineering techniques described in the earlier parts of this series [13] [14] [15] [16] [17], as well as those in [18], control E, H and EM-fields and so can be used to improve immunity to E- and H-fields from ESD events. Sometimes the techniques were described with examples of reducing emissions, and sometimes of improving immunity, but any technique that attenuates fields and/or conducted noise is equally effective for either purpose.

The fields from ESD events within a few metres can be very strong, making it necessary to take more care over the EM design, going into finer detail (e.g. using $\lambda/100$ gaps in seams instead of $\lambda/10$). Other sources of advice on good ESD design include [4] and [5].

6.1.9 Software techniques

Software is easily corrupted by transient voltages due to ESD, leading to a variety of possible errors, malfunctions and crashes. Where the hardware techniques in this series do not provide sufficient immunity to transient or short-term events such as ESD, or are too impractical or too costly, appropriate software programming techniques can be a huge help – and of course they generally add no cost in manufacture.

This series has described hardware techniques only, because this is where my experience and skills lie. I dare not write about software, because my ignorance in that area would soon be revealed, so instead I refer the reader to people who *do* understand software techniques for EMC, especially [20] [21] [22] [23] [24] [25], section 12.2.5 of [4], and Chapter 37 of [5].

Of course, software techniques cannot work if the devices the software runs on are damaged from ESD or other EM disturbances (e.g. surges). However, the use of multiple redundant processor ‘channels’ with voting and other operations on their independent outputs can be used to detect faulty digital processors, whether the errors are transient or permanent due to damage.

But it important to note that redundant hardware channels are often all exposed to the same EM disturbance in (almost) the same way at (almost) the same time, for example an E or H-field from a nearby ESD or lightning ground stroke, or an overvoltage surge on their common mains power supply.

So if all of the channels use the same technology and construction, they can all fail in the same way at the same time. This is a bad thing and is known as a ‘common-cause’ failure. It is best dealt with by using:

- Diversity of design (e.g. different types of microprocessor, different software languages, different PCB layouts, different designs of power converters, etc.), plus…
- Diversity of location and cable routing (e.g. not placing all the channels in the same cabinet, not routing all the cable sin the same trunking, etc.), plus…
- Diversity of power supply (e.g. more than one independent mains supply, battery backup, etc.).

For very high-reliability systems, such as those that control weapons, financial institutions, national security and safety-critical applications such as fly-by-wire passenger aircraft, a great deal of care needs to be taken with ensuring diversity of design. It can even require the different software programmes for the diverse channels to be written to different requirement specifications produced by different teams of people who have never shared the same university courses or employers.

6.2 Electromechanical devices and spark ignition

6.2.1 Important safety note

Electromechanical devices are often used on hazardous voltages (e.g. mains power), and of course spark ignition generates high voltages. When suppressing their emissions, it is very important that all safety issues (shock, fire, explosion, etc.) are fully taken into account.

This is usually achieved by fully applying the relevant safety standards (e.g. IEC/EN/UL 60950-1, IEC/EN/UL 61010-1, IEC/EN/UL 60601-1, etc.) to the suppression components, plus ensuring that the voltage, current, temperature and power ratings of the conductors and components are more than adequate for their worst-case operation (e.g. high ambient temperatures plus highest voltage, current or dissipation) taking into account their manufacturing tolerances, the effects of ageing, and reasonably foreseeable misuse (e.g. operation with covers removed).

6.2.2 Introduction to arcs and sparks

When a contact opens the current cannot change instantly because of the inductance in the circuit. Inductive energy is stored in any conductor (e.g. mains supply distribution and long DC cables), and is also stored in all coils and windings (e.g. solenoids, transformer windings, motor windings, etc.). As the stored magnetic field energy collapses, it creates what is known as a 'flyback' voltage, which can reach kV in a few microseconds and break down the contact's air gap, creating a brief arc (or spark).

The flyback phenomenon is used in motor vehicles to generate the high-voltage pulses for spark ignition petrol engines, but it is a nuisance in electromechanical switches – the arcs have a temperature higher than the surface of the sun, and so over time they corrode and erode the contact area, increasing unreliability, plus of course they generate bursts of emissions.

The arc or spark has a low impedance and so collapses the voltage and permits current to flow in the inductor once again so the field stops collapsing. The arc or spark then dies, the field starts to collapse again, and another flyback event terminating in an arc or spark occurs. Figure 6V shows this repetitive process, which continues until the stored magnetic field energy has decayed by so much that it cannot break down the air gap any more.

Real electromechanical contacts take a finite time to open to their full extent (1mm in figure 6V), so the flyback voltage does not have to reach very high voltages to break down the air gap when it is just a few tens of microns, and the rate of occurrence of the flyback arcs and sparks at this time can be MHz. As the contact spacing increases, the flyback voltage increases and the spark rate goes down, maybe to kHz for sparks of 1kV or more, depending on the stray capacitances that exist in the circuits.

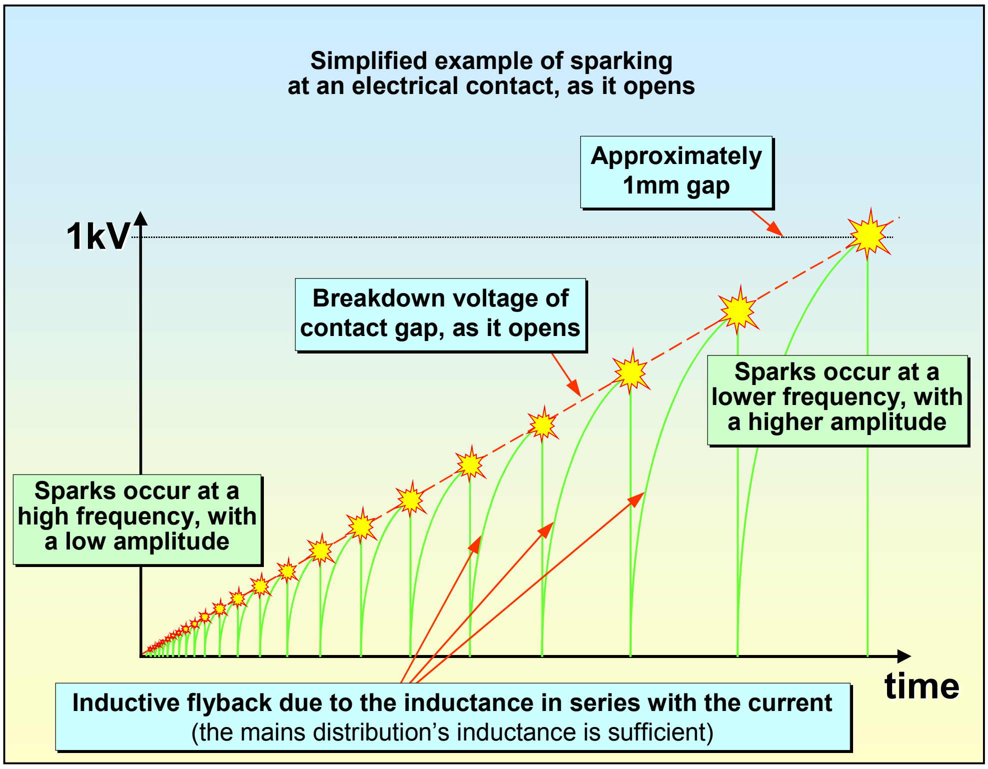


Figure 6V Simplified example of sparking at an electrical contact, as it opens

The frequency spectrum of each arc or spark is truly 'DC – daylight', with a fairly flat spectrum. The visible flash we see from an arc or spark is just a small part of its very wideband spectrum. The size of the arc or spark itself is so small that it is not an efficient radiating antenna at frequencies below a few tens of GHz.

Instead, it is the conductors and components that carry the flyback currents that cause the emissions, by conducting the RF noise into other circuits (e.g. via the power cables connected to the contacts); by induction into nearby conductors and components (e.g. via stray mutual inductances and capacitances, a near-field effect sometimes called crosstalk), and also by radiation of electromagnetic fields (a far-field effect).

All these coupling mechanisms are strongly affected by the resonant frequencies of the conductors and components connected to them, so in practice the significant conducted, induced or radiated emissions appear on a spectrum plot as bursts of noise centred on just a few frequencies, instead of the flat, very wideband spectrum of the arc or spark itself.

A common problem with arcs and sparks is that they will interfere with microprocessors and other devices within the same product. It is not at all unusual to find that the operation of an unsuppressed relay contact, on or near a PCB carrying a microprocessor, interferes with it.

A very pernicious problem is the borderline case where during most of the software's cycle the digital noise immunity is sufficient to reject the noise from the sparking contacts, but there are one or more software states, each lasting for possibly just a few microseconds, where the immunity is inadequate. The problem arises when one of these susceptible software states occurs just at the same time as a burst of noise from a sparking contact. Tracking down the cause of the resulting 'random' failures can be a very long and costly task, because it is so hard to repeat the exact situation in a controlled situation.

The effects of arcs and sparks are known as fast transient bursts, and IEC/EN 61000-4-4 is the basic test method called-up by the standards listed under the EMC Directive for testing immunity to this phenomenon. More information on fast transient bursts, where they come from and how they can cause interference, can be found in the guide on IEC/EN 61000-4-4 available from [7].

6.2.3 A problem with emissions and immunity standards

EMC standards are created by committees with strong representation from manufacturers' trade associations. Decades ago, based on human perception of radio and TV interference, the interference created by momentary arcs and sparks was deemed to be acceptable in residential environments – providing each burst of interference only occurs for a brief time (e.g. less than 10ms) and at a low enough rate (e.g. less than 5 times per minute). So the emissions limits for residential environments set no limits at all on such transient emissions.

Since radio and TV receivers were not features of the industrial environment, the emissions standards for industrial environments often had no special requirements for transient emissions at all. If arcs and sparks occurred in normal operation, they were measured using the normal techniques for continuous emissions, which use integrating detectors and so very significantly understate the amplitudes of such transients.

But people in developed countries now live in what they are pleased to call the modern world, and in this world very many domestic appliances, most consumer electronics, and almost all industrial equipment are controlled by microprocessors. Also, modern residential, commercial and industrial environments contain desktop, laptop, notebook, palmtop and other types of personal computers or personal digital assistants.

These now ubiquitous digital and computer technologies can easily be upset by a single transient EMC event lasting just a couple of nanoseconds, leading to a wide range of unpredictable malfunctions.

It can be argued quite strongly (and has been) that <u>all</u> of the current emissions standards listed under the EMC Directive (such as EN 55014-1, EN 55022, EN 55013, etc.) and similar national standards are unsuitable for the modern world, but the standards committees have somehow managed to ignore this important issue for the last 20 years at least.

So, when one is trying to do EMC engineering properly and cost-effectively, taking into account the possibilities for financial and other losses and safety issues relating to possible interference (see Part 0 of this series [13]), I recommend that all transient emissions due to arcs and sparks are adequately suppressed even where the emissions standards do not require it – if only to prevent them from interfering with other parts of the same product.

The basic immunity test standard for 'fast transient bursts', IEC 61000-4-4 attempts to simulate the electrical noise created by electromechanical contacts, but uses a fixed spark rate of either 5kHz or 100kHz, whereas as Figure 6V shows, real-life contacts create arcs and sparks at a rate that continuously varies from MHz to kHz, so passing tests to 61000-4-4 does not necessarily mean that products will be reliable enough in real life operation.

6.2.4 Suppressing arcs and sparks

It is always best to avoid the creation of arcs and sparks altogether, by using semiconductors that do not switch current instantaneously, such as power FETs, IGBTs, zero-crossing triacs, etc. Although this adds cost, it also increases reliability (providing the semiconductors are protected from overvoltages and overcurrents) and of course it completely eliminates the transient emissions.

Where it is decided to employ electromechanical contacts, for example in switches, relays, contactors, thermostats, etc., to comply with existing emissions standards for residential and commercial environments it is generally necessary to use snap-acting 'microswitch' types of contacts, and to switch less often than 5 times per minute.

Alternatively (or as well) the arcs and sparks can be usefully suppressed by what are called 'snubbers'. Figure 6W shows a traditional series-connected snubber, which bridges across the contacts. The capacitance has a low impedance at high frequencies, and so the snubber conducts the flyback current, dissipating its energy in the external circuits and loads. If designed correctly, the flyback voltages developed across the contacts and their snubber will be so low that arcs and sparks will not occur.

However, the flyback current still generates conducted, induced and radiated noises, although not with the huge wideband spectrum associated with arcs and sparks. The snubber should therefore be designed (usually by trial and error) to balance its cost against its EMC benefits.

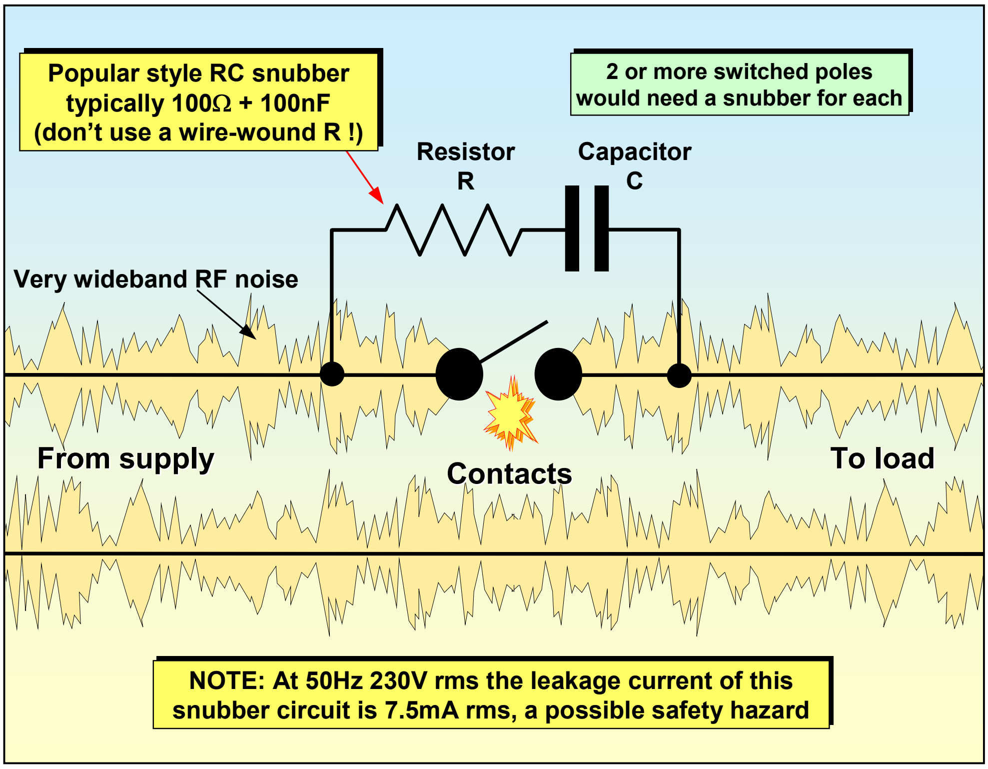


Figure 6W Typical series snubber

Snubber resistors should be low-inductance types, and snubber capacitors should have good performance over the frequency range to be suppressed. If in any doubt about the suitability of certain types of components for RF suppression, test them with signal generators and oscilloscopes (ideally with a Network Analyser) at frequencies of 1MHz or more. Sometimes it may be necessary to connect several low-value capacitors in parallel, to achieve the value of capacitance required with adequate high-frequency performance.

Snubbers must always be mounted in intimate proximity to the contacts they are suppressing. Any conductors (wires, cables, PCB traces, etc.) associated with snubbers should be very short indeed, so as to add very little inductance to the snubber. If these practical issues of component choice and assembly are not correctly dealt with, adding a snubber could just 're-tune' the emissions frequencies without reducing their amplitudes.

A significant problem with the series snubber technique in Figure 6W is that when the power or signal being switched is AC, or DC with a significant AC noise content, the snubber allows some current to 'leak' through, bypassing the contacts. I first met this problem when I put a series snubber on a set of relay contacts that were connected to a powerful alarm bell. When the relay contacts were open the bell continued to ring, although not as loud as when the contacts were closed.

Maintenance personnel and users who ignore the instructions to '*Isolate the mains externally to the equipment before removing the cover*' often assume that an open contact means the 'downstream' circuit is not hazardous. But with the snubber values in Figure 6W and 230Vrms 50Hz mains, the leakage current is about 7.5mA, not enough current to kill a person directly, but more than enough to cause uncontrolled muscular responses (i.e. violent twitching) that could lead to an injury (e.g. making someone fall off a ladder).

So I prefer to use 'low-leakage' snubbers, using the design of Figure 6X. In fact, I have generally found that for the contacts I have suppressed, this design is more effective at reducing emissions than the series snubbers of Figure 6W.

Although most snubber design is done by trial and error whilst EMC testing, there are some crude guidelines for choosing the component values, at least to have a starting point for the trial-and-error process: R (in Ω) = V / I_L , C (in μF) = $k(I_L)$, where I_L is the rms load current in amps, and V is the rms supply voltage in volts. The value of k is set by guesswork, usually it is set to 1, but it is not unusual for its final value to be between 0.1 and 2.

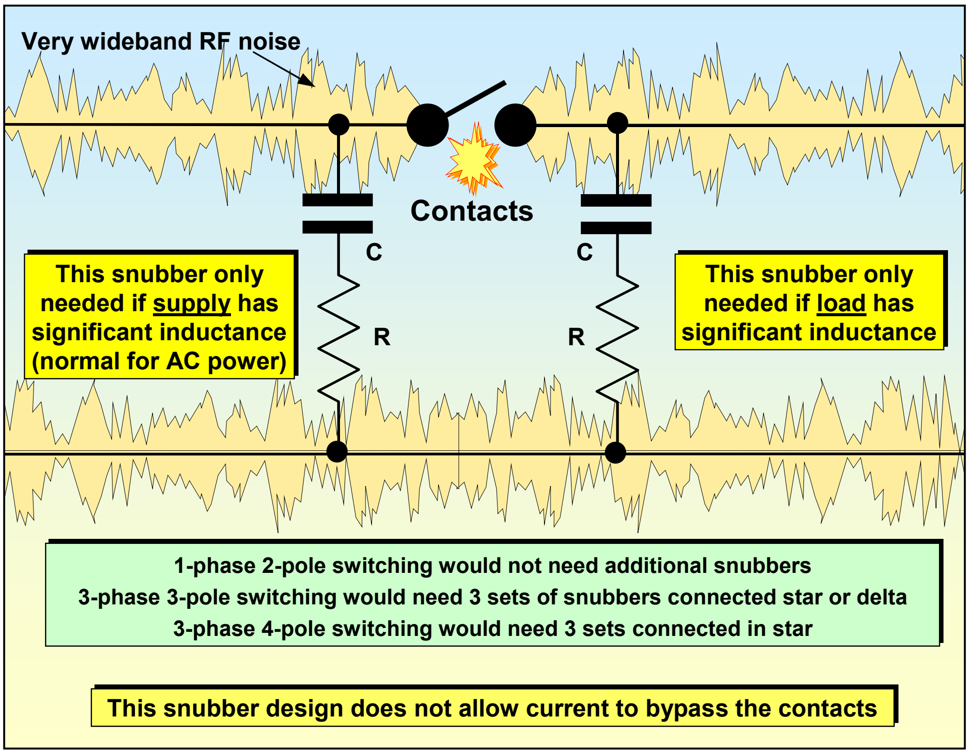


Figure 6X Low-leakage snubbers are safer, and often more effective

Where the RC time constant of the snubber is more than 50ms, the type of resistor used in the snubber could possibly be an inductive type. In some applications, snubbers can be subjected to severe voltage, current and/or power stresses and need to use quite large high-power components.

The above discussion has centred on RC snubbers, but non-linear devices can also be used to provide an alternative path for the flyback currents. Suitable devices include transient-rated diodes and rectifiers; and surge protection devices (SPDs) such as varistors or silicon avalanche devices (SADs), used either on their own or in combination with R and/or C components.

Figure 6Y shows an example of a rectifier snubber connected across an inductive load, such as a solenoid. DC powered loads can use rectifiers or unidirectional SPDs as snubbers, as in Figure 6Y, because the flyback voltage is always in the opposite polarity to the originally applied voltage.

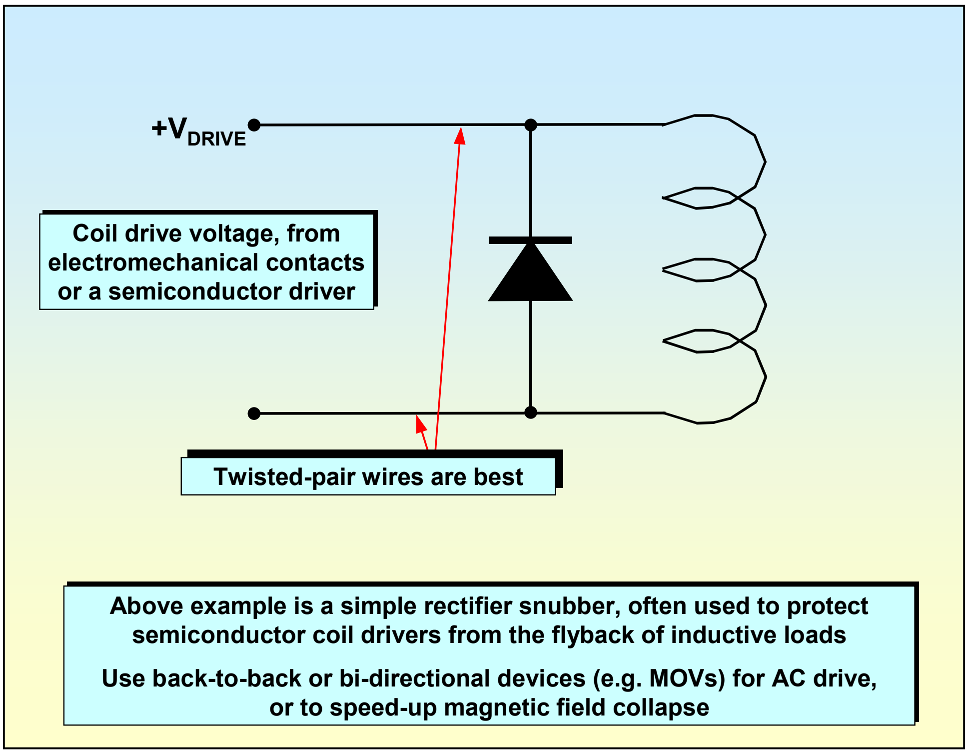


Figure 6Y Snubbing inductive loads reduces emissions

For reliability, diodes and rectifiers used as snubbers need to be surge-rated types, and even then they might need a series resistor of between 1 and 10Ω to limit their peak current to what they can reliably handle over the intended lifetime of operation, given the total number of surges they will have to withstand. It is no good using 1N4007 or 1N914, or zener diodes, or any of these 'regular' electronic components – although they may perform well on the test bench, using them in real products just guarantees high levels of dissatisfied customers and warranty costs.

Diodes and rectifiers used as snubbers will clamp the flyback voltage to around 1V, but such a low voltage means a low rate of field collapse ($V = LdI/dt$) and therefore a longer time to turn-off whatever the coil is actuating. Using SPDs (*not* zener diodes) instead of diodes or rectifiers increases the flyback voltage and hence the emissions – in exchange for a faster response time of whatever it is that is being controlled by the electromechanical device.

6.2.5 Suppressing commutator motors and generators

Some types of motors and electricity generators employ commutators, which spark and emit bursts or transient noise at a regular rate. If they interfere with an audio circuit, the result is usually a sharp-sounding buzz or whine that varies in pitch as the motor varies in speed – a noise that is very familiar to most people who have driven motor cars fitted with generators instead of the now-ubiquitous alternators.

The best way to deal with the problem of sparking commutators is not to use electric motors at all – use pneumatic or hydraulic motors instead (if pneumatic or hydraulic power is already available).

The next best method is to use electric motors that are based upon different technologies, for example:

- Brushless DC
- Stepper
- AC motors (squirrel cage, shaded pole, etc.) powered by an inverter if required to operate on DC supplies
- Switched reluctance
- 'Pancake' or 'PCB' motors

These do not spark, so do not create very wideband emissions, but each has its own EMC emissions problems over more limited frequency ranges so they might need some suppression applying anyway.

Pancake motors use commutators, but hardly spark at all because they have no iron in their rotors, and as a result have low inductance and low stored magnetic field energy. Their ironless rotors also have very low mechanical inertia, so they speed up and slow down very quickly, and they are, for example, used for disc drives in computers.

If it is decided to use a standard commutator motor, either powered by DC or rectified AC (a so-called universal motor), the best solution is to fit varistor (MOV) discs directly on the rotor. These dramatically reduce commutator sparking, and as well as reducing emissions will improve brush and commutator life.

There are several suppliers of varistor discs intended for the rotors of DC motors, but if the motor you want is not available already fitted with a varistor disc (e.g. the DC motors used in many CD and DVD players) motor manufacturers will probably require a large minimum order quantity to create a design of motor fitted with a disc. So having a special motor made with a varistor disc on its rotor is probably only a suitable solution for consumer, automotive or other high-volume products.

If none of the above methods are used, all that is left is to apply shielding and filtering techniques to suppress the emissions from the sparking commutators. I have suppressed many commutator motors, and have generally found it easier to suppress large industrial motors than smaller commercial ones.

The noisiest commutator motors are often the very small ones found in low-cost children's toys, and some of them would make good wideband noise generators for checking EMC test equipment and sites – if their emission characteristics were more stable.

I was once testing the emissions from a large road-going truck that had been fitted with a variable-speed electric motor drive, a large and very powerful X-ray generator, a bunch of computerised controls, Wi-Fi and had flashing beacons and a warning siren on its roof. The siren used a commutator motor to drive it, and I found that the motor in its warning siren, as it slowly cruised past my antennas under remote control, caused the highest emissions from the truck.

The motor and its brushes should be enclosed in a metal box, with only small ventilation holes, to act as a shield. This box could be a metal motor frame, or an external metal enclosure. But it can be difficult to use motor frames as shields, where their construction does not ensure a reliable electrical contact, over their lifetime, all around their joining areas between the metal parts of the structure.

If the rotor's metal shaft connects to external metal items, it may be necessary to use a rotary conductive gasket or other contacting method to 'RF bond' the metal shaft to the motor's metal box. I have never found it necessary to do this, but it is a possibility, since even steel roller or ball bearings are generally non-conductive during rotation due to their grease.

Finally, the motor cable must enter the metal box very close to the brushes, and ideally be shielded along its entire length with 360° shield bonding *at both ends*. Parts 2 and 4 of this series ([14] and [16]) describe how to do 360° shield bonding correctly. At the motor end the shield must bond 360° to the motor's shielded enclosure, so motors fitted with a terminal box must use a metal terminal box that is welded to the motor frame at numerous points around its perimeter. Plastic terminal boxes prevent the use of shielding techniques.

If the motor cable is not shielded as described above – for example, if the cable is a shielded type but its shield is only terminated at one end, or terminated at either end by a 'pigtail'(a short length of wire or connector pin) – then the motor should be fitted with a filter that is 'RF-bonded' to the motor's shield, right at the point where the motor cable enters it.

'Feedthrough' filters are the best, but costly. Any other types of filters must use very short 'earthy' connections to bond to the motor's box. Sometimes a simple capacitor filter (e.g. 1nF multilayer ceramic) on each brush lead is sufficient. If a brush lead is at the same potential as the box, it should be RF bonded directly to the box – no need for a filter capacitor. It is most important that the leads from such capacitors are as short as possible, as indicated by Figure 6Z, and the most difficult part of this suppression technique is designing a short-leaded construction that is also low-cost and easy to assemble.

If capacitors on their own don't provide enough suppression, chokes and other filter components can be added, as shown in the lower circuit in Figure 6Z. I have found that they work best if fitted on the 'other side' of the capacitors discussed above, but some guides show them in a different order, with a CM choke connected directly to the brushes instead of capacitors – a circuit topology that I have never found to be as good.

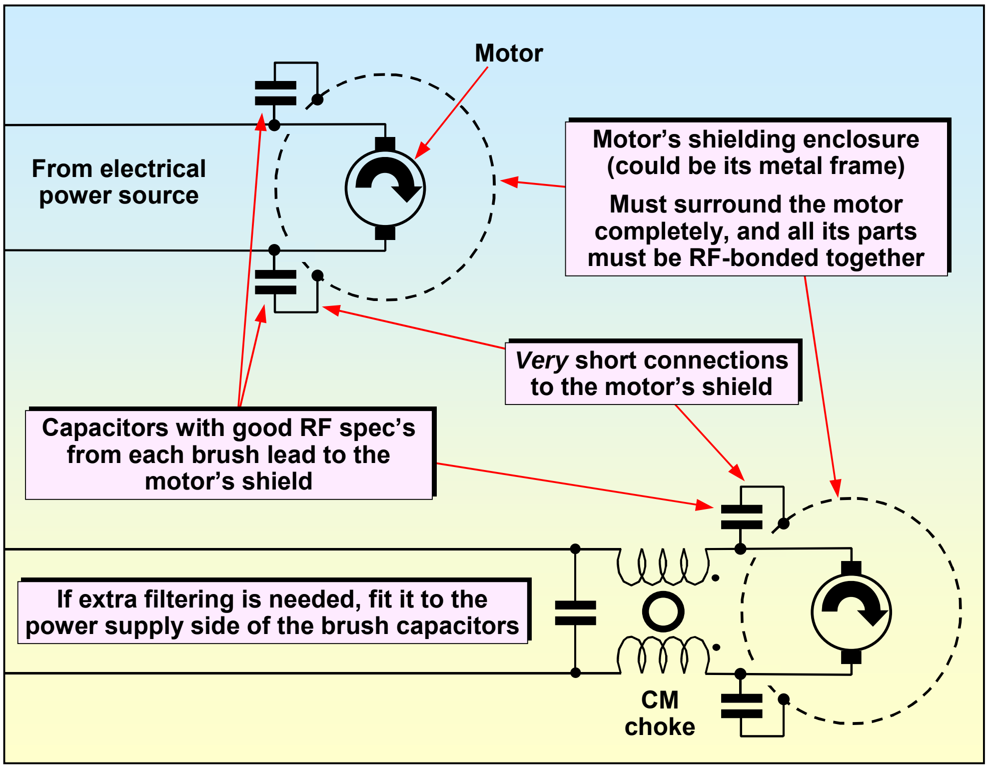


Figure 6Z Examples of filtering commutator motors

The X2Y™ Attenuator® is a recently developed type of device that has proven to be very effective indeed at suppressing DC motors. It is a three-plate 'balanced' capacitor with four terminals: A B and two 'ground terminals G1 and G2. Providing the current paths to the G1 and G2 terminals are identical (or nearly), the X2Y™ device has a much lower self-inductance than a two-terminal capacitor and so is more effective at suppressing RF.

X2Y™ devices are discussed as regards their benefits for decoupling applications on PCBs in section 5.2.10 of [18]. A great deal of useful information and test data on using X2Y™ devices for suppressing DC motors is posted at [27], some of which claims to show that a single X2Y™ Attenuator®, assembled correctly, provides more suppression than a good design of filter with as many as seven 'regular' components (chokes, capacitors, etc.).

Figure 6AA shows how to apply an X2Y™ Attenuator® to a motor's shield, and includes the important comment that the physical connections from the two ground terminals G1 and G2 must be physically identical if the full suppression benefits are to be achieved. As for the brush-connected capacitors in Figure 6Z – designing a practical assembly is the difficult part.

X2Y™ devices use ordinary capacitor manufacturing techniques, and are made under licence by several capacitor manufacturers. Types that are manufactured in high volumes, e.g. for automotive components such as suppressing windscreen wiper motors, cost little more than regular two-terminal capacitors of the same value, but types that are not made in high-volumes can be significantly more costly than regular capacitors. So to keep costs low it is best to discover which values have the lowest prices, and design with them.

However, if a single X2Y™ device can replace seven ordinary components, it could be cost-effective overall (e.g. by helping reducing board area) even if the device itself was significantly more expensive than an ordinary capacitor with the same value.

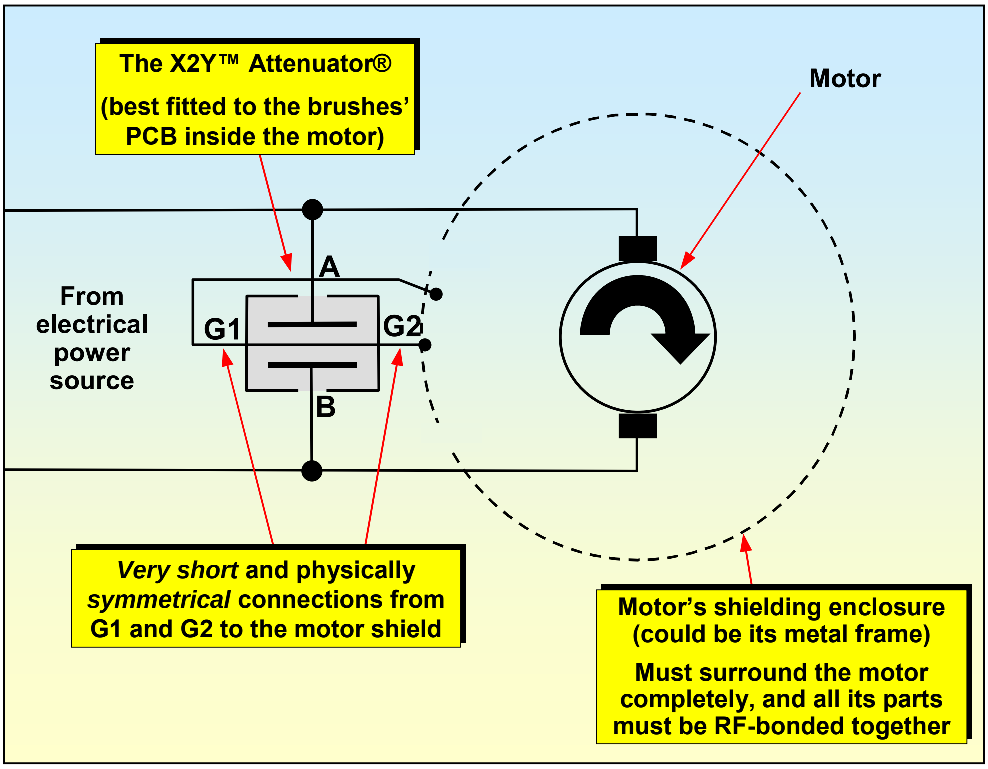


Figure 6AA Suppressing a commutator motor with an X2Y™ Attenuator®

6.2.6 Suppressing sliprings

Sliprings can be purchased with good performance for electronic signals up to several GHz, but passing power through sliprings often causes sparking, which of course is a source of EMI and also erodes the slipring's contact area and reduces its useful life.

The best approach is to replace power sliprings with an inductive coupling – essentially splitting a transformer core in half and making one half fixed while the other rotates. Another approach is to use an optical rotary joint – but at the moment devices for converting the laser light to electrical power are limited to just a few watts, so the technique is more appropriate where galvanic isolation of power is required.

If the angle of rotation is limited (e.g. a car steering wheel) it can be best to replace the slipring with a flexible cable (usually a flat cable) or flexible PCB wrapped around the shaft, often called a 'clock spring' connection. This requires no sliding contacts and can also be shielded.

But if it is decided to pass power through sliding contacts after all, it is best to use multiple brushes at each contact, with a very low inductance between the brushes, for example by mounting them on a short metal bar that acts as their common electrical connection. Then, if one brush experiences a high resistance contact, the current that was flowing in its circuit can divert to a brush that has a good low resistance without experiencing enough inductance to cause flyback and sparking.

Also, snubbers can be applied to the slipring. Obviously, the series snubbers shown in Figure 6W are unsuitable, but the 'low-leakage' snubbers shown in Figure 6X can be fitted to the circuits on one or both sides of the slipring.

6.2.7 Suppressing spark ignition

For gas/oil burners, it is best for EMC to replace their spark ignition with ‘hot-surface’ igniters. These are ceramic devices with resistance elements in them that cause them to quickly glow red-hot and ignite the fuel. Of course, they have no emissions, but many years ago they got a reputation for being unreliable, so most appliance manufacturers changed to using spark ignition which creates high levels of transient emissions.

However, I understand that modern hot-surface igniters are reliable, and using them instead of spark ignition should reduce complexity, save cost and eliminate sparks and their transient emissions altogether. Clearly, they are only suitable for gas ands oil burners, since they cannot respond quickly enough for applications where spark timing must be accurate within less than 1 second.

Where spark ignition is to be used, it is best to integrate its high-voltage generator in the same housing as its spark gap to minimise the length of the high-voltage conductors, making them much less efficient as ‘accidental antennas’ [14].

Another technique is to fit a suitable type of high-voltage filter very close to the spark gap, minimising the length of high-voltage conductor between it and the spark gap to minimise ‘accidental antenna’ effects. Snubbers as described in 6.2.4 are not suitable, because they aim to prevent the spark occurring. Filtering techniques might not be suitable where spark timing is critical.

If using long high-voltage conductors, their ‘accidental antenna’ efficiency can be reduced by using lossy (carbon) conductors, such as are normally fitted to the spark plug leads of motor vehicles that use a high-voltage distributor. The series resistance ‘dampens’ the RF resonances in the conductors, an effect that can also be achieved using ferrite cylinders of toroids around each *individual* conductor (not around both together, as is more common, for suppressing common-mode RF emissions).

Finally, all the usual wiring techniques described in [14] can be used to reduce the emissions from the high-voltage cables, for example:

- Twisted-pair wiring, maybe with ferrite cylinders or toroids around the pair as CM chokes
- Routing close against a CM return path along entire route
- Shielding (with the shield 360° terminated at both ends)

6.2.8 Electric bells and buzzers

The spark gaps in traditional electric bells and buzzers are a big EMC problem. Most of the emissions are conducted along the bell wires or induced by them into nearby conductors or components, as well as being radiated by the bell wires acting as ‘accidental antennas’ (see 2.2. of [14]).

The spark gap in the bell or buzzer can easily be filtered to keep the RF noise out of the bell wires, but the problem is that bells and buzzers are very low-cost components and effective EMC filters can cost more than the bells or buzzers themselves!

As for all of the arcs and sparks discussed in 6.2, the best approach (where it is feasible) is to get rid of the arcs and sparks altogether by using a different technology. In the case of bells and buzzers the sound generating element might be able to be replaced by a different technology altogether, such as a piezo-electric sounder.

If the bell clapper or buzzer plate are to be retained as the sound generators, the spark gap can be replaced by an astable oscillator – driving the coil that drives the striker or buzzer plate directly. Like most of the design techniques that eliminate arcs and sparks instead of suppressing their effects, this results in much more reliable low-cost electric bells because there are no sparking contacts to erode and wear out.

Assembly of such bells and buzzers could even be less costly, because there is no spark-gap adjusting screw to adjust and lock. Plus of course no need to adjust it to compensate for erosion of the spark gap over the life of the product.

I have a book entitled ‘Electric Bells’ that was published by Cassell and Company Ltd in 1901, and sold for the princely sum (at that time) of one shilling. Apart from a rectifier snubber the designs and construction of modern bells correspond exactly to the figures in that book. In these days of EMC emissions controls, it seems we might need to break with over 100 years of tradition.

6.3 Power factor correction (emissions of mains harmonic currents)

6.3.1 The particular problem of rectifiers with capacitive loads

All non-linear loads powered from a supply of AC electricity consume non-linear currents, which if analysed in the frequency domain can be seen to have some currents at even and/or odd harmonics of the supply frequency.

In the past, fluorescent lamps with traditional magnetic ballasts have created significant problems due to their high levels of harmonic currents, but in order to achieve higher energy utilisation efficiencies almost all loads are now becoming electronic. For example in the lighting industry alone, magnetic ballasts are being replaced by high-frequency electronic ballasts, and filament lamps by compact fluorescent or LED lamps, all of which are powered via what are essentially switch-mode power converters with rectifier-capacitor power inputs.

Most electronic loads start with a rectifier/capacitor circuit to generate an unregulated DC rail from the AC supply, and these are very non-linear circuits indeed, so one consequence of trying to save the planet (or at least reduce electricity costs) is an increase in harmonic current 'pollution' of the mains distribution networks.

Harmonic currents in supply networks have a number of undesirable effects, from waveform distortion to fire and smoke hazards [31], and [12] contains many real-life examples. As a result, limits have been imposed on emissions of harmonic currents into the mains supply, by standards (EN 61000-3-2 and EN 61000-3-12) listed under the EMC Directive for products consuming up to 75A from the 230/400V mains supply.

Much more information on mains harmonics, where they come from and how they can cause interference and other problems, can be found in the guides on IEC/EN 61000-3-2 (emissions) and IEC/EN 61000-4-13 (immunity) that are available from [7].

Figure 6AB shows a typical bridge rectifier feeding a capacitor, and notes how this basic circuit is used in both linear and switch-mode types of AC-DC mains power converters.

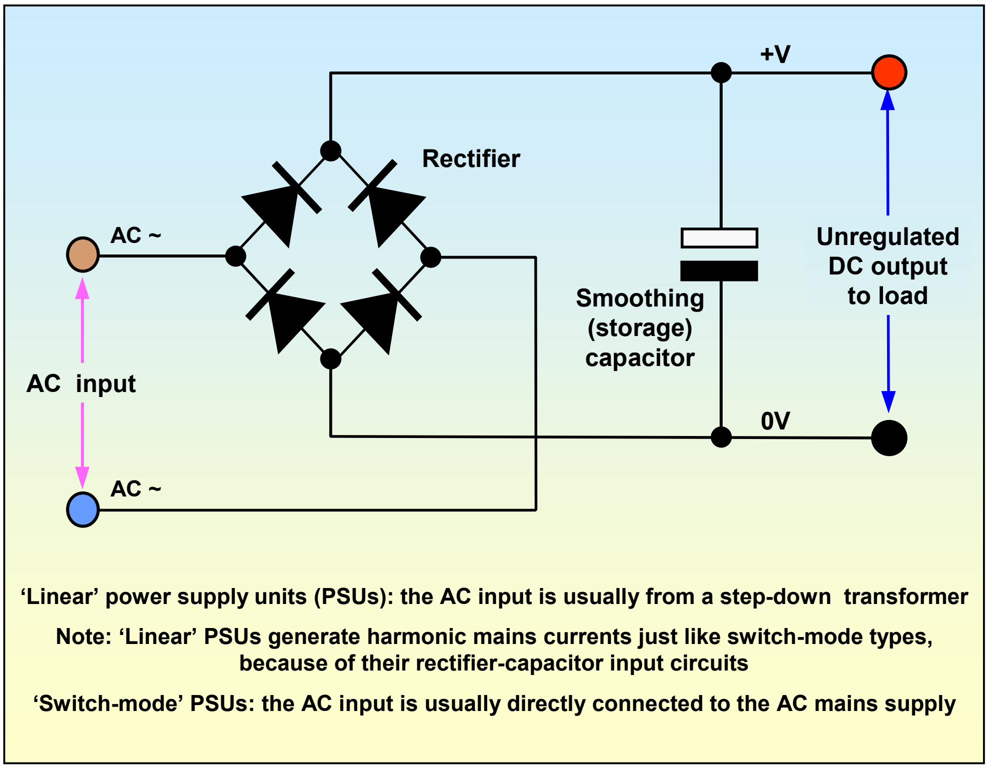


Figure 6AB Example of a bridge rectifier

Figure 6AC compares a pure sine-wave mains supply waveform, with the typical currents consumed by the circuit in Figure 6AB. (Pure sine-wave mains supplies have not been seen in real life for many years now, but it is the principle that matters here.) It is clear that because the capacitor holds a charge in order to 'smooth' the DC rail voltage, the rectifier's mains current only flows for the short periods when the mains voltage is higher than the capacitor voltage by at least one rectifier on-voltage.

Since all the load's power must be drawn from the mains supply during these brief periods, the peak mains current is much higher than it would be if the same power was drawn directly from the AC supply by a simple resistor. The more the DC rail is 'smoothed' by increasing the ratio of capacitance to load current, the shorter and higher the mains current pulses become.

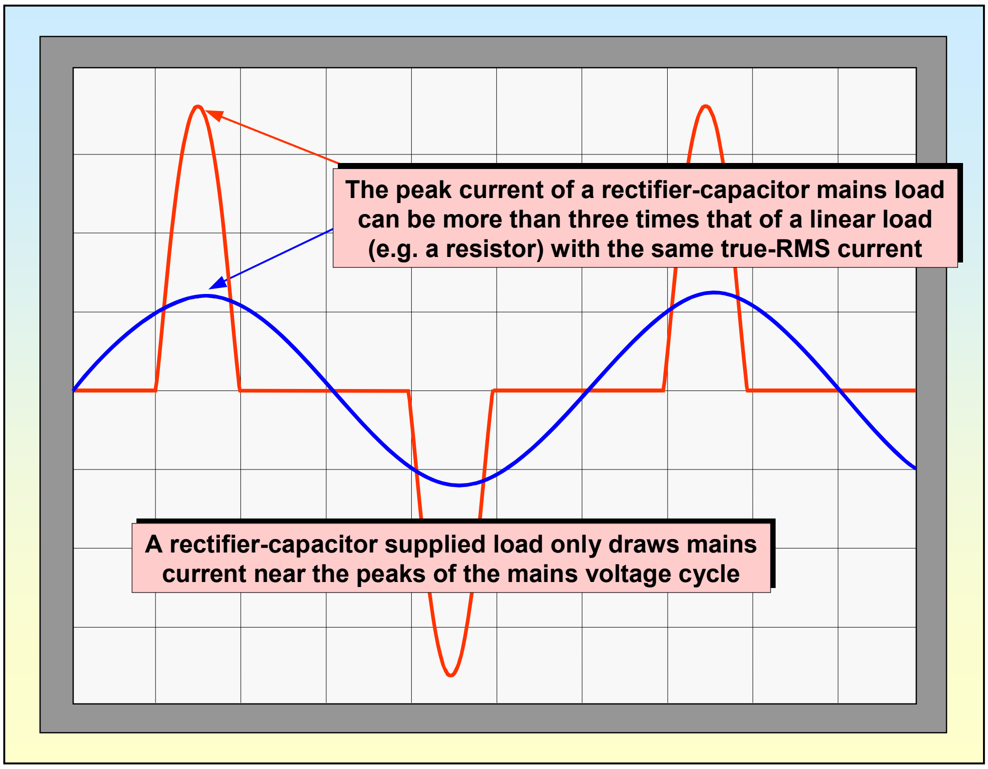


Figure 6AC Non-linear currents drawn by a bridge rectifier

Figure 6AD compares the current waveforms of a linear load and a rectifier-capacitor supplied load of the same wattage, and also compares their frequency spectra. The linear load has a single frequency component at the AC supply frequency, whereas the non-linear current also has frequency components at odd-numbered multiples (3rd, 5th, 7th, etc.) of the supply frequency.

The current in a single rectifier operating in half-wave (rather than a full-wave bridge rectifier) will have frequency components at even-numbered multiples (2nd, 4th, 6th, etc.) as well as the odd-numbered ones, and half-wave rectifiers will also have significant levels of harmonic currents when they feed into a resistive or inductive load, without a smoothing capacitor.

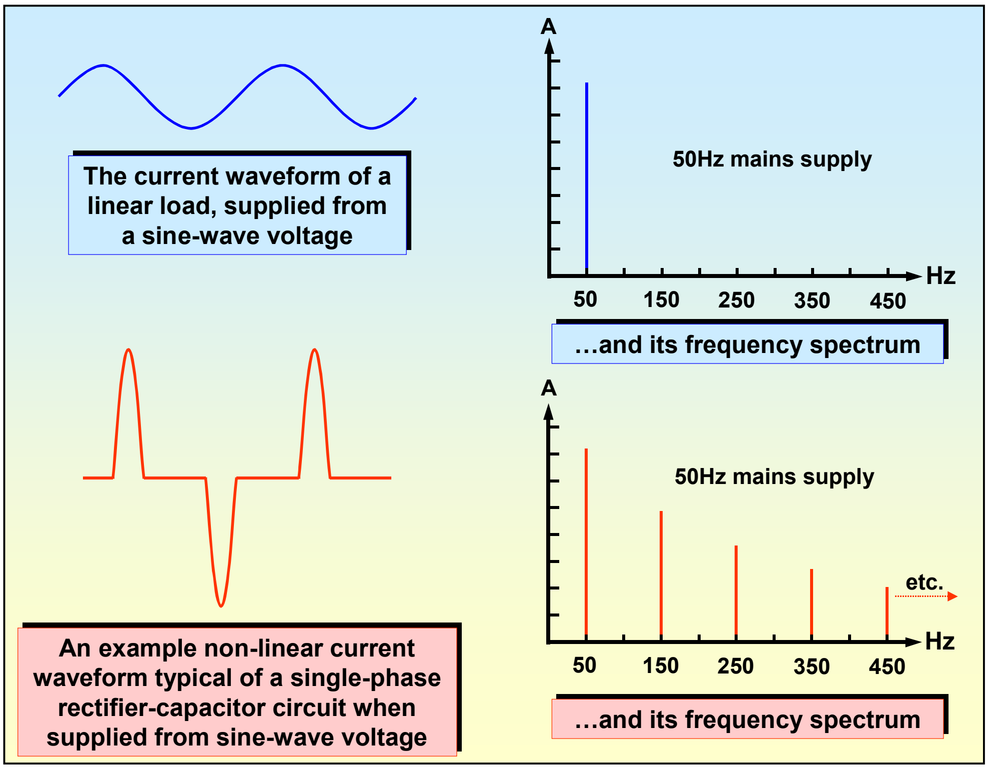


Figure 6AD Waveforms and spectra associated with rectifiers

Phase-angle-controlled silicon controlled rectifiers (SCRs) such as thyristors and triacs are also rectifiers, but their current waveforms vary depending on their phase angle of conduction, and so the frequency spectra of their mains currents also vary with phase angle, as shown in Figure 6AE. When set to less than full conduction they generate significant harmonic currents even when connected to a resistive or inductive load, without a smoothing capacitor.

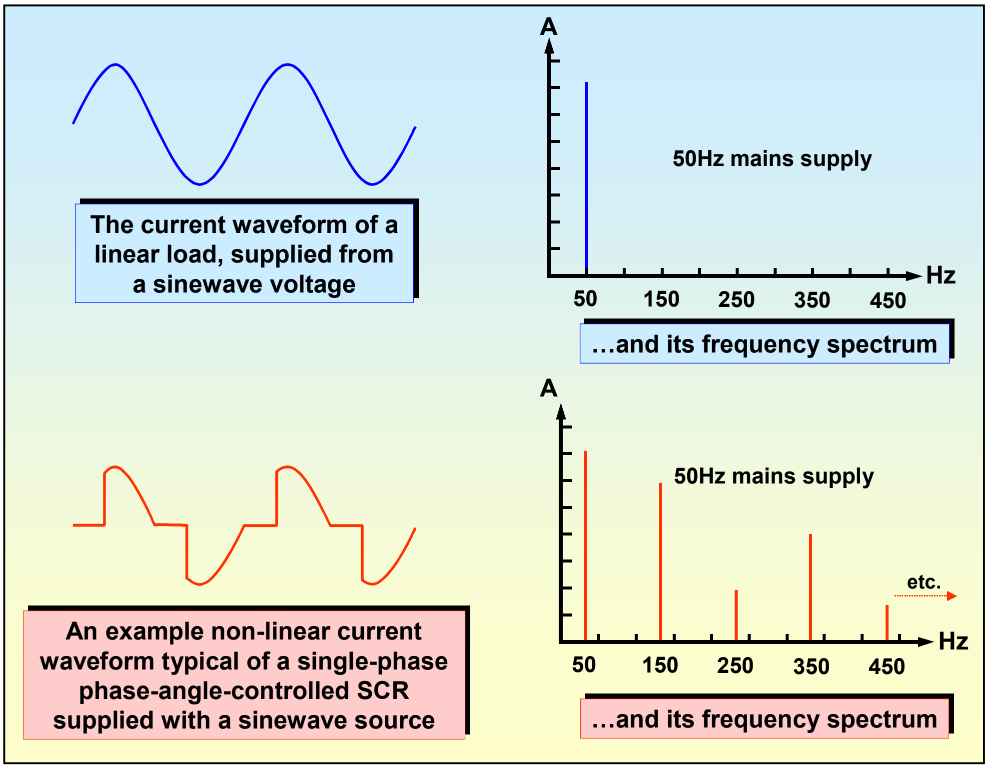


Figure 6AE Waveforms and spectra associated with SCRs

A solution to the harmonic currents generated by phase-angle-controlled SCRs, is to use burst-firing techniques instead, where practical. Burst-firing turns the SCRs on and off close to zero-crossings, so the current they draw depends solely on the load and supply voltage. Burst-fired SCRS cannot actually turn on until they have sufficient voltage, and automatically turn off when their current falls below a threshold, so they still cause RF emissions in bursts around each zero-crossing of the supply, which can usually be easily suppressed with capacitors. SCRs are not discussed any more in this article.

The Power Factor (PF) is the Total Watts divided by the Total VA, but harmonic currents are not in phase with the supply voltage, so they increase the VA but not the Watts. A PF of 1.0 is ideal, but increased consumption of harmonic currents reduces it below this magic figure. This is why designing to reduce emissions of harmonic currents is generally called Power Factor Correction (PFC).

It is important not to get confused with the term ‘Power Factor’ as traditionally used by electrical supply engineers, which is really ‘displacement power factor. Total Watts/Total VA is true power factor, because it is independent of waveshape, but displacement power factor is simply the cosine of the phase angle between a sine-wave supply voltage and its sine-wave load current.

Displacement power factor is still a relevant concept for AC mains distribution networks where the voltages and currents still resemble sine-waves, but is meaningless when considering the non-linear currents consumed by a rectifier such as Figure 6AB with waveforms like those shown in Figure 6AC and 6AD. Unfortunately, the technical press often do not differentiate between the two uses of the term ‘power factor’, so often publishes articles and advertisements about power factor correction that are really about correcting displacement power factor, for example by compensating for the effect of the inductance of a long power line by adding some capacitors at its end.

The rest of this section discusses design techniques for reducing emissions of harmonic currents from bridge rectifier – capacitor circuits. There are a number of methods that can be used, including…

- Using smaller values of unregulated DC smoothing capacitor
- Harmonic filtering
- Passive PFC using a series inductor
- Passive PFC using a charge pump
- Active PFC using a boost regulator

- Increasing 3ϕ rectifiers to 6ϕ
- 'Active Front End' (AFE) three-phase 'boost' rectifiers
- Anti-harmonic injection
- Other methods

Test equipment for measuring emissions of mains harmonics is available at reasonable cost for products consuming less than 1kW, and an example of such equipment is shown in Figure 6AF. The largest cost of any harmonics measurements system is in providing a source of high-power low-distortion low-impedance AC power. For those of a more 'do it yourself' (DIY) nature, some designs of alternative harmonic current measuring equipment are given in Part 7 of [6] – but the safety considerations in 6.2.1 must always be taken fully into account.

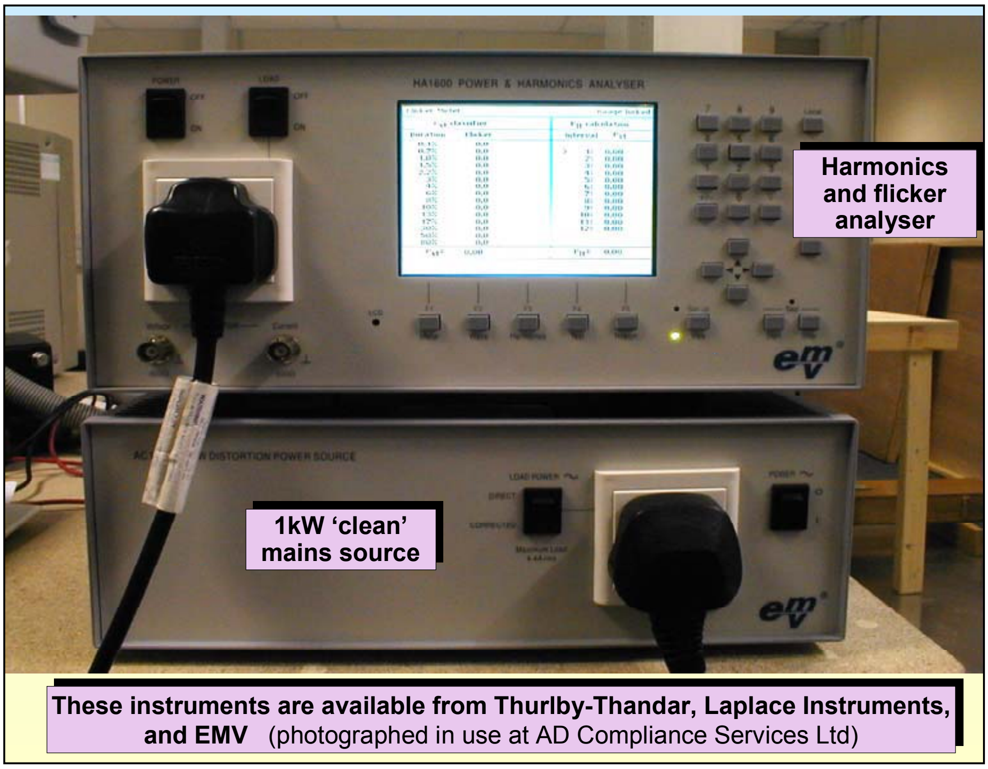


Figure 6AF Example of a low-cost harmonics tester

6.3.2 Using smaller values of unregulated DC smoothing capacitor

Lower values of smoothing (storage) capacitor cause it to discharge more completely into the load, during each mains half-cycle, and this increases the firing angle (on-time) of the rectifiers and so reduces the harmonic emissions, as shown in Figure 6AG.

The downside is that the ripple voltage on the unregulated supply will increase, and the load must be designed to operate correctly with such a large DC ripple. This technique has been successfully applied to some makes of variable-speed AC motor inverter drives.

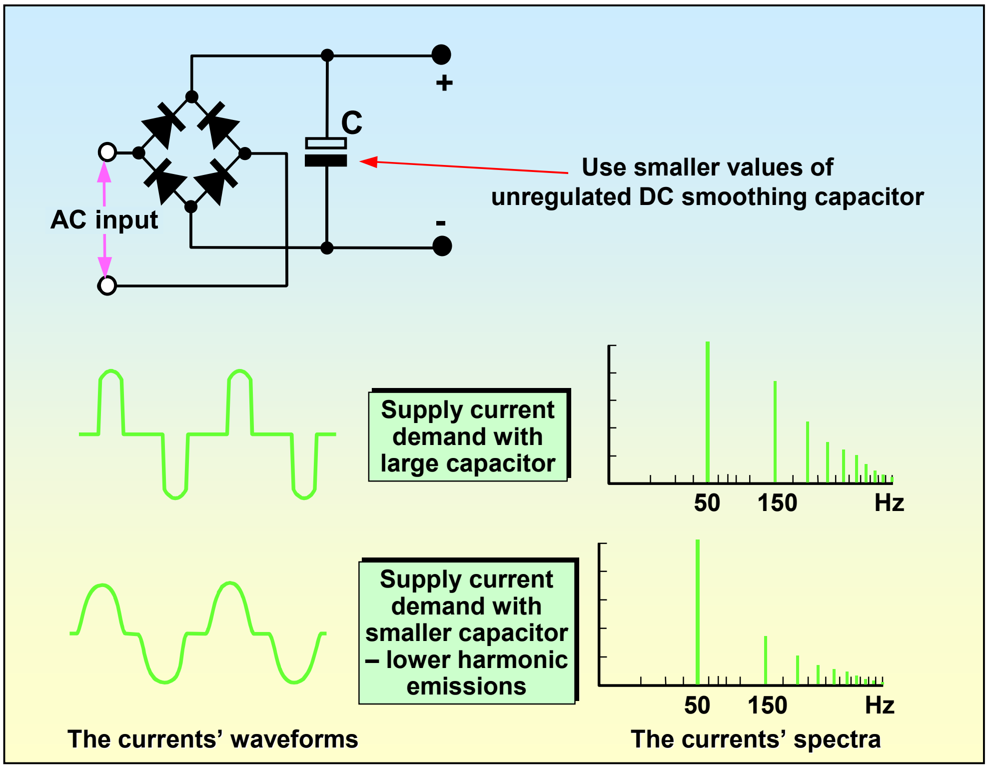


Figure 6AG Using smaller values of unregulated DC storage capacitor

6.3.3 Filtering mains harmonics

Harmonic currents can be treated as noise and simply filtered out using a low-pass filter. Unfortunately, even if the 3rd harmonic is the lowest to be filtered, this is still only 150Hz (or 180Hz for 60Hz mains) and achieving significant attenuation at such low frequencies with inductors and capacitors requires multi-stage filters using large values of components. Series inductors must handle the full mains current without dissipating too much heat, and shunt capacitors must handle the full mains voltage (and any surges and transients), which makes them large, heavy and costly components.

An alternative to the low-pass filtering approach is to use 'resonant trap' filters for harmonics up to the 6th or 7th, with single-stage low-pass filtering for all of the higher harmonics. One of the problems of shunt-connected filters is that they tend to absorb harmonic currents in response to distortion of the AC supply's voltage waveform, so filter designs should take this into account. Figures 6AH and 6AJ give some examples of the types of filters that are often used, and any good filter design manual will provide the necessary design equations for these and many other types.

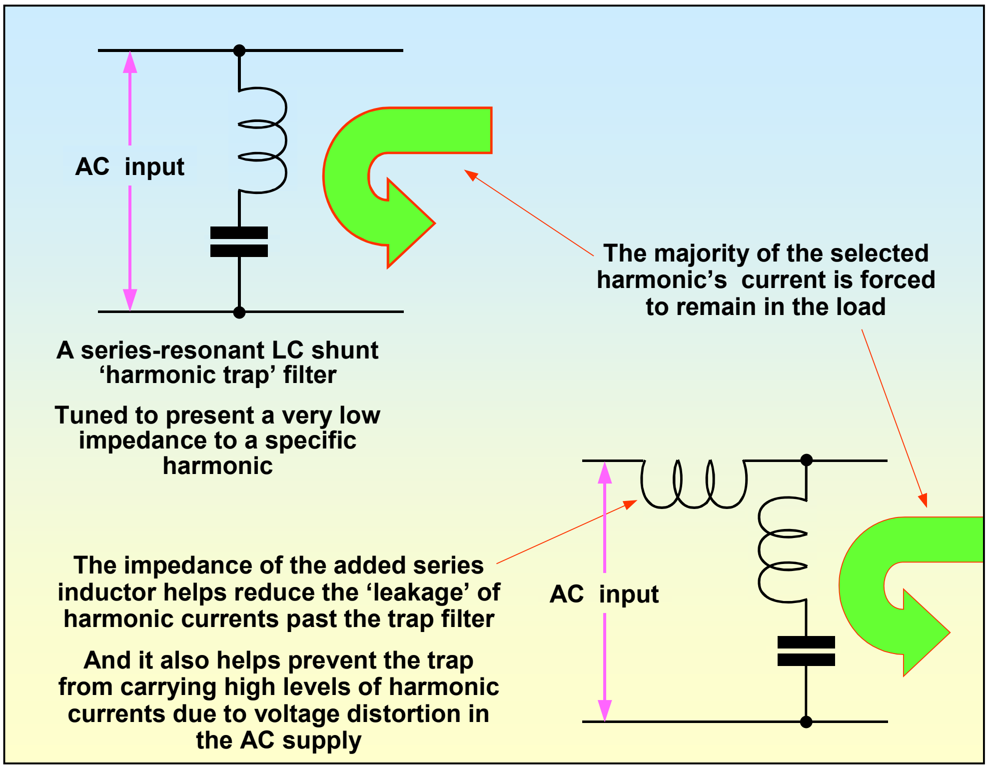


Figure 6AH Some examples of filtering techniques suitable for mains harmonics

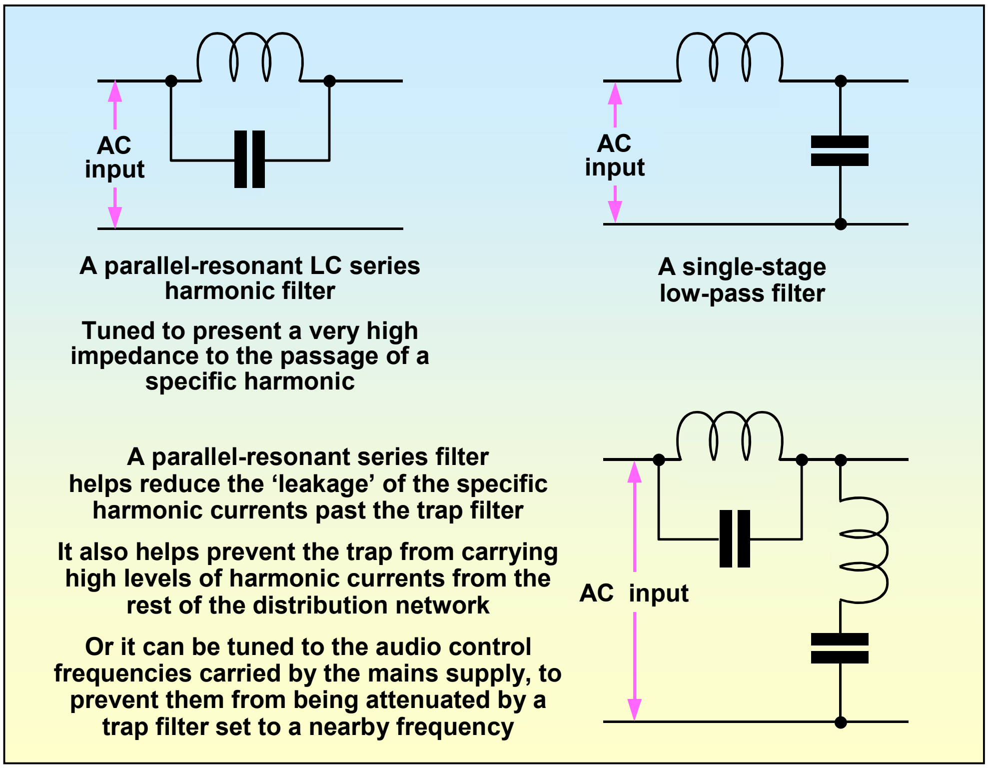


Figure 6AJ Some more examples of filtering techniques suitable for mains harmonics

6.3.4 Passive harmonic reduction with a series inductor

Adding an inductor (also known as a choke) in series with a rectifier-capacitor circuit forces the rectifiers to conduct for longer. Figure 6AK shows the series inductor added in series with the rectified mains, between the rectifier and the smoothing capacitor. The voltage of the unregulated rail will probably be different than without the choke, and due to the impedance of the choke it will probably vary by a greater degree with loads that have fluctuating current demands.

If the inductor has a large enough value, the current in it will not drop to zero during each cycle and the rectifier bridge will switch with close to a square-wave duty cycle – giving an excellent PF. The downside of this technique is that chokes in the DC path must not saturate, and so must be large and heavy – they can be about half the size of a mains transformer rated for the load power, or even larger.

It has been claimed that using a two-winding choke on a common magnetic core, with one winding in the positive output from the bridge, and the other in the negative output, can help meet RF emissions limits by reducing the switching noises from the rectifiers, but I have not tried this, and simply adding capacitors of 10-100nF in parallel with each rectifier seems to work just fine anyway.

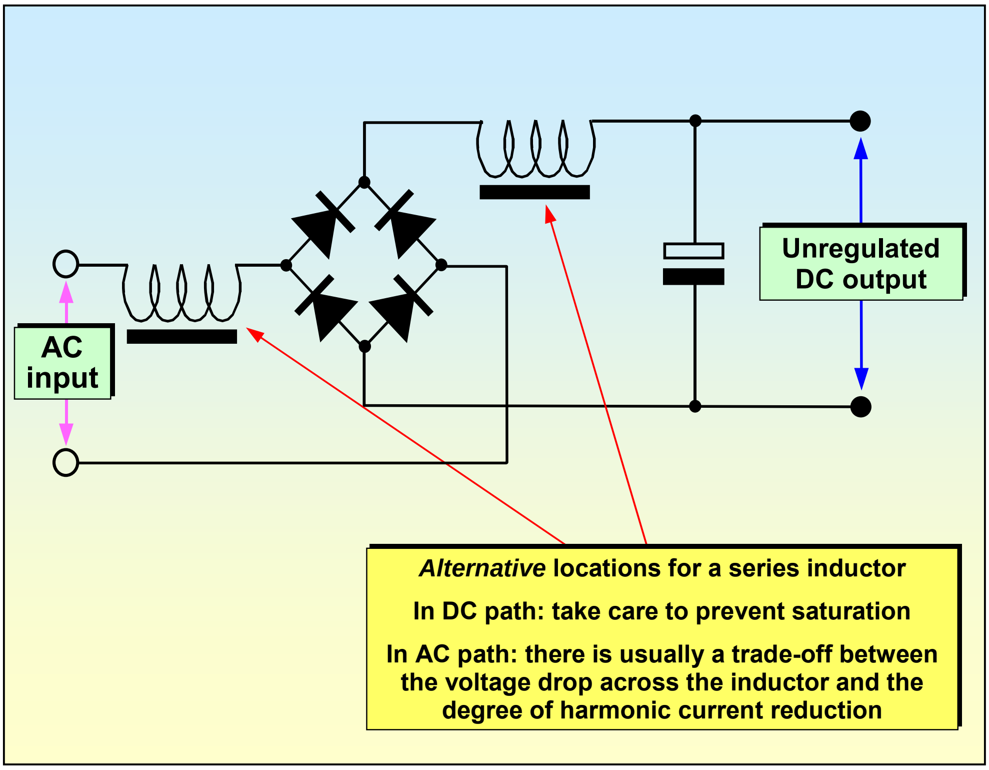


Figure 6AK Using a series inductor for PFC

Another, possibly less ideal location for the series inductor is in series with the AC supply itself, before the bridge rectifier. Several manufacturers supply such chokes (often called 'line reactors', or simply 'reactors') for installing in the mains supplies of products that have been found to emit too much harmonic current. (Adding chokes in the location shown by Figure 6AK would be better, but would invalidate the warranty, and possibly compromise the performance or reliability as well by altering the unregulated voltages.)

The oscilloscope traces in Figure 6AL show the effect of adding line reactors in series with the mains supply to a real-life three-phase variable speed motor drive.

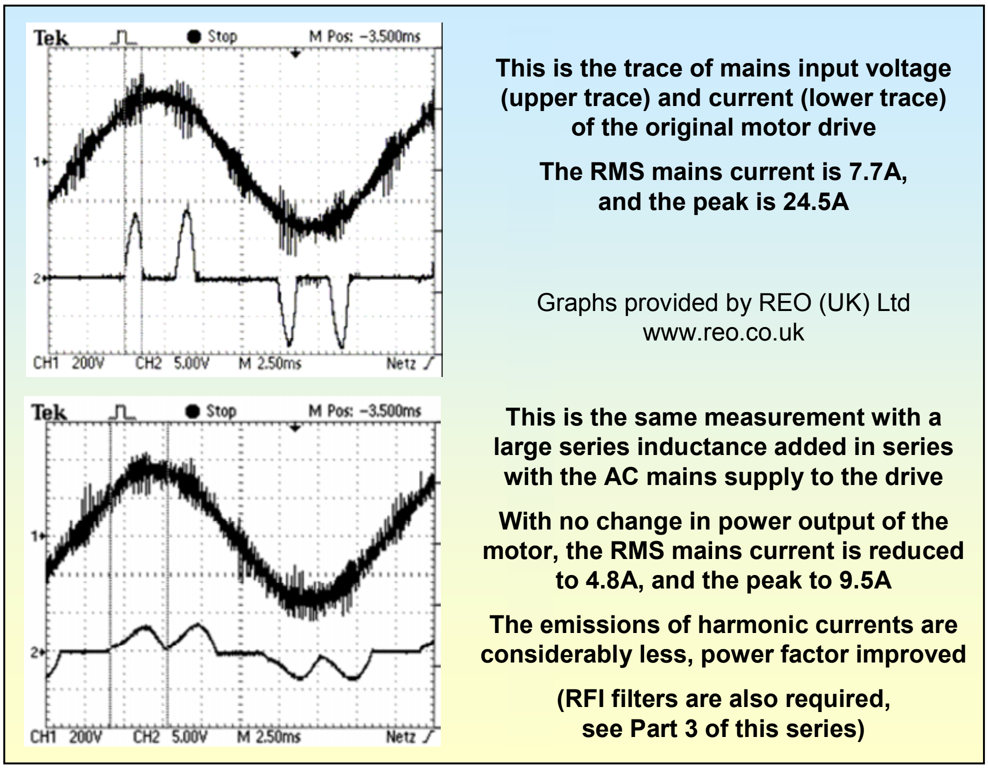


Figure 6AL Real-life example of adding a series inductor to a 3ϕ motor drive

6.3.5 Charge-pump PFC in a switch-mode converter

Some types of switch-mode power supply controller ICs make it possible to use 'charge pump' techniques to reduce harmonic emissions.

This is the lowest-cost method of PFC, and is widely used in TVs, VCRs, and similar mass-produced consumer products – but it only works with certain types of variable-frequency switch-mode controller ICs, for example the Infineon TDA 1684X family [28].

Figure 6AM shows the basic circuit, which is essentially just an ordinary switch-mode converter. The basic switch-mode circuit buzzes away at a few tens of kHz, and as the mains voltage decreases from its peak during each half-cycle of its waveform, the frequency of operation of the switching device is varied by the controller IC so as to maintain the desired DC output voltage (the feedback circuit is not shown in the figure).

When the storage inductor, rectifier and charge-pump capacitor within the dotted box are added and correctly dimensioned, the effect of the variable frequency switching is to vary the amount of charge injected via the capacitor in such a way that the mains current varies smoothly throughout the voltage cycle, simulating the sine-wave current that would result if the load on the bridge rectifier was a resistor instead of a capacitor. For details of how it works, including waveforms for all parts of the circuit, see [28].

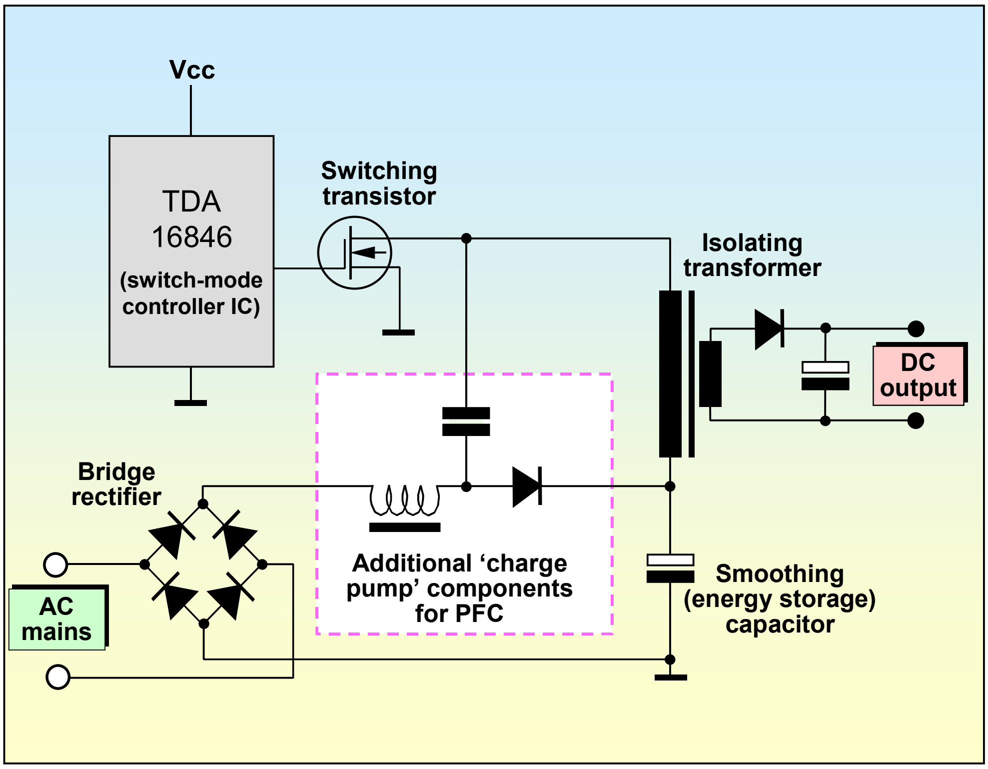


Figure 6AM Example of a charge-pump PFC circuit

6.3.6 'Active' PFC techniques

Both linear and switch-mode AC-DC power converters can use 'active PFC' techniques, which add a complete switch-mode boost converter between the bridge rectifier and the unregulated DC rail's smoothing (storage) capacitor. The converter is controlled by a suitable type of IC that causes it to simulate a resistive load on the bridge rectifier, hence a high value of PF and correspondingly low emissions of mains harmonic currents.

Figure 6AN shows the essential details of an active PFC circuit. The original bridge rectifier and smoothing capacitor (C1) are shown (compare with Figure 6AB), and a dotted box surrounds the active PFC boost converter.

With active PFC the value of C1 can generally be reduced, because instead of having to provide load current for nearly 10ms between 'top-ups' from the rectifier, its charge is topped up by the boost circuit every few microseconds.

Also, C1 generally needs to use a different type of capacitor – instead of a wound electrolytic, it probably should be a type that has lower impedance above 100kHz, so that it will respond appropriately to the high-rate of microsecond-duration current pulses from the boost converter.

The output of the active PFC circuit is still an unregulated DC rail, although because of the operation of the boost converter it tends to be smoother and more stable than would normally be the case with a simple bridge rectifier and smoothing capacitor. In fact, such boost converters can usually be designed to enable operation of the load with mains voltages over the range 100 to 250Vrms, a bonus for consumer products that might be sold or used anywhere in the world – all that is necessary is to supply them with the correct mains lead for the countries concerned.

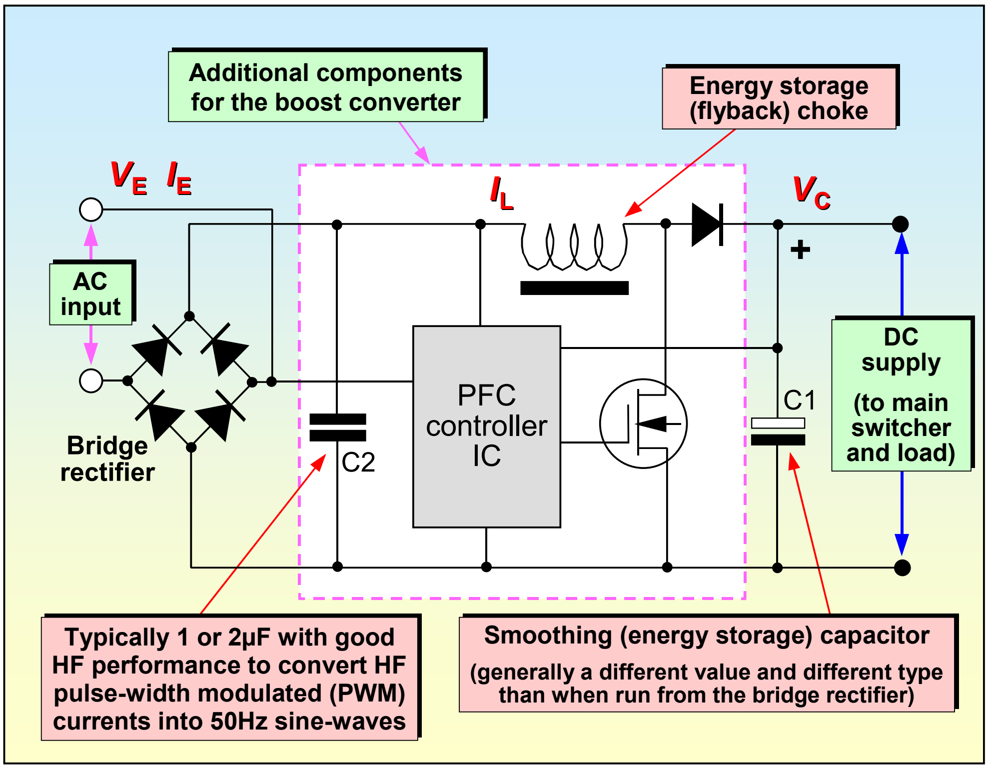


Figure 6AN Essential features of an 'Active' PFC circuit

Figure 6AP shows examples of the waveforms that would be expected at various points in Figure 6AN. The zig-zag waveform in Figure 6AP, I_L, is the current in the boost converter's storage choke, which is the integration of the pulse-width modulated (PWM) current in the hard-switched (either ON or OFF) switching transistor. When this current has been filtered by C2, it becomes a good representation of the sine-wave current that would result if the load on the bridge rectifier was simply a resistor. Hence the PF is very good – values of better than 0.9 are easily achieved.

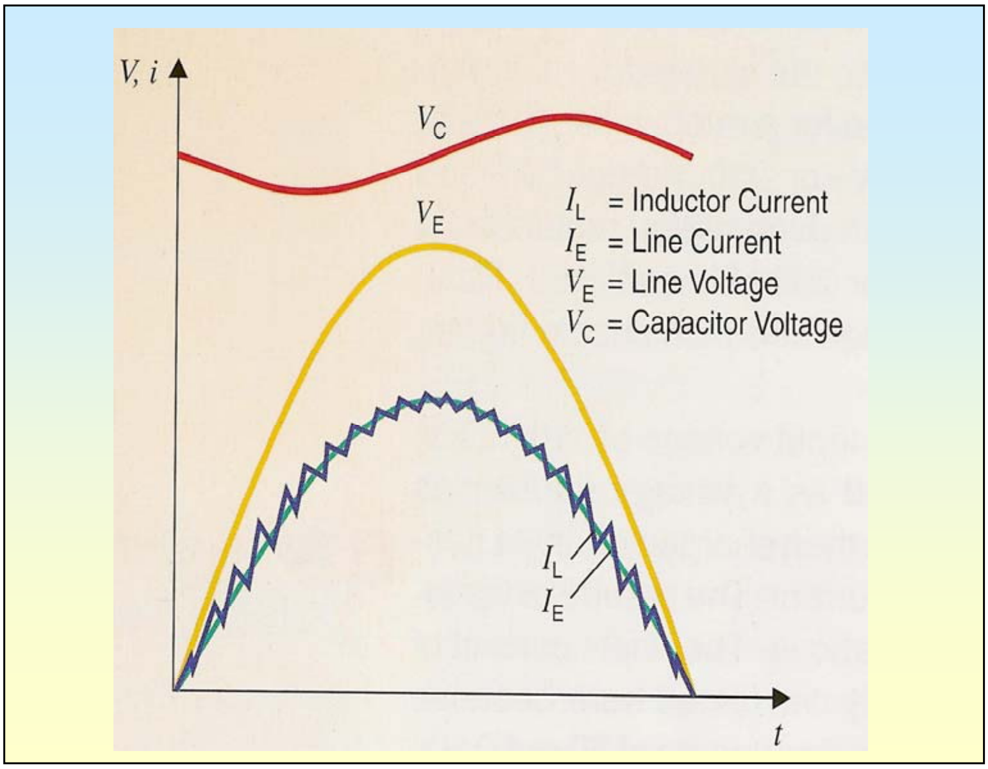


Figure 6AP Typical waveforms in an 'Active PFC' circuit

Figure 6AQ shows a practical realisation of an active PFC boost converter, copied from an EPCOS data sheet. All of the basic features shown in Figure 6AN are there, except for C2. Instead of C2 it relies on the mains filter to smooth the current waveform. In general, it is best to suppress a noise source as close as physically possible to it, and it might be that adding a capacitor from the flyback choke's input to the 0V (like C2 in Figure 6AN) might make it possible to simplify the mains filter and reduce the overall cost a little.

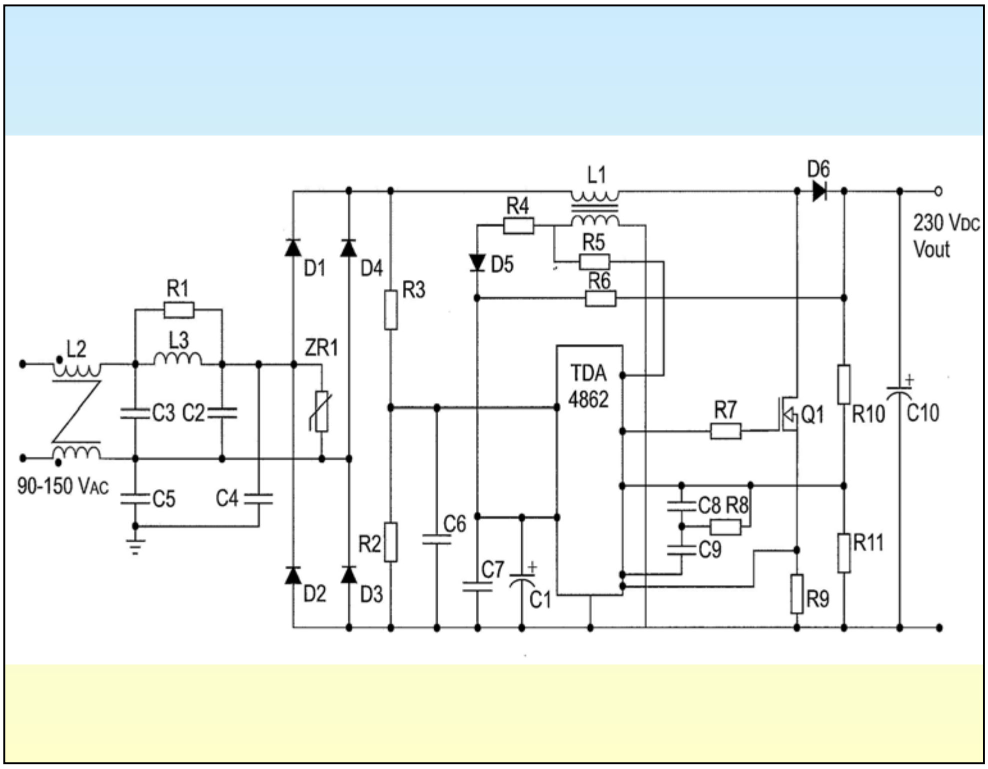


Figure 6AQ Example of an 'Active PFC' circuit from EPCOS

Texas Instruments publish many very useful application notes on designing PFC circuits [29], which were originally written by Unitrode. There are now many manufacturers of active PFC controller ICs, with competing circuit topologies that offer different features and claim to have advantages over each other, but such issues are beyond the scope of this article.

Active PFC circuits switch at tens or hundreds of kHz, maybe even MHz, and so can have high levels of RF emissions. Design techniques described in Parts 1, 3, 4 and 5 of this series ([13] [15] [16] and [17]) can be used to suppress these emissions, including…

- Soft-switching fast rectifiers
- Heatsink RF-bonding
- Segregation of PFC circuit from other circuits
- Suppression at all interfaces (shielding, filtering, overvoltage suppression, etc.)
- Use of planes to return currents
- Etc., etc.

6.3.7 Reducing emissions from 3-phase (3ϕ) equipment

3ϕ bridges (sometimes called '6-pulse rectifiers') fed from balanced supplies naturally have low levels of 3^{rd}, 9^{th} and other 'triplen' harmonics. Converting them to 6ϕ bridges ('12-pulse rectifiers') considerably reduces their emissions of 5^{th} and 7^{th} harmonics as well – providing they are supplied from a well-balanced 3ϕ supply with low waveform distortion.

There was a fashion for converting 3ϕ rectifiers to 6ϕ a few years ago, in industrial plant that had to comply with their electricity supply authority's regulations on emissions (e.g. [30] in the UK), but since they are less effective where power supplies are unbalanced or distorted – and since they require phase-shifting transformers, and copper prices have increased considerably – the trend now appears to be to use the techniques described in 6.3.8 and 6.3.9.

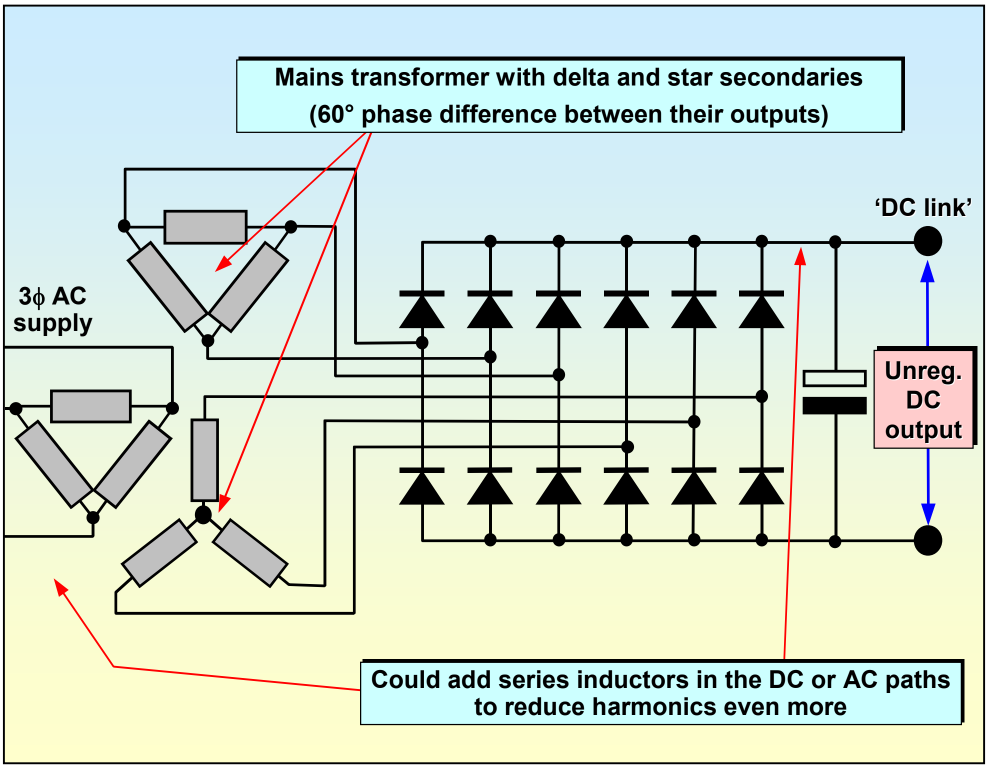


Figure 6AR Example of a 6ϕ bridge (12-pulse) rectifier

6.3.8 'Active Front End' (AFE) three-phase 'boost' rectifiers

These replace the 'dumb' rectifiers with pulse-width modulated (PWM) power transistors (usually IGBTs). When their PWM switching rate is much greater than the fundamental frequency of the supply, the switching functions act like their average values – so AFE techniques achieve PFC by using appropriate PWM switching patterns.

The AFE technique is of wide interest because by selecting different PWM switching patterns it can transfer power from its load to its source, so can be operated bi-directionally. As a result it can be useful, for example, for regenerative braking for machinery motors, and in alternative power generation equipment (solar cells, windpower, etc.).

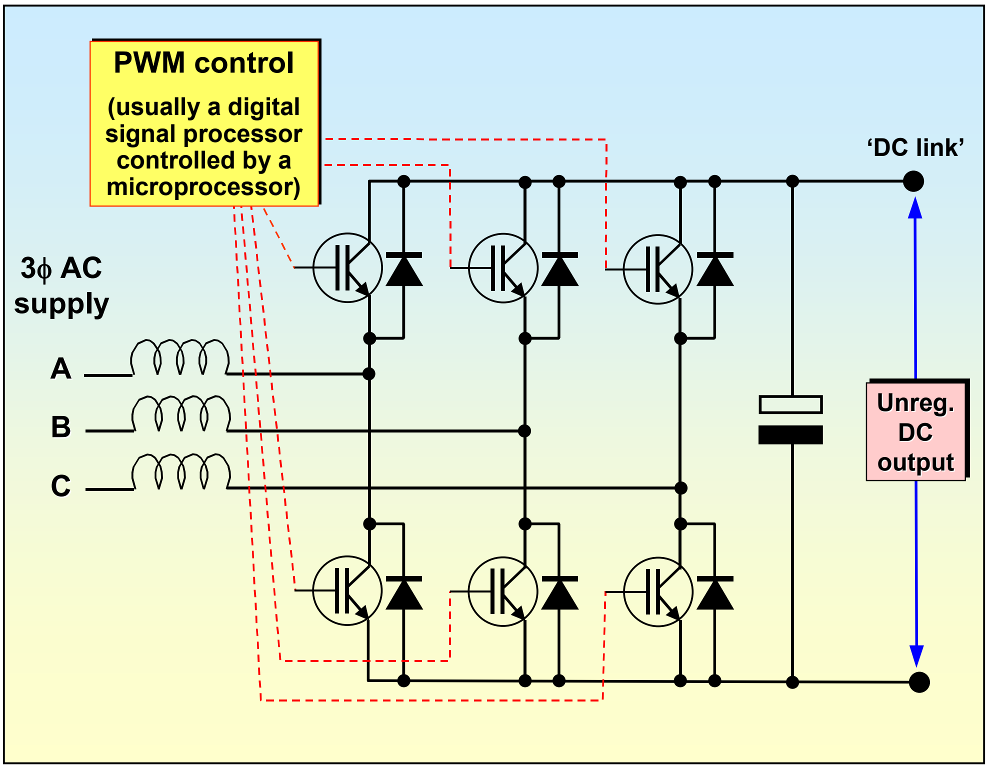


Figure 6AS An 'Active Front End' (AFE) three-phase 'boost' rectifier

6.3.9 Anti-harmonic injection ('active filtering')

Detecting the emissions of harmonic currents from a product makes it possible to generate antiphase versions of them. Injecting the anti-phase versions into the mains supply to the product can substantially cancel out the original harmonic currents, leaving an almost pure sine-wave current at the fundamental frequency with a PF of close to 1. Figure 6AT shows the general principle of operation.

This is a well-established method for cleaning up (i.e. reducing) the harmonic currents in mains power distribution networks in buildings of all types, and in industrial plant. I call this an 'anti-harmonic injection' technique, but the manufacturers of products intended for cleaning up mains distribution networks generally call them 'active filters' despite the fact that neither the technique nor their products have anything at all to do with filtering the mains supply.

This technique can be applied in a way that makes it relatively immune to phase imbalance and distorted supply waveforms – so it acts on the emissions from a product and doesn't try to correct the entire mains distribution network it is connected to. Although it is widely used in installations, this method can also be (and has been) incorporated into products.

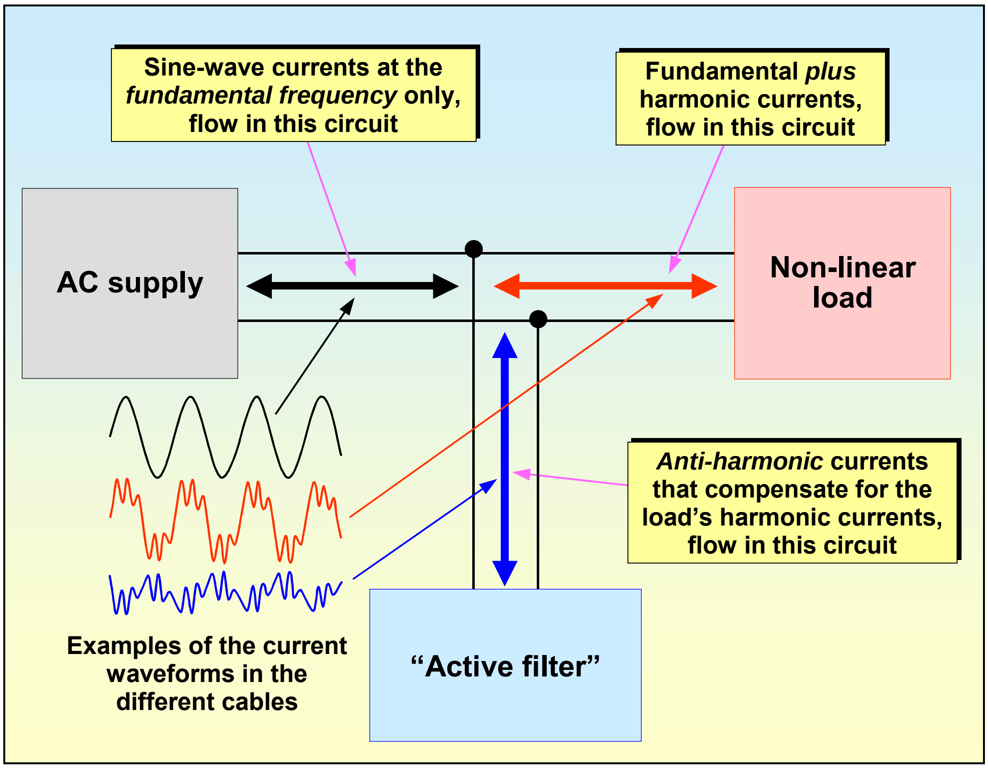


Figure 6AT Essential features of an 'active harmonic filter'

6.3.10 Other methods

This series is intended for designers of products and equipment, but there are many techniques for dealing with harmonics in power distribution networks themselves, described in [31]. Since they started to use electric thrusters, the mains distribution networks on ships have begun to suffer enormous levels of harmonic distortion, and special techniques have been developed for them, described in [32].

Products, especially larger items of equipment, might be able to use some of these other 'installation' techniques.

6.4 Emissions of voltage fluctuations and flicker

6.4.1 Causes of emissions of voltage fluctuations and flicker

This section addresses equipment powered by the 230/400V AC mains, for which there are standards (EN 61000-3-3 and EN 61000-3-11) listed under the EMC Directive that limit emissions of voltage fluctuations and flicker for equipment up to 75A. The same principles apply to limiting the emissions of voltage fluctuations and flicker into AC and DC electrical power supplies at any voltage (see 6.4.2).

As Figure 6AU shows, there is always impedance in an electrical power distribution network; so any fluctuating currents in a network will cause the supplied voltage to fluctuate accordingly. 'Flicker' is the term used for rapid fluctuations in a supply voltage.

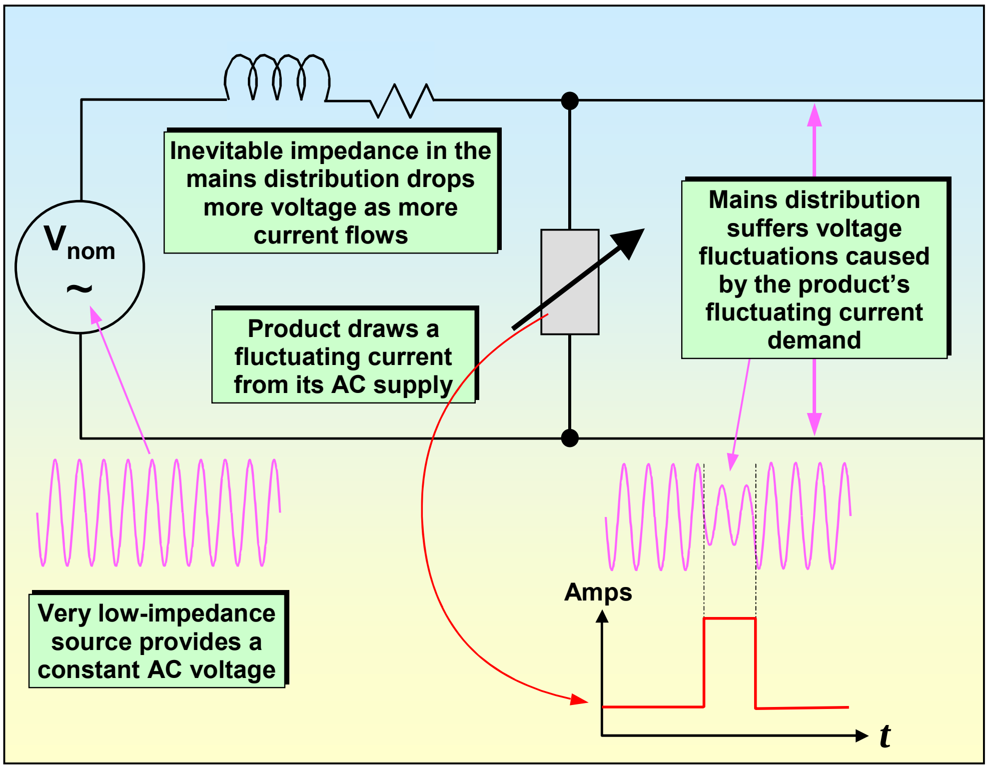


Figure 6AU How voltage fluctuations and flicker occur in the electrical supply

6.4.2 The standards and their limits

Historically, the main problem has been fluctuation in lighting levels, which can be very annoying to people and can even cause stress-related illnesses. Rapid fluctuations in lighting levels are known as flicker, and can quickly cause headaches, and even epileptic episodes in some people. The limits in the emissions standards are based on the human perception of the variations of luminous intensity from a mains-powered 60W filament light bulb – so they are not at all like the straight-line limits used by most other emissions tests.

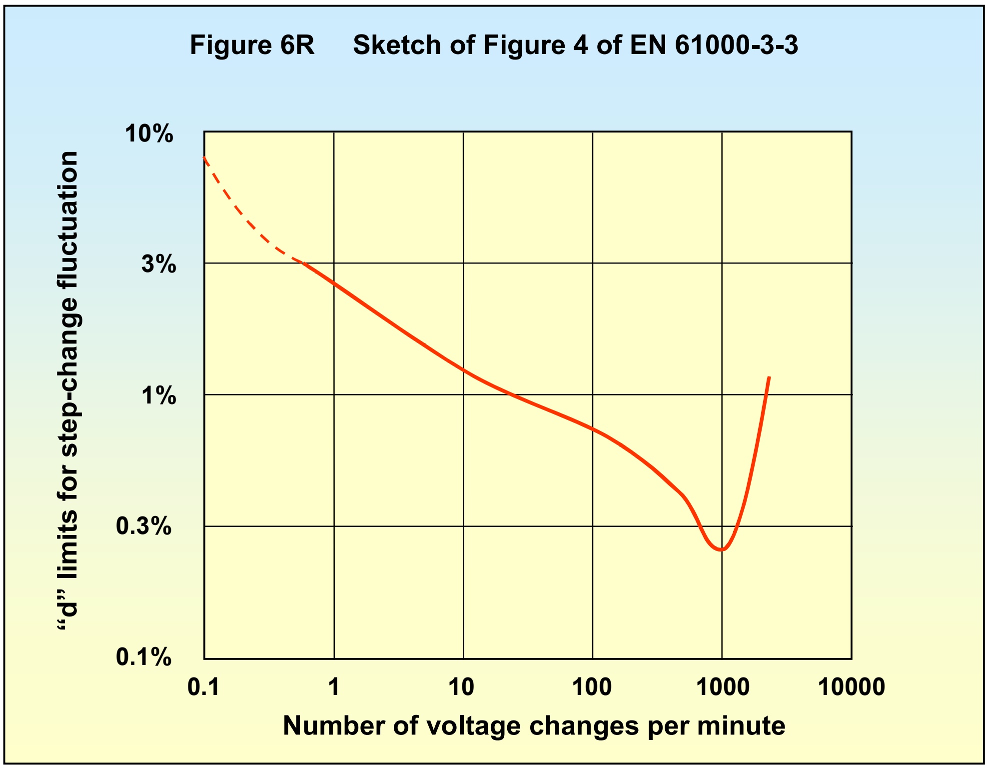


Figure 6AV An example of the emissions limits, from Figure 4 of EN 61000-3-3

Figure 6AV is an example for continuous 'square wave' load current variations; current fluctuations with different mark-space ratios will have different limits over the same range of frequencies. Irregular, transient and discontinuous current fluctuations will attract different limits still. Because of this complexity, all due to the psychometrics of human flicker perception combined with the time-constants of 60W filament bulbs, compliant measurements can only be made with a 'flickermeter' that uses digital processing techniques to determine pass or fail.

The standard for the processing involved in flickermeter measurements is IEC 61000-4-15. Low-cost test instruments are available for flicker measurements, for example the product shown in Figure 6AF combines the functions of measuring emissions of harmonics and voltage fluctuations and flicker. It is also possible to determine compliance in a rather rough and ready way by using calculations and/or simple test equipment. Those who are interested in DIY measurements of voltage fluctuations and flicker should read sections 7.5 and 7.6 in Part 7 of [6] – always taking the safety considerations in 6.2.1 fully into account.

Lighting flicker is mostly a problem for people reading or performing tasks illuminated by mains-powered filament lamps. Due to the 'smoothing' in AC-DC power converters, it hardly affects the illumination levels of mains-powered TV and computer screens.

In the modern world there are a great many things other than filament light bulbs that can be upset by voltage fluctuations and flicker on their mains electricity supplies. A particular problem is that dips in the mains voltage could cause a product's internal DC rails to drop below acceptable levels, causing errors, malfunctions or re-booting. Some discharge lamps will switch off due to dips, and not come back on again until they have cooled down sufficiently, which could take several minutes.

However, despite all this, and despite the fact that filament bulbs will soon not be legally available (for reasons connected with saving the planet from CO_2), the limits in the standards (EN 61000-3-3 and EN 61000-3-11) continue to be based on human perception of 60W filament bulb flicker.

Even where no legally mandatory or contractually-applied standards apply to products that connect to a DC or AC electrical power supply, there is still a good engineering practice requirement not to interfere with the operation of other devices, products, equipment, systems, etc., that share the same supply.

> Note: When talking to managers, always replace the phrase '*good engineering practice*' with '*practices that reduce warranty costs and financial risks*'. Of course in a properly managed company they mean the same thing, but the latter phrase expresses it in terms of the desirable financial outcome rather than the process by which it is achieved. Since it seems few managers care anymore about doing good engineering, but they all care passionately about saving money or making more of it, it is important for engineers to use language that will be understood.

6.4.3 Background to the suppression techniques

The standards measure the actual fluctuations in voltage in an electrical supply that has a specified impedance. In fact, what is really being measured (although indirectly) by their tests are the variations in the product's current demands from its electrical power supply – the variations in its load current.

So the design techniques for controlling emissions of voltage fluctuations and flicker centre on controlling the range and rate of variation in a product's supply current.

Note that some of the emissions standards permit greater fluctuations in a product's supply current, where the supply has lower impedance than usual. So sometimes it is possible to comply simply by specifying the characteristics of the electrical power supply that should be provided by the user. Of course, this must be reasonable – it would not be acceptable for the manufacturer of a coffee maker intended for domestic use, to specify that it must be connected to an industrial-strength 100A supply.

6.4.4 Reducing inrush current at switch-on

The inrush current at switch-on is a major cause of emissions of voltage fluctuations. The standards generally allow slightly higher values at switch-on (whether manual or automatic), and they generally do not apply any limits at all for the inrush currents during an uncontrolled power-up due to the resumption of mains power after an unanticipated interruption or failure of the mains supply.

Although the standards may not set limits for inrush following mains interruptions or failures, in practice it can be very important to limit them too. Consider the example of a branch of a mains distribution that is heavily loaded – an insulation failure somewhere on the branch will cause the overcurrent protection to trip, removing power to all the equipment.

But if the power is restored when all the loads on the network are switched on, their combined inrush currents can cause the overcurrent to trip again. It may be impossible to restart the mains power to that branch without going around and manually switching off many items of equipment, restoring the power and then going around switching them back on again one at a time. So unless automatic sequential mains switching is used (see later) on that branch, there could be significant benefits in limiting the inrush currents during uncommanded power-up events, even where not required by standards.

Most electronic equipment has a huge ‘spike’ of inrush current into their smoothing capacitors following their bridge rectifiers (see Figure 6AB), at the instant of switch-on. Even on power supplies rated at just a few watts, with normal current consumptions measured in tens of milliamps, the peak inrush current at switch-on can be tens of Amps, causing very high levels of voltage fluctuations at that instant.

However, flickermeters integrate voltage fluctuations over 10 millisecond periods, whilst charging the smoothing capacitors of low-power equipment might only takes a few tens of microseconds, so the very high but very brief voltage fluctuations caused by capacitor charging get averaged over 10ms and are generally measured as having much lower values.

Where the initial charging of capacitors would cause emissions to exceed the limits, Figure 6AW shows one technique for limiting the inrush current. At switch-on the relay contacts are open and the capacitors charge up more slowly, their peak charging currents limited by a suitable power and voltage-rated series resistor. After a short time (usually under two seconds) the capacitor should be substantially charged and the relay contacts (or triac) switched on to ‘short out’ the series resistor.

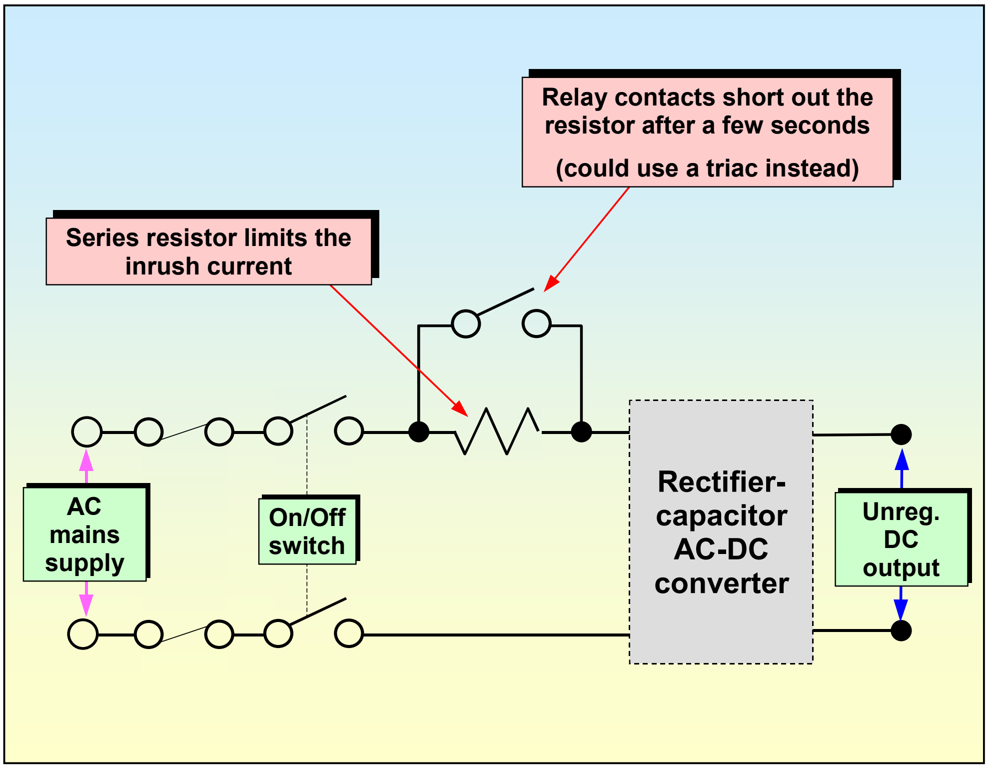


Figure 6AW An example of a technique for reducing the inrush current

In many real products, the electromechanical relay contacts shown in Figure 6AW are replaced by a triac. But triacs are not short-circuits, and in some applications their heating and/or emissions of noise around the zero-crossings might have to be dealt with.

For electronic loads it is usually very important to ensure that the load is not permitted to begin to operate until the unregulated voltage on the smoothing capacitor has ramped up to within specifications for correct operation of the load. In microprocessor circuits this is usually done with a combination of ‘power-on reset’ and ‘voltage monitor’ devices that hold all the devices in reset mode until they are both satisfied that the power supply conditions are acceptable. Many switch-mode controller ICs have soft-start functions, which also help reduce inrush currents at switch-on and so reduce emissions of voltage fluctuations.

Analogue circuits might need to actually monitor the DC power characteristics and switch DC power to the circuits using relay contacts, SCRs or power transistors. For example, power amplifiers that are connected to their voltage rails as they slowly ramp up to limit inrush currents, can often suffer instability and output false signals that might even damage their output transducers. In the case of audio systems, the false output signals can cause very loud and unpleasant noises.

Figure 6AX shows a similar scheme to Figure 6AW, but this time the relay contacts (or an SCR or power transistor) are installed after the bridge rectifier and before the capacitor, in the raw unregulated and unsmoothed DC supply. The operational principles are just the same.

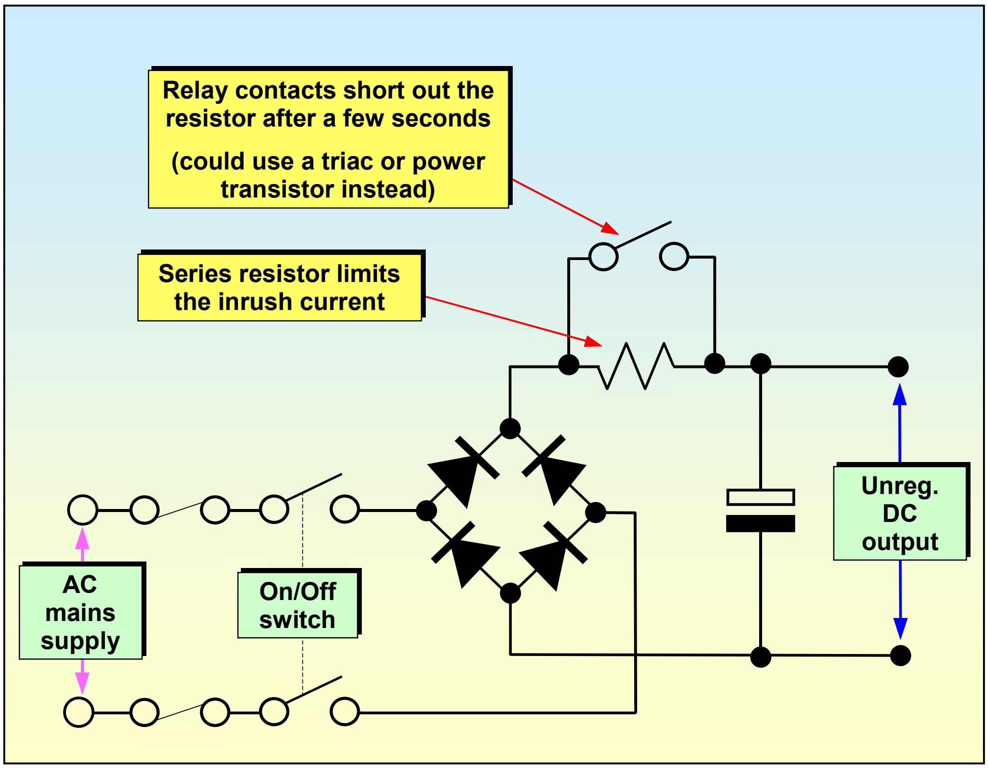


Figure 6AX Another example of a technique for reducing the inrush current

Figure 6AY shows the scheme of Figure 6AW with a negative temperature coefficient thermistor (or 'NTC') replacing the series resistor. NTCs are temperature-dependent resistors with a non-linear relationship between temperature and resistance. When they are at ambient temperature they have quite a high resistance, allowing the smoothing capacitors to charge up slowly and limiting inrush current. As charging current flows in their high resistance they heat up, and when they are hot enough their resistance very rapidly changes to a low resistance value. The NTC should be carefully chosen so that the flow of the normal load current through it is sufficient to keep it hot enough for it to remain 'switched on'.

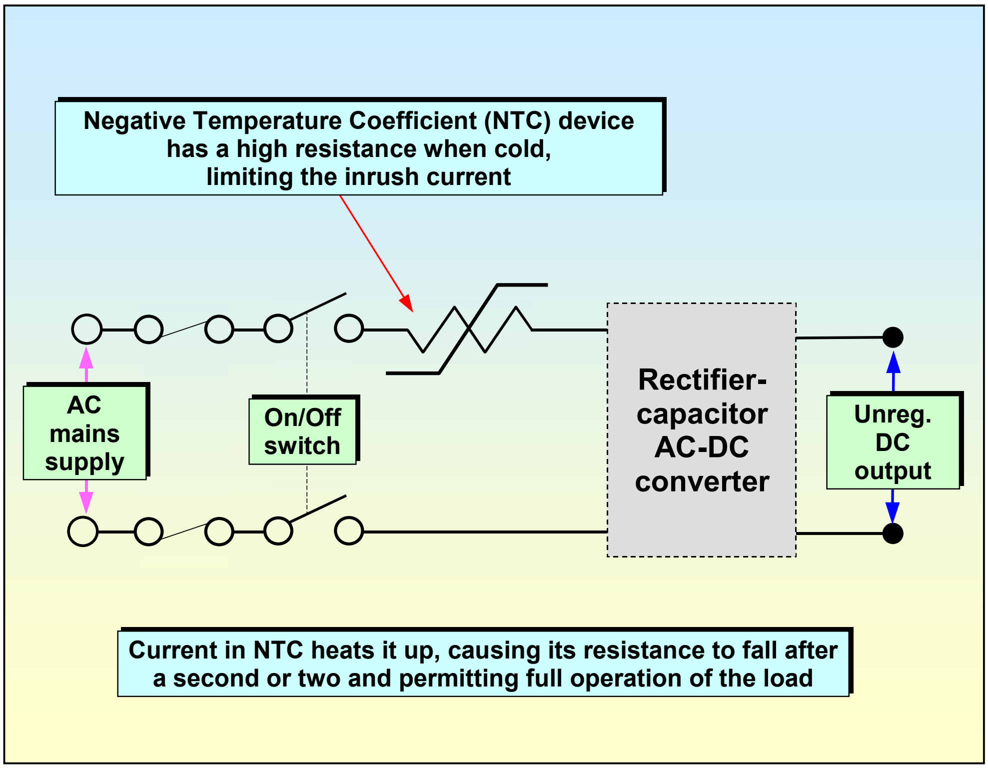


Figure 6AY Reducing inrush current with an NTC

NTCs run hot all the time when in normal operation – so it is necessary to design appropriate precautions to make sure they don't damage PCBs or nearby components, or melt a hole in a plastic enclosure. It is also important that they are protected from accidental contact so as not to burn service engineers who might have the covers removed.

It takes a number of seconds for an NTC to cool down by enough for its high-resistance state to be re-established, so if the power goes off and returns quickly they will not limit the inrush current.

Some designers have been known to take advantage of the use of inrush current limiting techniques to specify bridge rectifiers with lower surge current ratings to save space and reduce costs. When inrush is limited by NTCs, they can be caught out because short interruptions in the mains power – or users who switch off and then on again – can defeat the NTC, permanently damaging the bridge rectifier due to the peak inrush currents being much higher than it can handle.

Similar problems can occur for the inrush limiting schemes shown in Figures 6AW and 6AX, unless they are appropriately designed so they cannot be defeated by brief interruptions in mains power.

Large AC motors, transformers and other inductive loads can draw larger than normal inrush currents for many cycles after switch-on – when switched on at some point in the mains cycle that is not close to the voltage peaks. Switching on at zero-crossing causes the largest inrush currents.

The issue is the establishment of the load's steady-state AC magnetising current, which if allowed to overshoot by too much could saturate the magnetic circuit. Magnetic saturation reduces the impedance of the load to that of the resistance of the winding, effectively short-circuiting the mains supply and causing huge inrush currents. Figure 6AZ shows some examples of inrush currents in inductive loads.

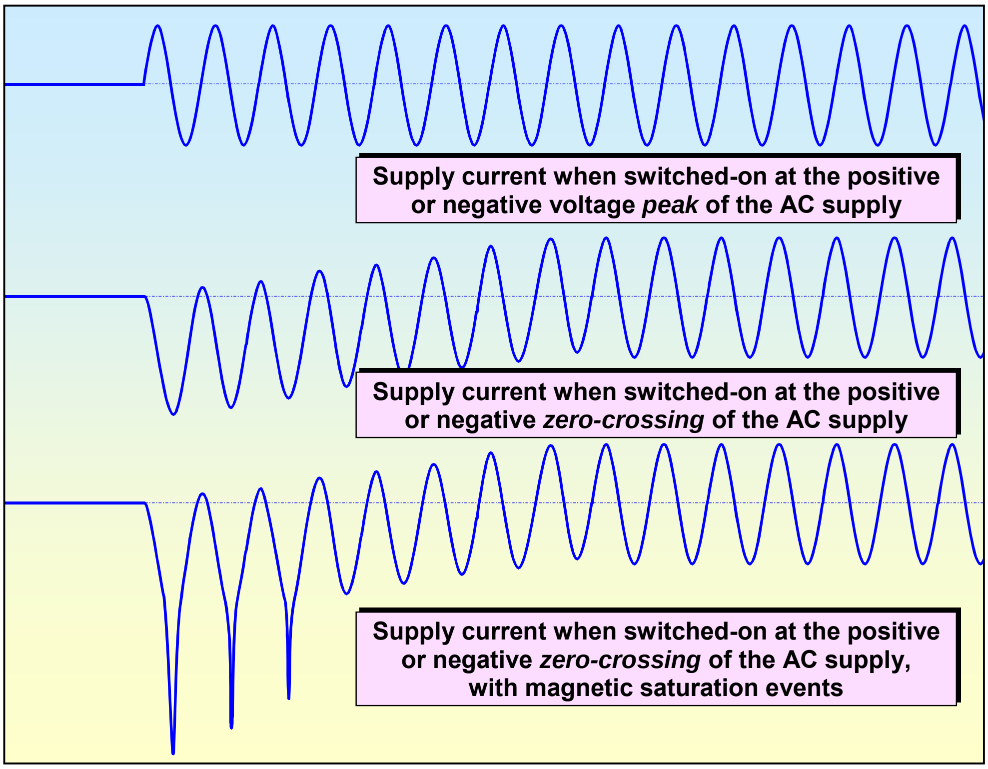


Figure 6AZ Examples of inrush currents into inductive loads

One obvious technique for reducing switch-on surge in inductive loads is to ensure that power is only applied at the instant when the AC supply is near a positive or negative voltage peak, and some manufacturers make triacs with the appropriate controls.

AC motors draw more current the greater their 'slip speed', so while they are spinning up their loads they can draw more current than is allowed by the emissions standards or is desirable for the power distribution network. Such motors and similar loads can use 'soft-start' techniques, which use phase-angle-controlled triacs with automatic ramping of their phase angle. Over several seconds, their conductive phase angle increases, increasing the average RMS voltage while the load slowly builds up to speed, until the full working voltage is reached.

It can also help meet emissions limits if the load current is reduced slowly instead of abruptly stopping at the instant of being switched off. Soft-start phase-angle controllers can easily be designed to also function as a 'soft-stop', slowly ramping the phase angle down to zero when the motor (or other load) is switched off.

There are many suppliers of soft-start/soft-stop SCR modules that can be added to industrial motors and other products, replacing their ordinary on/off switches. But few/none of them seem to be fitted with filters to attenuate the harmonics and RF emissions from the SCRs during ramping. The assumption seems to be that any interference will only be for a second or two once in a while, but whilst this might be permitted by emissions standards, it might not be acceptable in all applications for functional reasons.

Where products have their power or speed controlled by phase-angle SCRs, or similar methods using IGBTs, soft-start and soft-stop functions can easily be designing in. A very simple way to do this is to control the power with a rotary potentiometer that has the on/off switch mounted on the same shaft, so the potentiometer has to be turned down before it can be switched off, and when switched on it is always at low power.

Where several items of equipment are assembled in one unit, cabinet or system with a single master on/off power switch, their inrush currents will all occur simultaneously. The result can be emissions of voltage fluctuations that exceed the limits in the relevant standard, and/or practical problems of interference with other equipment. Sometimes, as mentioned earlier, the combined inrush currents will cause the overcurrent protection (fuse or circuit breaker) to open, although in such cases it is often possible to fit time-delay fuses or inrush-resistant circuit breakers.

One way to deal with the problem of simultaneous inrush currents is to power each item of equipment via a time-delay relay or contactor, a common industrial component, with the time delays all set to different values. Some manufacturers also offer mains distribution products ('socket strips') with built-in sequential switching, such as the units shown in Figure 6BA.

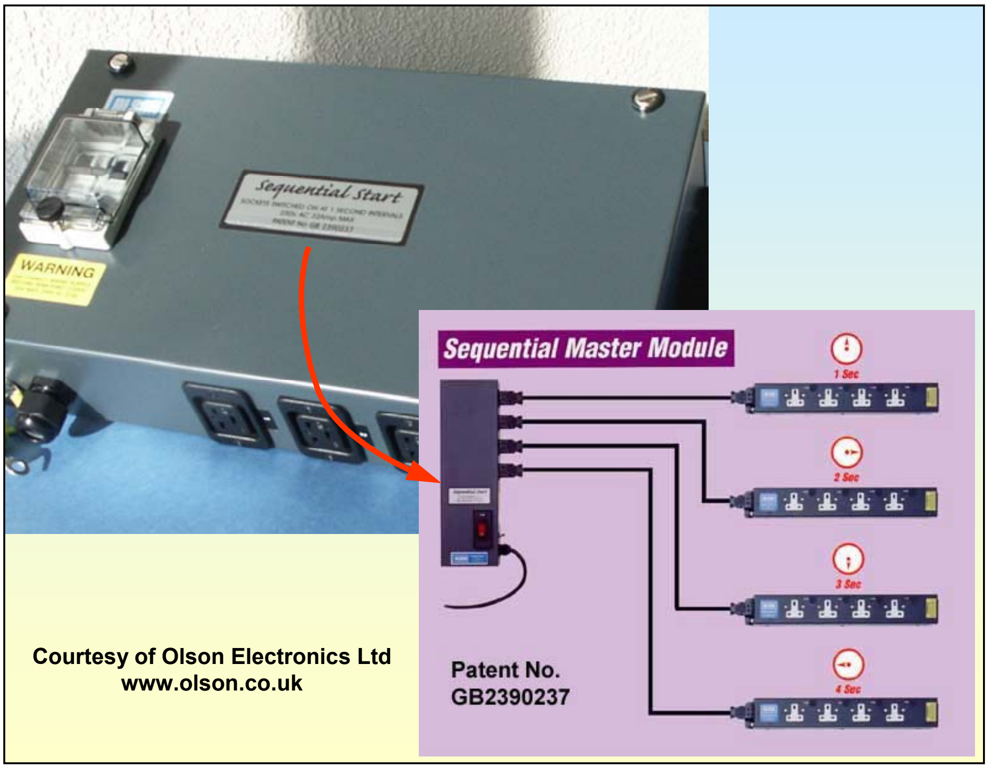


Figure 6BA Examples of sequentially-switched mains socket-strip

6.4.5 Reducing emissions of voltage fluctuations caused by varying AC loads

Time-proportioning on/off control is often used to provide power control of resistive loads such as heaters by varying the mark/space (on-time/off-time) ratio. It is sometimes called 'bang-bang control' because the load is switched on and off repetitively, and is a rather crude technique that is very unkind to the voltage that the distribution network supplies to other loads.

One way of reducing emissions of voltage fluctuations and flicker from bang-bang controlled loads is to split the load into two or more smaller loads, and switch them at different times, so there is a faster rate of smaller voltage fluctuations. Figure 5 of EN 61000-3-3 and its associated text gives some guidance on this technique. Another method is to use the soft-start/stop techniques described in 6.4.4.

It is very important to avoid using bang-bang control (or any other kind of power control) that results in voltage flicker in the range 100 to 2000 voltage changes per minute (1.7 - 33Hz) because this is where the human eye is most sensitive to lighting flicker from a mains-powered 60W filament bulb, so the flicker limits are much more severe, as can be seen in Figure 6AV.

However, the very best suppression of emissions of voltage fluctuations is achieved by replacing bang-bang control with some type of continuous power control, such as variable transformers or phase-angle-controlled triacs (or similar IGBT circuits). Variable transformers are a traditional remedy for controlling the AC power delivered to heating and similar loads, and although they are large, heavy and expensive they are also reliable, rugged, have no emissions, have very high levels of immunity, and when fitted with motors can be electronically controlled by analogue signals, or data from a computer.

All electronic circuits have other EMC problems, such as emissions of harmonics (see 6.3) and RF conducted and radiated noises, and also have EMC immunity issues. But – providing their maximum rate of change of power is set low enough – they will not cause significant emissions of voltage fluctuations or flicker.

6.4.6 Reducing emissions of voltage fluctuations caused by varying electronic loads

In a rectifier-capacitor AC-DC converter, increasing the size of the smoothing capacitor (unregulated storage capacitor) will reduce the ripple voltage caused by load fluctuations on the DC rails, and hence reduce emissions of voltage fluctuations and flicker. 'Supercapacitors' are now available with values measured in Farads, and peak current ratings measured in kA, which can provide huge energy storage and 'smooth out' the load's current demands very considerably.

Unfortunately, as discussed in 6.3, increasing the size of the smoothing capacitor increases the emissions of harmonics currents into the AC supply, making it more likely that 'power factor correction' will be required.

Adding series inductors to reduce harmonic emissions, as shown in Figure 6AK, will help 'round off' the edges of any very sudden fluctuations in load current, but the effect will probably be too small to have a significant effect on flickermeter measurements because they are integrated over 10ms periods.

Sometimes it is possible to design electronic loads so that their fluctuating current demands are not as severe, for example using Class A or AB analogue power amplifiers instead of Class B.

An excellent method of reducing emissions of voltage fluctuations and flicker due to variations in electronic loading, is to use an 'active PFC' boost circuit, such as described in 6.3.6 and Figure 6AN.

Active PFC controllers have a typical response time-constant to variations in load current of about 500 milliseconds, so for periods of time shorter than this they act like constant-current sources. Ripple voltages on the smoothing capacitor, due to variations in the current drawn by the electronic load, do not feed directly into mains current – they are smoothed out by the active PFC's time constant.

This helps reduce emissions of voltage fluctuations and flicker, but it is important to realise that the same slow response will cause ripple voltages on the unregulated DC rail to increase, and it may be necessary to either increase the value of the smoothing (storage) capacitor (C1 in Figure 6AN), or else design the electronic load to cope with the increased ripple.

The slow response time of the active PFC controllers can have another downside that needs to be guarded against. If the electronic load on the unregulated DC rail suddenly reduces by a large amount, for longer than 1 second, the active PFC will keep supplying the same current for at least 500ms and only then start to reduce it – so the voltage on the smoothing capacitor could rise so much that it would be damaged by overvoltage.

To prevent this from happening, active PFC controllers sense their output voltage and abruptly switch off their current before the capacitor's voltage rating is exceeded. Obviously, suddenly switching off the mains current causes a significant emission of voltage fluctuations and flicker – so with some types of loads and values of smoothing capacitance, active PFC circuits can *increase* emissions of mains voltage fluctuations and flicker. The solution is to design the circuits and dimension the components (especially the value and voltage of the smoothing capacitor) to make sure this protection mechanism doesn't happen, at least during normal operation with the worst-case load variations.

Active PFC boost circuits can be designed to provide many benefits…

- Comply with harmonic emissions standards (see 6.3.6)
- Achieve 'universal' operation from 84 to 260V AC rms, and DC to 400Hz, helping to sell the same product world-wide (only need ship it with appropriate mains lead)
- Reduce emissions of voltage fluctuations and flicker by acting as a constant current source
- Improve immunity to voltage variations, fluctuations and dips in the electricity supply (see 6.5)

6.5 Immunity to Power Quality issues

6.5.1 Introduction to power quality

Electrical power supplies, whether AC mains or DC (e.g. 48V for telephone exchange ('central office') equipment, blade servers, etc.), suffer from many types of high-frequency EM disturbances:

- Surges, spikes and other transients
- Bursts of transients

- Electrostatic discharge
- Common-mode (CM) and differential-mode (DM) RF voltages and currents

– all of which occur in all conductors (cables, wires, chassis, enclosures, PCB traces, etc.) and are dealt with by the earlier parts of this series [13] – [17].

This section covers 'non-RF' electrical power quality issues, at frequencies from DC to about 150kHz. It will generally refer to AC mains supply issues, but DC supplies suffer from many of the same power quality phenomena, so it is relevant for them too.

[33] is a guide to Power Quality issues from the point of view of systems and installations engineers. This article is aimed at product designers, but the descriptions of the various power quality phenomena in [33] will be just as useful for them too.

The quality of the delivered mains power can be measured in a variety of ways, but proper tests use instruments that comply with IEC 61000-4-30. The use of standardised and repeatable measurements can be an important issue when dealing with customer specifications or complaints – if both parties are measuring power quality in a different way there is endless scope for misunderstanding, wasted resources, and loss of customer goodwill.

In general, poor mains power quality generally causes more problems for electronic products when the real-life RMS mains voltage differs from the nominal supply voltage the product is expecting. So it is best to ensure that a product's mains input is set for the correct nominal voltage for the real-life mains supply. For example, products designed to run on 220V rms mains supplies have been known to overheat when run on 240V supplies in the UK. The official pretence that all of Europe has a common mains power rail at a nominal 230V rms does not help overcome this sort of problem.

The IET will sell you a wallchart listing the mains supply voltages (and many other details) for most of the countries in the world, and I have one of these. Unfortunately, it is pretty much useless because it only states the *official* specifications for nominal voltage and frequency and their tolerances, and they often differ from the real ones.

It is still quite easy to find rural areas of developed countries like the USA, Australia and Spain (to name just a few) where the normal range of mains voltages is much wider than the usual ±6% or ±10% specifications. In parts of rural Spain, during the late 1990s, the nominally 230V mains supply would fall to as low as 180Vrms during the afternoon. Parts of Australia are still supplied by single-wire mains, with the neutral current being returned through the soil, giving very poor power quality indeed.

The situation is often worse, or at least worse over wider areas, in less developed countries. For example, in 2005 in Nigeria the effective RMS value of the mains commonly varied from 140 to 300V. In India many people have their own standby electricity generators, so that when power fails they can keep operating. But when the mains power returns there is very little load on the distribution network and the nominal 230V mains voltage supply can rise to well over 300V for several seconds.

Small distribution networks with limited generation capability are very prone to significant power quality problems. An extreme example occurred on a North Sea oil exploration platform in the 1970s where the 230V mains supply from its 10MW diesel generator had frequency variations of about ±90%, lasting for several seconds. When the 10MW drilling motor was switched on, the diesel generator almost stalled and the mains voltage dropped to about 50V at about 5Hz. When the drill motor was switched off, the diesel generator would overspeed, and the mains voltage would rise to about 430V with a frequency of about 95Hz. This would happen several times each day.

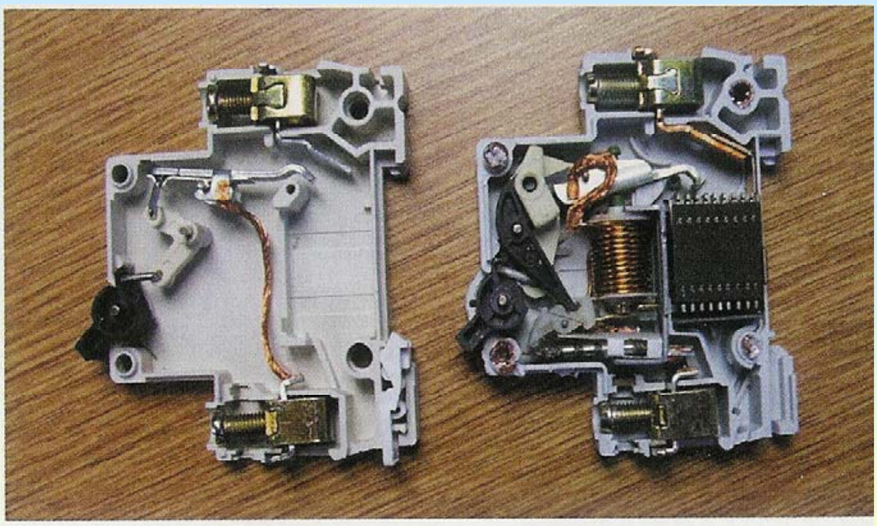
A big problem for offshore and marine vessels these days is the use of electric thrusters, which are variable-speed AC motor drives often rated at 100kW or more, which cause their mains supplies to suffer severe distortion, often as much as 30%. For many more details on power quality phenomena see [31] and [33].

6.5.2 Important Safety Considerations for Mains Circuits

All components and wiring used in mains circuits must be rated for safe use on the highest anticipated mains voltage, including overvoltages and surges.

There are appropriate safety standards that should have been applied by component and cable manufacturers, who should make third-party Safety Approvals certificates available to their customers. Customers should check that the certificates are valid, by contacting the issuing authorities, and not use components that have anything suspicious about any details.

Stringent measures should be taken to avoid using counterfeit components, like the counterfeit circuit breaker shown in Figure 6BB alongside a genuine one, a photograph used to help promote the new "Electrical Industry Installation Charter" scheme launched by BEAMA, EDA, ECA and SELECT. People have died and premises burnt down because these safety precautions were not taken – make sure it is not your product that is the cause.



Counterfeit and genuine circuit breakers – the difference is potentially fatal

Photograph used in the launch of the "Electrical Industry Installation Charter" by BEAMA, EDA, ECA and SELECT, in "Electrical industry fights fakes" by Lorna Sharpe, IET Engineering & Technology Magazine, 12 April - 25 April 2008, Page 3, www.theiet.org/engtechmag

EIEMA also have an anti-counterfeiting scheme, www.eiema.org.uk

Figure 6BB Comparing a counterfeit circuit-breaker with the one it was imitating

6.5.3 Overvoltages (swells)

Swells are when the supply voltage is higher than normal limits, for a while (e.g. a few seconds), and are generally assumed to have very slow rise and fall times, such as a few seconds. They can exceed the normal tolerance of the mains supply voltage and cause overvoltage or overheating damage, and/or can cause surge protection devices (SPDs) to overheat and be damaged.

A relevant immunity test standard is EN/IEC 61000-4-11, and a guide on its application is included in [7]. Some low-cost but non-compliant tests that can be done by anyone with sufficient competence are described in [6]. [7] includes more detailed descriptions of 'swell' phenomena, including what causes them, what they can affect and how.

To protect products from swells, it is best to simply design (or choose) AC-DC power converters that have mains input circuits that use higher-voltage devices and circuits, so that they operate within their rated limits during anticipated swells, without damage for as long as the swells last. Their bridge-rectifiers and off-line switching power FETs might need to be rated up to 1200V or more, and their unregulated storage capacitors up to 600Vdc or more.

Before the days of switch-mode power converters, a range of electronic products were sold worldwide and proved to be very reliable despite the very wide range of voltages and waveforms that they were powered from. They used linear power supplies in which the mains transformers had multiple tappings with automatic tap selection, as shown in Figure 6BC. It is still a viable technique these days, especially for larger products, systems or installations.

The primary winding of the transformer in Figure 6BC has to cope with all of the power quality problems discussed in this section, so will need insulation suitable for the swells and distortions; enough turns to ensure that swells, low frequencies, and any DC components do not saturate the core, causing excessive magnetising currents and overheating; and a core size large enough to prevent overheating due to harmonic distortion of the mains waveform.

Where it is not feasible to design (or choose) mains power converters with a swell capability that will cope with the worst-cases that can occur in some countries and/or situations, the mains input circuit should be protected from damage during such events.

How it is protected depends upon the application – whether the product must keep functioning; whether it is acceptable for it to stop during the swell but restart automatically later on, or whether it is acceptable for a fuse or circuit-breaker to open, requiring manual intervention to restore correct operation.

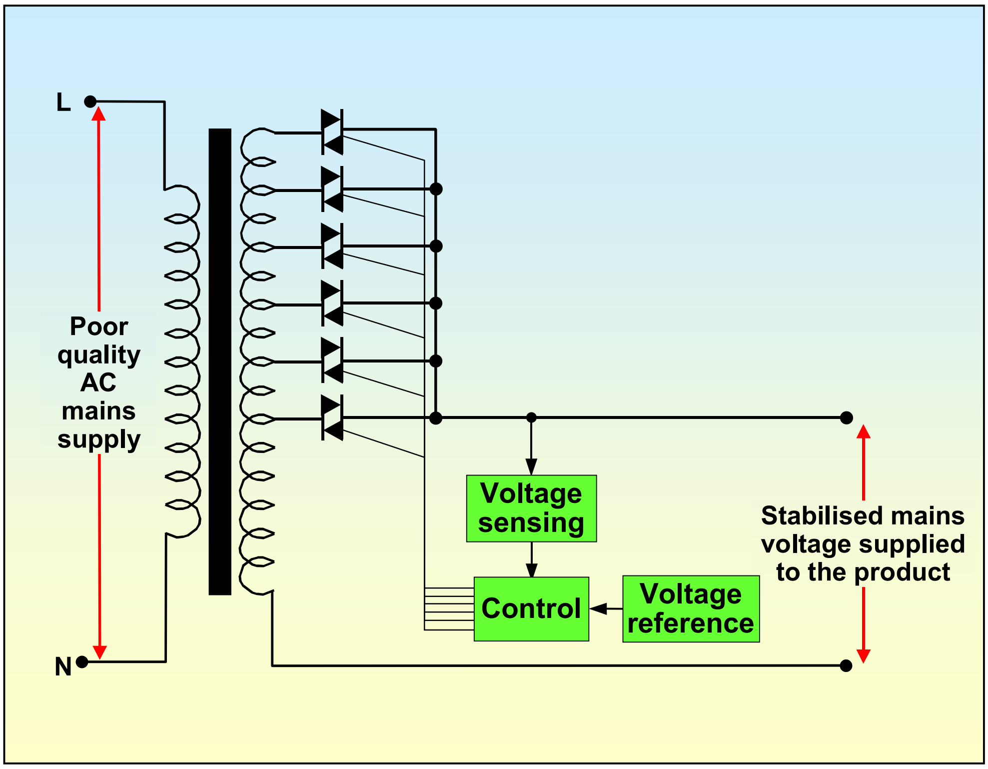


Figure 6BC Multi-tapped mains transformer with automatic tap selection

All of these options could use an overvoltage protection device (OVPD) such as a metal oxide varistor (MOV) or gas discharge tube (GDT), described in section 3.5 of [15], to protect the product's mains input devices from damage. A series element is employed between the mains supply and the OVPD, as shown in Figure 6BD, to limit the power dissipated in the OVPD.

Alternatives to using OVPD devices such as MOVs or GDTs are shown in Figures 3AG and 3AJ of [15], and might be useful.

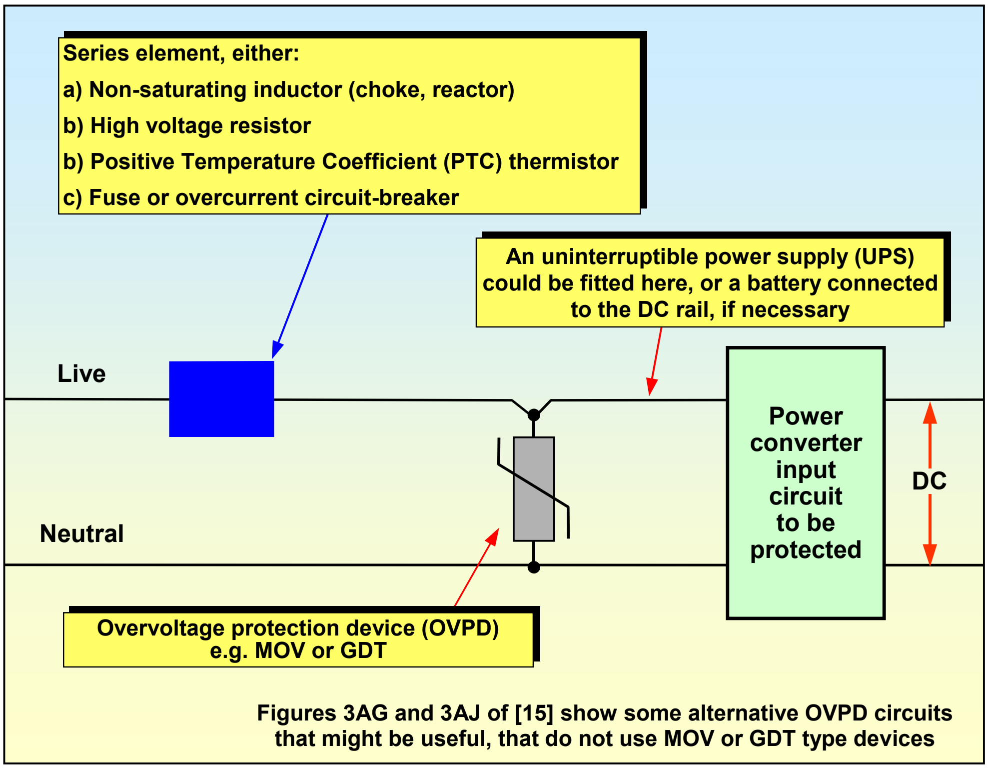


Figure 6BD An example of 'swell' protection

There are several choices of component, each with different design compromises:

- An inductor (choke) that does not saturate

- A resistor
- A PTC (positive temperature coefficient) thermistor
- A fuse or circuit-breaker

The inductor and resistor, when used with an MOV type of OVPD, ensure that the mains supply peaks are clamped (clipped) below the maximum level that the power converter input circuit will withstand, so the product keeps on functioning during the swell.

Inductors provide an impedance that limits the current, and those used for this purpose in industrial applications are often called 'reactors'. Resistors must be high voltage surge/pulse rated types, and may need to be 'fusible' types that open-circuit safely if overloaded beyond their ratings.

Power dissipation in the series elements and OVPDs are serious concerns, and have safety implications. Inadequate ratings will result in short product life, dissatisfied customers and increased warranty costs – even if they do not result in smoke and fire hazards. Section 3.5 of [15] discusses surge protection, and even quite small SPDs can handle very large pulses of transient power lasting a few microseconds. However, here we are talking about overvoltages that can last for several seconds, probably comparable with the thermal time-constant of the device itself, so the OVPDs must be rated for continuous power dissipation at the levels expected during the swells.

Where capacitive energy storage is used instead of OVPDs (see Figures 3AG and 3AJ of [15]) the capacitor value must be large enough for it not to suffer overvoltage damage due to absorbing the energy of the swell. Supercapacitors with values measured in Farads, might be suitable, but batteries are generally unsuitable because they cannot handle the very large charging currents.

A big advantage of using capacitive energy storage instead of OVPDs, is that there is less thermal cycling and so it can be easier to design for a longer, more reliable lifetime (taking account of the propensity of electrolytic capacitors to dry out and lose capacitance over time, especially if they are operated at high ambient temperatures).

SAFETY NOTE: Protection must be provided for if/when the OVPD fails low-resistance, so there should always be a co-ordinated overcurrent protection using a fuse, OPTC thermistor or circuit-breaker as well as the inductor or resistor. This safety feature is not shown on Figure 6BD.

The PTC thermistor, fuse or circuit-breaker, used with MOV or GDT types of OVPDs, will remove the power from the equipment during a swell that would otherwise exceed the ratings of the power converter.

PTC thermistors are often called 'resettable fuses' – their resistance increases suddenly when they heat up beyond a certain temperature, removing the mains power from the OVPD and from the power converter. When they cool down, below the critical temperature their resistance suddenly falls so they allow the full mains current to flow once more.

When PTC thermistors, fuses or circuit breakers are used and the product does not have a UPS or battery with sufficient energy storage, it must be acceptable (e.g. safe) for the product to stop in an uncontrolled manner. For some applications it will be acceptable for the product to start up again immediately upon the replacement of the fuse or resetting of the circuit breaker, whereas some others will require that a manual restart is also employed (see 6.5.12).

6.5.4 Frequency variations

A relevant immunity test standard is EN/IEC 61000-4-28, and a guide to its application is included in [7]. This guide also describes the 'frequency variations' phenomenon, including what causes it and what it affects and how. Obviously, DC supplies do not suffer from variations in frequency.

Mains frequency variations can cause problems for circuits that rely upon the mains frequency for timing, and large frequency drops can cause problems for mains transformers, direct-on-line (DOL) AC motors, relays, solenoids and contactors. The problems caused include magnetic saturation, excessive mains currents and overheating. Saturation also has the effect of reducing the transformer ratio, causing electronic loads to run on a lower DC voltage than would be expected from the RMS value of the mains voltage, possibly malfunctioning as a result (also see 6.5.8). Some of these problems have occurred when equipment designed for 60Hz mains (e.g. USA) was operated on 50Hz (e.g. Europe).

Design solutions for timing accuracy include using stable reference oscillators, such as the 32kHz oscillators that are standard for digital wristwatches. For the highest precision, products can use off-air atomic clock time signals, from GPS (satellite), MSF (terrestrial, Rugby, UK) or DCF (terrestrial, Frankfurt, Germany) for example. I have an inexpensive wristwatch that corrects its own time using off-air terrestrial broadcasts, so these solutions are clearly low-cost and small.

For inductive components such as transformers or AC motors, it is best not to design them to run close to saturation on their nominal supply. Use larger cores and/or more turns on their windings to reduce the core flux density so that they still operate out of saturation, and their magnetising currents are not excessive, during anticipated frequency variations.

For relays, contactors and solenoids with AC coils: choose types that have lower ‘drop-out’ or ‘hold-in’ voltages, as they should cope better with lower frequencies. Typical low-cost relays can drop-out at 78% of nominal supply, whilst better types will remain held-in down to 50% or less. ‘Coil hold-in’ devices (e.g. ‘KnowTrip’, ‘Coil-Lock’, etc.) can also be used, some of which claim to keep coils energised when the supply is as low as 25% of nominal. They appear to power each coil individually from a small AC-AC converter with capacitor energy storage, essentially a small UPS (see 6.5.11) that should suppress the influence of frequency variations.If we are prepared to make greater changes, we notice that the typical rectifier-capacitor AC-DC rectifier used as the front-end of switch-mode power converters is insensitive to mains frequency (as long as the storage capacitor is large enough) – so we can replace all off-line mains transformers with switch-mode power converters, AC-AC or AC-DC as appropriate. We can also power all AC motors and AC coils from switch-mode AC-AC inverters, such as UPSs (see 6.5.11) instead of direct-on-line (DOL), or replace them all with DC motors and DC coils powered from rectified mains.

All solutions that involve ‘adding electronics’ can increase the harmonic emissions into the mains so might need power factor correction (see 6.3 in [36]), and they can increase other emissions and suffer immunity problems that the originals did not suffer from.

6.5.5 3-phase unbalance

A relevant immunity test standard is EN/IEC 61000-4-27, and a guide to its application is included in [7]. This guide includes descriptions of the ‘3-phase unbalance’ phenomena, including what causes it and what it affects. 3-phase unbalance can be due to voltage and/or phase differences between the three mains phases, and unbalanced loading or faults in the mains distribution network cause them. Obviously, DC supplies do not suffer from such problems.

Unbalance causes big problems for larger three-phase motors, which can destroy themselves quickly (and expensively) when they lose a phase even momentarily due to a fault in the mains distribution network. Industrial control manufacturers guard against this by using special ‘motor control contactors’ (MCCs) (see 6.5.12) that detect excessive phase unbalance (and other potential problems, such as undervoltages, see 6.5.8) and remove power from the motors to protect them.

As for frequency variations above, it is also possible to overcome phase unbalance problems by powering three-phase AC motors from switch-mode inverter drives (instead of DOL), or replace them with DC motors powered from rectified mains.

6.5.6 DC in AC supplies

This is not often a problem for modern LV supplies, because the adoption of harmonic emissions standards that prohibit half-wave rectification (in most cases) have reduced the amount of even-order harmonics (hence DC) in the mains networks. But some process plants use high-powered half-wave rectifiers, maybe rated up to 0.5 MW or more, which distort the local distribution by adding a DC component (actually, even-order harmonic distortion, see 6.5.10) to it.

This problem has exactly the same deleterious effects as the low frequency mains discussed in 6.5.4, and the solutions are the same too.

6.5.7 Common-mode (CM) low-frequency voltages

A relevant immunity test standard is EN/IEC 61000-4-16, and a guide to its application is included in [7]. This guide includes descriptions of the phenomena, including what causes it and what it affects, in very much greater detail than this article does.

Currents flow out of equipment and their interconnecting conductors and into their safety earth/ground via a number of routes, including capacitive and inductive stray coupling, and also due to any ‘Y’ capacitors in their mains filters that are connected between phase or neutral and earth.

Equipment operation produces currents from DC to several tens of kHz, depending on the equipment, but the dominant frequencies are usually at the mains frequency (50 or 60Hz) and its harmonics. Insulation breakdown and similar earth-faults in equipment and mains distribution cabling also inject mains (and its harmonics) currents into the earth (ground). Surge protection devices (SPDs) connected to earth (see [15]) also inject currents into earth during their operation, and in the case of ‘crowbar’ devices, such as gas discharge tubes, they continue to inject a ‘follow-on’ current for some time after the surge is over, at least for the remaining part of the mains cycle before the next zero-crossing.

Because of the impedance in the earth (ground), all these currents create voltage differences between the earths (grounds) of items of equipment that are connected to different points in the earth structure of a site.

These voltages appear as CM ‘earth/ground noise’ voltages on their interconnecting conductors (mains, signals, data, control, etc.), with continuous voltages generally in the mV-Volts range.

Earth-faults and SPD operation (and its follow-on) in the LV mains distribution can create up to the full mains voltage, for up to a few seconds, and similar events in the MV or HV distribution networks can create kV of earth/ground noise, for up to a few seconds. The designers of the MV and HV networks are keenly aware that most equipment intended to be powered from the LV mains supply will not survive CM voltages of much more than a couple of kV – and that their failure would result in severe safety problems such as fire and electrocution – so they design their networks to provide the necessary protection. However, I do not know how reasonable it is to assume that such protection is provided in every country in the world, or in every offshore or marine installation.

Off-line mains power converters, whether linear or switch-mode, are protected against CM low-frequency voltages by complying fully with the relevant electrical safety standard, for example EN/IEC 60950 (information technology, IT, and telecommunications), EN/IEC 60335-1 (household appliances and portable tools), EN/IEC 60601-1 (medical equipment), EN/IEC 61010-1 (equipment for measurement, control and laboratory use), etc.

All of these have very similar requirements for dealing with the problem of short-term kilovolt CM disturbances, either:

a) Use a safety-earthed metal chassis with mains wiring and components insulated to safely withstand CM voltages of around 1500Vrms continuously, or…

b) Use double or reinforced insulation and a safety-isolating mains transformer all rated to safely withstand CM voltages on the mains of around 3kV continuously.

c) Use overvoltage protection similar to that described in 6.5.3.

The actual values of voltages vary from one standard to another, but it is important to realise that they are all based on certain assumptions, and one of them is that the equipment is used in a building in a city or other built-up area.

I had recently to deal with some agricultural electronic equipment that was situated in fields far from any building, and subject to whatever the local farmers thought was good electrical installation practices. The very visible damage to the mains input stages indicated that the mains power converters (purchased from a Chinese manufacturer whose data sheet claimed that safety compliance to domestic safety standards was ‘pending’ (but hey, they were cheap!)) were being subjected to much higher mains voltages than they could safely handle.

Even when the CM voltages on the mains input are dealt with safely, the CM noise can pass through the interwinding capacitance in the mains transformer, putting noise on the DC rails and possibly interfering with signals. This is not so much a problem for 50 or 60Hz, as it is a problem for higher-order harmonics or non-mains-related CM frequencies, and it can be dealt with by using a mains isolating transformer with increased CM attenuation. This can be achieved by adding an earthed interwinding shield, and/or by reducing primary-secondary capacitance by winding them on different limbs of the core.

Another technique is to use CM filtering on the mains supply, at the troublesome noise frequencies, but such filters can be large and costly due to the high mains currents and voltages.

Signal inputs and outputs can be designed to protect against CM low-frequency voltages, and for example, in professional audio it has been normal for many decades to use galvanic isolation transformers for inputs and/or outputs, often replaced these days by electronically balanced-and-floating input and output amplifiers. Electronic technologies that started out being used on the small scale, such as video, often suffer from CM earth/ground noise when connected to longer cables to form larger systems. The electronic designers never designed the input or output amplifiers to be able to cope with such disturbances, because they were not significant for small-scale systems.

Another approach is to use CM filtering to remove the troublesome earth/ground noises, for example in video systems it not unusual to use large and heavy CM hum chokes, usually purchased as ‘ground loop eliminators’. (The general assumption is that it is the earth/ground current flowing in the shield of the signal cable that cause the problem, hence the term ‘ground loop’ or ‘hum loop’ – but in fact it is easy to show with simple tests that it is almost always the CM voltage difference between the chassis of the two items of equipment that causes the noise, not the equalising current that flows in the shields of the signal, data or control cables. See [34] and [35] for more on this topic.)

But the above approaches will not cope with the high voltages that can occur from time-to-time, due to earth/ground-faults for example. The galvanically-isolating transformers used in professional audio in previous decades were not usually rated to safely withstand at least 1500Vrms continuously, although they could have been, and of course electronically balanced-and-floating amplifiers cannot withstand such voltages.

Ethernet transformers were traditionally rated to withstand 500Vrms, but to comply with EN/IEC 60950 for 'safety-earthed' equipment they should withstand 1500Vrms, and several manufacturers now offer such components.

Galvanic isolation rated at least at 1500Vrms continuously for use with 'safety-earthed chassis' equipment, or rated 3kVrms continuously for use with 'double insulated' equipment, is (in my view) the best way to protect against high levels of CM earth/ground voltage differences. (As was mentioned earlier, in some applications higher voltages than these may be necessary.)

Appropriately-rated signal or pulse transformers have already been mentioned as a possible solution, but there are many others including opto-isolators/couplers, fibre-optics, wireless (e.g. Wi-Fi, Bluetooth, ZigBee, etc.), infra-red, guided microwaves, free-space laser, etc. Fibre-optics are preferred for high-bandwidth signals/data, for reasons discussed in [14] although some of the newer wireless communication methods (e.g. 60GHz radio systems, UWB) might one day be able to handle several hundreds of MB/s using low-cost modules.

As an alternative to galvanic isolation, electronically-balanced amplifiers can be protected from overvoltages lasting a few seconds – providing loss of signal for that period is acceptable – by overvoltage protection similar to that described in 6.5.3. The typical method of protecting semiconductors connected to telephone cables that extend outside of a building uses high-voltage fusible resistors or PTC thermistors as the series elements, and SCR-based OVPDs. Many similar protection circuits exist to suit most common types of signal/data input and output circuits, and manufacturers such as Harris, Raychem (from Tyco), Texas Instruments, STMicroelectronics (used to be SGS-Thomson) and Bourns make a wide variety of special protection devices for use in them, often with special names. Chokes and fuses are also possibilities as series elements, but not commonly used.

The capacitive energy storage technique shown in Figure 3AG of [15] may also be suitable, and should be easy to design to be more reliable (although physically larger) than using OVPDs as shown in Figure 6BD.

Another alternative is shown in Figure 6R in section 6.2 of [36] – using reverse-biased transient-rated diodes or rectifiers to dump the excess energy into the 0V and/or power rails. Design issues that are not important for ESD, but *are* important for effectively handling high levels of CM earth/ground noise – are that the series impedance must be high enough to limit the current to what the PCB traces will handle, and there must be enough decoupling capacitance to prevent the DC power rail voltage from rising so high that the semiconductors relying on it for power suffer overvoltage damage.

6.5.8 Undervoltages (sags, brownouts, dips, dropouts and interruptions)

Undervoltages are caused by load current fluctuations, and by faults (and fault clearance) in the electrical power distribution. The series impedance of the distribution network means that load currents flowing in it create voltage drops. At various points in AC networks there are automatic tap-changing transformers that maintain the mains voltage within certain limits at their location, so in such networks it is usually the impedance 'downstream' of such devices that cause the undervoltages in response to the load current.

During an insulation fault, only the supply impedance limits the mains current taken from an ordinary single-phase mains plug, since the phase voltage collapses to 0V at the fault. In an ordinary house the currents can reach or exceed 1000 Amps RMS, and commercial and industrial sites with lower-impedance supplies can source correspondingly higher fault currents. (In fact, a common measure of supply impedance in the electrical distribution industry is the 'short-circuit current'.) This current flows until the protective fuse or circuit-breaker opens, which is generally well under a second.

Another cause of undervoltage is when the generated power is inadequate for the load. In AC networks the generator will slow down and eventually stop, so the undervoltage will also be associated with a significant fall in frequency (see 6.5.4). In highly developed and integrated networks, such as the UK's mains distribution network, loads (e.g. whole towns) will be shed (have their mains power cut off) before the frequency drops dramatically, to protect the generators from stalling, but this is not necessarily the case for all mains networks worldwide or for local or portable generators.

I have a 2kW portable generator with a circuit breaker on its output, but find that if I overload it, sometimes it keeps running and the circuit breaker opens, but sometimes it stalls without the circuit breaker opening.

Other causes include when the electrical power is switched off, for example by the operation of a protective overcurrent device (e.g. fuse, circuit breaker, etc.) in an LV or MV network, or by 'arc quenching' devices in the HV network in response to a flash-over arc caused by lightning strikes, and of course by load-shedding by the electricity network to protect their generators from damage by overheating or stalling when the power demand exceeds the generated supply.

Dips are short-term reductions in supply voltage e.g. 30% down for 10ms, 60% down for 100ms, see Figure 6BE, and usually happen abruptly as a result of inrush currents and/or insulation breakdown in other equipment connected to the same network. If the equipment is located nearby on the same branch of the network, the dip depth increases and when it exceeds 95% it is called a dropout or 'short interruption', different words for the same thing.

Dips, dropouts and 'short interruptions', see Figure 6BE, generally last less than 1 second, because inrush currents do not last that long, and because safety standards require equipment to be designed so that faults that draw excessive mains current will blow fuses or open circuit-breakers within 1 second, to minimise fire risks.

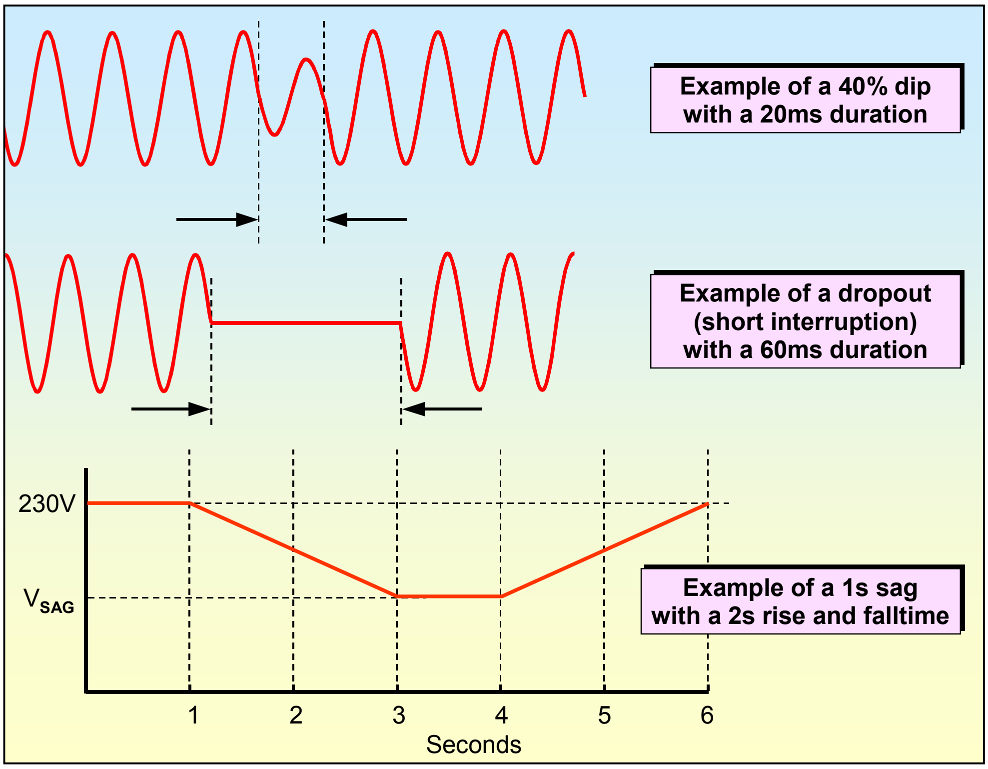


Figure 6BE Examples of types of undervoltage events

Note that the dips and dropouts in Figure 6BE started and stopped at the zero-crossings of the mains waveforms. This is the way they are tested by the EN/IEC 61000-4-11 test method – but not like real life, where the start and stop phase angles occur at random.

Sags (often called brownouts in the USA) are reductions in supply voltage that last longer than 1 second. They are generally caused by the accumulated loading on a network, and so are assumed by the tests in EN/IEC 61000-14-11 to build up and recover slowly, however in an industrial location where there is a very high-power load, real-life sags could occur instantaneously and recover abruptly too. A sag of 50% for 8 hours has occurred in the UK (see No. 21 in [12]), although this is unusual.

'Long interruptions' are caused by fault clearance can last for seconds, minutes, hours, days, etc., and are caused by removal of power to protect the network, or by damage to the network (e.g. trees falling across roadside power lines) in which case after an extreme weather event or earthquake they can last for weeks.

For equipment powered by DC, for example telecommunications equipment and blade servers at 48V, a test method for immunity to dips, dropouts and interruptions is EN/IEC 61000-4-29, which is described in a guide available from [7]. This guide also describes what causes these phenomena, what kinds of effects they have on equipment, and how to test to improve real-world reliability.

For AC supplies, a relevant immunity test standard is EN/IEC 61000-4-11, and a guide on it is also available from [7]. Some low-cost but non-compliant tests that can be done by anyone with sufficient competence are described in [6]. [7] includes more detailed descriptions of undervoltage phenomena, including what causes them, what they can affect and how. EN/IEC 61000-4-11 includes a large range of tests, only some of which are called up by the generic and product immunity standards listed under the EMC Directive.

Figure 6BF is a block diagram of a make-it-yourself tester that could even be designed to comply with the generator specification in EN/IEC 61000-4-11.

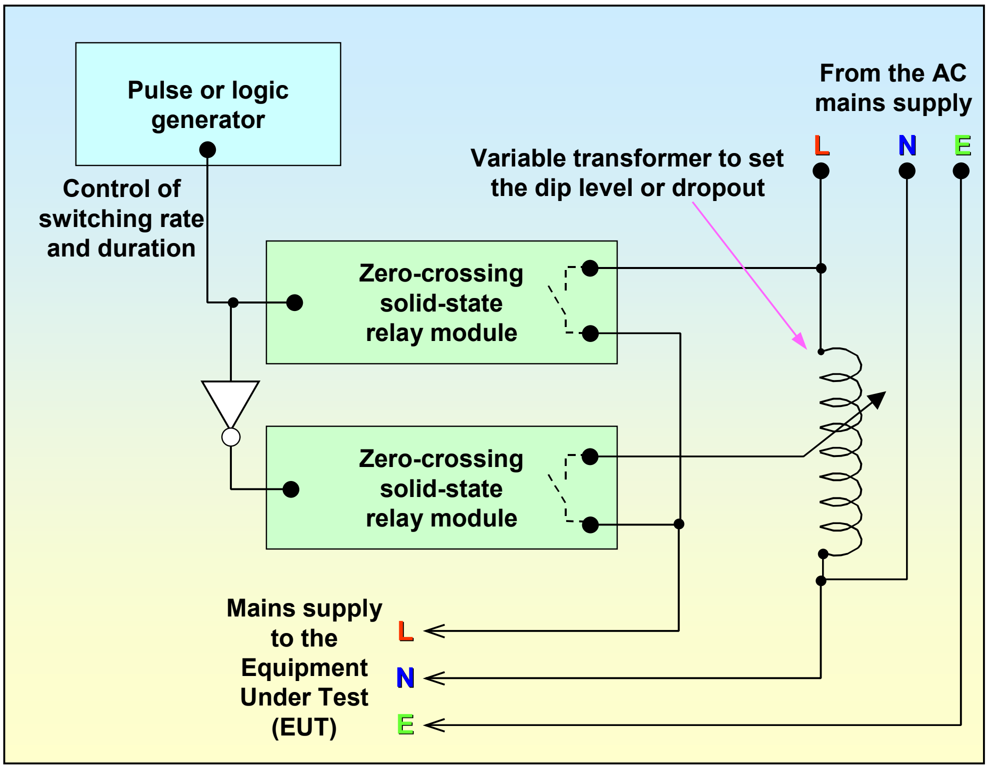


Figure 6BF Block diagram of a do-it-yourself dips and dropouts tester

The definitions of dip, dropout, short interruption, long interruption, sag and swell (see 6.5.3), are all derived from the IEC EMC standards, but it important to be aware that other industries might use the same terms to mean different phenomena, and this might also depend upon the country.

For example, in the USA it is common to use the term ‘sag’ to mean a dip, and the UK’s electricity supply authorities have decided that a ‘short interruption’ in the electricity they supply to consumers, is any interruption lasting less than two minutes. They do not regard short interruptions as being as serious as long ones, although for most electronic equipment a power interruption of greater than one second is as serious as one lasting a minute or more.

Figure 6BG shows how undervoltages can cause problems for electronic circuits, by causing the unregulated DC voltage rail to drop below the minimum input voltage of the voltage regulator, causing its regulated output voltage to fall below specification. ICs require a certain minimum DC supply voltage for their correct operation to be guaranteed, and when their DC supply falls below this value, for example due to a dip or dropout, they can behave in very strange ways. For example, a NAND gate might behave as a different kind of gate, microprocessors and other programmable devices might over-write their data or program memories with garbage, and software operation might be corrupted by ‘looping’, ‘hanging’ or ‘crashing’

Undervoltages can also cause problems for electromechanical devices such as solenoids, relays, contactors, solenoids, AC and DC motors. Solenoids can lose control of their loads, relays and contactors can drop out. Motors might slow down or overspeed depending on their type and control method, but during a sag they can lose so much power that they stall, which causes their magnetising currents to increase significantly, in turn causing overheating that can lead to fire (if not interrupted in time by a protection device) – or at least damage their winding insulation so that reliability is significantly impaired and electric shock risks increased.

If a relay or contactor that is held-in by a normally-open contact will not recover to its original state afterwards. If a relay or contactor that is held-in at reduced voltage (to reduce power consumption and heating) is held-in by a normally-open contact, it might not pull back in again when the supply recovers – depending on the type of coil power supply.

Until recently, most safety systems in industry relied upon so-called ‘hard-wired’ devices such as solenoids, relays and contactors. But few designers of such systems ever bothered to consider the effects of undervoltage events, or the fact that the responses of solenoids, relays and contactors to them varies depending on their temperature, and on how old they are.

Since the publication of IEC 61508 in 2000, many industrial safety systems have started to use computer techniques and fieldbusses instead of hard-wiring, but in all the very many glossy advertisements for this new equipment I have yet to see anything that indicates that the designers have taken undervoltages or other power quality issues into consideration.

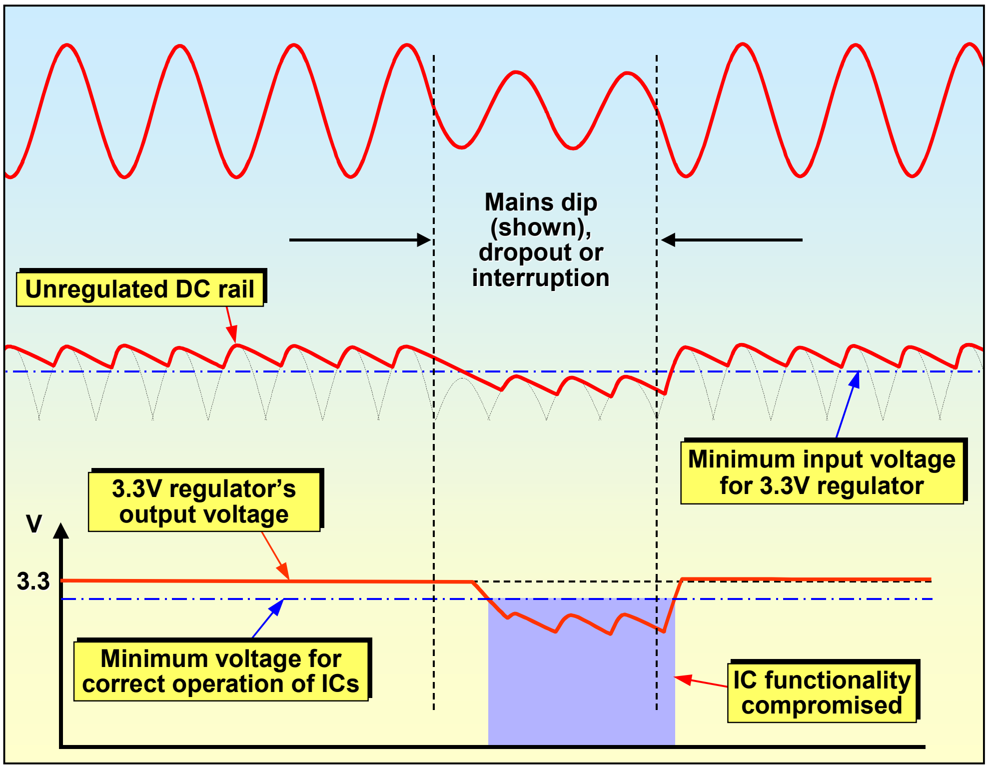


Figure 6BG How undervoltages can cause problems for electronic circuits

Perhaps it is a requirement of their manufacturers that these new computerised safety system controllers will be operated from continuous on-line double-conversion UPSs (see 6.5.3 and 6.5.11) with sufficient energy storage at least to maintain safety whilst shutting down the equipment under control? But if this is so – the advertisements do not state it – perhaps it is in the small print somewhere in the sales terms and conditions or user instructions.

Equipment can be designed with increased protection against undervoltages, for example by operating DC equipment from AC-DC (or DC-DC) power converters that have a very large input voltage range. For example, the typical 'universal input' power converters used for charging the batteries of laptops and cellphones are rated for 100-240Vrms inputs, so that their manufacturers can sell the same product all around the world, only having to ensure that each shipment is packed with the correct type of mains cord. Using such converters on nominally 230V mains supplies provides total protection against dips and sags down to 43% of nominal (a dip or sag of 57%).

AC-DC power converters that automatically select either 115 or 230V nominal inputs are not suitable, and not recommended. For example, if the mains voltage was 180Vrms and the AC-DC power converter device selected a 115V input, the DC output would be damagingly high, and the converter might be damaged too. But if is selected 230V the DC output would be too low and the equipment would be powered at too low a voltage for correct operation. Power converters with a continuous range of input voltage are strongly recommended.

To protect against dips and sags that go below the input range of the power converter, or of course against dropouts and short interruptions, requires the power converter have an adequate 'hold-up' time, for the load it is powering. This requires sufficient energy to be stored in electrolytic capacitors, supercapacitors, batteries, fuel cells, flywheels, etc., and the energy storage is generally provided at the unregulated DC rail. Figure 6BH shows some examples of energy storage devices.

It helps make best use of the stored energy in the unregulated DC rails' capacitors, if the DC regulators and/or equipment being powered from it are designed to cope with very large fluctuations in the unregulated DC voltage.

If using capacitive energy storage, operating the unregulated DC storage capacitors at higher voltages requires less physical volume (size) for a given stored energy, because capacitive energy storage is proportional to the capacitor voltage squared (stored energy in Joules = ½CV^2). Figure 6BJ shows a block diagram of a power converter that uses this effect.

Figure 6BH Some examples of energy storage devices

The step-up (boost) stage could be an active power factor correction circuit (see 6.3.6 and 6.4.6 in [36]). If aiming to double the unregulated voltage, for operation from nominally 115V mains supplies, the switching devices and storage capacitors need only be rated the same as for a normal power converter running on nominal 230V mains (at least 600V and 350Vdc respectively). But for operation from nominal 230V mains the switchers would probably need to be rated for 1200V and the capacitors for 800V – which might limit the availability of low-cost components, so maybe a smaller boost percentage might be more cost-effective.

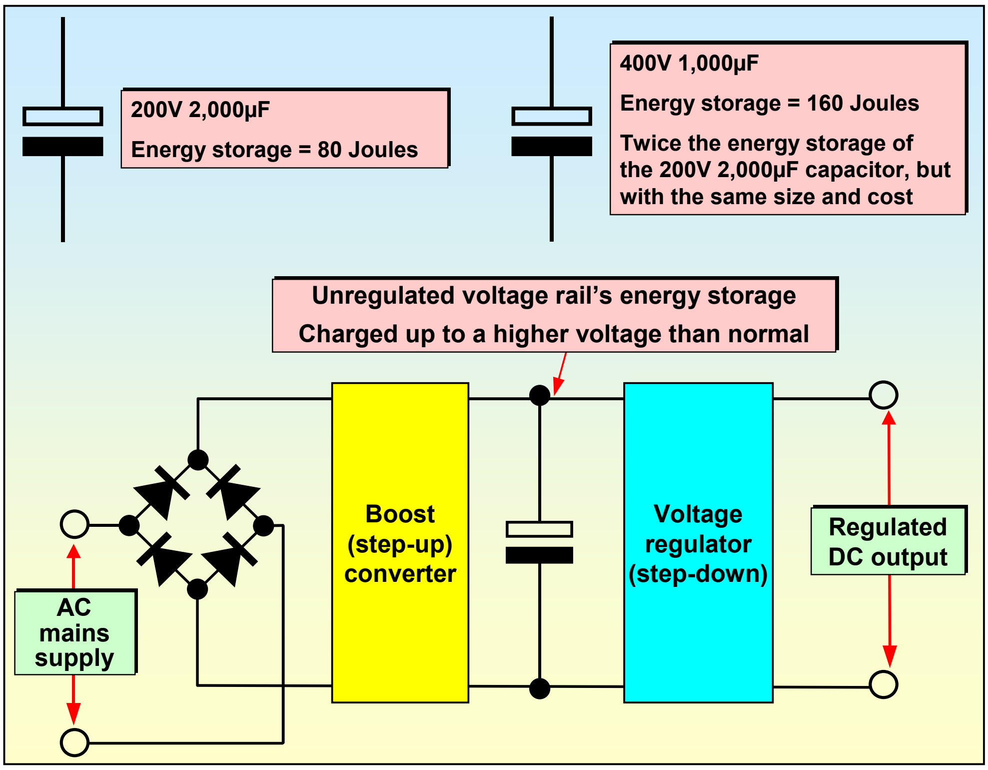


Figure 6BJ Storing more energy in unregulated storage capacitors by using a higher voltage

Where very large energy storage is required, large battery systems or fuel cells may be required, perhaps just while back-up generators get up to speed, and Figure 6BK shows the example of a 'battery room' used for such a purpose, and such rooms are typical of large UPS installations.

To reduce the size and cost of the energy storage, it is sometimes possible to identify certain parts of the product that must keep operating at full specification during the undervoltage event, whilst others can be operated at lower power or even switched off completely. For example, the brightness of a backlight or other illumination could be allowed to reduce, microprocessor clock speed reduced, etc.

Before the energy storage runs out completely and the product must cease to operate, it is important to initiate a 'controlled shutdown' to prevent loss of data, damage or safety incidents. However, in some critical applications (such as life-support) shutdown can never be permitted, and such products will need to have a guaranteed source of electrical power (e.g. generators, and sufficient fuel for them).

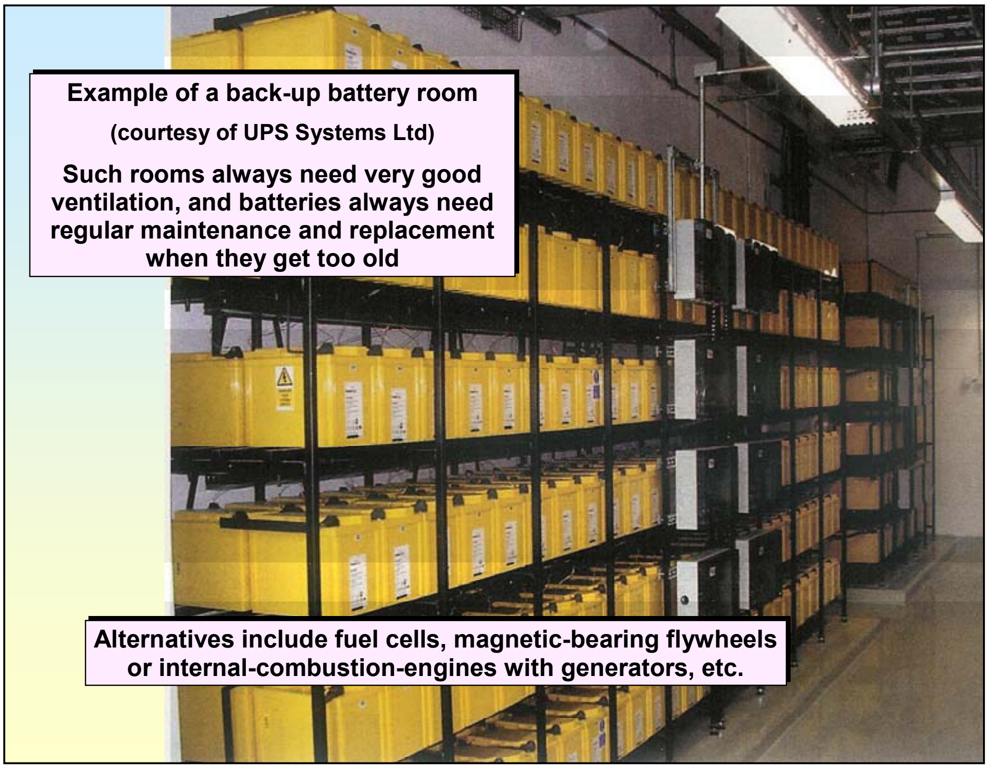


Figure 6BK Example of a large battery room

All digital devices, including microprocessors, microcontrollers, etc., should be protected by voltage monitor devices, often called 'brownout detectors' or 'brownout monitors'. These detect an out-of-specification DC voltage (ideally, one that is *about* to become out-of-specification) and freeze RAM and programmable ROM, terminate disc writes, etc, so that malfunctioning ICs can't destroy data or alter programs. They are available with a range of accuracy specifications, and of course the more accurate ones are generally the better ones to use, and they cost more.

Non-volatile RAM can be used to store the operating state so that operation as before can resume after the undervoltage event is over – but only in appropriate applications where such self-recovery is acceptable, for example does not increase safety risks.

Some circuits sample the mains voltage (usually to control heat or other parameters), and they can often use a large value capacitor, or rely on a digitally-stored value, to cope with short-term variations in mains voltage.

Relays, contactors and solenoids can often use DC coils (instead of AC coils) and be powered by a AC-DC power converter which has a very large input voltage range and/or sufficient hold-up time (energy storage) as discussed earlier. An advantage of DC power is that it is relatively easy to operate the devices from batteries that are 'float-charged' from the regular mains supply, the way that a normal laptop PC is operated.

If using AC coils, choose AC relays, contactors and solenoids with lower 'drop-out' or 'hold-in' voltages. As mentioned earlier, typical low-cost relays can drop-out at 78% of the nominal supply, whilst better types are available with hold-in voltages down to 50% (or less) of nominal. So-called 'coil hold-in' devices (such as 'KnowTrip' or 'Coil-Lock') are available, some of which claim to keep coils energised when the AC supply is as low as 25% of nominal.

Whilst a lower hold-in voltage will increase protection from dips and sags, it will not protect against dropouts and interruptions in the AC power, for which AC coils will need a source of AC power from a UPS (see 6.5.11) that has sufficient energy storage.

6.5.9 Voltage fluctuations

Voltage fluctuations – according to the IEC 61000-4 series of standards – are rapid sequences of voltage dips and/or voltage increases or alternating dips and increases, and an example is shown in Figure 6BL. For AC supplies, a relevant immunity test standard is EN/IEC 61000-4-14, and a guide on it is available from [7]. This guide includes more detailed descriptions of voltage fluctuation phenomena, including what causes them, what they can affect and how, and how to test to help improve real-world reliability. Some low-cost but non-compliant tests that can be done by anyone with sufficient competence are described in [6].

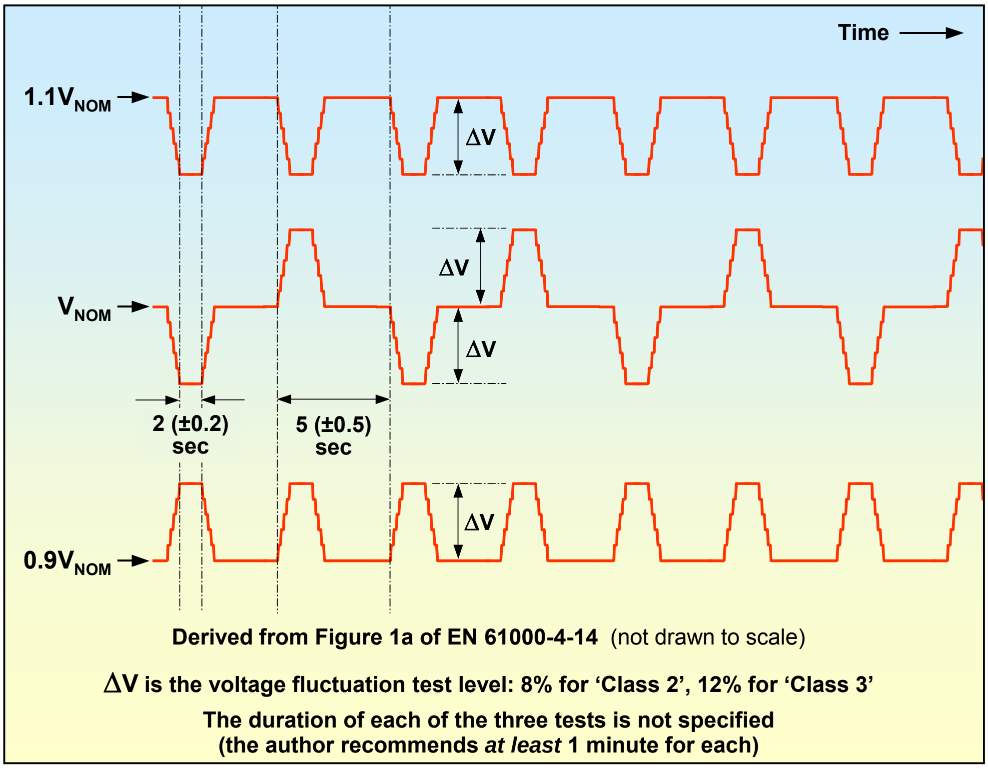


Figure 6BL The three types of voltage fluctuation test in EN/IEC 61000-4-14

Tests appropriate for *individual* mains voltage dips, sags, swells, etc. are covered by EN/IEC 61000-4-11, see 6.5.3 and 6.5.8 and the relevant guides from [7]. However, these tests do not cover individual abrupt voltage increases (the opposite of dips), even though they can occur.

For DC supply networks, a relevant immunity test standard for voltage fluctuations is EN/IEC 61000-4-29, which it calls "variations" instead of fluctuations, and a guide on this standard is available from [7]. This guide also goes into what causes the phenomena, what they can affect and how, and how to test to help improve real-world reliability.

Voltage fluctuations are dealt with by designing equipment to protect against both undervoltages (see 6.5.8) and overvoltages (see 6.5.3). Note that the frequency of the fluctuation will appear as a ripple on the unregulated DC rail, and circuits that are not powered via a DC regulator operated within its correct input range, will experience this ripple directly, and it might appear as noise in the passband of the circuit.

6.5.10 Waveform distortion (harmonic and/or interharmonic)

Harmonic distortion is when a spectrum analysis shows that the AC electricity supply contains frequencies that are integer multiples of the fundamental mains frequency (e.g. for 50Hz: 100Hz, 150Hz, 200Hz, 250Hz, etc., known as the 2^{nd}, 3^{rd}, 4^{th}, 5^{th}, etc., harmonics respectively). When viewed on a 'line-triggered' oscilloscope, the supposedly sine-wave mains waveform will be seen to be distorted, with the distortions 'phase locked' to the mains waveform.

Harmonic distortion is caused by non-linear loads, such as the rectifier-capacitor AC-DC converters typical of almost all AC mains-powered electronic products. Products using thyristor/triac power control are also non-linear loads. As more and more direct-on-line motor and heating loads are replaced by electronic controlled loads (e.g. variable-speed AC motor drives) the waveform distortion is generally worsening. Typical values are under 4% total harmonic distortion (THD) and the electricity supply authorities in Europe have agreed that they must keep it below 8% because it is commonly observed that typical electronic equipment often malfunctions with THD above this level.

As was discussed earlier, many offshore vessels now use electronically-controlled 'thrusters' that represent such a large proportion of their generator capacity that it is not unusual for their on-board mains waveforms to suffer THDs of up to 30%. Figure 6BM shows an example of a mains waveform recorded in a domestic house in Israel in 2000.

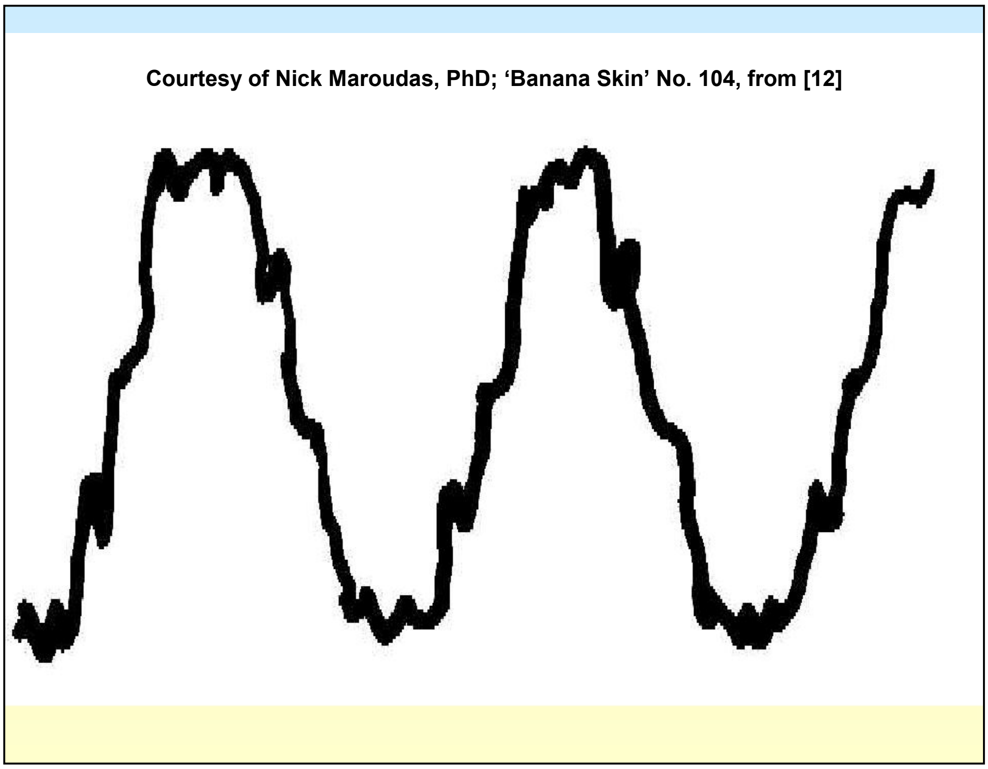


Figure 6BM An example of a mains waveform in a house in Israel in 2000

Interharmonic distortion is when a spectrum analysis shows that the AC electricity supply contains frequencies that are not integer multiples of the fundamental mains frequency (e.g. 39Hz, 105Hz, etc.). When viewed on a 'line-triggered' oscilloscope, the shape and amplitude of the supposedly sine-wave mains waveform will be seen to wobble, or even be blurred, because the distortions are not phase-locked to the mains waveform.

Interharmonic distortion is created quite differently from harmonic distortion – it is simply the voltage fluctuations caused by rapidly fluctuating loads like those discussed in 6.5.9. The chief source is powerful switch-mode frequency-converters, such as variable speed AC motor drives, the use of which is rapidly increasing to save power consumption and hence help reduce the rate of warming of our planet. Another source of interharmonic distortion can be high-power frequency converters used in the AC power distribution networks themselves.

A switch-mode frequency converter operating at, say, 39Hz has AC mains current demands at 39Hz *and* at its harmonics, and these non-mains frequency currents flowing in the impedances of the mains supply network result in voltage fluctuations at those frequencies – known as interharmonic waveform distortion. The situation is actually much more complicated than this, because switch-mode power converters act as 'frequency mixers' that cause the mains frequency and its harmonics to intermodulate with the inverter frequency and its harmonics.

Figure 6BN shows a measurement of the spectrum of the current into a 700kW variable-speed AC motor drive running at an output frequency of 39.375 Hz. The X-axis markings are probably too small to read in this Figure, but include (for example) frequencies such as 131.25Hz created by the intermodulation of the 6^{th} harmonic of the 50Hz mains and the 11^{th} harmonic of the motor frequency. When this complex current waveform flows in the impedance of the mains distribution network, it will give rise to voltage waveform distortion at all of the frequencies shown in Figure 6BN.

Neither of the above waveform distortions apply to DC electrical power, of course, but DC power can (and usually does) carry frequencies other than DC, caused by the same rapidly fluctuating load currents that would create interharmonics in an AC supply.

A relevant immunity test standard is EN/IEC 61000-4-13, and a substantial guide on it is available from [7]. This guide includes more detailed descriptions of waveform distortion phenomena, including what causes them, what they can affect and how, and how to test for them to help improve real-world reliability. Other useful guides, more appropriate for systems and installations engineering, are [31] and [33].

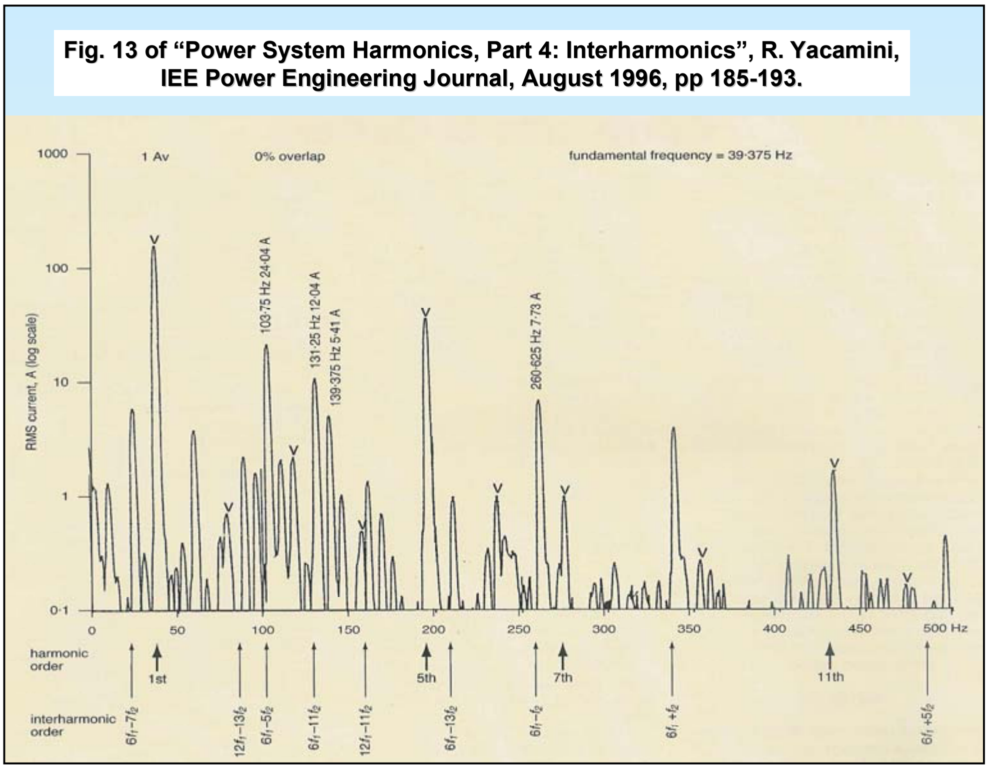


Figure 6BN Example of harmonic and interharmonic currents from an inverter-fed AC motor

Harmonic distortion can result in mains waveform peaks that are lower than the √2 Vrms expected from a pure sine wave, and an extreme example is a 230Vrms square wave supplied by single-phase mains sockets, which has occurred in China. Since the typical bridge rectifier-capacitor AC-DC converter used in almost all electronic equipment (other than thyristor power control) charges up to the peak of the supply waveform, a THD of X% means that the peak can be up to X% lower than expected.

A power converter for use on mains supplies of 230V ±10%, if supplied from a supply with 4% THD, could be running at -4% of its unregulated DC voltage even when the supply is at its nominal RMS value. So the unregulated rail could become too low for correct operation of the product when the mains voltage falls below 230Vrms -6%. If the THD was 8% (unlikely when powered by a national grid network except in some industrial sites, more likely when running on local generation) then the product might malfunction below 230Vrms -2%.

The same levels of THD, but with the harmonic components in different phases, can instead result in peaks higher than the expected √2 of the nominal RMS voltage, which could cause overvoltage damage. For example, a power converter for use on mains supplies of 230V ±10%, if supplied from a supply with 4% THD, could be running at +4% of its unregulated DC voltage even when the supply is nominal as measured on a true-RMS meter. In the UK, the nominal mains voltage is in fact 240V (230Vrms +4.3%), so with a supply of 240V +4% THD, the product could be damaged by operating above its maximum unregulated DC voltage when the mains rose above 240V + 1.7%, which it often does.

Figure 6BP shows the example of adding 8% third harmonic to a pure sinewave. When added in-phase it causes the peak voltage to be 8% higher than √2 Vrms, and when added in antiphase it causes the peak voltage to be 8% lower than √2 Vrms.

Interharmonic distortion causes the same overvoltage and undervoltage problems, but they beat with the fundamental frequency so although they can cause instantaneous overvoltage damage they are only likely to cause undervoltage problems when the beat frequency is low, say less than 10Hz.

Both kinds of distortion can cause errors in the zero-crossing point, causing problems (such as double-zero-crossings) for circuits that use the zero-crossings of the mains supply to control power switching, timing or other functions. Thyristor/triac power control circuits that use simple gate drive circuits to derive their firing signals from the mains waveform can misfire causing malfunctions, even catastrophic damage.

Distortion frequency components can exceed 2kHz, and because the impedance of a capacitor reduces as frequency increases, such high frequencies can cause very significant increases in the current flowing in capacitors in EMI filters, and in the displacement power-factor correction capacitors in luminaires and electricity distribution networks. These high currents can cause overheating damage, and such damage to capacitors is not uncommon, for example see No. 7 in [12].

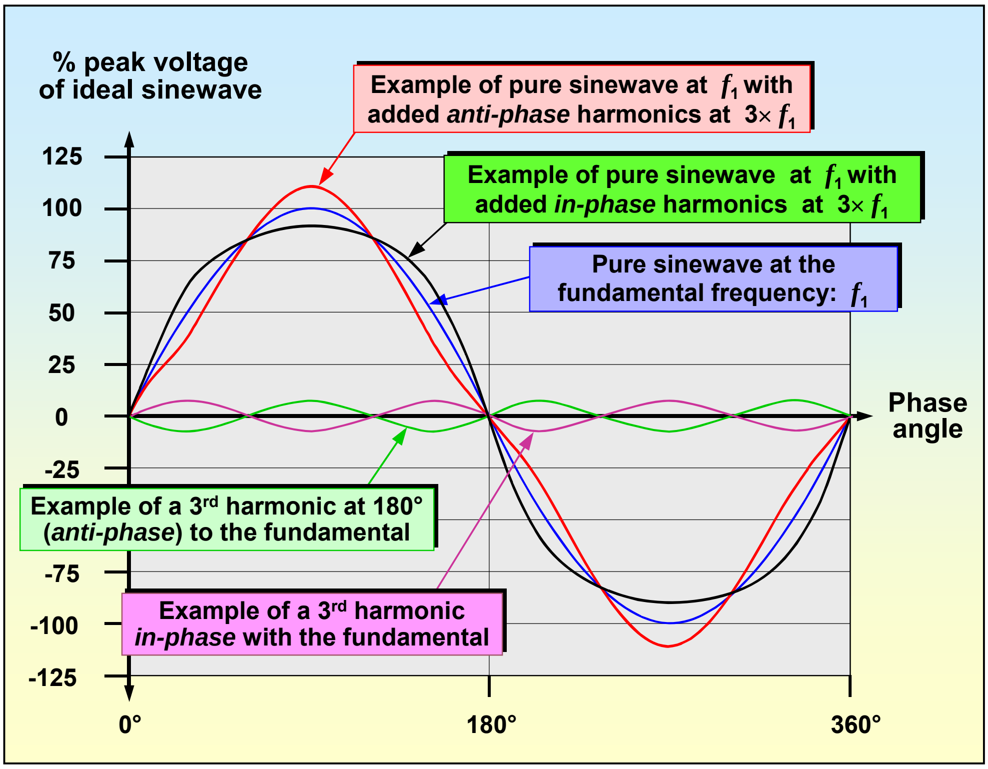


Figure 6BP An example of effect on peak voltage of the phase of a harmonic

Mains waveform distortions and their associated currents can also cause motors, transformers and cables to overheat, and can cause severe interference with wired telephones. They also produce ripple on the rectified DC voltages of AC-DC power converters, at the frequencies of the interharmonic noises, which could interfere with circuits powered from those DC rails.

To design products to protect against waveform distortions, appropriate overvoltage and undervoltage design techniques described previously should both be applied. Thyristor/triac power control circuits should be designed to cope with all foreseeable timing errors due to distorted mains waveforms. Timers and real-time clocks should use stable reference oscillators (e.g. as used in wristwatches) or off-air frequency references from terrestrial or satellite transmitters (see 6.5.4). Filters or voltage regulators may be needed in some applications, to remove in-band noise from unregulated DC power rails.

6.5.11 Improving the quality of the mains supply itself

There are numerous ways of obtaining a better quality of mains supply for a product, including the following:

- Obtaining a better quality mains supply
- On-site generation of an AC supply
- Passive or 'active' mains filtering
- Constant voltage transformer (CVT)
- Motor flywheel - generator sets
- Multi-tapped triac-switched transformers
- Servo-motor variable transformers
- Uninterruptible Power Supplies (UPSs)
- Dynamic voltage restorers (DVRs)

All the above techniques can be applied at system or installation level, which is covered by [33] but is not the subject area of this series of articles. However, the word 'product' encompasses a huge range of possible equipment, and some types of products may be able to incorporate some of these techniques within themselves.

Obtaining a better quality mains supply

Powering a single-phase mains-powered product from phase-to-phase mains voltage, instead of phase-neutral, reduces the amplitude of any phase-to-neutral or phase-to-earth dips, dropouts, interruptions, sags, swells and voltage fluctuations. The rate of occurrence of such problems is reduced (but not eliminated) and *average* power quality is improved. This technique will generally also reduce the effect of any DC in AC networks, but will usually have little or no effect on distortion, frequency errors, and CM voltages.

If it is the load currents consumed by other equipment connected to the same mains distribution network that is causing the power quality problem, the product could be connected to a different branch of the network from that used by those loads. It might even be necessary to connect it directly to the 230/400V distribution transformer feeding that mains network.

Going further, the product could be connected to a 'point of common connection' (PCC) that is upstream of that distribution transformer, and therefore operates at a higher voltage, with a lower impedance that is less disturbed by the load currents of the other equipment, and therefore provides a better power quality. This will generally require a high-voltage transformer to connect to the PCC (for example 3.3kV, 11kV, 33kV, etc.).

Generating a 'clean' AC supply on-site

Generating your own electricity supply, for example using an internal combustion engine driving a generator, can cure all of the problems caused by the poor quality of the normal mains power distribution.

But it is important to understand that electricity generators have significantly higher impedances (approximately 3 times) than a mains distribution transformer of the same kVA or kW rating – so fluctuating or distorted load currents could cause significantly larger effects in a locally generated supply, than when powered from the normal mains supply network.

So it could happen that where a national electricity supply is plagued with (say) dips and dropouts, and a local generator is used instead, the non-linear nature (say) of one of more of the loads causes the local generator output distortion to rise to unacceptable levels – exchanging one set of power quality problems for another.

One solution is to ensure that the generator is rated for a much higher output than the load will consume, ideally three times higher, so for example a 100kW load would use a generator of 300kW or more. Greater cost-effectiveness can be achieved whilst also improving power quality if the likely effect of the loads on the generator is analysed in sufficient detail, taking the generator's output impedance from DC to at least 5kHz into account.

Some generators use automatic voltage regulators (AVRs) that cause voltage transients and/or waveform distortion, so it can be important to check the power quality provided by the generator.

Often, local generation is used in 'stand-by' mode, to pick up the load when the normal mains supply fails. But changing over the supply from mains to generator (and back again) can give rise to very significant undervoltages, overvoltages, fast transients, and surges, that can cause problems for some types of loads (see No. 55 of [12] for a hospital example).

One solution to the above problems inherent with local electricity generation is to operate all sensitive or critical equipment from high-reliability continuous on-line double-conversion UPSs (see later).

Passive or 'active' filtering to reduce distortion

Passive filtering at/above the 7th harmonic often uses low-pass (LP) filter techniques, but for the 6th harmonic and below LP filters often have excessive thermal losses at 50/60Hz so 'resonant trap' filters are mostly used, each one tuned to a specific problem harmonic frequency. The design of such mains filters for use in an installation is not trivial, and unless you are an expert in doing just this, I strongly recommend that you employ experts.

So-called 'active' filtering does not actually use filter technology, it is just a marketing term invented to try to make people who are used to traditional passive mains filtering feel more comfortable with this new electronic technology.

'Active mains filtering' uses 'anti-harmonic injection' techniques that employ switch-mode AC-AC power inverter technologies, and Figure 6BQ shows its basic operating principles.

Active filters monitor the non-linear currents consumed by the load, and inject antiphase harmonic currents into the mains distribution network so that, upstream of the injection point, the network is only required to provide sinewave current at the fundamental frequency and so its intrinsic impedance does not give rise to waveform distortion.

Active filters can be sized just to deal with one load, and can even be incorporated into products to act as power factor correctors, see 6.3 in [36]. They can also be sized to cope with multiple loads, for example using one active filter per floor of a tall office building, so that the main risers providing power to the floors do not carry harmonic currents and any voltage distortion just arises within each floor.

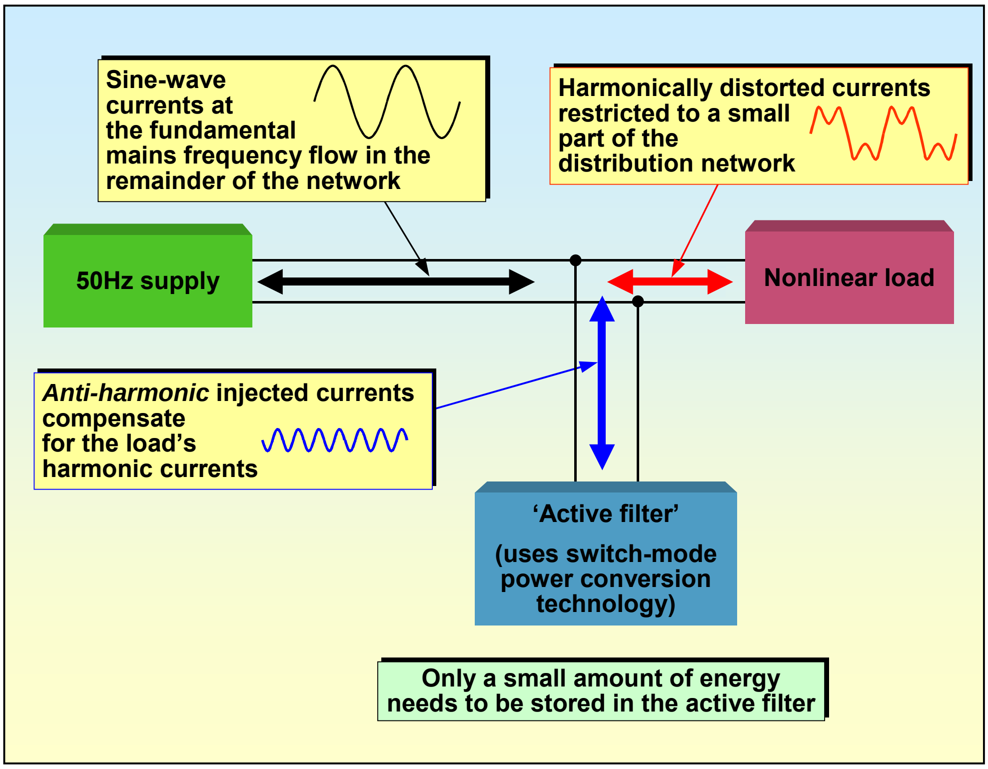


Figure 6BQ 'Active' mains filters – operating principles

Constant Voltage Transformers (CVTs)

This venerable technology operates the secondary winding of an isolating transformer in saturation, as part of a 50/60Hz resonant circuit, so it is inefficient and runs hot. Its output waveform is generally not a very good sinewave but it effectively suppresses other mains waveform distortions, sags, brownouts and swells.

There is stored energy in the resonant circuit, so if sufficiently oversized it can continue to provide power even during dips and dropouts, often called 'hold up' or 'ride-through' – and if this is required the general recommendation is to rate the CVT at 2.5 times the power of the load, or more.

CVTs are large, inefficient, and run hot, when compared with modern solid-state technologies. But because they contain no semiconductors they are very robust and reliable, and what little maintenance and repair they might ever need is easily provided using standard electrical knowledge and tools. In some applications, they may well be more cost-effective than their modern alternatives.

Motor-flywheel-generator sets

The motor is powered from the poor quality electricity supply (which could be AC or DC) and drives a generator to provide a 'clean' supply to the protected equipment. The motor has automatic speed control to set the output frequency, and the generator has automatic voltage regulation to set the voltage, and the flywheel provides some energy storage. If correctly dimensioned and competently designed and constructed, they can solve all power quality problems other than long interruptions, and provide CM and DM isolation from DC to many GHz. They can also convert from one frequency to another.

The 'hold-up' or 'ride-through' time for longer interruptions depends on the size and rotational speed of the flywheel, which can be designed to store a great deal of energy. Modern types use lightweight non-metallic flywheels rotating at huge speeds to safely store very large amounts of energy in quite small volumes.

It is important that the motor is rated to withstand the poor quality of the mains power supplied, which generally means increasing its size and power rating to prevent overheating due to waveform distortions, low frequencies, DC in AC supplies, undervoltages, etc., and it should have insulation that will cope with the anticipated overvoltages (swells, see 6.5.3).

As described earlier for engine-driven generators, the generator will have an impedance that is approximately three times that of an HV distribution transformer of the same VA rating powered from a national mains supply network, so it is important to ensure that the fluctuating and/or non-linear loading on it does not result in worse power quality overall, than the original poor-quality electricity supply that it is supposed to be protecting against.

Multi-tapped transformer with triac switching

This technology was discussed in 6.5.3, and its block diagram was shown in Figure 6BD. It generally takes a few tens of milliseconds to correct a voltage change in the supply, and will not compensate for dips or voltage fluctuations that occur on shorter timescales. Having no energy storage it cannot compensate for dropouts and interruptions of any duration. It does not suppress the distortion of the waveform passing through it, but (as mentioned earlier) if its voltage sensing circuit is designed to respond to the peak of the mains waveform rather than its RMS or Average, it will operate so as to maintain the peak of its output voltage at a constant level, which is just what most rectifier-capacitor AC-DC converters require, and so for such loads it could be considered to be compensating for waveform distortion.

Servo-motor controlled variable transformers

This technology is very similar to the multi-tapped transformer with automatic tap-selection using triacs, discussed above, but instead of a multi-tapped winding it uses a continuously variable tapping via the sliding wiper of an autotransformer. Because the tapping point is infinitely variable, the output voltage can be 'stabilised' almost exactly at the desired voltage, but the mechanical movement required means that it can take a few seconds to correct for a voltage change, so it will not correct for short-term swells and undervoltages.

Figure 6BR shows the basic principles in block diagram form, and includes a photograph of a commercially available three-phase unit that shows that even very large powers can be controlled by just a small motor, driving the wiping contacts through a reduction gearbox.

As for the multi-tapped method, it has no energy storage so cannot compensate for dropouts and interruptions, and it has no effect on the distortion of the mains waveform, but if the voltage sensing works on the peak rather than on the RMS or Average value, it will maintain the peak voltage output at a constant level and thus compensate for waveform distortion as far as rectifier-capacitor AC-DC converters are concerned.

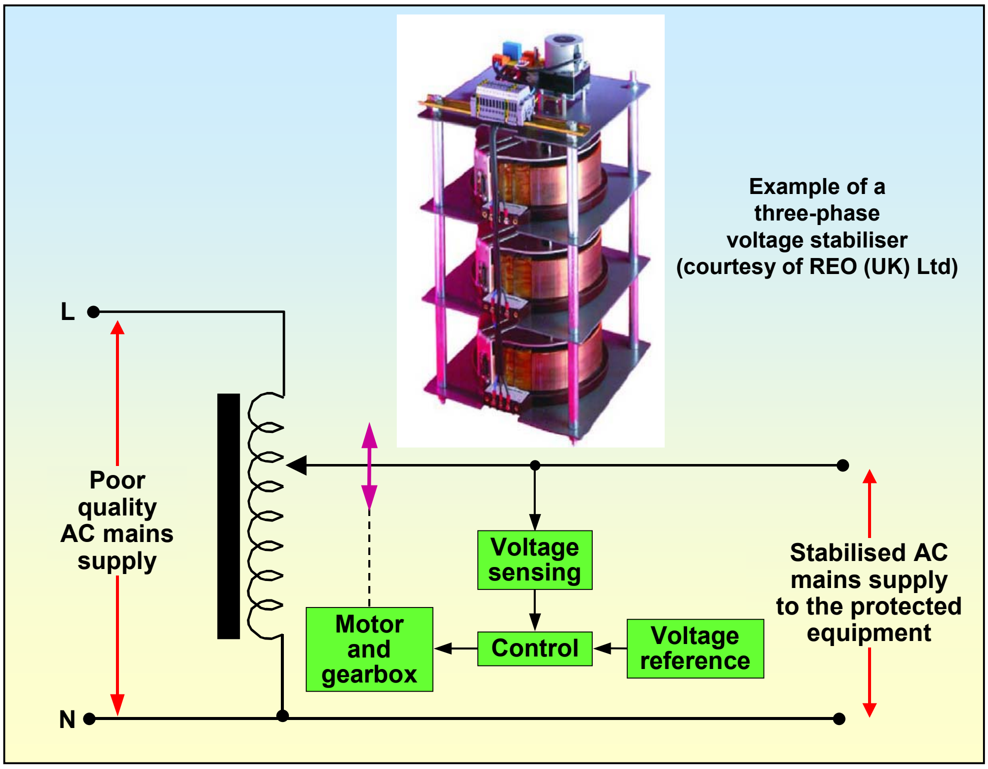


Figure 6BR Servo-motor controlled 'voltage stabiliser' principles

Uninterruptible power supplies (UPSs)

These are AC-AC (actually AC-DC then DC-AC) switch-mode power converters, often called 'inverters' with their output set to the required mains frequency. They can convert from one mains frequency to another, or can be used as DC-AC converters to generate an AC supply from a source of DC electrical power.

'Continuous-on-line double-conversion' types can cure all power quality problems, and are conceptually similar to a motor-generator set with flywheel storage. The poor quality power supply is used to charge their energy storage (e.g. battery, flywheel, etc.), and their energy storage is used to supply power to the protected load circuit, as shown in Figure 6BS.

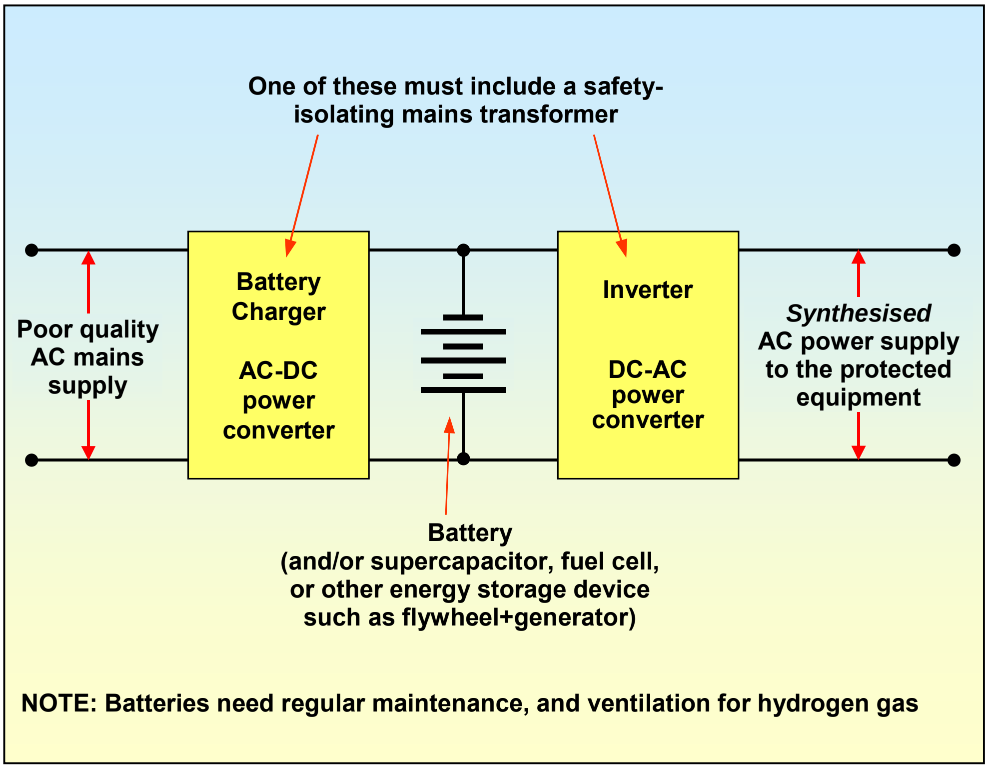


Figure 6BS Principles of continuous-on-line double-conversion UPS

Their mains-powered charging circuits must, of course, be able to withstand the expected voltage swells, sags, distortion, etc.

I have seen examples of UPSs specified by their manufacturers as providing at least 80dB of attenuation for all mains-borne disturbances from DC to 1GHz, from their input to their output. (Motor-generator sets can also be designed and manufactured to achieve such excellent EMC specifications.)

Many types of lower-cost UPSs are available, but can cause more power quality problems than they solve for their protected loads. For example, a common type powers the load from the mains supply, and only switches over to UPS mode when it detects that some aspect of the input supply's power quality (e.g. RMS voltage) has dropped below a preset threshold. These types do not protect against all power quality problems, and can cause dips/dropouts and transients when they switch-over, or when they switch the load back again to the normal mains power.

Reliability is another cost-related issue, and some models of UPS (even continuous double-conversion types) have been known to expose their loads to more supply interruptions (due to failure of the UPS) than the mains supply they were supposed to be protected from.

So take great care, when purchasing a UPS, to make sure that it really *will* provide the power quality improvements that that are required.

Dynamic voltage restorers (DVRs) (sometimes called dynamic sag restorers)

These use similar switch-mode power conversion technology to the 'active' filters described earlier, but instead of injecting currents in parallel with the mains supply, they inject voltages in series with it, usually with the aim of maintaining an adequate mains voltage during a dip, sag, dropout or short interruption. They need significant amounts of energy storage (supercapacitors, batteries, etc.) depending on the load power, and on the dip/sag depths and durations that are to be protected from.

Number 53 of [12] describes one successful application, called a 'voltage dip protector'.

6.5.12 Tripping-out techniques

Special protection devices, often called ‘protection relays’, are available to detect a wide variety of mains power quality problems and remove the power completely from the protected equipment by operating a circuit-breaker or triac.

‘Protection Relays’ can protect against under/over voltage, current or frequency; phase unbalance or failure; unbalanced load currents, and Figure 6BT shows some commercially available units.

Examples of protection relays that protect against under/over voltage, current or frequency; phase unbalance or failure, or unbalanced load currents

Courtesy of Eltime Controls Ltd

Figure 6BT Some examples of protection relays providing a variety of trip functions

Loss of power and/or trips can occur at unpredictable times, so it is important to ensure that they do not cause unacceptable damage, financial loss, or safety incidents depending on the application. For example, the industrial processing of webs of material (paper, plastic, tyre rubber, etc.) generally employs numerous motors that all need to keep rotating in step so as not to tear the web, which can require hours to repair at the cost of a great deal of lost production.

So, if the electrical power supply ceases either due to an interruption or trip, UPSs or other energy storage techniques are generally required to provide sufficient operating time for a controlled power-down of the motors that does not tear the web (although a length of it may be spoilt).

The electrical power supply will also be restored at unpredictable times, and it is important to consider all of the consequences. For example, it might be restored during the controlled power-down of an industrial process, in which case the process might be required to continue ramping down to a stop and await manual restart, or ramp back up again and continue production, but in either case the process must be controlled at all times.

Where a product is controlled by digital processing, it is important to ensure that all the register contents are set appropriately on restart. If a ‘cold start’ is required, they should all be reset to zero. If a ‘warm start’ (continuing as before the interruption) is required instead, they should all be loaded with the appropriate data.

Power amplifiers can misbehave during power down or power up, so to protect them and their transducer loads from damage their control circuits should protect them against all power down/up situations. Testing with the full range of dropouts and interruptions in EN/IEC 61000-4-11, or more, is recommended.

How the product should be designed, to respond to power interruptions and/or restorations, depends on its application, especially if there are any possible safety implications.

6.6 Conclusion to the series

This series is now concluded, and I hope you have found it interesting and useful. It has spanned over a dozen issues and over more than two years, and it is all available from the EMC Journal’s website at www.theemcjournal.com.

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